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MICROWAVE RECEIVERS

Edited by

S. N. VAN VOORHIS
ASSOCIATE PROFESSOR OF PHYSICS
UNIVERSITY OF ROCHESTER

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Foreword

THE tremendous research and development effort that went into the development of radar and related techniques during World War II resulted not only in hundreds of radar sets for military (and some for possible peacetime) use but also in a great body of information and new techniques in the electronics and high-frequency fields. Because this basic material may be of great value to science and engineering, it seemed most important to publish it as soon as security permitted.

The Radiation Laboratory of MIT, which operated under the supervision of the National Defense Research Committee, undertook the great task of preparing these volumes. The work described herein, however, is the collective result of work done at many laboratories, Army, Navy, university, and industrial, both in this country and in England, Canada, and other Dominions.

The Radiation Laboratory, once its proposals were approved and finances provided by the Office of Scientific Research and Development, chose Louis N. Ridenour as Editor-in-Chief to lead and direct the entire project. An editorial staff was then selected of those best qualified for this type of task. Finally the authors for the various volumes or chapters or sections were chosen from among those experts who were intimately familiar with the various fields, and who were able and willing to write the summaries of them. This entire staff agreed to remain at work at MIT for six months or more after the work of the Radiation Laboratory was complete. These volumes stand as a monument to this group.

These volumes serve as a memorial to the unnamed hundreds and thousands of other scientists, engineers, and others who actually carried on the research, development, and engineering work the results of which are herein described. There were so many involved in this work and they worked so closely together even though often in widely separated laboratories that it is impossible to name or even to know those who contributed to a particular idea or development. Only certain ones who wrote reports or articles have even been mentioned. But to all those who contributed in any way to this great cooperative development enterprise, both in this country and in England, these volumes are dedicated.

L. A. DUBRIDGE.

Preface

THE receivers and circuits treated in this volume have sprung almost entirely from radar techniques. It is felt, however, that many of the features are applicable in other services where high sensitivity and excellent transient behavior are required. An attempt has been made to reach a sufficiently fundamental standpoint in the presentation to permit such use in allied fields.

The long and uninterrupted record of unstinting cooperation on the part of other laboratories and organizations, both industrial and government, has continued throughout the preparation of this book. Particular thanks are due to the Raytheon Manufacturing Company for the use of illustrative material for Chap. 13 and to the Hazeltine Corporation for the use of reports on which some of the discussion of Chap. 20 is based.

Entire credit for whatever coherence exists between this book and the others of the series as well as much of the credit for any internal coherence must go to the Technical Coordination Group.

S. N. VAN VOORHIS.

CAMBRIDGE, MASS.,
July, 1946.

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CHAPTER 1

INTRODUCTION

By S. N. Van Voorhis

The receivers used with radar systems, particularly those operating in the microwave region, are usually characterized by two special properties. In the first place it is important to be able to deal with the weakest possible signal, which may be comparable to or even weaker than the noise in the system. The level of atmospheric noise at these frequencies is very low, so that it becomes necessary to do everything possible to reduce the internal noise in the receiver and to minimize losses in the input circuits. Considerable gain is then required to bring this noise level and signals that are comparable to it up to a usable level. In the second place the signals with which the receiver must deal are pulses covering an extremely wide dynamic range. Accordingly, wide-band circuits are required, and extraordinarily severe requirements are placed on the transient behavior of the whole receiving system.

Almost every common type of receiver has found application at some point. These include

1. The superheterodyne.
2. The superregenerative.
3. The tuned radio frequency (trf).
4. The crystal video.

In many cases the choice among types depends on the relative performance in picking up a signal that is comparable in level to noise. It is well, therefore, to begin with a brief review of the ultimate limitations on receivers imposed by fluctuation noise and of the methods of measuring the performance of actual receivers in terms of such ultimate limits.

1·1. Noise Figure.—To be useful, any receiver must be connected to a source of signals such as an antenna. If this antenna could be placed inside a large absorptive enclosure or black body at a uniform temperature T, it would be in equilibrium with the thermal or black-body radiation at this temperature. At the terminals of the antenna there would appear a voltage equivalent to the Johnson noise generated in a resistance equal to the radiation resistance of the antenna at a temperature T. Therefore, even if a receiver could be produced that had no internal sources of noise whatever, noise would still be introduced into the system from the

antenna, and weak signals would have to compete with this noise. If the receiver does introduce additional noise, the signal must be correspondingly stronger. These concepts may be made quantitative by the use of *noise figure* or *noise factor*.[1] This may be defined with the aid of Fig. 1*a* as follows:

$$F = \frac{\dfrac{S}{kTB}}{\dfrac{S_0}{N_0}}, \tag{1}$$

where F = noise figure of network,
S = available signal power from signal source,
S_0 = available signal power from network,
N_0 = available noise power from network,
k = Boltzmann's constant,
T = absolute temperature of signal source,
B = noise bandwidth of network.

The available power represents the maximum power that may be drawn from a network with a properly chosen load impedance. It is obtained when the load impedance is the complex conjugate of the internal impedance of the network. It will receive a more detailed treatment in Chap. 4. It will also be shown there that the quantity kTB represents the available noise power from a resistor of arbitrary value at a temperature T and that this is merely a restatement of the usual expression for Johnson noise.

The qualitative meaning of Eq. (1) is now clear. A signal, no matter how it arises, has associated with it at least a minimum amount of noise, kTB. If it passes through a network that either amplifies or attenuates both noise and signal without adding additional noise, the ratio of signal power to noise power at the output will be the same as at the input and the noise figure F will be unity. If, however, the network adds additional noise, F will be greater than unity. Equation (1) may be rewritten as

$$F = \frac{S}{S_0}\frac{N_0}{kTB} = \frac{1}{\mathcal{W}}\frac{N_0}{kTB}, \tag{2}$$

where $\mathcal{W}$ = available power gain of network.

A situation that occurs sufficiently often to merit explicit treatment is the case of two networks in cascade shown in Fig. 1·1*b*. The noise figure of the combination F_{12} may be shown to be

$$F_{12} = F_1 + \frac{F_2 - 1}{\mathcal{W}_1}, \tag{3}$$

[1] D. O. North, *RCA Rev.*, **6**, 332, January 1942; H. T. Friis, *Proc. IRE*, **32**, 419, July 1944.

where F_1 = noise figure of Network 1,
F_2 = noise figure of Network 2, under the condition that it is fed from a source whose impedance is equal to the output impedance of Network 1,
$\mathcal{W}_1$ = available power gain of Network 1.

Note the dependence, which is implicitly admitted here, of the noise figure on the internal impedance of the source. It is also assumed that the bandwidth of Network 2 is less than that of Network 1.

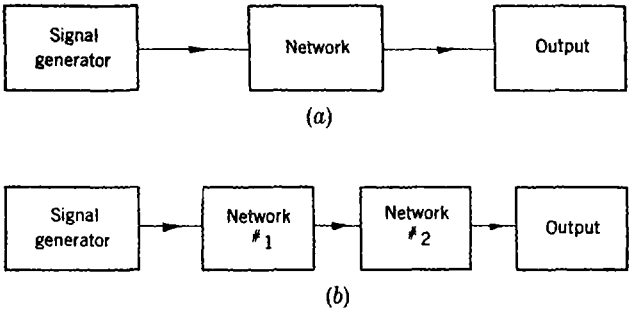


FIG. 1·1.—(a) Setup for definition of noise figure; (b) noise figure of networks in cascade.

By defining a new quantity T_1 from the relation

$$N_0 = kT_1B, \tag{4}$$

Eq. (2) may be rewritten as

$$F = \frac{1}{\mathcal{W}} \frac{T_1}{T} = \frac{t_1}{\mathcal{W}}, \tag{5}$$

where

$$t_1 = \frac{T_1}{T}.$$

T_1 may be called the "equivalent noise temperature" of the network in the sense that the available noise power from the network is equal to the available noise power from a resistor at a temperature T_1. With this change Eq. (3) may be written as

$$F_{12} = \frac{t_1 + F_2 - 1}{\mathcal{W}_1}, \tag{6}$$

where t_1 now refers to Network 1. This is the form generally used in the discussion of crystal units for mixers (see Chap. 2).

Another situation of practical interest may be studied in terms of Fig. 1·1*b* as follows. Suppose that Network 1 consists of an amplifier of an arbitrary number of stages and relatively good noise figure while Network 2 represents some unit of much poorer noise figure. It is desired to know how much gain must be provided in the amplifier, Network 1,

in order that the over-all noise figure be within some specified multiple of F_1. Since noise figures are power ratios, they are often expressed in decibels by the ordinary rules. The preceding requirement might then be stated that F_{12} be not more than x db above F_1. Equation (3) may be solved for the gain under this condition as follows:

$$\mathcal{W}_1 \geqq \frac{F_2 - 1}{F_1} \frac{1}{10^{x/10} - 1}. \tag{7}$$

To illustrate the use of this relation, suppose that a superheterodyne receiver is to be designed at a frequency such as 200 Mc/sec where r-f amplification is feasible. Suppose that the noise figure of a practical r-f amplifier is 7 db and of the mixer is 15 db and that it is desired to have the over-all noise figure not more than $\frac{1}{2}$ db worse than that of the r-f amplifier. The gain required is

$$\begin{aligned} \mathcal{W}_1 &\geqq \frac{31.6 - 1}{5.01} \frac{1}{10^{1/20} - 1} \\ &\geqq 50.1 \text{ or } 17 \quad \text{db.} \end{aligned}$$

From this the number of stages required can be immediately determined.

The noise figure of a receiver may be related to other methods of specifying sensitivity by noting that for room temperature ($T \approx 300°\text{K}$) and a bandwidth of 2.5 Mc/sec, kTB has a value of 10^{-14} watt. Suppose that a receiver of this bandwidth has a noise figure of 10 db and an input impedance of 50 ohms. Then a signal voltage of 2.23 μv developed across the input will produce an output power just equal to the noise power at the output.

Strictly speaking, this calculation is not quite correct, since it assumes that the receiver is matched to the signal source. It will be shown later in Chap. 4 that the best noise figure is usually obtained with some degree of mismatch between the receiver and antenna. The full expression is

$$V_R = \frac{2\sqrt{R_R R_S}}{R_R + R_S} \sqrt{F_R R_R k T B},$$

where V_R = signal voltage developed across input to produce power output equal to noise power output,
R_R = input impedance of receiver,
R_S = internal impedance of source,
F_R = noise figure of receiver with source of internal impedance R_S.

1·2. Noise Figure as a Function of Receiver Type and Operating Frequency.—A procedure has been indicated in the preceding section for making the noise figure of a superheterodyne receiver come within some specified increment of the noise figure of the r-f amplifier alone. For the bandwidths of interest here, of the order of some megacycles per

second, it is profitable from the point of view of noise figure to use r-f amplifiers in this way up to about 1000 Mc/sec. Above this point the noise figure of the amplifier is no longer better than that of the mixer. The range of frequency from about 150 to 1000 Mc/sec may be taken to represent the region in which vacuum-tube amplifiers, while still useful, present special problems differing from those met at low frequencies. It also covers the transition from lumped to distributed circuit parameters. Some of the special problems met in constructing the high-frequency portions of such receivers will be discussed in Chap. 5.

Instead of the mixer and i-f amplifier of a superheterodyne circuit, a detector and video amplifier might follow the r-f amplifier, making a trf receiver. Somewhat similar arguments may be used here to determine the amount of r-f gain required to make the noise figure approach that of the r-f amplifier alone. It must be noted, however, that a noise figure cannot be defined as before for a nonlinear device such as the detector. Instead, a signal level S_1 is found such that the output signal power is just equal to the output noise power. The quantity S_1/kTB may then be used instead of the noise figure in the calculation of necessary r-f gain. It is found that S_1 lies in the neighborhood of 10^{-8} watt or higher for any of the useful detectors, and therefore considerably more r-f gain is needed in a trf receiver. For the range 150 to 1000 Mc/sec, the same r-f amplifiers discussed in Chap 5 may be used ahead of a detector. At frequencies higher than this the complexity and poor performance of r-f amplifiers make trf receiver design in general impractical.

There are cases in which the relatively low sensitivity of the detector-video-amplifier combination is acceptable. In such cases the detector used is almost always a silicon crystal. This has caused this type of receiver to be named "crystal video" which is the term by which it will be referred to in this book. These receivers have been used up to frequencies as high as 10,000 Mc/sec and may find some applications at still higher frequencies. Some of their features are covered in Chap. 19.

The superregenerative receiver occupies a place midway between the crystal video and the superheterodyne as far as sensitivity is concerned. By the use of velocity-modulation tubes, superregenerative receivers have been built for frequencies as high as 10,000 Mc/sec. Such receivers can be made to be small and to have low power drain, but they are limited in the fidelity of reproduction that they can give. Chapter 20 contains a discussion of superregenerative receivers for this frequency range.

1·3. Bandwidth Requirements.—The fact that the signals with which radar receivers must deal are pulses that must be reproduced reasonably well sets a lower limit to the bandwidth of the circuits—r-f, i-f, and video. For pulse lengths of 1 μsec or less, the i-f bandwidth required is usually

at least 1 Mc/sec. Amplifiers with bandwidths of this magnitude usually employ high-transconductance pentodes such as the 6AC7. The coupling-circuit impedance is low compared with the internal impedance of the tube, and the ultimate limit on the product of gain and bandwidth set by the quotient of transconductance and shunt capacitance is a serious restriction requiring constant attention. The problems in the design of i-f amplifiers are discussed more fully in Chap. 6, and those peculiar to radar video amplifiers are treated in Chap. 8. Chapter 18 contains a treatment of some of the methods for obtaining unusually wide i-f amplifiers to give coverage of r-f bands as great as 100 Mc/sec.

CHAPTER 2

DUPLEXERS, MICROWAVE MIXERS, AND LOCAL OSCILLATORS

By M. C. Waltz

The microwave superheterodyne receivers under consideration here differ from those used in lower-frequency applications in two respects: (1) the mixer is almost invariably a crystal instead of a vacuum tube, and (2) no r-f amplification is used ahead of the mixer. In radar receivers an additional factor enters because of the necessity for protecting the receiver from the high-power transmitter. The portion of the radar system that connects transmitter and receiver to a common antenna and directs both transmitted and received power to the proper destination is known as the *duplexer*. In many cases much of the r-f circuit of the mixer is inextricably mingled with the duplexer so that some discussion of the latter is necessary here.[1]

2·1. Microwave Duplexers.— The switching operation from transmission to reception must be completed in a matter of microseconds at most. This requirement suggests the use of some form of gaseous-discharge device. Figure 2·1 illustrates the principle of operation of most of the systems in common use. It uses two spark gaps or switches, one known as the *transmit-receive* (TR) *tube* or *switch* which effectively disconnects the receiver during transmission, and the other known as the *antitransmit-receive* (ATR) *tube* or *switch* which disconnects the transmitter during reception. The high-power pulse from the transmitter breaks down the ATR gap, and the power flows out toward the antenna. The TR gap in the receiving branch likewise breaks down and (if it is designed so that negligible power is required to maintain the discharge) puts a short circuit across the line to the receiver, thereby protecting the delicate input circuits of the receiver. Since the short circuit is a quarter wavelength from the T-junction, the impedance in parallel with the

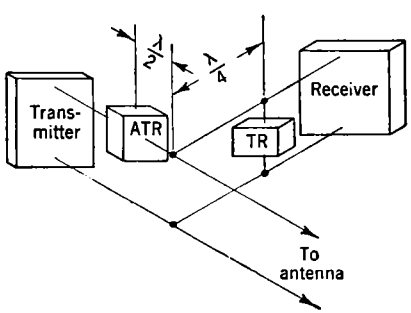


Fig. 2·1.—Duplexing system—two-wire transmission line.

[1] An extensive treatment of duplexers is given in Vol. 14, Radiation Laboratory Series.

antenna line at the junction is very high and does not affect the power traveling toward the antenna. At the end of the transmitted pulse the discharge across the gaps goes out and the system is ready to receive echo signals. The impedance at the T-junction looking toward the transmitter is infinite because there is an open circuit half a wavelength away. Looking toward the receiver there is a matched line, so all the power goes into the receiver.

The requirements for satisfactory transmission of the outgoing pulse are rather easily met, but the requirements for reception are much more stringent.

1. During the transmitted pulse, the power getting past the TR tube into the receiver must be less than 0.1 watt to avoid damage or burnout of the crystal.
2. The TR tube must fire in less than 0.01 μsec, or a preignition "spike" (see Sec. 3·7) of transmitter power will be let through and may burn out the crystal.
3. The gap must deionize in a few microseconds after the end of the transmitted pulse so that echoes from near-by objects will not be unduly attenuated. A typical specification might call for an attenuation of less than 3 db by a time 6 μsec after the end of the transmitted pulse.
4. The received signal must see a reasonably good match into the receiver, and the losses must be kept to a minimum.

Some refinements in the rudimentary system of Fig. 2·1 are necessary to meet the above requirements. Since the fired TR-tube gap is not a perfect short circuit, there will be some voltage V developed across it. If Z is the impedance looking toward the receiver measured at the gap terminals, the leakage power going to the receiver will be V^2/Z. The voltage may be reduced by having the discharge take place in a gas at a low pressure (of the order of a few millimeters Hg), but a still greater reduction of leakage power is necessary. This reduction may be accomplished by a stepup transformer to the gap and an identical stepdown transformer to the receiver line. In the unfired condition the standard line impedance is maintained on either side of the TR tube, but in the fired condition the line impedance Z seen at the gap is much higher and less power goes to the receiver.

The practical method of accomplishing this is by a resonant cavity. Figure 2·2 shows a section through a 1B27 TR tube and associated cavity, with input and output couplings. This assembly is designed for operation near 3000 Mc/sec. The gap across which the discharge takes place is formed by two reentrant cones on the axis of symmetry of the approximately cylindrical cavity. Tuning is accomplished by pushing one cone

in and out on a flexible diaphragm. The unloaded Q of the cavity[1] is about 2000, and with normal input and output loading the loaded Q is about 350. The ratio of power lost in the cavity to input power is nearly equal to the ratio of loaded Q to unloaded Q. Thus in the present case 17.5 per cent of the input power is lost; the TR tube may be said to have a gain of -0.85 db.

Both input and output coaxial lines end in coupling loops, which play the role of the stepup and stepdown transformers. They may be thought

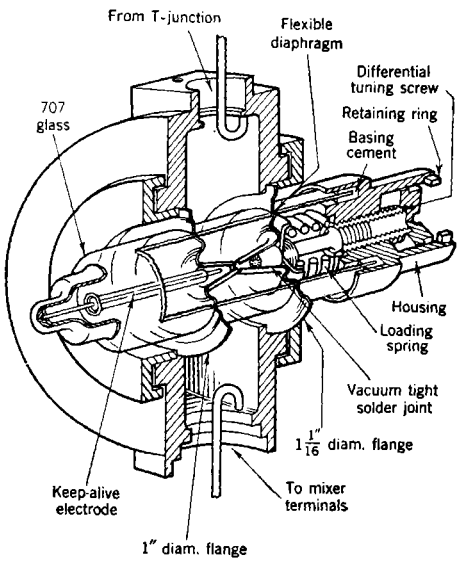


FIG. 2·2.—1B27 TR tube and cavity assembly, loop coupling.

of as single-turn windings which, in proportion to their area, loop more or less of the magnetic field in the cavity. The smaller the loop the higher the stepup ratio and the higher the loaded Q. In a great many cases the output loop is a part of the mixer assembly. Figure 2·5*b* shows such an assembly complete with coupling loop.

It will be noted that the leakage power through the TR tube is more or less inversely proportional to the square of the loaded Q and the loss during reception is directly proportional to the loaded Q. Therefore a compromise value must be reached, such as the value 350 mentioned above. This value corresponds to a bandwidth between 3-db points of about 10 Mc/sec. Accordingly, it will be necessary to provide a tuning adjustment for the cavity. By proper design, this same tuning adjustment

[1] See Vol. 11, Chap. 5, Radiation Laboratory Series, for a discussion of the Q of cavity resonators.

may be made to take care of variation in mixer impedance resulting from change of crystals. If the r-f impedance of the mixer is measured at the crystal, it will be found that there is more of a spread in the resistive component than in the reactive one. If uncorrected, this would result in a change in the impedance presented at the TR cavity that would change the loaded Q as well as introduce undesired reflections and power loss. By placing the crystal an odd number of eighth wavelengths from the cavity, these resistance variations may be made to appear as reactance variations at the cavity and may be tuned out in the adjustment of the cavity.

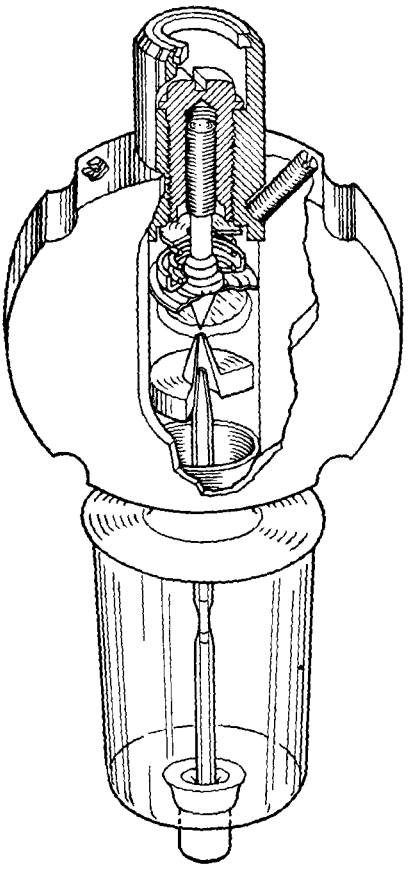


FIG. 2·3.—Cutaway view of 1B24 TR tube.

To ensure rapid breakdown of the TR tube at the beginning of each transmitted pulse, a supply of ions is maintained in the gap by a continuous auxiliary discharge inside one of the cones. This discharge requires an extra electrode, known as the *keep-alive* electrode, which draws about 150 μa from an 800-volt supply. A ballast resistor drops the voltage to about 400 volts at the tube itself. Even with this arrangement there is a considerable spike on the signal transmitted by the TR tube, and this spike plays an important role in the action of AFC systems (see Sec. 3·7.) The keep-alive current must be maintained within the range of about 100 to 200 μa. Too high a current will introduce excessive loss during reception and will cause cleanup of the gas in the tube as well. Too low a value of current will cause failure to protect the crystal.

Figure 2·3 shows a cutaway view of another type of TR tube, the 1B24, which is designed for frequencies in the region of 10,000 Mc/sec. In it the whole cavity is a region of reduced pressure, with a gas reservoir on one side to increase the total volume of gas and so prevent too rapid cleanup. Input and output coupling is done by round windows or irises

that are closed by glass seals. The loaded and unloaded Q and the transmission loss are much the same for the 1B24 as for the 1B27. The bandwidth is somewhat greater, since the operating frequency is higher, but a tuning adjustment for the cavity on the 1B24 must still be regarded as an operating control. As before, the TR tuning may be used to compensate for variation in mixer impedance.

2·2. Noise Considerations.—For consideration of the influence of mixer and duplexer performance on the over-all noise figure of the receiver F, the most suitable expression is the following:[1]

$$F = \frac{(t_c + t_{LO} + F_2 - 1)}{\mathcal{W}_c \mathcal{W}_d}, \tag{1}$$

where t_c = ratio of effective noise temperature of crystal, when supplied with noiseless LO power, to room temperature,

t_{LO} = ratio of increase of effective temperature of crystal, when supplied with actual LO power, to room temperature,

F_2 = noise figure of i-f amplifier,

$\mathcal{W}_c$ = conversion gain of mixer (usually less than unity),

$\mathcal{W}_d$ = gain of duplexer assembly (also less than unity).

Both F's and $\mathcal{W}$'s are expressed as numbers, not in decibels. This expression assumes that the bandwidth of the i-f amplifier is less than that of the r-f circuits and of the circuit coupling the mixer to the i-f amplifier. Of these quantities, t_c and $\mathcal{W}_c$ are determined largely by the crystal. Ordinarily, each type of crystal has minimum values specified for these parameters.

2·3. Crystals.—The most sensitive mixer for microwaves is the crystal type in which a contact on a semiconductor results in rectification of radio frequency. The semiconductor may be one of many different materials such as silicon or germanium, with very small quantities of added impurities. The most common material is silicon. For mechanical protection the semiconductor and metallic "catwhisker" are mounted in a cartridge unit, which is almost always made nonadjustable.

A cross section of a cartridge is shown in Fig. 2·4*a*. The silicon A is soldered to the screw D. A tungsten whisker shown at B is pointed and embedded in the metal at C. The whole assembly is held together by the brass endpieces E and F and the ceramic shell G. The pressure on the contact is maintained by the springlike bend in the whisker and is adjusted in manufacture by the adjusting screw D. The space H is filled with a powder or wax to hold the pieces in place.

Another type of cartridge developed more recently is shown in Fig. 2·4*b*. In this cartridge the silicon C is soldered to the plate A, which is a

[1] This will be recognized as a restatement of Eq. (1·3) with minor changes in notation.

force fit in the tube B. The whisker D is welded to the conductor E, which in turn is held in place by the ceramic washer F. The enlarged section G on the conductor is an r-f transformer for matching purposes. This unit is adjusted during manufacture by carefully forcing in plate A until the cartridge has the correct characteristics.

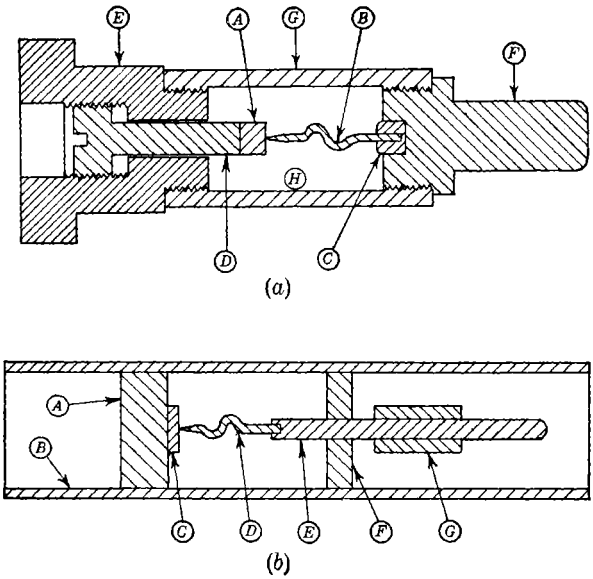


FIG. 2·4.—Cross section of (*a*) crystal cartridge and (*b*) coaxial crystal cartridge.

2·4. Microwave Crystal Mixers.—To make up a complete mixer, a mechanical arrangement is required that will provide r-f circuits, i-f terminals, and suitable means for plugging in the crystal cartridge. Two such mixers for use at 2500 to 3500 Mc/sec are shown in Fig. 2·5*a* and *b*. The one shown in Fig. 2·5*a* is a broad-band mixer with no tuning and is meant to be used without a duplexer. The mixer crystal is shown at B. It is inserted by breaking the transmission line at F. The signal input is at A, thence through the broad-band stub support[1] C to the crystal. The LO power is introduced at H with a termination consisting of a resistor disk indicated at J to the probe L. The coupling of the local oscillator may be varied by the knob K, which varies the depth of penetration of the probe into the main coaxial line. The distance from the sleeve M to the knob K should be about one quarter wavelength. The i-f signal appears at the fitting G. The quarter-wave cup or choke D

[1] See Vol. 9, Chap. 4, Radiation Laboratory Series for additional material on broad-band stubs.

effectively prevents the signal and LO power from getting into the i-f amplifier.

The stub C is needed in the mixer to furnish a return to ground for the i-f and the d-c components developed by the crystal. In order to make

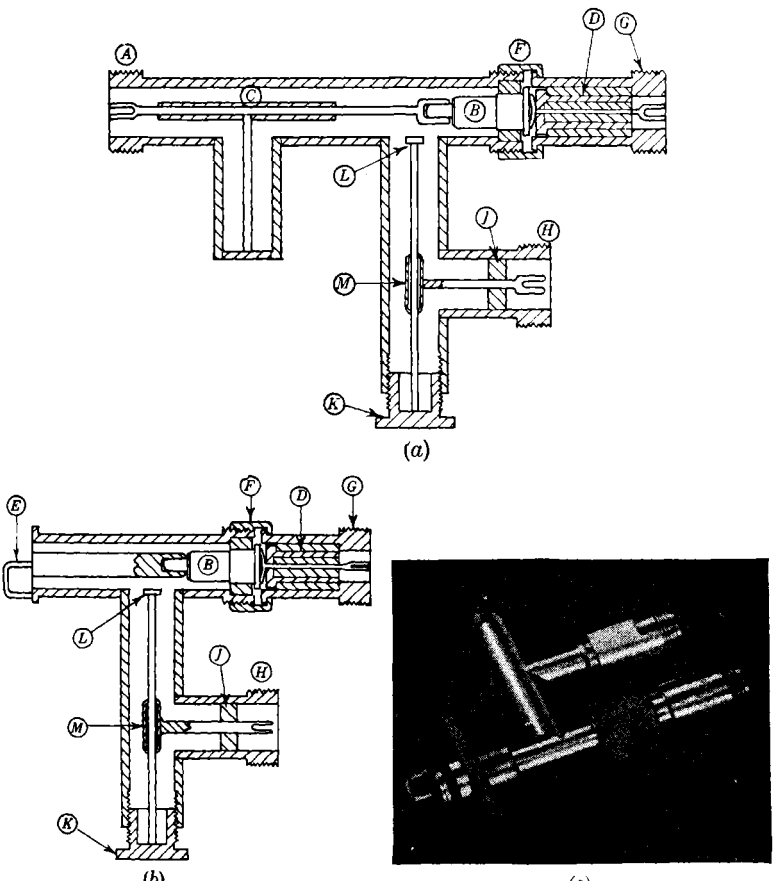


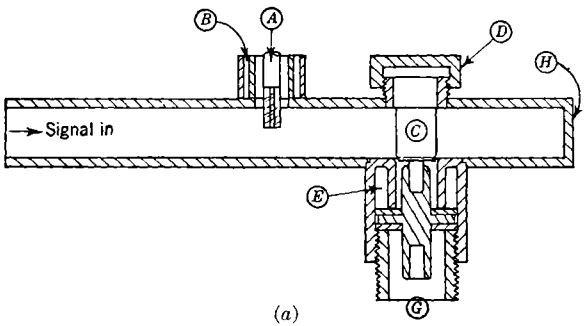
FIG. 2·5.—(*a*) Cross section of broad-band 3000-Mc/sec mixer; (*b*) cross section of 3000-Mc/sec crystal mixer; (*c*) photograph of 3000-Mc/sec crystal mixer.

the termination in the LO line effective, the electrical distance from the end of the probe to the terminating resistor should be some multiple of a half wavelength.

In Fig. 2·5*b* and *c* is another form of mixer for the same frequency as Fig. 2·5*a* but designed to work out of a tuned cavity. In this case the signal is fed through the tuned cavity, is picked up by the coupling loop

E, and thence is applied to the crystal. The loop also furnishes the return path for intermediate frequency and direct current. The LO injection is the same as above.

The r-f impedance of a 1N21 crystal is such that it is a match for the signal if the input line has an impedance of about 50 ohms and if the d-c crystal current due to local oscillator is about 0.5 ma. The i-f impedance seen on looking in at G will be about 300 ohms but will vary with the



(*a*)

(*b*)

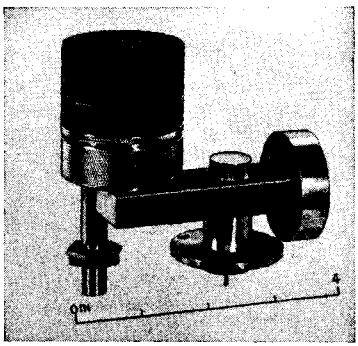


FIG. 2·6.—(*a*) Cross section of typical 10,000 Mc/sec mixer; (*b*) photograph of typical 10,000 Mc/sec mixer.

impedance that the crystal sees on looking out into the r-f line. This result is to be expected if the mixer is thought of as a lossy transformer.

At higher frequencies it becomes more convenient to use waveguide instead of coaxial transmission lines, and as a consequence the mixer changes its form.

A cross-sectional view of a typical mixer for use in the vicinity of 10,000 Mc/sec is shown in Fig. 2·6*a*. A photograph of a similar mixer is shown in Fig. 2·6*b*. In it the local oscillator is behind the crystal instead of in front. This is a section of waveguide with the mixer crystal C placed across the guide. The axis of the crystal is parallel to the electric

field. The signal comes in through the waveguide, and the LO power is injected in the side of the guide with a probe A, the probe in this case being the output probe of the LO tube. The amount of LO power injected is varied by changing the depth of penetration of the probe. The signal is prevented, by the chokes B and E, from going out the LO line and the i-f line respectively. The i-f signal appears at G. The position of the end plunger H is chosen to make the crystal look like a match for the signal.[1]

Another type of mixer using the coaxial crystal cartridge is shown in Fig. 2·7. The direction of travel of the signal in the waveguide D is perpendicular to the page. The signal is picked up on the crossbar C and fed to the crystal shown at A. No r-f signal appears at the ends of the crossbar so that no chokes are necessary in the i-f line B. A mixer

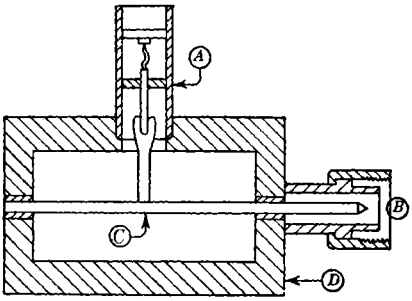


FIG. 2·7.—Cross section of mixer using coaxial crystal cartridge.

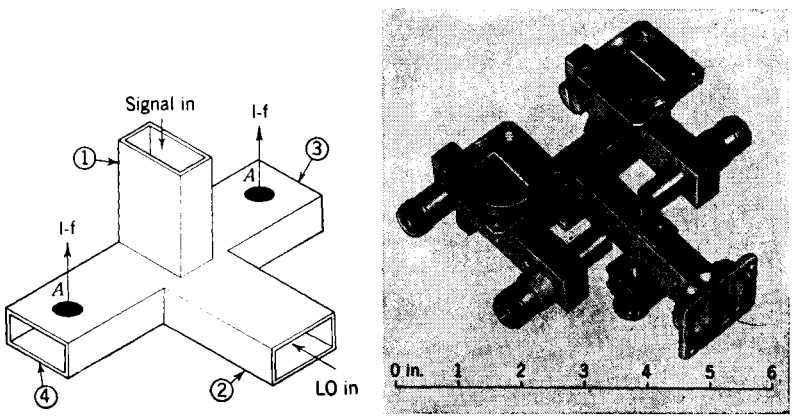
of this type can be designed to operate at all frequencies at which waveguides are used. As in the mixer shown in Fig. 2·6, the local oscillator is injected into the waveguide along with the signal by either a probe or window coupling.

In all the above mixers any modulation sidebands of the local oscillator that are spaced away from the LO frequency by an amount equal to the intermediate frequency will be detected by the mixer as signals and will come through the receiver as such. One of these types of modulation of the local oscillator is noise coming from shot effect and similar phenomena in the oscillator tube. This modulation is equivalent to amplitude modulation of the local oscillator. The noise sidebands so produced fall off in intensity as the frequency departs from the proper LO output frequency, at a rate that is determined by the Q of the oscillator circuits. Therefore, contributions from this source of noise will be less for a relatively high-Q oscillator, such as the type 417, and will also be less for a higher intermediate frequency. This is the origin of the term t_{LO} in Eq.

[1] There is an extensive discussion of the design of such mixers in Vol. 16, Radiation Laboratory Series. Somewhat more complicated designs are shown there which give more uniform and reliable performance.

(1). Also if the local oscillator is not well enough shielded, it is possible that it may be modulated by i-f signals leaking out of the i-f amplifier and may thus cause instability in the receiver or even over-all oscillation of the receiver in some cases.

One method of preventing modulation of the local oscillator from disturbing the receiver is to use a balanced mixer. A form of balanced mixer using a magic T is shown schematically in Fig. 2·8. The magic T is described elsewhere,[1] and its operation explained. It will suffice to say that r-f power fed in Branch 1 (Fig. 2·8*a*) will go into Branches 3 and 4,



(*a*) (*b*)

FIG. 2·8.—(*a*) Schematic diagram of a balanced mixer using a magic T; (*b*) photograph of a balanced mixer using a magic T.

and will arrive at the mixer crystals AA, which are spaced equidistant from the junction, with 180° phase difference. Also, r-f power fed in Branch 2 will go into Branches 3 and 4 and will arrive at the crystals with no phase difference. Thus in the two crystals the LO power with its sidebands is applied with the same phase, and the signal is applied out of phase. The i-f signals will thus appear out of phase in the two crystals. If the two i-f output terminals of the mixer go to a push-pull input circuit in the i-f amplifier, the modulation of the local oscillator will be canceled and the true signals will be retained. The use of a balanced mixer, by largely eliminating the LO noise, will improve the over-all sensitivity of a typical receiver[2] by about 2 db at 10,000 Mc/sec or about 5 db at 25,000 Mc/sec.

[1] See Vol. 9, Radiation Laboratory Series.

[2] The receiver is assumed to have an intermediate frequency of 30 or 60 Mc/sec, and a bandwidth of 2 to 8 Mc/sec and to use a local oscillator such as the 723A/B or 2K25 for 10,000 Mc/sec or the 2K33 or 2K50 for 25,000 Mc/sec.

In the receiver that employs double-mixer AFC (see Sec. 3·8) the mixer takes still another form. In this case it is necessary to have two mixers, one for the regular channel with signal input through the duplexer and another with signal input through an attenuator for the AFC channel. Both of these mixers must get their LO power from the same source. The top schematic view of such a double mixer is shown in Fig. 2·9. It is composed of three sections of waveguide fastened together along their narrow sides. The LO probe projects into the waveguide at *E*. This section of guide is terminated in one direction by the resistive material *F* placed a quarter wavelength from the end of the guide. In the other direction this section of guide feeds two resonant windows *C* and *D*, which are independently adjustable to vary the LO power to the two mixers. The regular mixer shown at *A* is the same as that of Fig. 2·6. The AFC mixer at *B* is modified by having a cutoff waveguide attenuator *G* connecting it to its source of signals. The amount of attenuation required here is discussed in Chap. 3.

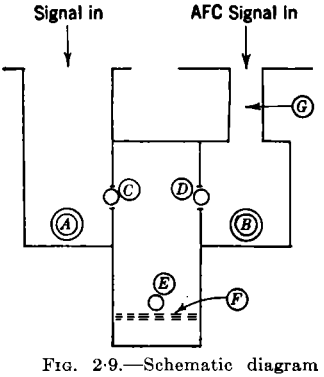


FIG. 2·9.—Schematic diagram of top of double mixer for receivers having AFC.

In use, the two resonant windows are detuned to opposite sides of the LO frequency to increase the cross-attenuation and therefore the isolation between the two mixers.

Another way of applying LO power to the two mixers and still getting isolation between the two mixers is to use a magic T. This method is used when both the regular and AFC mixers are of the balanced type, as is the case in Fig. 2·8*b*.

The sensitivity of a crystal mixer is dependent upon several factors. (1) It should absorb all the signal power incident upon it, because any power reflected is lost. (2) The impedance for the image frequency is influential in affecting the gain of the crystal for the following reason. When the signal and LO frequencies are applied to the crystal, the sum and difference frequencies appear in the mixing. Further mixing of the i-f and LO signals will give rise to a voltage at the image frequency. If this image frequency is returned to the mixer in the correct phase, it will, upon conversion, add to the i-f output. The phase at which it returns depends on the distance between crystal and TR tube and on the impedance of the TR tube at image frequency. If the phase is incorrect when it is reflected to the crystal, it will subtract from the i-f output. The i-f impedance of the crystal is also dependent upon the impedance seen by the image frequency.

In order that the mixer may work properly, it is necessary that it have ample available LO power. This does not mean, however, that the more LO power applied to the crystal the better. Figure 2·10 shows the variation of crystal temperature ratio t_c and crystal conversion loss $\mathfrak{L}$ (the reciprocal of $\mathcal{W}_c$) as a function of the crystal current. Also shown is the over-all noise figure F calculated on the assumption that the i-f amplifier has a noise figure of 3.0 db. As can be seen, the best noise figure is obtained with this particular crystal at a crystal current of approximately 0.4 ma. This is a representative case. If the crystal current is a little higher than this value, not much sensitivity is lost; but if the crystal current is lower, there is a loss in sensitivity. It is standard practice to run the crystal current at about 0.5 ma. About 0.5 mw of LO power is required to give a crystal current of 0.5 ma.

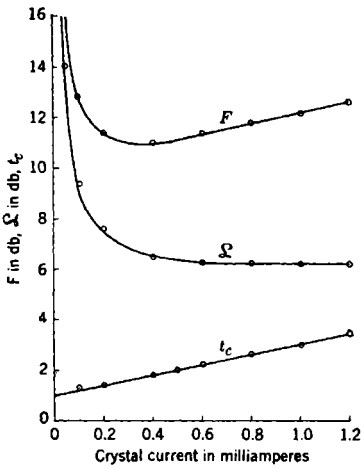


FIG. 2·10.—Curves showing crystal temperature ratio, conversion loss, and over-all noise figure of receiver as function of crystal current.

The available power from the local oscillator should be much larger than the above amount. The more power the local oscillator is capable of giving the more decoupling can be used between the mixer and the local oscillator. This decoupling allows adjusting the load seen by the local oscillator so that the local oscillator is stable. It also prevents loss of signals down the LO line.

The power required in a typical case is shown in Fig. 2·11. To obtain this curve, the power from the local oscillator was reduced by the use of lossy line. The available power plotted on the abscissa is the power that could be obtained at the end of the lossy line into a matched load. As would be expected, there is a very slow variation of noise figure as long as the available power

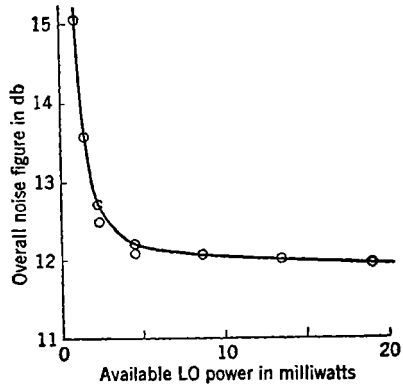


FIG. 2·11.—Curve showing over-all noise figure of receiver as a function of available LO power.

is more than about ten times that required by the mixer. As the available power approaches that required for the mixer, however, the loss of signal down the LO line becomes appreciable. In this case the loss is about 3 db when the available power gets to 1 mw.

Another interesting crystal mixer is one in which a harmonic of the local oscillator is generated in the crystal and then mixed with the signal to give the intermediate frequency. To prevent the receiver from responding to the fundamental, some selectivity, usually in the form of a section of waveguide that will pass the harmonic and not the fundamental frequency, is required. A choke system is also necessary in the LO line to prevent loss of signal. A cross section of such a device is shown in Fig. 2·12. The LO power is fed by the crossbar *B* through the coaxial

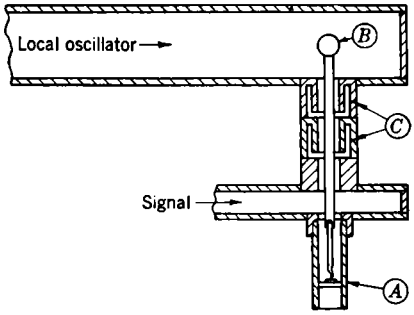


FIG. 2·12.—Cross section of harmonic crystal mixer.

chokes *C* to the crystal *A*. Here the harmonic is generated and mixed with the signal to give the intermediate frequency which is taken off the crossbar. The chokes are tuned to the signal frequency and have very little effect at the local oscillator and intermediate frequencies.

Not many measurements of noise figure have been made on this harmonic mixing. A series of measurements whereby the local oscillator at 5000 Mc/sec was used for signals in the vicinity of 10,000 Mc/sec gave a conversion loss of about 12 db and noise temperature ratio of about 2. Thus over-all noise figures of around 17 db could be expected.

2·5. Thermionic Mixers for Microwaves.—Although crystal mixers have been found to give the best performance in the frequency range from about 2500 Mc/sec up, particularly for cases where good noise figure is of primary interest, thermionic mixers are not completely ruled out. Velocity-modulation tubes may be used as mixers at almost any frequency at which analogous tubes may be made to oscillate,[1] although they are somewhat noisy. They will not be treated here. More conventional tubes

[1] W. C. Hahn and G. F. Metcalf, "Velocity Modulated Tubes," *Proc. IRE*, **27**, 106, February 1939.

such as diodes and negative-grid triodes do find some use up to about 4000 Mc/sec. The operation of such tubes at somewhat lower frequencies is covered in Chap. 5. For the frequency range 2500 to 4000 Mc/sec, transit-time effects becomes so important as to preclude almost any treatment other than a purely empirical one.

In the design of tubes suitable for use in this frequency range two requirements must be satisfied. (1) Transit times must be kept small and uniform. This requirement demands close spacing of elements and a fine mesh for the grid, if one is used. (2) The connection from the actual tube elements to the external circuits must be short and have low impedance. The planar-disk-seal construction used in the so-called "lighthouse" tubes is one approach to these requirements. Both diodes and triodes, such as the types 559 and 2C40, have been made with this construction.

Another method of assembly is used in the CV58, a British diode, which has found some use at 3000 Mc/sec. In it both anode and cathode are cylindrical, with flat ends that are the active areas. The anode is a solid tungsten rod sealed through the glass bulb, and the cathode is a thin nickel cylinder with one end closed by a flat disk. The oxide coating is applied to this end. A helical heater is placed inside the cathode cylinder. The spacing between cathode and anode at operating temperature is about 0.001 in. Differential expansion of the various components of the tube makes the achievement of such a spacing both complicated and difficult.

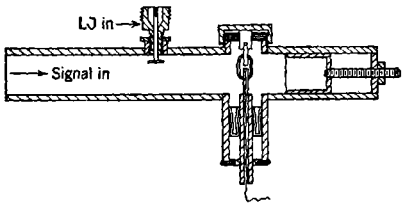


FIG. 2·13.—Cross section of CV58 mixer.

A cross section of a mixer using this tube is shown in Fig. 2·13. There are various methods of tuning the device and injecting LO power. At 3000 Mc/sec, the noise temperature ratio is about 3 and the conversion loss is approximately 13 db, giving an over-all noise figure of about 20 db. Because the diode does not burn out as a crystal does, the duplexer can be much simpler and have a somewhat lower insertion loss.

The LO power required is about 3 mw, which gives a rectified current of about 1 ma.

For some purposes where only medium sensitivity is required, triode mixers have been used. One tube, the 2C40, has already been mentioned. Another tube used as a triode mixer is the WE708A, a grounded-grid tube with cylindrical electrodes. The filament leads are brought out through coaxial leads, and two connections to the plate are also brought out through coaxial leads. The grid is connected to the main metallic base of the tube. In this tube coaxial tuned circuits are attached to the fila-

ment leads, and LO and signal power fed into these lines. Some gain is realized in the plate circuit of this mixer, but the noise output power is greater than that given by crystals or diodes. Over-all noise figures of about 24 to 25 db were measured on receivers using these mixers.

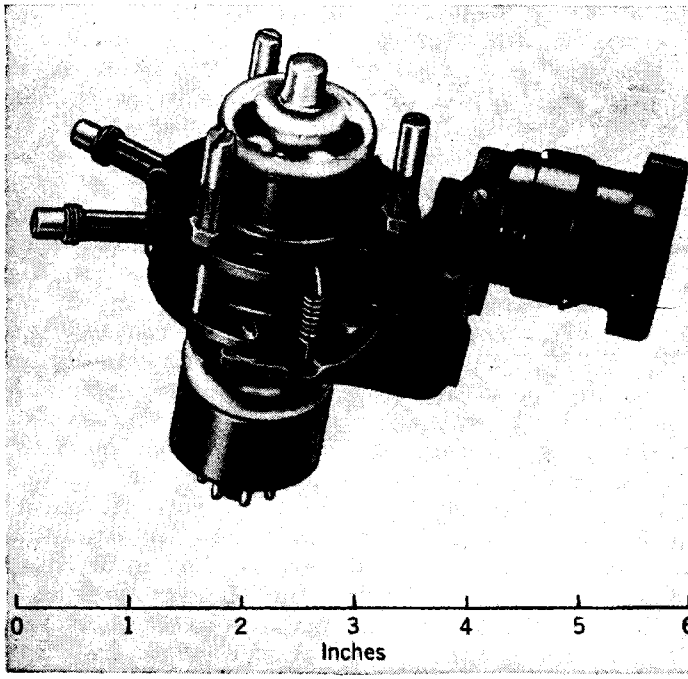
2·6. Local Oscillators.—To be suitable for use in a superheterodyne receiver, an oscillator must possess the five following qualities. (1) It must have adequate power output. As has been seen, the power output must be at least about ten times that actually required by the mixer in order to prevent excessive loss of signal into the local oscillator. For any of the mixers that are likely to be used in the microwave region, 25 mw of LO power is adequate. (2) It must possess adequate frequency stability. If automatic frequency control is to be used with the receiver, this requirement can be relaxed considerably. (3) It must possess sufficient tuning range. Most of the radar receivers described in this book have been designed to cover a relatively narrow frequency range, usually covering not more than 5 per cent of the operating frequency. Accordingly, the oscillator tubes have been built to serve such a purpose. (4) The oscillator must have reasonable power-supply requirements; that is, it should not take an unreasonably high voltage or require an unreasonably high power input. (5) If the receiver is to have automatic frequency control, a fifth requirement may be imposed, namely, adaptability to electronic tuning. It is, of course, always possible to apply an automatic mechanical control to the tuning of a local oscillator, but such systems are likely to become too complicated to be of good serviceability.

Tubes for LO service may be of either the conventional negative-grid type or the velocity-modulation type. At frequencies up to about 1000 Mc/sec the negative-grid type is almost universally used. Such tubes are described in Sec. 5·21. From about 1000 Mc/sec up to perhaps 4000, the lighthouse tube 2C40 or similar tubes can be used. Above about 4000 Mc/sec, the velocity-modulation tubes offer almost the only practical tube for receiver use. They are suitable for use over the entire range from 1000 to 25,000 Mc/sec.[1]

2·7. Reflex Klystrons.—Velocity-modulation tubes may be divided into two principal classes: (1) those in which an electron beam goes through two resonators in succession and (2) those in which there is only one resonator, the electron beam being reflected back through this a second time. Tubes of the first class are rather more difficult to tune, since two resonators must be ganged together or otherwise set to the same frequency. Accordingly they have been used rather little in most

[1] An extensive discussion of oscillator tubes suitable for use in receivers of both velocity-modulation and negative-grid types will be found in Vol. 7 of the Radiation Laboratory Series. See also Vol. 17, Chap. 16, for a listing of many of the types with the frequency range that they cover.

receivers and will not be discussed here. Tubes of the second class, usually known as reflex klystrons, have been by far the most widely used tubes for receivers. They may be subdivided still further according to whether the tuned cavity is entirely within the vacuum envelope or is partially outside the vacuum. As examples of the first class, the types 417 and 726A, both suitable for operation at about 3000 Mc/sec, may be listed, together with the 723A/B or 2K25 suitable for operation at 10,000



(*a*)

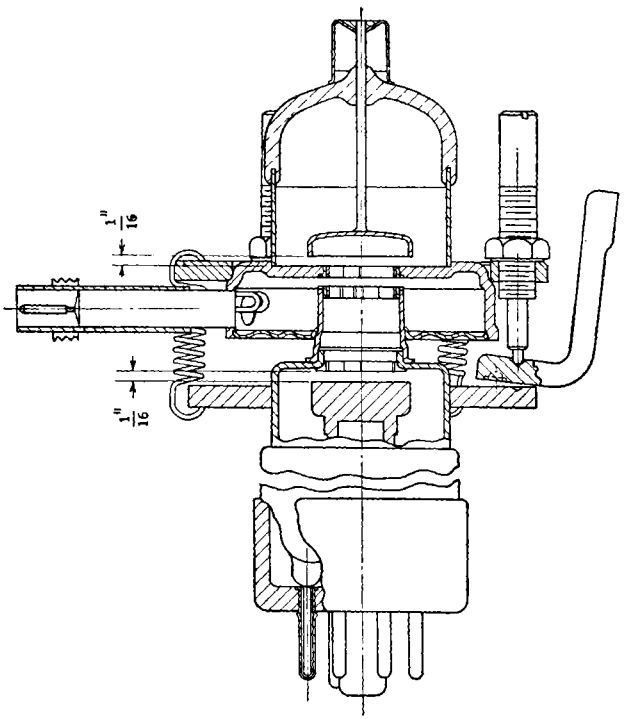
FIG. 2·14.—(*a*) Photograph of type 417 klystron;

Mc/sec. As examples of the second class, the 707A or 707B or the 2K28 may be mentioned. Many of the features of operation of such reflex tubes are described by E. L. Ginzton and A. E. Harrison.[1]

Figure 2·14*a* and *b* shows a photograph and cross section of a type 417 klystron. In this tube tuning is accomplished by mechanical distortion of the cavity. This distortion changes the spacing between the grids of the cavity, which is roughly equivalent to changing the capacity of the equivalent lumped circuit. Major changes in spacing are made by

[1] E. L. Ginzton and A. E. Harrison, "Reflex-Klystron Oscillators," *Proc. IRE*, **34**, 97, March 1946.

adjusting the three supporting screws, and fine changes may be made with the tuning knob provided. Typical power-supply circuits for use with this tube are shown in Fig. 2·15. The resonator voltage may be about +300 volts, and the reflector voltage variable between −50 and −200 volts. Both of these voltages are measured with respect to the



(*b*)

FIG. 2·14.—(*b*) cross section of type 417 klystron.

cathode. The resonator current is usually about 25 ma. The reflector ordinarily draws a negligible amount of current. In many of the velocity-modulation tubes the resonator constitutes a large part of the outside of the vacuum envelope. It is accordingly inconvenient to operate it at a potential 300 volts above ground. For this reason, if possible, a separate power supply is used in which the positive end is grounded. If it is necessary to use a power supply in which this cannot be done, some means for insulating the cavity, tuning controls, and output leads must be taken. See the receiver described in Chap. 14 for an example of this sort.

If a reflex klystron is operated with a fixed resonator voltage, a fixed

load, and variable reflector voltage, and if measurements are made of power output and operating frequencies, a set of curves much like Fig. 2·16 may be drawn. It will be noticed that there are discrete regions of reflector voltage in which the tube does not oscillate at all. The regions in which the tube does operate are usually known as modes. For successive modes the transit time of the electrons, as they go from the resonator to the point at which they are reflected and back to the resonator, differs by just one cycle of the radio frequency. Within any one

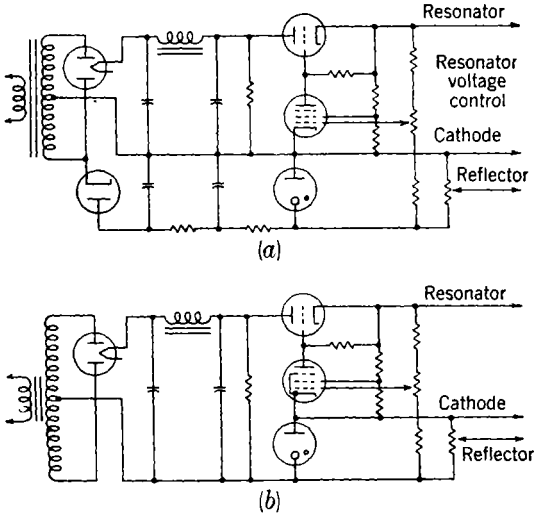


FIG. 2·15.—Typical electronically regulated supplies for reflex oscillators.

mode this transit time is changing continually and is responsible for the shift in operating frequency within the mode. The change in transit time may also be produced by a change in resonator voltage, and with it a corresponding change in frequency. Therefore, regulated power supply is provided for the resonator. Several parameters of importance in receiver design may be obtained from the curves. The first of these is the so-called *electronic tuning coefficient.* This is the slope of the curve of frequency vs. reflector voltage and is expressed in megacycles per second per volt. The second parameter is often called the *width* of the mode between half-power points. It is the frequency spread between points on a given mode at which the power output has fallen to half its maximum value. It is usually taken as a measure of the frequency range over which the tube will furnish useful output with electronic tuning. The third parameter is the width of a mode between zero output points. It is of importance in the operation of the AFC system because

it determines whether or not a receiver may be tuned to the wrong sideband (see Sec. 3·10).

Both the half-power width and the zero-power width of a mode depend on the resonator voltage, both increasing as the voltage is increased. The zero-power width also depends on the loading applied to the oscillator. If a reactive load is applied to the oscillator, it will also change the operating frequency. With the very high operating frequencies of some thousands of megacycles per second, it is easy to have a condition where the load is a number of wavelengths away from the oscillator, the intervening connection being either a coaxial cable or waveguide. This condition may at times lead to a peculiar effect known as a "long-line effect," in which

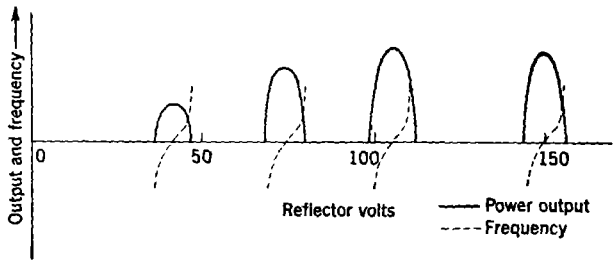


FIG. 2·16.—Modes of oscillation of reflex klystron.

the oscillator will refuse to work over a band of frequencies. If an attempt is made to tune it electronically through this band, the frequency will jump from one edge to the other. In extreme cases, the operation may actually be unstable, the oscillator jumping back and forth from one frequency to another in a random manner.[1]

Certain of the newer tubes such as the 2K45 and 2K50 utilize a different means of mechanical tuning. These tubes are known as *thermally tuned* tubes. In them the resonator is entirely within the vacuum envelope. As before it is tuned by mechanical distortion. This distortion is produced, however, by the expansion or contraction of a strip of metal as it is heated or cooled. This strip of metal ordinarily forms the anode of a separate triode placed within the same vacuum envelope. Accordingly, control of the potential of the grid of this auxiliary triode suffices to tune the main oscillator over its entire operating range. Since the operating frequency of these tubes changes rapidly with changes in any of the supply voltages, they are ordinarily useful only where some form of AFC is provided.

2·8. Triode Oscillator.—A type of triode oscillator that is useful at 3000 Mc/sec used the lighthouse tube in what is called a *reentrant-cavity* oscillator. In this type of oscillator a folded coaxial transmission

[1] A discussion of this point will be found in Vol. 7, Radiation Laboratory Series.

line is used to connect the plate to the cathode. A schematic cross section of such an oscillator is shown in Fig. 2·17. In the diagram, a sleeve *B* is supported by the grid disk of the lighthouse tube *A*. A resistor *G* serves to connect the grid to ground for direct current. A rod *E* is connected to the plate of the lighthouse tube and is slid on and off the plate line by the screw *S*. A movable plunger with a choke joint on the plate line *D* is used to close the line. The feedback line is from the plate through the coaxial line between *B* and *E*, then around the end and

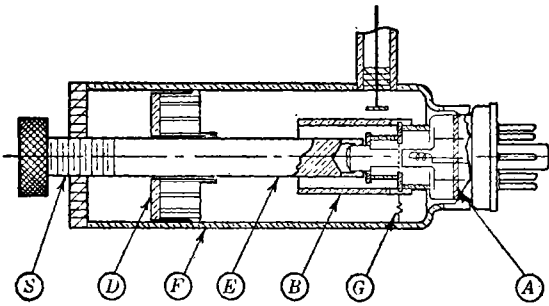


FIG. 2·17.—Cross section of reentrant oscillator.

toward the cathode by way of the coaxial line between *B* and *F*. Pulling the plate lead off the tube changes the oscillator frequency over narrow limits. Large changes in frequency are accomplished by changing the length of the grid sleeve *B*. The position of the plunger *D* determines the feedback.

An oscillator of this type has a good frequency stability as a function of voltage variations on the heater and plate. It is easily tuned by a single knob. The power output with 250 volts on the plate is about 50 mw, which is ample for all LO uses.

CHAPTER 3

AFC SYSTEMS AND CIRCUITS

By G. H. Nibbe, F. E. Towsley, and E. Durand

3·1. Introduction.—The primary consideration of automatic frequency controls,[1] AFC, is the frequency stabilization of some source of r-f energy. This chapter will discuss the various types of AFC that have been used in receivers to maintain the correct tuning relations for optimum performance. These may be roughly divided into two classifications: (1) *difference-frequency* systems and (2) *absolute-frequency* systems. A difference-frequency system of AFC is one in which the difference frequency obtained by mixing two signals is maintained at a constant value, irrespective of the absolute stability of either source. An absolute-frequency system is one in which the receiver is tuned to a specified frequency and maintained there without regard to the frequency of any incoming signals.

All the AFC systems to be described are used with superheterodyne receivers and operate by control of the frequency of the local oscillator in the receiver, which has a conventional fixed-tuned i-f amplifier. The method used to maintain this control or to obtain the desired stabilization is to compare two frequencies and obtain an error voltage that is dependent on the deviation from the desired frequency. This error voltage is used to correct the LO frequency in a manner similar in principle to a servomechanism. As in a servomechanism, there must always be a small error to provide a correction voltage, but the error required may be decreased by increasing the amplification in the system.

Systems using the propagation of r-f energy for the conveyance or gathering of information may be divided for present purposes into three categories:

1. Normal communications systems. The intelligence is impressed on the r-f power sent out by a transmitter by means of some form of modulation and is recovered at a distant receiving point by an appropriate process of demodulation.
2. Propagation study systems. The transmitter serves as a source of constant power, and variations in the signal received at a distant point are interpreted in terms of variations in the medium through which the energy has been transmitted.

[1] The entire AFC problem with primary emphasis on the radar aspect is presented in Vol. 16, Chap. 7, Radiation Laboratory Series.

3. Echo or radar systems. Again the condition of the medium of transmission, in particular the presence of scattering or reflecting objects, is determined by effects on the received signal, but, in contrast to the second category, the transmitter and receiver are located at the same point.

Whether or not the transmitter whose signal to be received is near the receiver in question plays an important part in the choice of an AFC system. As indicated above, radar systems involve a local transmitter. Since much of the work on which the present discussion is based has been concerned with radar systems, the applicable types of AFC will receive the major emphasis.

DIFFERENCE-FREQUENCY AFC SYSTEMS

In one of the first systems for AFC, the local oscillator was maintained at the correct frequency by tuning it electronically with an error voltage obtained from the i-f amplifier.[1]

The method was used with a selective receiver and achieved the dual result of simplifying the mechanical tuning of the receiver and reducing the stability requirements to be met by the local oscillator. The same principles were applied in another AFC system used to maintain the center frequency of an f-m transmitter at an assigned value.[2]

The increased use of very high frequencies and microwaves has placed great demands upon AFC systems as a means of obtaining constant i-f signals, since both the received signal and local oscillator often have poor frequency stability. The use of AFC becomes even more important in the case of a radar receiver, where the transmitted signal frequency may shift or be *pulled* at a fairly rapid rate as a result of varying loads due to the antenna rotation reacting upon the unstable transmitters that are used.

SYSTEMS OPERATING ON THE RECEIVED SIGNAL

AFC systems operating on the received signal can function only after the receiver has been tuned to an incoming signal by other means. A block diagram of such an AFC system is shown in Fig. 3·1.

The discriminator supplies an error voltage indicating the degree of mistuning, and the control circuit changes the LO frequency in the correct direction to reduce the error voltage nearly to zero.

Since the mixer and i-f amplifier of the signal channel are often used for the AFC signal, a conventional a-m receiver may be provided with AFC by the addition of only a discriminator and a control circuit. How-

[1] C. Travis, "Automatic Frequency Control," *Proc. IRE*, **23**, No. 10, October 1935.

[2] I. R. Weir, "Field Tests of Frequency- and Amplitude-modulation with Ultra-high-frequency Waves," *Gen. Elec. Rev.*, May 1939.

ever, the discriminator for the AFC channel sometimes follows a separate i-f amplifier stage bridged across the input to the last i-f stage in the receiver. This arrangement affords isolation between the discriminator and signal detector. Because f-m receivers already have discriminators, only the control circuit need be added. An output voltage vs. frequency characteristic typical of the discriminators used for AFC is shown in

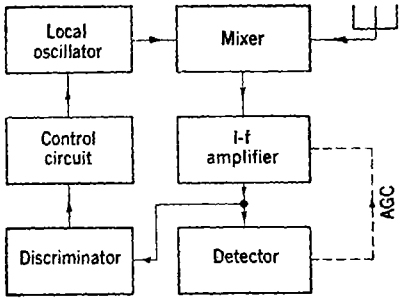


FIG. 3·1.—Block diagram of receiver with AFC.

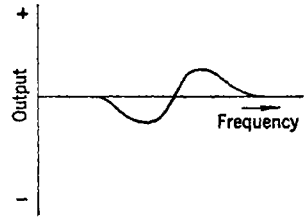


FIG. 3·2.—Discriminator output as a function frequency.

Fig. 3·2, and the circuit diagram of a commonly used discriminator is shown in Fig. 3·3. The i-f signal input to the discriminator can be maintained at a nearly constant value with automatic gain control, AGC, of the i-f amplifier.

The function of the control circuit is to convert the error voltage output of the discriminator into a frequency correction of the local oscillator.

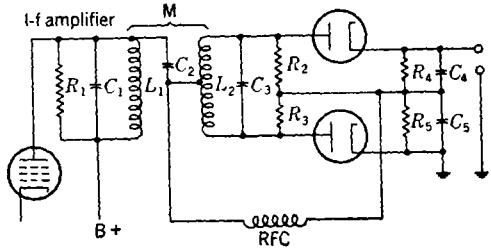


FIG. 3·3.—Foster-Seeley discriminator.

The type of AFC desired, the discriminator characteristics, and the type of local oscillator used set the requirements for the design of the control circuit.

3·2. Control Circuits for Feedback Oscillators.—At high and very high frequencies, ordinary negative-grid oscillators are commonly used for the local oscillator in the receiver. The frequency of oscillation is largely determined by the constants of the associated tuned circuits, and, to effect a frequency change, one of the tuned-circuit parameters

must be varied. It is possible, for example, to vary mechanically the capacity, inductance, or resistance of the tuned circuit with a motor that is part of a servomechanism controlled by the amplified discriminator output voltage. Electronic variation of the tuned-circuit parameters is a more common means of control, however, and numerous approaches to this problem have been suggested.

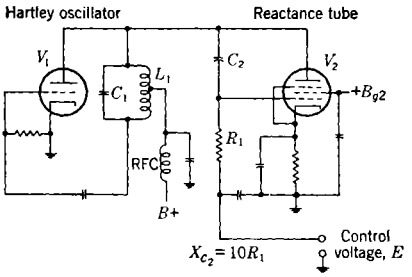


FIG. 3·4.—Frequency control by means of a reactance tube.

A widely used method for obtaining a frequency variation from a voltage variation is the injection of a reactive current into the tuned circuit with a "reactance tube."[1]

A circuit diagram of a Hartley oscillator and one type of reactance tube is shown in Fig. 3·4. Here C_2 and R_1 act as a voltage divider and phase shifter, so that the grid voltage of V_2 leads the plate voltage by almost 90°. The plate current of V_2 is then almost 90° from the voltage appearing across the tank circuit L_1, C_1. The net effect is equivalent to a shunting capacity across L_1, C_1, whose value is $C_i = g_m R_1 C_2$.[2]

A pentode is usually used for V_2 in order to obtain a high value of g_m and thus a large variation in injected capacity. The injected capacity can then be varied over a considerable range by changing the bias voltage E, in Fig. 3·4, thereby changing g_m. Therefore, a frequency variation is obtained that corresponds to the bias-voltage variation. The performance of such a circuit may be shown by a curve of frequency vs. control voltage, such as Fig. 3·5. (Frequency is chosen as the abscissa to facilitate a subsequent use of the curve.) This curve is seen to cover a range of frequencies extending from f_1 to f_2. The end of the curve at f_1 corresponds to the highest usable value of g_m and is marked by the drawing of grid current in V_2. The limiting frequency f_2 corresponds to g_m equal to zero, with complete cutoff of V_2; therefore the frequency of the oscillator no longer depends on the control voltage. Obviously the range $(f_2 - f_1)$ represents the greatest possible frequency coverage of the

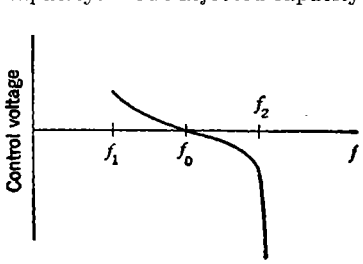


FIG. 3·5.—Oscillator frequency vs. reactance-tube grid voltage.

[1] D. E. Foster, S. W. Seeley, "Automatic Tuning, Simplified Circuits, and Design Practice," *Proc. IRE*, **25**, No. 3, March 1937.

[2] A. Hund, *Frequency Modulation*, McGraw-Hill, New York, 1942, p. 166.

system. The frequency f_0 corresponds to the operating frequency with no applied control voltage. It may be varied by changing the mechanical tuning of the oscillator. This variation essentially displaces the whole curve to the right or left. Another parameter of importance is the tuning coefficient in megacycles per second per volt, which is the reciprocal of the slope of the curve as drawn. Exactly analogous quantities, tuning coefficient and maximum frequency range, will be important properties of any of the other oscillator systems considered.

The operation of the complete AFC system may now be considered graphically by combining the curves of Figs. 3·2 and 3·5, as shown in Fig. 3·6. The zeros on the frequency scale will differ for the two curves by an amount f_s, the frequency of the signal, which is here assumed to be a constant. As drawn, the curves apply to the case where the LO frequency is above the signal frequency.

Figure 3·6*a* corresponds to the case where the mechanical tuning of the oscillator is such as to produce the desired frequency without any help from the control circuit. The discriminator then provides no output, and the entire system therefore remains at this point. Figure 3·6*b* corresponds to the case where the mechanical tuning is off on the high-frequency side. For any frequency between f_0 and f', the output voltage of the discriminator is more than enough to bring the oscillator to that frequency. Therefore f' is the equilibrium operating point for the system. The operating error is then $f' - f''$, and the error that has been corrected is $f_0 - f''$. The ratio $(f' - f'')/(f_0 - f'')$ is thus a measure of the efficacy of the system. It may be expressed in terms of the tuning coefficient T_0 of the oscillator system expressed in megacycles per second per volt and the slope D of the discriminator output expressed in volts per megacycle per second

$$\frac{f' - f''}{f_0 - f''} = \frac{1}{1 + T_0 D},$$

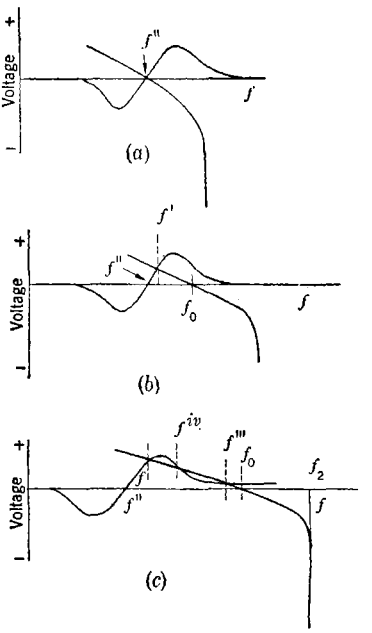


FIG. 3·6.—AFC action. (*a*) Oscillator frequency correct without control; (*b*) oscillator frequency somewhat too high in absence of control; (*c*) oscillator frequency still higher in absence of control.

assuming that both curves are linear over the range involved. This equation is analogous to the expression $1/(1 + \mu\beta)$ for the reduction of error in an inverse-feedback system.

Consider now the situation shown in Fig. 3·6c. Here the mechanical tuning is off still farther on the high-frequency side. There are therefore two new points of intersection of the two curves labeled f''' and f^{iv}.

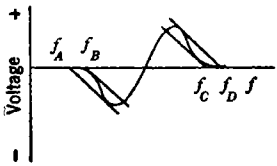


FIG. 3·7.—Pull-in and hold-in range for AFC.

For any frequency between f_0 and f''' the discriminator output voltage is more than sufficient to bring the oscillator to that frequency, but for frequencies between f^{iv} and f''' the discriminator output is insufficient to maintain the frequency at that value. Therefore f''' will be a point of stable equilibrium, as will f'. On the other hand, f^{iv} will be a position of unstable equilibrium. If the system is put in that state, a slight disturbance will cause it to go to either f' or f'''. For this condition of mechanical tuning, the operation is satisfactory as long as the received signal is maintained provided the system has initially been put in the state f' by some means. However, if the signal is interrupted, the oscillator frequency will shift to f_0; and if the signal is then restored, it will

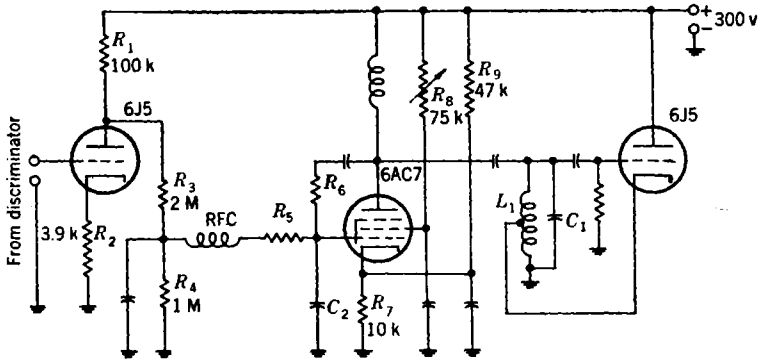


FIG. 3·8.—Direct-coupled amplifier, reactance tube, and Hartley oscillator. At the frequency of operation unmarked capacitors have low reactance. $X_{C_2} \approx 10R_6$; $R_5 \approx 10R_6$.

return only to f''', where operation is definitely unsatisfactory. This behavior may be described by saying that the system will not *pull in* from the condition shown in Fig. 3·6c but it will *hold in* if properly set to begin with.

Two sets of limiting conditions may now be defined for the operation of the AFC system; these limits may be called the *pull-in* range and the *hold-in* range. To avoid useless complication, the maximum range of the oscillator–control-circuit combination will be assumed to be much

greater than either of these ranges, and the tuning curve over the range of interest will be considered to be a straight line. The conditions at the limiting cases may be represented as in Fig. 3·7, where f_A, f_B, f_C, and f_D represent four different mechanical settings of the LO frequency. Since the system will pull into the correct operating frequency under all conditions if the mechanical tuning is between f_B and f_C, this may be called the *pull-in* range. On the other hand, it will hold itself at the proper point, once set there and barring interruptions of the signal, if the mechanical tuning is varied over the range from f_A to f_D, so that this may be called the *hold-in* range. If it is set outside the range from f_A to f_D, it will not lock at the correct point under any conditions.

A direct-coupled amplifier can be added to the system between the discriminator and the reactance tube to increase the loop gain. A circuit diagram of the elements of such an AFC system is shown in Fig. 3·8. The advantages of increased loop gain are greater range, both pull-in and hold-in, and more accurate tuning. In terms of Fig. 3·7, this circuit increases all ordinates of the discriminator-output curve without affecting the oscillator-tuning curves. Since the direct-coupled amplifier inverts the discriminator characteristic, it is necessary to reverse the action of one of the elements if the discriminator characteristic is as shown in Fig. 3·2. The local oscillator can be operated on the low-frequency side of the signal frequency, for example, or the reactance tube can be connected as shown in Fig. 3·8 so that it will inject inductive reactance rather than capacitive reactance. The injected inductance will then be[1]

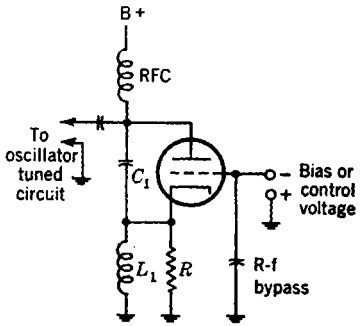


FIG. 3·9.—Reactance-tube circuit.

$$L_i = \frac{C_2 R_6}{g_m}.$$

The reactance-tube circuits shown in Figs. 3·4 and 3·8 depend on a phase-shifting network to supply the grid with a voltage that leads or lags the plate voltage. At higher frequencies the interelectrode capacities of the tube load the phase-shifting network, and transit-time effects add an additional phase lag between grid voltage and plate current. These factors tend to make the design of such reactance-tube circuits difficult. As a result of these effects, a resistive component of plate current as well as a reactive component is injected into the tuned circuit.

[1] *Ibid.*

The resistive component may load the oscillator so heavily that it becomes difficult to maintain oscillation, especially when the resistive component is varied by the controlling voltage.

A reactance-tube circuit[1] that is more suitable for operation in the vhf region is shown in Fig. 3·9. If $X_{C_1} = X_{L_1} = R$, it has been shown that a reasonable value of resistive injection, which remains substantially constant over the useful range of operation, can be obtained. Other reactance-tube circuits that have advantages over the simple forms shown in Figs. 3·4 and 3·8 have been described[2] in the literature.

3·3. Control Circuits for Reflex Oscillators.—Reflex velocity-modulation oscillators such as the 707B, 2K28, 723A/B, and 417 are com-

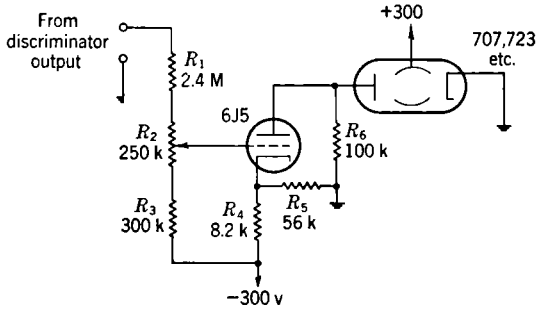


FIG. 3·10.—Control circuit for reflex oscillator tube.

monly used as local oscillators in microwave receivers. Inasmuch as a variation in reflector voltage in these tubes produces a large variation in frequency, it becomes easy to provide AFC for receivers using this type of local oscillator. The reflector is usually operated at a potential from 100 to 200 volts negative with respect to the cathode. A direct-coupled amplifier is often used between the discriminator and LO reflector electrode to obtain greater loop gain. The proper operating potentials are maintained by connecting the amplifier cathode lead to a −300-volt supply. A circuit diagram of this type of control is shown in Fig. 3·10. The proper operating range for the control circuit is selected by adjusting the potentiometer R_2, which varies the reflector voltage from −70 to −270 volts by varying the grid bias of the 6J5 direct-coupled amplifier. There is some variation in gain, of course, as R_2 is

[1] W. R. Rambo, "Notes on a Common-grid Reactance-tube Circuit," RRL Report No. 411-TM-40, Harvard University, Jan. 15, 1944.

[2] M. G. Crosby, "Reactance-tube Frequency Modulators," *RCA Rev.*, **V, 89**, 1940; B. E. Montgomery, "An Inductively Coupled Frequency Modulator," *Proc. IRE*, **29**, No. 10, October 1941; C. F. Sheaffer, "Frequency Modulation," *Proc. IRE*, **28**, No. 2, February 1940.

varied. Many of the circuits to be described later for radar AFC may be used with slight modification.

3·4. Discriminator Circuits. *Discriminators for C-w Systems.*—The discriminator circuit illustrated in Fig. 3·3 has been widely used for the demodulation of f-m signals as well as for a source of error voltage for AFC systems. The theory of the circuit has been presented, and expressions for the bandwidth in terms of the tuned-circuit parameters have been obtained.[1]

If the discriminator is used both for demodulation and as a source of error voltage, as in an f-m receiver, the bandwidth required is determined by the modulation index and maximum modulation frequency. In a receiver for the reception of a-m signals the discriminator bandwidth should be approximately equal to the bandwidth of the i-f amplifier.

The resistors R_4 and R_5 in Fig. 3·3 are usually large compared with the diode impedance in order to maintain high rectification efficiency. In the case of a detector for f-m signals, C_4 and C_5 are made large enough to act as a low-pass filter to eliminate the transmission of an i-f signal component. If the demodulated signal voltage were applied to the reactance-tube grid or to the reflector of a reflex oscillator tube, it would cause frequency modulation of the local oscillator. If the polarities of the system were correct for AFC, the net effect would be to reduce the modulation index or apparent frequency swing of the received signal. If the LO frequency modulation is not linear, the apparent signal modulation will be distorted. For this reason the discriminator output should go through a low-pass filter before application to the control circuit. This filter removes the modulation components from the output voltage, leaving a voltage that is due to the carrier frequency alone.

Discriminators for Pulse-signal Receivers.—Discriminators for operation with pulse signals are similar to ordinary discriminators; but if good efficiency is to be obtained, the frequency response of the output circuit must be such as to let it handle the video-pulse signals properly. The same principles that apply to diode detectors (see Chap. 7) for pulsed signals apply to discriminators, since a discriminator can be divided into two parts, each consisting of a coupling network and a diode detector. The equivalent coupling networks act as though tuned to either side of the desired crossover frequency, and the output voltages of the two sections are subtracted after detection to give the positive and negative characteristic that is desired. Whether the staggered-frequency effect is obtained from a complex network such as that shown in Fig. 3·3 or from two tuned circuits actually staggered is unimportant, except for

[1] H. Roder, "Theory of the Discriminator Circuit for Automatic Frequency Control," *Proc. IRE*, **26**, No. 5, May 1938.

the effect that such coupling networks may have on the impedance of the source of pulsed signal.

A receiver of pulses will have an over-all bandwidth determined by the length of the pulses for which it is designed. Since the pulse has an energy spectrum covering a range of frequencies, the discriminator bandwidth must be large enough to average this spectrum to obtain the center frequency of the pulse. The peak-to-peak separation of the discriminator should therefore be approximately equal to 1.5 times the reciprocal of the pulse length.

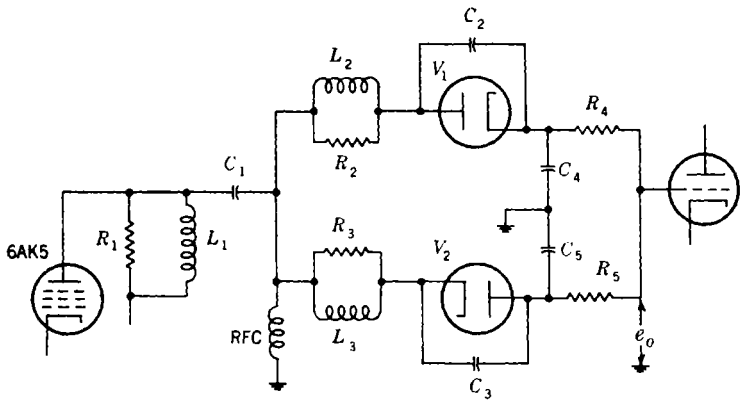


FIG. 3·11.—Peak-reading discriminator circuit, crossover frequency, 60 Mc/sec; V_1 tuned circuit resonates at 65 Mc/sec; V_2 tuned circuit resonates at 55 Mc/sec; bandwidth, 10 Mc/sec; V_1 and V_2, 6AL5's.

The low value of duty ratio found in most pulse applications plays an important part in determining the design of both discriminator and control circuits. For pulse lengths of a few microseconds or less it is practically impossible to achieve frequency correction within the length of a single pulse; therefore, the error voltage and resulting control action are averaged over several pulses. If a simple RC-circuit is used in the discriminator output to do this averaging, the value of the output voltage will be the peak output of the discriminator multiplied by the duty ratio. Since duty ratios as small as 1/1000 are not uncommon, the average voltage out of the discriminator can be very small and so require a large amount of amplification.

To the extent that the discriminator circuit can be made a peak-reading device, this difficulty will be minimized. For the circuit shown in Fig. 3·3, some progress can be made in this direction by making resistors R_4 and R_5 large. The circuit of Fig. 3·11 is a still closer approach to a peak-reading discriminator. At the same time it has excellent gain when used with large bandwidth. The size of the resistors R_4 and R_5

determines the pulse characteristics. If R_4 and R_5 are small, approximately 5000 ohms, the output will be a pulse with short rise and fall times. As R_4 and R_5 are increased, the fall time becomes longer and longer, until, as the resistors approach 1 megohm, the output is essentially flat between pulses if the repetition rate is of the order of 2000 cps. This latter condition is obtained only if the circuit connected to the discriminator has a high impedance, such as does the grid of an amplifier. The action of the diodes will provide a d-c return for the grid of the following stage.

The operation of this discriminator may be analyzed with the aid of the equivalent circuit shown in Fig. 3·12. Assume that the parallel combination L_2, C_2, and the capacity of the diode V_1 resonates at a higher

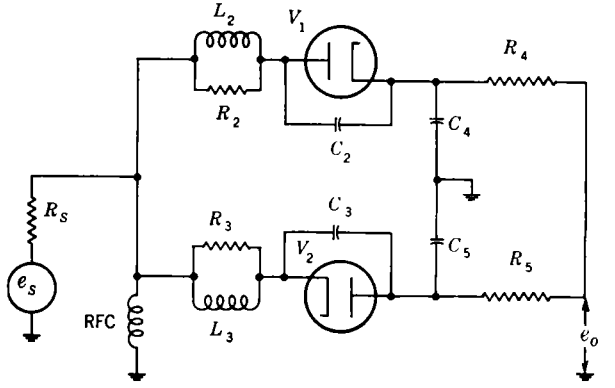


FIG. 3·12.—Equivalent circuit for discriminator shown in Fig. 3·11.

frequency than the crossover frequency whereas the combination L_3, C_3, and diode V_2 resonates at a lower frequency than the crossover frequency. Suppose that these frequencies are 65 and 55 Mc/sec respectively. If a pulsed 62-Mc/sec signal e_s is applied to the two branches, the voltage at C_4 will be a positive pulse of magnitude e_1, and the voltage at C_5 will be a negative pulse e_2. The positive pulse will be larger than the negative. At the end of the pulse the diodes will become nonconducting. Since C_4 and C_5 will be charged to the peak value of the voltage, the only path of discharge is then through R_4 and R_5. The voltage e_0 appearing at the junction of R_4 and R_5 will be $e_0 = (e_1 + e_2)/2$. Capacitor C_4 will then discharge toward e_0, while C_5 discharges toward that same potential; but at the time when the potential of C_5 becomes zero, V_2 will become conducting and C_4 will then discharge toward ground potential. If R_4 and R_5 are large enough, V_2 will not become conducting until after the next pulse, and e_0 will then consist of constant direct voltage indicating the deviation of the signal frequency from the cross-

over frequency. The value of e_0 will be the same for a c-w signal and a pulsed signal having equal peak-to-peak amplitudes, provided that the pulse is long enough to charge C_4 and C_5 up to the full value.

The discriminator characteristic is the same as that shown in Fig. 3·2 for pulsed or c-w signals. The bandwidths of the individual series-resonant circuits are made approximately 0.4 of the desired peak-to-peak separation, and the tuned circuits are tuned to the opposite sides of the crossover frequency by 0.5 of the desired peak-to-peak separation. At the crossover frequency, to which L_1 and C_1 are tuned, the discriminator reflects zero reactance to L_1, and therefore C_1 resonates with L_1 at the crossover frequency. The resistor R_1 loads L_1 and C_1 to approximately the same bandwidth as the discriminator peak-to-peak separation.

Another discriminator circuit is shown in Fig. 3·13. The output circuit is arranged in the same manner as the Foster-Seeley type, but the reversed diode type of connection shown in Fig. 3·11 could be used. L_1 and C_1 are resonant at the desired crossover frequency, as is L_2 with C_4 plus the series capacity of C_2 and C_3. The difference between C_2 and C_3 determines the peak-to-peak separation within the limits of the bandwidth of the tuned circuits. For a fixed difference in capacity between C_2 and C_3, the crossover frequency is determined by L_2 and C_4 with very little interaction from C_1 and L_1.

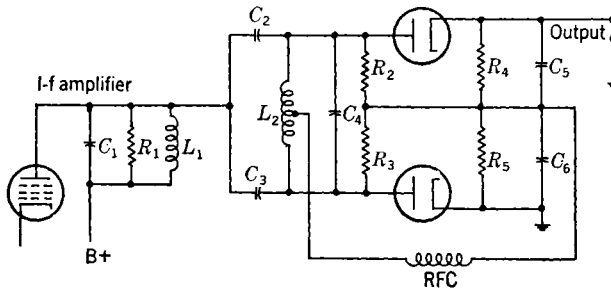


FIG. 3·13.—30-Mc/sec discriminator circuit.

The discriminator circuits mentioned above for use with pulsed signals are also suitable for c-w signals. The principles involved in obtaining good response to pulses apply directly to obtaining good transient response and for that reason should be adhered to in the design of discriminators for use as demodulators for f-m signals. In particular, it is sometimes not realized that C_2 is in shunt with C_5 in the circuit shown in Fig. 3·3 and that as a result C_5 should be smaller than C_4 by the value of C_2.

[1] A mathematical analysis of this circuit is found in Vol. 16, Sec. 7·8, Radiation Laboratory Series.

SYSTEMS OPERATING ON A LOCAL TRANSMITTER

3·5. General Requirements.—In this class of system, the signal transmitted directly from the local transmitter to the receiver (spill-over signal) is used for frequency-control information. Signals reaching the receiver after reflection (echo signals) from radar targets at extreme ranges are so low in power that the i-f signal voltage output from the mixer is equal to or slightly more than rms noise voltage at the same point. Although these signals of the same magnitude as noise furnish useful information at the radar indicator, noise interference effectively prevents the use of such signals for AFC information. Large echo signals, well above noise, are obtained from short-range targets, but these large echo signals cannot be depended upon, since short-range targets are not always available. The spill-over signal, in contrast, is always present and is much greater than receiver noise. Also since it is nearly constant in power, its use for AFC has the added advantage of eliminating the necessity for AGC or limiter action in the AFC amplifier.

AFC Requirements.—There are two requirements to be met by the AFC system and the local oscillator.

1. There is a band of frequency over which the system must operate without manual readjustment. This band is determined by the frequency shifts expected from the transmitter and local oscillator as a result of variation of temperature, voltage, load, etc. As frequency is shifted across this band, LO output to the mixer must remain substantially constant in order that receiver sensitivity will not be changed by AFC action. A typical limit is ± 1.5 db in the power received by the mixer from the local oscillator.
2. The control method must be of such nature that frequency shifts can be "followed" with sufficiently small time lag. If transmitter frequency increases, LO frequency must increase along with it in order to keep the mixer output at a constant intermediate frequency.

These two requirements must, of course, be met by any AFC. In radar AFC the second requirement is relatively more important than it is in other types of AFC, since rapid transmitter-frequency changes are likely to be encountered.

Radar AFC has been developed predominantly in the frequency band from 3000 to 30,000 Mc/sec. There have been two principal factors that have been influential in this development.

1. As the operating frequency of a superheterodyne receiver is increased, it is usually necessary to improve the frequency stability of the oscillators employed in the system. The allowable fre-

quency variation in cycles per second is set by the channel bandwidth, which depends on the service (speech, music, or picture transmission; radar; etc.). Accordingly, since the channel bandwidth remains the same, the allowable *percentage* frequency variation decreases at the higher frequencies, and better frequency stability is required.

In the microwave region resonant elements of very high Q are available for improving oscillator stability. The use of such circuit elements has resulted in sufficient frequency stability for radar applications up to about 3000 Mc/sec, since the bandwidths are fairly great. Above this range, however, with the velocity-variation oscillators used at the present time, temperature and voltage changes result in frequency variation so large that some form of AFC is necessary in many applications in order to eliminate continuous retuning adjustments by the radar operator.

2. Small mechanical imperfections or small mechanical distortions produced in the antenna system as the radar scans are not serious when the longer wavelengths are used. However, at the shorter wavelengths, particularly in the 1- to 10-cm range, electrical path-length changes between the transmitter and antenna occurring during the radar scanning cycle may produce a transmitter-load variation that causes a change in transmitter frequency. AFC is then needed in the receiver in order that this frequency change may be followed by the local oscillator.

Below 3000 Mc/sec, high-frequency triodes with mechanical tuning are often used as local oscillators. It is possible to shift frequency by changing voltage in these oscillator circuits, but power output changes so rapidly with voltage that electronic tuning is not useful for radar AFC; that is, AFC requirement 1 is not realized.

Electronic and mechanical tuning are available with the velocity-modulated tubes used as radar local oscillators at 3000 Mc/sec and above, and the electronic tuning is adequate to meet both the AFC requirements. Electronic AFC, involving only electronic elements in the control circuits, is simpler in application than mechanical AFC, in which an electromechanical linkage must be used for conversion of the electrical error signal into a mechanical correction. Most of the radar AFC methods in the 1- to 10-cm region are electronic rather than mechanical in nature. The present discussion will be confined to systems applicable to pulse radar systems.

3·6. R-f Problems.—The proper operation of radar AFC depends in large measure on the correct design of the r-f circuit components, such as the mixer, the local oscillator, and the waveguide connections and

attenuators. Since these are considered in detail elsewhere, the emphasis in this chapter will be placed on the requirements and specifications that must be realized in the r-f circuits in order to obtain correct AFC operation rather than on the construction and design details for these components.

LO Shielding.—Faulty AFC operation due to leakage of transmitter r-f power or of modulator video pulse power into the LO system, with resulting extraneous frequency shifts, is possible if the oscillator shielding is not done carefully. The size of holes in the shielding for cooling or tuning is usually reduced to the smallest possible value consistent with the necessary heat dissipation and the desired ease of mechanical tuning. The shield must be solidly grounded. The wires connecting the local oscillator to its power supply must be well bypassed and shielded. In the low- and medium-power radars, with transmitter pulse powers ranging up to 50 kw, pickup on connecting wires is low enough so that the local oscillator and its power supply may be mounted in separate chassis. With sets making use of higher transmitter power levels, or where there is excessive power leakage from unavoidable gaps in the transmitter-to-antenna line, the local oscillator and power supply are mounted together in one well-shielded chassis in order to eliminate wiring pickup troubles.

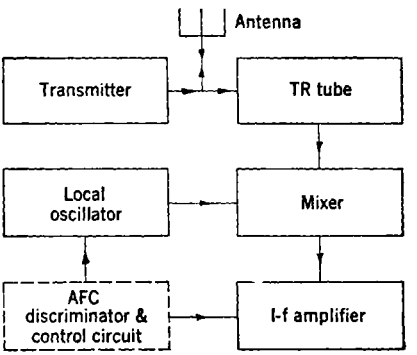


FIG. 3·14.—Block diagram of radar system with AFC.

Mixers.—Two serious problems occur in the design of the mixer. These are *error signal distortion* and *i-f harmonic production*, and they must be suppressed or avoided in order that the AFC may lock the oscillator frequency at the correct value. The description and solutions of these problems are contained in the following discussion of the various mixer systems used in radar AFC.

In the 1- to 10-cm radar systems, the same antenna is used both for transmission and reception (Fig. 3·14). The transmitted pulse power is in the range from 1 kw to 1 Mw. Echo pulse powers may be as small as 10^{-13} watt. The mixer and receiver must be protected from overload and burnout due to the large amount of power in the pulse received directly from the transmitter and yet must be able to detect the very small echo pulses. The TR tube described in Sec. 2·1 operates to furnish the necessary protection during transmission. There is a small amount

of TR-tube coupling, however, and some of the initial pulse power does get through to this mixer. This pulse is the r-f signal used for AFC in early radar systems. Since one mixer is used both for AFC and for echo signals, such systems have been called "single-mixer" AFC, in contrast to those using separate mixers for AFC and echo signals.

3·7. The Single-mixer System.—Since the input signal for AFC comes in through the TR tube, the mechanism of breakdown in the gas discharge has an important part in determining the characteristics of the signal and so of the AFC operation. Studies show that the envelope of the r-f power transmitted by the TR tube has a form more or less like Fig. 3·15. The sharp spike on the front edge is the result of direct transmission of the transmitter power before the arc forms. It is exceedingly short, a very small fraction of a microsecond, and its exact shape depends on a large number of factors: rate of rise of transmitter pulse, Q of TR cavity, conditions inside TR tube, etc. Furthermore the energy in the spike is fairly high. Whether or not LO power is applied to the mixer, there will be a signal produced by rectification in the mixer crystal that will have essentially the same shape as Fig. 3·15. Since the spike is so short, its energy spectrum will extend up to frequencies perhaps as high as some hundreds of megacycles per second, and the distribution of energy in this spectrum will be affected by all of the things that affect the shape of the spike.[1]

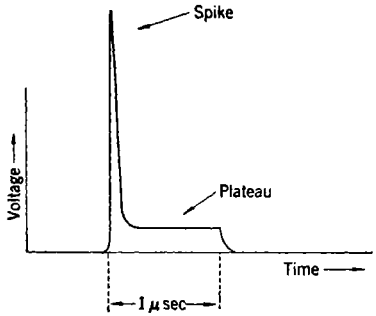


FIG. 3·15.—Video pulse output from TR-tube leakage.

If such a signal is applied to the discriminator, the output signal will be the result of integrating the product of the discriminator transmission function and the signal frequency spectrum over all frequencies. Two

[1] It will be recalled that the frequency spectrum of a single rectangular pulse of length τ, given by the conventional Fourier transform, is of the form

$$E(\omega)\, d\omega \propto \frac{\sin^2 x}{x^2}\, d\omega,$$

where $E(\omega)\, d\omega$ is the energy in the angular frequency range $d\omega$ and $x = \frac{1}{2}\omega\tau$. This function has zeros at $x = n\pi$ ($n = 1, 2, 3, \cdots$). The principal part of the energy is contained in the region between the origin and the first zero or in a range of frequencies between zero and the reciprocal of the pulse length. If, instead of a simple rectangular pulse, a high-frequency wave is modulated by such a pulse, similar expressions hold, with energy distribution extending a like amount on each side of the frequency corresponding to the original unmodulated wave.

typical possible cases are shown in Fig. 3·16. In the first case the spike is assumed to be so short that the energy distribution in the vicinity of the intermediate frequency is practically constant, whereas in the second case a somewhat longer pulse has caused a zero in the energy spectrum to fall near the intermediate frequency. Suppose now that the output from the discriminator is observed for these two cases in the absence of local oscillator. In the first case there will be practically no output, since the regions on each side of the center frequency contribute nearly equally and so cancel. However, in the second case there will be a relatively considerable positive output, in other words, the same sort of output as if the local oscillator had been running and had been producing too low a difference frequency. This output will then act just the same as a proper error signal and will cause the LO frequency to be shifted incorrectly. If LO power is now applied to the mixer, the proper error signals will be produced in addition to the spurious signals due to the spike. If these spurious signals are small compared with the proper ones, the operation of the system may be fairly good; but if the two are comparable in size, considerable disturbance or complete inoperation is likely to result.

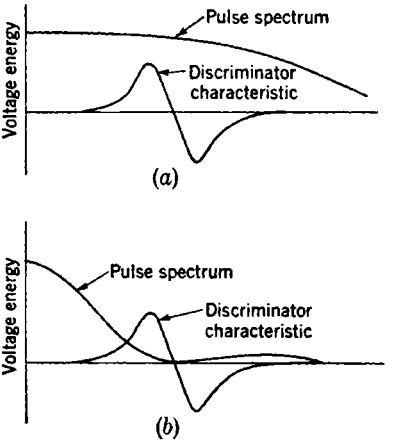


FIG. 3·16.—I-f spectrum of video pulse from TR-tube leakage. (*a*) Short pulse; (*b*) long pulse.

Various palliative measures have been used, in some cases with reasonable success. Probably the earliest such measure was the intentional distortion of the discriminator characteristic. If in Fig. 3·16*b* the low-frequency peak of the discriminator curve is made considerably smaller than the other, it may be possible to reach a condition where there is little or no error-signal output in the absence of local oscillator, and yet with local oscillator applied the correct error signals will be produced. As long as the details of the shape of the spike remain unchanged, operation will be reasonably good, but any change in transmitter or TR tube is likely to throw things completely out. A second method of more general utility is the elimination of the spike signal. Since the spike, of necessity, comes at the front edge of the transmitter pulse and is over considerably sooner, it may be possible to apply a blanking pulse to the AFC system so that the spike will not be transmitted but the last part of the transmitted pulse will be. For this to be successful, the band-

width of the system up to the point where the blanking pulse is applied must be wide enough so that the duration of the spike signal after passing through the system is still much shorter than the transmitter pulse.

The net result of these difficulties has been to eliminate single-mixer AFC systems almost completely from consideration in new radar system designs.

3·8. The Double-mixer System.—The next logical step in improvement of AFC operation is to use a separate mixer for AFC. A total of two converters is thus used in the radar, one for echo-signal mixing and one for obtaining the AFC signal. This type of system is usually called double-mixer AFC, although separate-mixer AFC would be more precise.

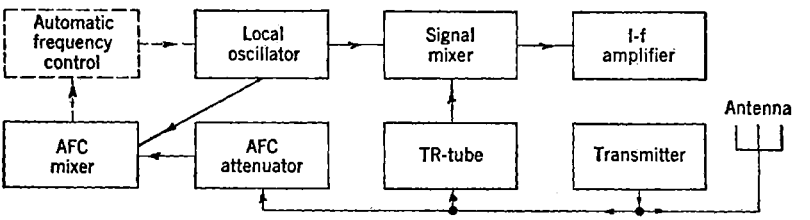


FIG. 3·17.—Block diagram of double mixer AFC.

In Fig. 3·17 is shown a diagram of the double-mixer connection. The AFC chain is now entirely separate from the echo-signal mixer and amplifier. Accordingly, since the AFC chain does not work on echo signals but makes use only of the main pulse, a fixed attenuator section may be inserted before the AFC mixer. There is no need for a TR tube, since the fixed attenuator furnishes protection for the AFC mixer, and the AFC signal distortion due to the TR-tube spike may therefore be eliminated.

The local oscillator must now supply power for two mixers rather than one, but this requirement is comparatively easy to accomplish. For example, the oscillators commonly used, such as the 707B, 2K28 and 723A/B tubes, have normal power outputs of the order of 25 to 50 mw. In the operation of the signal mixer, the conversion gain and the noise figure are of primary importance in determining how much LO power should be made available at the mixer. This adjustment is discussed in Chap. 2. The oscillator power at the crystal that produces maximum signal-to-noise ratio, that is, the greatest sensitivity for the small echo signals returned from targets at extreme range, is approximately $\frac{1}{2}$ mw, and this power corresponds to rectified crystal current of 0.5 to 1.0 ma.

In the AFC mixer the principal criteria are the harmonic ratio and the suppression of leakage interference. The influence of these factors on correct AFC action also results in an optimum mixer power contribution from the local oscillator of about $\frac{1}{2}$ mw. This is convenient for field

operation in that the coupling to both mixers may be adjusted to give the same value of rectified crystal current in both circuits.

The relatively large ratio between oscillator output power and mixer requirements permits loose coupling between the two, which is helpful in avoiding loss of signal power into the oscillator. With tighter coupling, resonances in the mixer and r-f circuits react on the oscillator and affect its operation. For instance, with some of the lower-power 723A/B's used at 10,000 Mc/sec where the normal coupling must be increased to obtain the necessary ½ mw at the crystal, peculiar effects are produced, as shown in Fig. 3·18*b*. Irregularities such as these may be tolerated if they are outside the frequency band that the local oscillator is required to cover. If a large irregularity occurs near the operating frequency, however, AFC requirement 1 is certainly not realized, and correct functioning of AFC will be difficult if not impossible, since the operating frequency may drift into the irregularity during operation. It is sometimes possible to shift these resonances by mechanical tuning and thus to remove them from the vicinity of the operating frequency, but this solution is not always feasible. This effect is, of course, made negligible by broad-banding the mixer system, that is, by reducing the resonances in the frequency-coverage band.

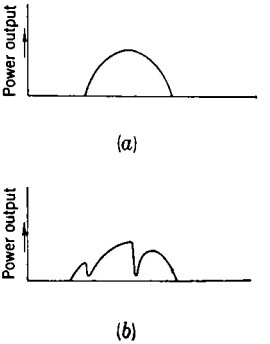


FIG. 3·18.—Effect of load on oscillator modes. (*a*) Normal tube with loose coupling; (*b*) low-power tube with tight coupling.

AFC Attenuator.—A small fraction of the transmitter pulse energy is obtained for the AFC mixer through a sampling connection in the line between the transmitter and antenna. An r-f attenuator is inserted between this connection and the mixer. The attenuation is usually produced by a section of cylindrical waveguide below cutoff. The action is essentially like that of a high-pass filter, which passes energy above the critical frequency but attenuates the applied power below this point. The diameter of the cylinder is made small enough so that the critical or cutoff frequency is greater than transmitter frequency. The attenuation in decibels is proportional to the cylinder length, and this length is adjusted to give the required fraction of transmitter power at the mixer.

The attenuation value is made very large, from 75 to 90 db, in order to reduce the transmitter power, ranging from 35 to 1000 kw, down to 1 or 2 mw at the crystal. All leakage paths must accordingly have more attenuation than this value. A dissipative pad inserted at the mixer may be used to provide 20 to 40 db of the total. This pad serves to reduce the effect of leakage into the cylindrical attenuator connections, since only about 50 db is then needed in the cylindrical section.

Harmonics of the transmitter output may be above the cutoff frequency of the attenuator and are then passed with negligible attenuation. This harmonic energy produces, at the attenuator output, a much larger ratio of harmonic to fundamental energy than exists in the transmitter pulse itself. The LO input to the crystal may also contain harmonic energy, or both transmitter and LO harmonics may be generated by the nonlinear action of the crystal. In practice, considerable harmonic content is found in the crystal output. There are then several values of LO fundamental frequency at which an i-f pulse of a given frequency may be produced. If, for example, 30 Mc/sec is the desired intermediate frequency, a mixer output at this frequency may occur not only when the LO frequency is 30 Mc/sec away from the transmitter frequency but also at separations of 15, 10, $7\frac{1}{2}$, . . . Mc/sec. If the second-harmonic 30-Mc/sec pulse is large enough in amplitude to work the AFC circuits, the local oscillator may be maintained or locked at a 15-Mc/sec interval from the transmitter instead of the 30-Mc/sec interval for which the signal channel is designed. This situation must be avoided, since sensitivity and hence range will be greatly reduced.

Transmitter harmonics may be suppressed before the AFC signal reaches the crystal by the insertion of resistance strips, mode-damping fins, or polyiron slugs into the cylindrical attenuator. In the proper positions and orientations in the guide, these act like low-pass filters, in that the harmonics are attenuated more than the fundamental component in the transmitter pulse.

Harmonic generation at the crystal must also be suppressed. The second harmonic crystal output is, next to the fundamental, the strongest component and therefore the most likely to produce erratic AFC operation if measures are not taken to suppress it. The ratio

$$\frac{\text{Second harmonic crystal output voltage}}{\text{Fundamental crystal output voltage}}$$

is called the "harmonic ratio," and the figure is an important criterion in the mixer power adjustments. The harmonic ratio is a function of both LO power at the mixer and the transmitter power reaching the crystal through the AFC attenuator (transmitter sample).

Experiments have been made at 10,000 and 30,000 Mc/sec to determine what total power should be applied to the mixer during the pulse (see Vol. 16, Sec. 7·6). It was found that the harmonic ratio generally increases but slowly with total power in the range from 1 to 2 mw, where partial saturation occurs in the crystal. This slow rate of change with power provides only a slight improvement, that is, decrease, in the ratio as power is decreased below the 1-mw level, whereas interfering leakage power soon becomes intolerable. Accordingly, operation at the 1- to

2-mw level gives about the best harmonic ratio obtainable with sufficient reduction of leakage interference.

With these values of input power, the crystal is essentially a linear detector. Then, if the transmitter sample and LO contribution are equal, the harmonic production will be greater than if the two inputs are unequal.[1] Since the interfering effect produced by a given amount of leakage power decreases as the transmitter sample is increased, this input power is made the larger fraction of the total power. However, the difference-frequency output with linear detection is largely independent of the amplitude of the stronger input signal and is nearly proportional to the strength of the weaker signal;[2] consequently the LO power contribution must not be reduced too far.

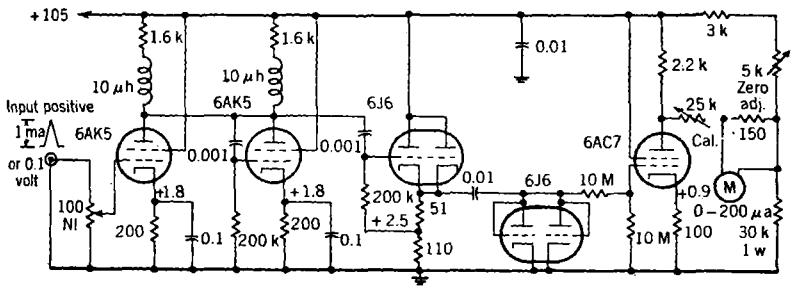


FIG. 3·19.—Circuit of peak-reading voltmeter.

A factor of 3 difference in the two power inputs produces in the radar AFC mixer a harmonic ratio of −20 db without too severe a loss in i-f signal output, and this power ratio is commonly used in the AFC mixer. Thus the sample power is set at about 1.5 mw; the LO contribution at $\frac{1}{2}$ mw; and with this arrangement harmonic locking seldom occurs.

Since the crystal current produced by the transmitter pulses has a very low average value, it is convenient to have a peak-reading device for measuring it. A circuit that has been found very useful is shown in Fig. 3·19. It consists of a video amplifier with a cathode-follower output to provide low impedance, followed by a diode peak voltmeter and d-c amplifier. It may be calibrated by the use of a source of pulses of known size or by attenuating a pulse large enough to measure on an oscilloscope. An r-f probe and crystal may be connected to the input by a length of cable for use in detecting leakage of transmitter power at joints, etc.

One source of leakage into the AFC mixer is minimized by using tight joints in the AFC attenuator line connections. Another source of annoying leakage is the coupling between the signal and AFC mixers. Since

[1] F. E. Terman, *Radio Engineering*, McGraw-Hill, New York, 1937, pp. 448–449.
[2] *Ibid.*

these are supplied with power by a common oscillator, there is definite coupling between them. The plateau portion of the transmitter pulse power leaking into the signal mixer is of the order of 20 mw; therefore the cross attenuation between signal mixer and AFC mixer would have to be at least 19 db in order to reduce leakage power at the AFC mixer to less than $\frac{1}{4}$ mw, which is about the maximum that may be tolerated. Power in the TR-tube spike at the signal mixer is much greater than 20 mw, and cross attenuation of the order of 30 db is needed to avoid harmful spike feedover.

Magic-T Mixer.—The magic-T mixer (Chap. 2) provides further improvements in AFC mixer operation. The LO and transmitter sample channels are connected to operate two crystals in push-pull. As a result of push-pull operation even-order harmonics generated by the crystals are canceled in the i-f output circuit, reducing the problem relating to a high harmonic ratio encountered in double-mixer AFC. Only harmonics present in the original r-f signals remain. Customarily two magic T's are used, with two push-pull crystals for the AFC channel and two for the signal channel. All four crystals are fed by one local oscillator. Another advantage of the magic T is that the cross attenuation between signal and AFC mixers obtainable with this mixer assembly is greater than that found in simple double-mixer systems.

Because of the higher cross attenuation and because possible contributions of LO noise to the receiver output are balanced out, tighter coupling between the local oscillator and the mixers is allowable. Proper mixer design prevents the effect of Fig. 3·18. Therefore even a low-powered oscillator is ample to provide $\frac{1}{2}$ mw to each of the four crystals.

3·9. I-f Amplifiers and Discriminators.—The discriminator circuits used in radar AFC are identical with those already described in Secs. 3·2 and 3·4. With double-mixer AFC, however, an i-f amplifier must also be provided.

The peak-to-peak separation of the discriminator is usually made equal to or slightly less than the main i-f amplifier bandwidth. Since there is some production variation in the maximum response frequency of various i-f amplifiers and in the crossover frequency (see Fig. 3·2) of the discriminator, due to manufacturing variations in the tuning coils and condensers, provision is usually made for the ready shifting of the crossover point over a small interval corresponding to the anticipated maximum departures from design-center values. For instance, in a typical set using a 60-Mc/sec amplifier, the crossover may be shifted ± 2 Mc/sec to compensate for tuning variations of ± 1 Mc/sec in the main i-f amplifier and ± 0.5 Mc/sec in the crossover. The crossover control may be adjusted in operation with the local oscillator locked in by AFC for maximum signal output. The crossover control may be either a variable

condenser or a slug-tuned coil in the resonant-circuit network of the discriminator. In Fig. 3·3, showing the Foster-Seeley discriminator, C_3 is commonly made variable for crossover control in radar AFC. In Fig. 3·13, L_2 is made slug-tunable for the same purpose.

Care should be taken to preserve the symmetry of the discriminator characteristic. If the discriminator curve is asymmetrical, zero output is not produced at crossover but at some other frequency. Since dependable operation of the discriminator may be seriously impaired by such a shift of the zero-output point, asymmetry should be avoided. Furthermore, the injurious effects of a pulse with a poor lopsided spectrum may be aggravated.

With approximately 1.5 mw of transmitter sample power and 0.5 mw of LO power at the crystal, the i-f output lies between 0.25 and 0.5 volt rms. This voltage is maintained at a nearly constant level in operation, since only small variations occur in the transmitter pulse and these variations are reduced by the action of the crystal, as is explained in Sec. 3·6. Accordingly, from one to three i-f stages, depending on the bandwidth, will be needed ahead of the discriminator. Since the i-f stages are in cascade with the discriminator, the i-f bandwidth must be at least large enough so that the over-all peak-to-peak separation is not decreased from the value determined by the main amplifier requirements. In other words, the i-f amplifier must not "bottleneck" the discriminator bandwidth. In practice, the i-f bandwidth may be made considerably greater than the peak-to-peak separation in order that discriminator symmetry be unaffected when the crossover is shifted in tune-up operations. As an example, in one system where the pulse length is 0.5 μsec and discriminator bandwidth is 3 Mc/sec, the i-f amplifier in cascade is 6 Mc/sec in width, to the half-power points. The amplifier pass band under these conditions is essentially flat in the frequency range covered by the discriminator and thus does not cut down on the amplitude of either peak when the crossover is changed.

For 1- or 2-μsec pulse lengths, one amplifier stage is sufficient. As the pulse length used decreases, bandwidth must be increased in proportion, and additional stages to give the requisite gain are needed. Two stages are customarily used with $\frac{1}{4}$- to 1-μsec pulses, three or more with shorter pulse lengths.

The output voltage level at the discriminator should be at least 1 volt and preferably larger with diode detectors, since hum and pickup levels in the diode outputs may be about $\frac{1}{10}$ volt.

Discriminator output may consist of video pulses, or the output time constants may be large enough to produce an essentially d-c voltage. The ratio of this peak-pulse or d-c voltage output to the rms i-f voltage at the mixer output is indicative of gain for the amplifier-discriminator

system. When the input frequency is at either of the discriminator peak frequencies, the gain is maximum. The ratio at these maximum response points is commonly referred to as the *gain* of the amplifier-discriminator system. This figure usually lies in the range between 5 and 20, where diode discriminator detectors are used. Plate, or anode-bend, detectors may also be used, as well as infinite-impedance cathode detectors. With plate detection, extra gain is obtained and the gain figure is correspondingly greater.

Some extra gain in the i-f amplifier may be realized in the coupling network between the crystal and the first grid by use of a double-tuned circuit with loading only on the mixer side. This gain between converter and first grid is very important in the signal channel, where sensitivity is increased by the use of such a circuit (Sec. 4·9). However, along with the large gain is an increased likelihood of bandpass asymmetry due to crystal and grid-to-ground impedance variation. Since symmetry is of especial importance in the over-all discriminator characteristic, the coupling circuit for the AFC channel may be loaded by a resistor on the secondary side to obtain a symmetrical pass band reasonably independent of impedance variations. Loads permitting a voltage gain from crystal to grid of about 1.5 with coupling-circuit bandwidths in the range of 10 to 15 Mc/sec are usually satisfactory in preserving symmetry. However, single-tuned coupling circuits are often used for simplicity.

The i-f amplifier in the AFC chain is a convenient unit for the application of various switching operations. For example, the AFC may be turned on or off by switching B + voltage for the i-f amplifiers.

3·10. General Properties of Control Circuits for Pulsed Systems.—The control circuit, which is the final link in the AFC chain, has the function of converting the error voltage from the discriminator into an LO-frequency correction. Four distinct circuit types for accomplishing this purpose have been used in radar AFC, the principal differences among them being due to (1) variation in the method of obtaining the correct average reflector voltage and (2) variation in the characteristics of the LO-tuning mechanism.

Follow Rates.—In the case of electronic control (for example, reflector control or reactance-tube control), the response of the oscillator to changes in the control voltage is essentially instantaneous. Provision must therefore be made to hold the control voltage reasonably constant during the interval between transmitter pulses when no error signal is available. A condenser that limits the excursions between successive pulses to a fraction of the receiver bandwidth is generally used for this purpose. There is therefore a limit on the *follow rate*, or speed with which the oscillator can follow a frequency disturbance.

An AFC system is an example of a feedback or servomechanism, as

has been mentioned previously. The rate at which any such device may be made to follow the input information is limited by considerations of stability.[1] In this particular case one such limit is apparent: the AFC system must not change the operating frequency by an amount as great as the bandwidth of the receiver in the interval between transmitted pulses. In general, the speed of operation cannot be made great enough so that frequency correction is possible within the length of a single pulse.

In the case of electromechanical (motor-driven) systems, mechanical speeds and inertia are the limiting factors, although by the generous use of power and amplification they may be overcome to any desired degree within the limits set by stability considerations.

Finally, in the case of thermally tuned oscillators, the thermal time constant of the tuning assembly usually provides the limit, although in cases of narrow-band pulsed systems with low repetition rates additional restrictions on the follow rate may have to be provided by the control circuit.

When the radar transmitter pulls, that is, changes its frequency due to changes in the standing-wave ratio during the antenna scan, the frequency change during one pulse is negligible. The variation over a few repetition cycles may be equal to or larger than receiver bandwidth, however, and it is this variation which requires fast following compensation. During pulling, the discriminator input consists of a series of i-f pulses, the frequency of consecutive pulses varying in steps proportional to the pulling rate. The circuits must be such as to use this type of information for the generation of the control voltage.

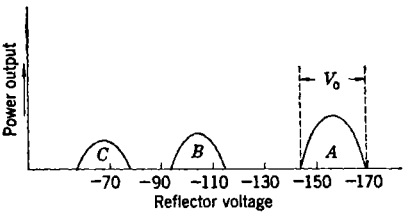


FIG. 3·20.—LO reflector-voltage modes—note, mechanical tuning is fixed.

Control Range.—A design consideration of importance in circuits controlling reflector voltage at the local oscillator is the voltage range over which control is needed. An idea of the problem involved may be obtained by considering the characteristics of one of the reflex local oscillators frequently used. Figure 3·20 shows a typical curve of power output vs. reflector voltage for the 723A/B tube.

The three regions of power output are the LO modes (see Chap. 2). Radar operation is confined to one of these modes, usually the one at the most negative reflector voltage, since this generally provides the highest

[1] See, for example, H. W. Bode, *Network Analysis and Feedback Amplifier Design*, Van Nostrand, New York, 1945; L. A. MacColl, *Fundamental Theory of Servomechanisms*, Van Nostrand, New York, 1945; Vol. 25, Chap. 10, Radiation Laboratory Series.

power output. The mode positions (measured by the voltage at peak output) are not the same in different tubes. Manufacturing variations at present are such that the band center of mode *A* varies in position between −147 and −185 volts in a representative sampling of tubes.[1] The maximum voltage difference between the half-power points in this mode is about 22 volts. Since AFC is normally designed to operate out to the half-power points, it is apparent that the voltage range, from −136 to −196, required for the AFC to operate with any 723A/B at this particular mechanical frequency setting is 60 volts. For operation at other frequencies within the transmitter scatter band, mechanical tuning and reflector voltage must both be varied, since the electronic tuning range is small. This adds another 30 volts to the 60 for a typical 10,000-Mc/sec band. This total voltage requirement will, for present purposes, be called the "mode range." In Fig. 3·20, V_0 is the reflector-voltage range over which a specific local oscillator operates at one mechanical setting. It is obviously much smaller than the mode range.

Wrong-sideband Operation.—Another factor in the reflector-control circuit design is the elimination of *wrong-sideband* locking. The correct

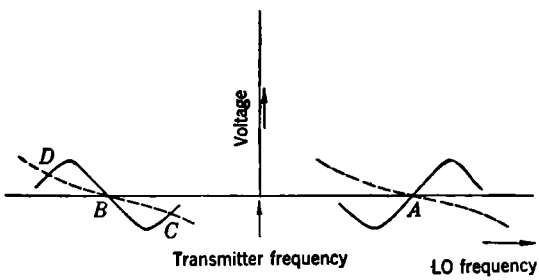


FIG. 3·21.—Illustration of wrong-sideband effect.

difference frequency to which the main receiver channel is tuned may be obtained at two settings of the local oscillator, one below and one above the transmitter frequency. The usual AFC system will operate correctly at only one of these two settings. Thus in Fig. 3·21 discriminator output is plotted as a function of LO frequency over a range including the transmitter frequency. The LO tuning curve is shown in dashed lines at the two crossover points (see Sec. 3·2 and Fig. 3·6).

Correct AFC action has this important characteristic: when LO frequency is changed, the voltage change in discriminator output must be such as to *oppose* the LO-frequency variation. This characteristic requires that the slope of the discriminator curve, at and adjacent to crossover, must be opposite in sign to the tuning-curve slope. At *A*

[1] D. N. Sands and F. S. Bailey, "Characteristics of Recent 723A/B Tubes," RL Report No. 570, May 18, 1944.

this is true, and correct locking is obtained. At B the slopes have the same sign; that is, both are negative, and the local oscillator is *forced away from* point B. At the extremes of the B discriminator curve, however, the slope signs are opposite, and the local oscillator will thus lock *outside* the discriminator peak-to-peak interval at C or D, with the result that the receiver is mistuned even though the local oscillator is locked.

Some individual oscillator tubes have such a broad electronic tuning range that output is available over a frequency interval greater than twice the intermediate frequency. That is, with one setting of mechanical tuning the oscillator may be electronically tuned either at A or B (Fig. 3·21) with different reflector voltages V_1 and V_2.

The voltage swing $(V_2 - V_1)$ required to cover this wide range is generally no larger than that required to tune an average tube over its range, since the broad tuning range is always accompanied by a high value of electronic tuning coefficient (megacycles per volt). Since the control circuit must have adequate range to operate with all tubes, it is difficult and in some cases impossible to exclude V_2 or V_1 from the operating range.

Furthermore, even with tubes having a narrow electronic tuning range, wrong-sideband locking may occur either through faulty adjustment of the mechanical tuning or through frequency drifts. The latter may be so large that a tube which locks on the right sideband when the set is tuned up (cover removed, room temperature ambient) may drift and lock on the wrong sideband during actual operation (cover in place, high ambient temperature).

Usually, however, it is necessary only to tune at the correct point (A, in the figure) to start with. Then the local oscillator ceases oscillation at some point between A and B; there is no output at the wrong-sideband region; and locking cannot occur at C or D.

Types of Control Circuits

The four control circuit types are listed below, in the order in which they will be discussed. In each case, the name customarily applied to the AFC system making use of the circuit is given.

1. D-c amplifier AFC. The discriminator output is integrated over a number of pulses, producing a slowly varying d-c voltage. This voltage is amplified and applied to the LO reflector in the proper polarity for frequency correction.
2. Gas-tube AFC. Two gas-filled tubes are used in connection with a resistor-condenser network to produce the appropriate voltage at the reflector. One of the stages provides control during locking. The other tube, with the associated circuit, is essentially a resetting

device, especially useful where the AFC must function unattended for long periods of time.

3. Diode-phantastron AFC. This circuit also provides both the locking and resetting functions, the action being obtained with two high-vacuum tubes and associated network.
4. Thermal AFC. This is the most complex of the four control-circuit types and is commonly employed for the 2K45 and 2K50 local oscillators, used respectively at 10,000 and 25,000 Mc/sec. These are both thermally tuned tubes (see Chap. 2).

3·11. Direct-current Amplifier AFC.—Systems of this type operating on the received signal were discussed in Secs. 3·2 and 3·4, and the pull-in and hold-in ranges were defined for the general discriminator and tuning curves. A diagram of circuit connections for a simple d-c amplifier system was shown in Fig. 3·10. For pulsed radar operation, one of the pulse-integrating discriminators discussed in Sec. 3·4 is added.

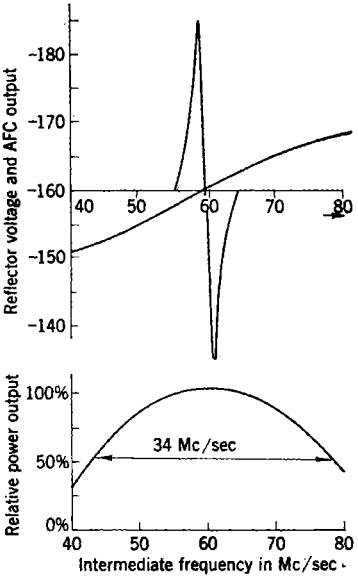


FIG. 3·22.—Operating curve for *X*-band AFC.

The d-c amplifier should provide sufficient gain so that close control may be obtained. When the amplifier is added in cascade with the discriminator, the ordinates of the discriminator output curve shown in Fig. 3·2 are multiplied by the amplifier gain factor. The d-c amplifier system is a *proportional control*, in which the correction voltage output is proportional to the error signal in the region between discriminator peak frequencies.

Representative curves, analogous to those in Fig. 3·6*a*, for a system at *X*-band using a 723A/B local oscillator, are shown in Fig. 3·22*a*. Figure 3·22*b* shows LO output on the same frequency scale, where the mechanical tuning is set to give peak output when the intermediate frequency is at crossover (60 Mc/sec).

The design parameters taken from these curves are the discrimination slope, $D = 20$ volts per Mc/sec, and the tuning coefficient, $T_0 = 2$ Mc/sec per volt. The feedback or error-reduction factor is then

$$\frac{1}{1 + T_0 D} = \frac{1}{1 + 40} = \frac{1}{41}$$

(see Fig. 3·6 and text, Sec. 3·2).

If, after this particular AFC is set up, temperature changes occur such that the LO frequency would be shifted 10 Mc/sec away from the correct value in the absence of AFC, the circuit will correct or bring back the oscillation frequency to a point where the intermediate frequency is $\frac{10}{11}$ or roughly $\frac{1}{2}$ Mc/sec from crossover.

The net effect of drift in either transmitter tuning or *LO uncorrected tuning* is equivalent to a mechanical tuning change at the local oscillator. Thus, as the mechanical tuning is displaced from the design-center value corresponding to 60 Mc/sec in the figure, the intermediate frequency, controlled by AFC, will remain close to 60 Mc/sec until the end of the hold-in range is reached.

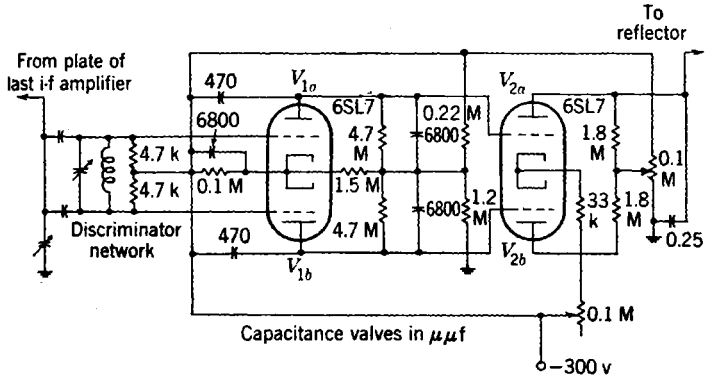


FIG. 3·23.—AFC system using d-c amplifier.

There is a decrease in loop gain during this process, since LO output goes down as the mechanical tuning is changed from the peak power point. The result is a smaller i-f pulse produced at the mixer, and the end effect is a contraction of the AFC output ordinates in Fig. 3·22*a*. A slightly smaller hold-in range than would be predicted by inspection of the curves (assuming the tuning curve extends for a sufficiently large distance in both directions) results. The actual hold-in range achieved is somewhat greater than the frequency interval between half-power points in the LO output curve, or 34 Mc/sec in the present example. The pull-in range is much smaller, however, being about 10 Mc/sec.

A dependable wide-range d-c amplifier AFC system[1] is shown in Fig. 3·23. The circuit has several interesting features. The discriminator circuit is similar to the one shown in Fig. 3·13, but anode-bend detectors are used in place of the diodes. Since the output voltages from both detectors are negative, it is necessary to reverse the polarity of one of them to provide the proper output characteristic. The output from

[1] One example of the use of this circuit was in an airborne radar set built by the Bell Telephone Laboratories.

the upper detector V_{1a} is directly coupled to the d-c amplifier V_{2a}. Output from the lower detector V_{1b}, however, is cathode-coupled into V_{2a} through the cathode-follower V_{2b}. Negative output from V_{1a} produces positive output at the plate of V_{2a}. Negative output from V_{2b} produces a negative signal at the cathodes of V_{2a} and V_{2b}, which is equivalent to a positive signal at the grid of V_{2a}. Plate output is therefore negative. By including a plate-load resistor in the cathode-follower circuit, the system is made symmetrical and degenerative for changes in supply voltage.

Most of the pulse integration is performed in the detector plate circuits. If a large i-f signal appears at the detector grid, its plate resistance is lowered and may become as low as 5000 ohms to provide for the quick discharge of the 6800-$\mu\mu$f plate condensers. When the signal is removed, however, the plate resistance is very high, and the condensers recharge through the 4.7-megohm plate-load resistors. The time constant RC is 3.2×10^{-2} sec. Now the lowest PRF used in the system is about 400 cps, so that the longest interpulse interval is 2.5×10^{-2} sec. Therefore, only about half the charge is lost between pulses.

The sawtoothed ripple voltage at the plate of V_{2a} that results from this alternating discharging and charging process is removed by the low-pass filter formed by the 1.8-megohm plate-load resistor and the $\frac{1}{4}$-μf condenser across the reflector. Note that this filter causes a great reduction in the rate at which the system can follow frequency disturbances.

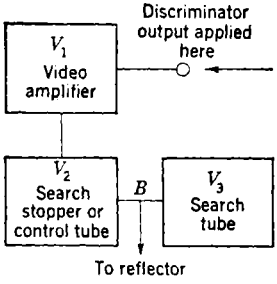


FIG. 3·24.—Block diagram of gas-tube AFC.

The circuit is very dependable in performance, although its pull-in range is small. Its operating range is set for the particular oscillator tube by adjustment of the 0.1-megohm *range set control* potentiometer in the amplifier plate circuit.

3·12. Gas-tube Control Circuits.—The block diagram of Fig. 3·24 illustrates the action of a typical gas-tube control circuit.

The search tube is a sawtooth generator much like the X-axis sweep generator used in cathode-ray oscilloscopes. The sawtooth-wave period is from 0.1 to 1 sec, and the voltage sweep is more than enough to cover the range between the half-power points in the LO output. In operation, the mechanical tuning is set at the desired frequency—at an interval equal to the amplifier intermediate frequency, above or below transmitter frequency, depending on which one gives "right-sideband" action (see Sec. 3·10)[1]—and the sawtooth sweep is

[1] In microwave receivers, the r-f components are usually such that it is immaterial which sideband is chosen, so that a more or less arbitrary choice must be made in the initial design.

centered at the reflector-voltage value corresponding to maximum LO output.

If V_2 is not operating, the search alone may be considered, and the situation is as shown in Fig. 3·25.

In the center region of the sweep, the beat frequency passes through the discriminator region, and the output pulse envelope follows the discriminator curve. The output pulses are inverted by the single-stage video amplifier.

The function of the search stopper is to interrupt the sweep by injecting a negative pulse at B in Fig. 3·24. This pulse is initiated, as will be seen, by the positive triggers that appear at the output of the discriminator after the crossover frequency has been passed It pushes the LO frequency back through crossover momentarily until the circuit at B begins to recover from the injected pulse, at which time the frequency again starts through crossover in the original direction of the sweep. As soon as the critical point past crossover is again reached, another pulse is injected and the cycle continues in this fashion.

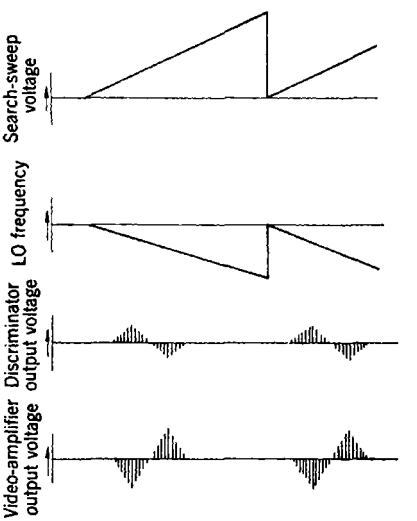


FIG. 3·25.—Waveforms in search function of gas-tube AFC.

Thus LO frequency swings between two extremes, or hunts, during AFC locking, in contrast to the proportional control case where the whole AFC chain and LO frequency are at stable equilibrium during locking.

The schematic diagram in Fig. 3·26 shows the details in the gas-tube circuit. The notation used in the following circuit analysis is as follows: $\bar{V}_F$ is time average for V_F under equilibrium conditions. $(\bar{V}_F)_{max}$ is maximum average V_F. $(\bar{V}_F)_{min}$ is minimum average V_F. $(V_R)_{max}$ and $(V_R)_{min}$ are the end points of the mode range. V_1 is a video, or pulse, amplifier; V_2 is the search stopper, a gas-filled tetrode, in which firing is nearly independent of plate voltage (the 2050 and miniature 2D21 are examples of the type); V_3 is a gas-triode, usually an 884 or a 6D4. The following relations should hold: $R_0C_0 \gg R_1C_1$; $C_0 \gg C_1$; $R_0/R_1 \approx \frac{1}{2}$ to 3. Typical values are $R_1 = 1$ megohm, $R_0 = 1$ megohm, $C_0 = 1\ \mu$f, and $C_1 = 0.005\ \mu$f.

The sweep action of V_3 requires no detailed explanation. The sweep speed is determined by $(R_0 + R_1)C_0$ and $E_1 - E_2$, approximately.

Sweep voltage swing is a function of control-grid bias. If there is no discriminator pulse output (for example, if the transmitter is not on), the sweep action is unimpeded and reflector voltage V_R passes through a range including the value required for correct LO tuning. However, as soon as the transmitter is turned on and the resulting discriminator output pulses are applied through the video amplifier to the control grid of V_2, plate voltage at V_3 never reaches its firing potential because V_2 fires first. That occurs, as shown in Fig. 3·25, somewhere near the middle of the sweep, just past the reflector voltage corresponding to crossover.

The system gain is such that the pulse amplitude at a very small frequency interval from crossover is large enough to fire V_2. Condenser C_1 discharges through the arc in V_2 when it fires, and V_F drops rapidly toward cathode potential. At a level about 15 volts above the cathode, the arc extinguishes. This plate surge downward takes place in about 1 to 5 μsec. After the arc has gone out, V_F rises as C_1 is charged up again. The time constants are arranged so that C_1 charges rapidly, the time constant being of the order of R_1C_1 sec. During the process, the voltage across C_0 changes, but only by a very slight amount, since R_0C_0 is so much larger than R_1C_1, or, in other words, the R_0C_0 network acts as a filter against the trigger pulse at the plate of V_2. In operation, the voltage change across C_0 is of the order of $\frac{1}{100}$ of the V_2 plate swing.

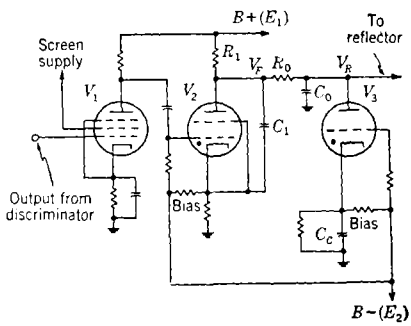


FIG. 3·26.—Circuit diagram of gas-tube AFC.

As soon as V_2 fires, the current in R_0 reverses, since V_F is no longer *above* V_R but below it. Accordingly C_0 discharges toward the negative voltage at the plate of V_2.

After the V_2 arc goes out, however, at some time during the subsequent charging of C_1 in the positive direction, the voltages V_F and V_R are equal, and thereafter V_F is more positive than V_R. At this instant C_0 ceases to *discharge* and thereafter proceeds to charge positively. It is apparent that the reflector voltage starts down immediately when V_2 fires, reaches a *minimum* value some time later, then drifts back in the positive direction. As soon as a sufficiently positive voltage is reached, another pulse fires V_2 and the action is repeated. Under locking conditions, V_R is a d-c voltage with a small "ripple" component superimposed on it. The corresponding LO frequency ripple must be small compared with receiver bandwidth. The action is illustrated by the waveforms shown in Fig. 3·27*a*.

The only d-c connection to the reflector is the resistor R_0, which leads to the plate of V_2. Therefore, at equilibrium the d-c voltage on the reflector will equal the average voltage $\bar{V}_F$ at the plate of V_2 as shown in Fig. 3·27*a*.

If the transmitter frequency shifts so that a more negative reflector voltage is required to produce the correct difference frequency, $\bar{V}_F$ must decrease. This decrease is brought about automatically by more rapid

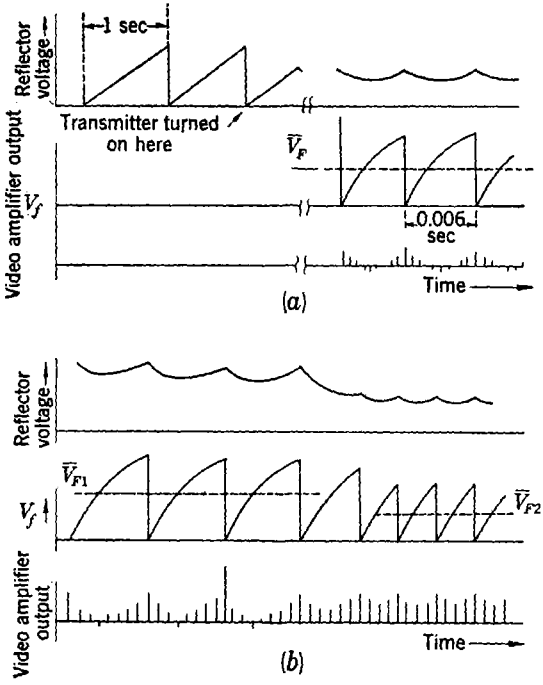


FIG. 3·27.—Search and locking waveforms—gas-tube AFC. (*a*) Initial locking; (*b*) shift of frequency to follow transmitter shift.

firing of the control tube V_2 (refer to Figs. 3·27*b*). $\bar{V}_{F1}$ is the original value of $\bar{V}_F$ and $\bar{V}_{F2}$ is the value after the transmitter frequency shift.

When the control tube is fired at more frequent intervals, C_1 collects less charge per interval, and the peak and average values of V_F decrease. This is an example of *firing-rate control* ("frequency principle") of the average reflector level, in contrast to the *proportional control* ("amplitude principle") of the d-c amplifier AFC.

A more detailed examination of the search-stopping or trigger pulse is of advantage here. The C_1 charging curve is of the form

$$V_F = E' \left(1 - e^{-\frac{t}{R'C_1}}\right),$$

where R' and E' are the effective charging resistance and voltage respectively in the circuit. (V_R is assumed to be constant, since its change is negligible compared with that of V_F.) The average value of V_F for a given firing rate of V_2 may be obtained by suitable manipulation of this expression.[1] The result is shown in Fig. 3·28. The abscissa is T, the time allowed for C_1 to charge (interval between successive firings of V_2).

It is evident that $\bar{V}_F$ increases rapidly with T at first; but as the firing interval becomes larger, that is, as the firing rate for V_2 decreases, the slope decreases. Accordingly at trigger-pulse intervals large with respect to $R'C_1$, an increase in V_F is obtained at the expense of an inordinately large increase in T and hence in the length of the injected pulse at B. Since R_0 and C_0 form a low-pass filter, they will not be so effective in reducing ripple with the longer pulse length.

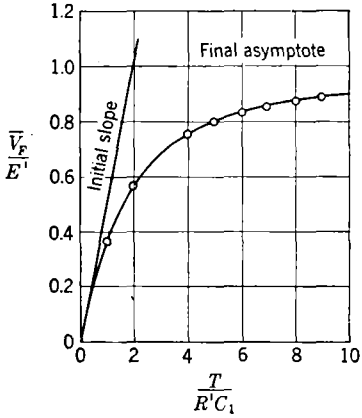


FIG. 3·28.—Average V_F vs. firing interval.

The advantage of choosing an operating range as near as possible to the origin of the curve of Fig. 3·28, in the linear region, is evident. As E' is increased, the linear portion extends over a larger voltage interval, so that with a given mode range to be covered it is desirable to have a value for E' considerably larger than the mode range. A larger E' may be obtained by an increase in E_1 or a decrease in E_2. Accordingly the voltage difference between B^+ and B^- should be large compared to the mode range.

It is also important to have a relation between $R'C_1$ and the system pulse-repetition period that will provide operation in the linear region of the curve. In practice, this amounts to picking C_1 so that when V_R is at the center of the mode range, V_2 fires on every third or fourth system pulse. When V_2 fires on the average at nearly every system pulse, $(V_R)_{\min}$ is attained; and when V_2 fires on about every eighth to tenth pulse $(V_R)_{\max}$ is attained. The $(V_R)_{\min}$ case is not carried to the extreme of firing exactly at every system pulse, however, because of follow-rate considerations.

The follow rate required depends on the transmitter pulling rate, which is ordinarily less than 300 Mc/sec per sec. For pulse-repetition

[1] An exact theory for the calculation of this and other important quantities is found in Vol. 16, Sec. 7·11, Radiation Laboratory Series.

frequencies greater than 300 cps, the frequency shift between successive pulses will therefore not exceed 1 Mc/sec. The region over which a discriminator gives useful output is somewhat greater than the peak-to-peak separation. Thus if a discriminator has a spacing of 2 Mc/sec between peaks, the maximum frequency shift between successive pulses will not carry the system out of the operating range of the discriminator, and the limitation on performance must be sought in other parts of the circuit.

The designer is interested only in the case of the maximum pulling rate, since if the AFC will follow this, it will follow slower changes. Under maximum pulling conditions, there are two possible cases to consider, depending on the direction of transmitter frequency shift. In one direction, the i-f departure from crossover is such as to produce negative pulses at the control-tube grid; in the other, consecutive positive pulses are applied. In the one case, the control tube stops firing; in the other, it fires at every pulse. If the control tube is not firing, V_R rises exactly as in the search sweep. The current in R_0 and R_1 is

$$\frac{E_1 - V_R}{R_1 + R_0}.$$

This current is the rate of change of charge in C_0 (C_1 is neglected, since $C_1 \ll C_0$), and the charge $q = C_0 V_R$. Then

$$\left(\frac{dV_R}{dt}\right)_+ = \frac{1}{C_0}\frac{dq}{dt} = \frac{E_1 - V_R}{(R_1 + R_0)C_0}.$$

The plus sign denotes the rate for V_R changing in the positive direction. On the other hand, if the control tube is firing at each pulse, the lowest $\bar{V}_F$ average is obtained. This is $(\bar{V}_F)_{\min}$[1] and by the same reasoning as above, the rate of change for V_R is

$$\left(\frac{dV_R}{dt}\right)_- = \frac{V_R - (\bar{V}_F)_{\min}}{R_0 C_0}.$$

The minus sign indicates that the voltage following rate is in the negative direction. Multiplying these voltage change rates by the electronic tuning coefficient T_0 gives the frequency follow rates obtainable in the AFC system.

T_0 varies among the different tube types and also with different tubes of the same type. The lowest follow rates occur when T_0 has its smallest value. This value will be denoted by $(T_0)_{\min}$.

Since V_R may be anywhere in the mode range, it is evident that not only are the positive and negative rates usually unequal but also each

[1] Strictly speaking, $(\bar{V}_F)_{\min}$ is not a constant. The effect of V_R on $(\bar{V}_F)_{\min}$ is, however, secondary and can usually be neglected. See Vol. 16, *loc. cit.*

rate varies considerably over the range. It is also apparent that if C_1 is adjusted to have $(V_F)_{min}$ equal to $(V_R)_{min}$ when the control tube fires at every pulse, the negative voltage following rate at $(V_R)_{min}$ will be zero. Accordingly C_1 is adjusted to provide sufficient margin between $(V_F)_{min}$ and $(V_R)_{min}$ so that adequate following may be obtained at this end of the control range.

R_0 and C_0 form a filter effective in reducing ripple amplitude. If R_0C_0 is made smaller, ripple increases. R_0C_0 may not be reduced beyond a certain point for improvement of the follow rate (increasing it) without allowing excessive ripple. An appreciable improvement in follow rate with only a very slight increase in ripple is obtained, however, by using a higher negative voltage for E_2, making possible a more negative value for $(V_F)_{min}$. With present equipments, an adequate follow rate with reasonably low ripple is obtained with -250 volts. If closer control or faster following is required, a relatively cheap method for obtaining them is to use -300 volts.

The simple circuit of Fig. 3·26 must sweep over a very wide range to be able to accommodate all oscillators. The spread of mode position from tube to tube is so great that, referring to Fig. 3·20, mode *B* for one local oscillator may occur at the same voltage as mode *A* for another tube. Now it may be desirable to exclude the possibility of locking in mode *B*, which, in certain tube types, generates more excess noise than mode *A*. To accomplish this, the voltage sweep at the reflector must be reduced to a value not much greater than the mode width for a particular tube (25 to 40 volts). Then, a *range-set* control must be provided to center this swept range at the mode center for the particular tube.

One arrangement is to install in the cathode circuit of V_3 (Fig. 3·26) a potentiometer that allows the cathode voltage to be set anywhere in a 50-volt range. Since the searching sweep at V_3 is limited to 30 to 50 volts, the voltage range covered by the sweep and range set combined is 80 to 100 volts. A disadvantage is that in the field the sweep must be adjusted independently of manual tuning. Another method that overcomes this disadvantage is incorporated into the receiver described in Chap. 13.

If there is a long lead from the AFC output to the reflector, there may be excessive video pickup. This pickup is cut down by using an *RC* filter right at the reflector, that is, a series resistance in the lead and a bypass condenser from reflector to ground. The time constant *RC* must be approximately $\frac{1}{20}$ of the system pulse-repetition period in order not to affect AFC following.

The most serious effect of video pulse pickup at the reflector arises because the reflector voltage may deviate regularly from its interpulse value just at the time when a sample of the LO frequency is being used

to operate the AFC circuits. Thus the AFC will hold this momentary value of the oscillator frequency at the desired point, whereas the interpulse frequency may be far from correct.

In designing the gas-tube circuit the practical approach is to use values giving sufficient range and follow rate and then to measure the ripple. Although an expression for ripple voltage as a function of the circuit parameters has been obtained,[1] it is inconvenient to use, whereas the ripple amplitude is easily measured on a cathode-ray oscilloscope.

Experimental observations as well as calculations based on Rochester's theory have led to a rough working rule that gives the relation between ripple and follow rate. The accuracy is ample for most design purposes.

The rule is applicable when normal values of the parameters are used, that is, when $2R_0 > R_1 > R_0/2$ (Fig. 3·26) and when C_1 is chosen so that midrange voltage will be maintained with every third transmitter pulse firing the search stopper V_2. The rule is as follows: The maximum ripple occurs when the search stopper is fired by about every tenth pulse or less. Its peak-to-peak voltage is twice the maximum amount of down-pull voltage between successive pulses, which occurs when the reflector voltage is at the top of its range and conditions are such that the search stopper fires at each transmitter pulse.

The design considerations may be briefly stated: R_1 must be large enough to assure extinguishing of V_2. With 2D21's and 2050's, 1 megohm is about the lowest safe value. R_1 should not be much larger or the positive follow rate is unnecessarily reduced. The difference between E_1 and E_2 should be large for adequate follow rate and low ripple. C_1 is set to provide firing at every third or fourth system pulse when V_R is at the mode-range center. When V_2 fires at every pulse, $(\bar{V}_F)_{\min}$ must be appreciably lower than $(V_R)_{\min}$. The difference between these two may sometimes be as much as 40 volts.

With given E_1, $(\bar{V}_F)_{\min}$, $(V_R)_{\max}$, and $(V_R)_{\min}$, it is easily determined where the lowest follow rate will occur [either at $(V_R)_{\max}$ or $(V_R)_{\min}$]. The time constant R_0C_0 may be obtained by setting this lowest voltage follow rate times $(T_0)_{\min}$ equal to the maximum expected pulling rate. It is unfortunate that large variations occur in the follow rate. Since improvements in ripple and follow-rate performance are directly opposed, that is, improving one means impairing the other, a loss is taken in ripple performance over most of the range when the lowest follow rate is set equal to the maximum pulling rate.

Under these conditions the frequency ripple when reflector voltage is held at $(V_R)_{\max}$ should be small with respect to receiver bandwidth. If the ripple is excessive, some compromise must be made, either in volt-

[1] N. Rochester, Sylvania Electric Products Co., Boston, Mass. This theory is developed in Vol. 16, Sec. 7·12, Radiation Laboratory Series.

age range, ripple, or follow rate. This situation may be improved in the future by reduction in the variations in mode range, firing potential, etc., from tube to tube.

Considerable detail has been included in the gas-tube AFC section in an effort to show the nature of the AFC problem for fast pulling radars. Where low pulling rates are expected the design problem is simple. R_0C_0 is simply made large enough to reduce ripple to negligible values. A low or negligible pulling rate also makes it possible to fire the control tube at very low frequencies, on the order of power frequencies. This is a distinct advantage with beacon AFC, an absolute-frequency-control method taken up later in this chapter (Sec. 3·16).

The search sweep, which provides the recycling or resetting function, is often extremely useful. It provides a pull-in range equal to the hold-in range. Therefore, if input AFC signals are interrupted momentarily, the AFC will lock on again when signals return.

3·13. Hard-tube Control Circuits.—At the Radiation Laboratory, the first use of high-vacuum tubes[1] in the control circuit was made by Strandberg.[2] The problem at hand was to provide a hunting sweep time in excess of 1 min, as required by the A-5022A thermally tuned oscillator, and the use of an "amplified capacity" (see later, this section), possible with hard tubes, seemed desirable. Later, the difficulties associated with the variable firing potential of the gas triode led to the development of another circuit, also using hard tubes, in which tube variability was overcome through the use of feedback.

The original Strandberg circuit is probably of historical interest only, since the new one performs the same functions more simply. It utilized a gas-tetrode search-stopping tube in a circuit identical with the one of Fig. 3·26 except that the resistor R_0, instead of being attached to V_3, was attached to the control grid of a *phantastron* oscillator. One can thus think of Strandberg's circuit as a transition between the all-gas-tube circuit and the all-hard-tube circuit, since it uses the search-stopping principle of the one and the searching principle of the other.

Before proceeding to the *diode-phantastron* control circuit, the phantastron itself should be considered. One form of this circuit was introduced by the British as a precision ranging-sweep generator and has since been used extensively in both British and American indicators.[3] As used in ranging, it requires an external trigger to start the cycle, which consists of a single linear downsweep of the plate voltage, followed by a

[1] See also Vol. 16, Sec. 7·13.

[2] M. W. P. Strandberg, "Some Automatic Frequency-control Circuits," RL Report No. 687, Mar. 10, 1945.

[3] See Vol. 19, Chaps. 9 and 14, Radiation Laboratory Series.

quick recovery and a quiescent period while awaiting the next trigger. For AFC purposes, it is necessary to have a self-triggering generator, and some modifications are necessary. The final circuit, including also the diode search stopper, is shown in Fig. 3·29. The basis of operation is the negative transconductance that exists between the suppressor grid g_3 and the screen grid g_2 of a pentode. If one end of the grid resistor R_1 is tied to some potential at or above the cathode potential, current will flow, tending to cause the grid voltage e_g, which is negative at the start, to approach the cathode voltage E_c. The rate of rise of e_g will be less than is indicated by the values of R_1, C_1, and the voltage drop across R_1, because of the large "amplified" or Miller[1] capacity appearing at the input. In other words, as the grid potential tries to rise, the plate potential falls rapidly; this fall, coupled through C_1, inhibits the rise in grid potential. These and the subsequent effects are illustrated in Fig. 3 30.

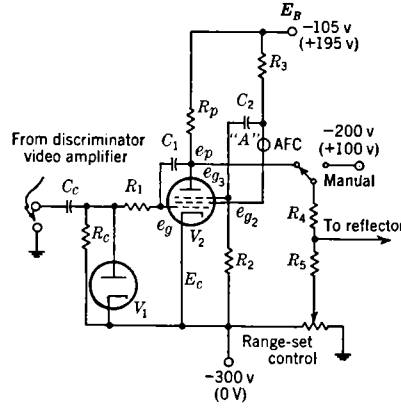


FIG. 3·29.—Diode-phantastron control circuit.

However, e_g does rise, and the plate potential e_p does fall, and ultimately the latter comes close to the cathode potential E_c. At this time, a regeneration that is the first part of the recycling process sets in. For as e_p approaches E_c, the normal distribution of current between plate and screen is disturbed and an ever-increasing current flows to the screen, whose potential e_{g_2} therefore starts to fall. This fall is coupled by C_2 to the suppressor grid g_3; and when it reaches a large enough value, g_3 starts to cut off the current to the plate The plate voltage then rises, carrying with it e_g, thus increasing the cathode current and hence the screen current. The regeneration s complete in 1 or 2 μsec, when g_1 starts to draw current. This action occurs at time A (Fig. 3·30). From this time on, e_p can rise only as fast as C_1 can be charged through R_p.

C_2 is made large enough so that the suppressor will not recover until e_p has come close to the supply voltage E_B. Presently, however, e_{g_3} comes close enough to E_c to allow some plate current to flow. The second regeneration immediately occurs (time B, Fig. 3·30). The downward plate-voltage motion, coupled through C_{a_1}, tends to drive e_g to

[1] J. M. Miller, "Dependence of the Input Impedance of a Three-electrode Vacuum Tube upon the Load in the Plate Circuit," *Bur. Standards Sci. Paper* 351. See also F. E. Terman, *Radio Engineers' Handbook*, McGraw-Hill, New York, 1943, p. 468.

cutoff. This in turn reduces the screen current, and the resultant rise in e_{g_2} and e_{g_3} causes a further drop in e_p, and so on. At the end of the process, e_p is down by some 10 to 15 volts from E_B, e_g is near cutoff, and the slow downward sweep of e_p sets in.

If no error signal comes through, because, for example, the transmitter is turned off, the phantastron plate will execute a series of sawtooth downward sweeps, separated by moderately long recovery periods. If, however, during one of the sweeps, the transmitter is turned on so that pulses appear at the discriminator output, the sweep will be stopped as soon as the discriminator crossover frequency is passed. For at this time, positive pulses appear at the output of the discriminator video amplifier. These pulses are coupled to the shunt diode detector and cause it to develop a negative voltage. As soon as the voltage across the diode becomes equal to the voltage on the phantastron control grid, the current flow in R_1 ceases, and the downward sweep of the plate therefore stops. Thus, the system is locked. The exact amplitude of triggers will automatically set itself in such a way that the frequency is held to the required value, since any deviations tend to change e_g so as to restore the frequency to the right value. As long as e_g is maintained at a negative value, the recycling described above will not occur.

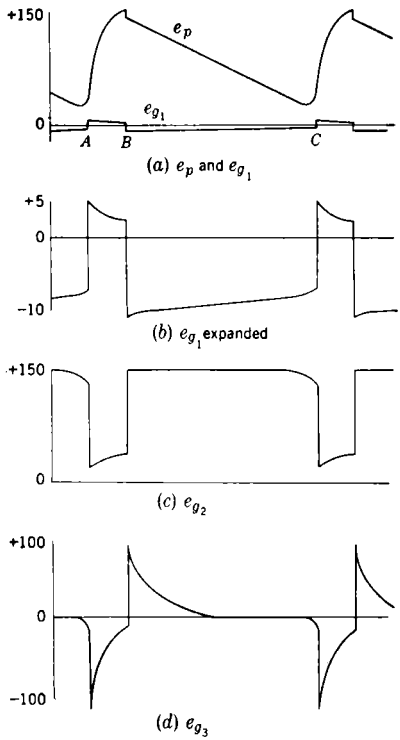


FIG. 3·30.—Voltage waveforms in the phantastron circuit.

In the design of a practical circuit, several factors determine the proper choice of constants. The voltage sweep rate at the plate of the phantastron is given by[1]

$$\frac{de_p}{dt} = \frac{E_c - e_g}{(R_c + R_1)C_1}. \tag{1}$$

Of this, a fraction $R_5/(R_4 + R_5)$ appears at the reflector.

[1] See Vol. 16, Eq. (7·36), Radiation Laboratory Series.

In the usual design the linear-sweep range of e_p comes out to be about two-thirds of the total supply voltage E_B. The range desired at the reflector is determined by the tube properties and should be slightly in excess of the largest voltage swing required to carry any oscillator over the desired mode (see Fig. 3·20). For many tube types a sweep of 40 to 50 volts is required; usually, therefore, $R_4 \approx 2R_5$ (Fig. 3·29) for $E_B = 195$ volts.

The actual voltage reached by the plate during the recharging of C_1 is reduced below E_B because of the current drawn through R_5 and R_4. These resistors are therefore usually kept fairly high, 1 and 2 megohms respectively being common values. The loss in effective supply voltage depends on the setting of the range-set control and may vary from an actual gain to a loss of some 15 volts. The available swing is correspondingly reduced, and allowance should be made in the choice of the R_4/R_5 ratio. R_p should be large to secure high amplification and to reduce the current demand of the circuit. Values ranging from 0.2 to 0.5 megohm are common.

R_3 is critical in that it determines the point in the downward sweep at which regeneration sets in. If it is too large, a relatively small change in screen current will cause enough drop to start the cutoff of the plate current by the suppressor. If too small, the recycling will not take place. If the sweep tube is a 6SJ7 or similar tube, values from 5000 to 20,000 ohms should be used. If the new 6AS6 miniature type, developed especially for transitrons and phantastrons, is used, R_{g_3} must be very low, and the screen drain will be excessive unless a trick is used.

The difficulty with the 6AS6 comes from the very high transconductance between g_3 and the plate. A small drop in e_{g_3} will start the regeneration. The trick then is to divide down the screen swing before coupling it to g_3. In Fig. 3·29, a resistor is inserted at point A. Values of 47,000 ohms for the new resistor and 10,000 ohms for R_3 are satisfactory and reduce the maximum screen current to less than 5 ma. The phantastron can therefore be held for a long time in the state where the screen is drawing full current, without excessive screen dissipation, for example, by applying an external negative bias to g_3. This feature is used in one of the later circuits (Sec. 3·14).

The value of R_2 may be conveniently chosen as 2 megohms; C_2 is then chosen so that $R_2C_2 \geqq 3R_pC_1$, to provide adequate time for plate recovery.

Even when long sweep times are not required, R_c and R_1 are usually made equal and as high as possible (2 megohms) because the effective voltage developed across C_c is approximately proportional to R_c unless the length of the pulse at the video amplifier is long.

To provide for some integration, R_cC_c is made equal to several inter-

pulse intervals. In a typical case, $C_c = 0.05$ μf. Since the typical-system pulse length is of the order of a microsecond, only a few volts are developed across C_c by any one pulse. The video amplifier should draw as much current as possible and should have a large g_m. For small signals, the charge transferred to C_c is proportional to g_m; for very large signals, the largest current that can flow into C_c is the standing current in the load resistor, the amplifier being fully cut off.

Unlike the gas-tube case, ripple is seldom the limiting factor in determining the sweep rate. During the search cycle, before locking has taken place, the sweep velocity is high, corresponding to the steep part of the curve of Fig. 3·30. Furthermore, usually more than one trigger is required to stop the sweep because of the small charge that can be transferred to C_c even by a full-amplitude pulse. The sweep must, therefore, be slow enough so that several positive pulses appear during the initial transit of the discriminator peak; otherwise the system will fail to lock even though, if locking were once established, it would be maintained.

The diode-phantastron combination is essentially a d-c amplifier system. Pulse integration is accomplished partly by the diode network R_cC_c and partly by the phantastron input network R_1C (amplified). When locked, the pentode acts as a normal d-c amplifier with a gain of about 50. The difference from the previously mentioned d-c amplifier circuits is that the pull-in range is equal to the hold-in range because of the search and recycle provisions (see Secs. 3·2 and 3·4).

The circuit will work well with any tube of the specified type. The voltage at which the sweep starts is essentially constant, being equal to the supply voltage less the small grid cutoff voltage. A proper choice of R_3 will ensure that every tube will sweep to within not more than 50 volts of the cathode.

In addition to the increased sweep time afforded by the capacity amplification, a further increase is available because of the fact that the condenser-charging resistor is located in the grid circuit and the load (the reflector divider) is in the plate circuit. The former may be made very high if necessary, and the latter is kept at a low enough value to ensure against reflector "runaway," in which the reflector becomes positive and emits electrons, so preventing a high-resistance voltage source from bringing it back to the required negative operating region.

The long cycle that may be obtained with small condensers is particularly useful in connection with thermally tuned tubes, where the time required for the oscillator to drift from one end of the band to the other may be as long as 15 sec even at full power. The method of application of the tube to the circuit is, as will be seen in Sec. 3·14, very different, however.

The diode-phantastron circuit has been used to advantage in reflector

AFC, the chief benefits coming from its independence from tube characteristics and the fact that the search sweep rate is relatively constant over the whole control range. In the gas-tube circuit, ripple at the steep end of an exponential determines the upper limit of sweep rate; the rate elsewhere in the range may be one-third to one-quarter of the maximum value. In the above circuit, however, the sweep rate is constant to within about ± 10 per cent.

Another interesting property of the circuit is the fact that the ultimate sweep rate per transmitter pulse is much larger than the interpulse ripple because of the double integration of the diode and the phantastron. If the system is allowed to reach equilibrium and then the source of control information is abruptly turned off, the reflector will not start to drift abruptly, as in the case of the gas-tube circuit, but will have an "acceleration" in the drift, as shown in Fig. 3·31. This acceleration is due to the fact that the charge on the diode condenser C_c must all leak off before the phantastron drift velocity assumes its maximum value. Since every pulse is effective, instead of every third or every tenth, it is the slope near the origin that determines ripple, but the slope far from the origin determines the ultimate follow rate that can be reached. Although a double time constant would achieve the same result for a gas-tube control circuit, the final effect would be unsatisfactory because of the ripple frequency. That is, a filter that would remove the low-frequency components present when only every tenth pulse fires the search stopper would cause the system to become very sluggish, even though the ultimate control rate might be very high. The real advantage, then, comes from the utilization of every pulse, with control proportional to amplitude.

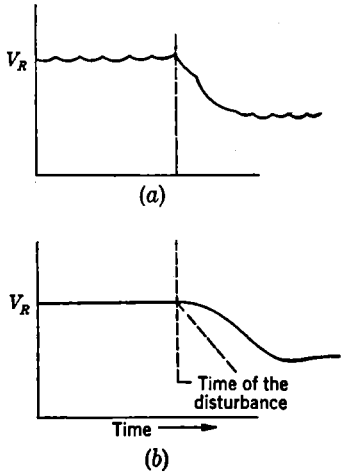


FIG. 3·31.—"Inertia" effect of diode-phantastron. (a) Gas-tube circuit; (b) diode-phantastron AFC.

3·14. Control Circuits for Thermally Tuned Oscillators.—Thermally tuned oscillator tubes are similar to other reflex oscillator tubes except for the method of tuning the cavity to the desired frequency. A small strut in the cavity is distorted by heat, and the strut in turn distorts the configuration of the cavity and varies its resonant frequency. In the newer tubes, the strut is actually the plate electrode of a small triode tube built into the shell along with the oscillator proper but electrically

independent of it. The amount of heat in the strut is varied by changing the grid bias of the triode section, thus changing the amount of plate current and heat dissipation in the strut. Such a construction affords an extremely wide tuning range controlled by electronic means.

The wide tuning range of thermally tuned tubes requires a different approach to the problem of AFC because the tube will tune through either side of the transmitter frequency. As a result, the AFC system, if it is to use information from a conventional discriminator following an i-f amplifier, must use entirely different control circuits from those used for reflector-tuning AFC systems to avoid looking to the wrong sideband (see Sec. 3·10). Since the tuning is done by a thermally sensitive element, the temperature stability of the tube is very poor. Thus good AFC is a necessity.

In all of the previously considered control circuits, there existed an instantaneous "one-to-one" correspondence between the voltage on the frequency-control electrode (for example, the reflector or the reactance-tube control grid) and the LO output frequency. It is therefore necessary, as was pointed out in Sec. 3·10, to introduce into the control circuits networks that limit the follow rates.

In thermally tuned oscillators, however, the heat capacity of the tuning mechanism introduces a delay between the application of a voltage to the frequency-control electrode (tuner triode grid) and the reaching of the corresponding output frequency by the oscillator. Consequently, there is an inherent limitation on the achievable follow rates which is usually, but by no means always, lower than the follow rates that would be permissible under the bandwidth and PRF conditions discussed in Sec. 3·10.

The response of such thermal tubes to changes in control voltage is approximately exponential. That is, the rate of change of frequency is directly proportional to the difference between the instantaneous frequency and the equilibrium frequency corresponding to the instantaneous value of the control voltage. Therefore, the highest rates of drift in the two directions will be achieved when the tuning power is the maximum allowable or zero, respectively.

Most proposals for thermal AFC involve "on-off" control. That is, searching is provided by turning the power alternately fully on and fully off (heat reversal) at intervals long enough to allow the oscillator to sweep from one edge of the band to the other between heat reversals. Locking is accomplished by turning the power on and off so rapidly that a negligible drift or *ripple* of the frequency occurs between heat reversals. The average power is then determined by the *duty ratio*, or ratio of time spent with power on to time spent with power off.

When a severe demand is made of such a system, the power will remain

fully on or fully off as the case may be until the required compensation is effected, thus providing the highest possible follow rates.

The maximum follow rates depend on the position within the band. Close to either band edge, the follow rate toward band center is high, whereas the follow rate away from band center is low.

Typical values are as follows: The 2K45 (10,000-Mc/sec) oscillator has a time constant of about 8 sec and covers a 2000-Mc/sec band. Its maximum follow rate at band center is therefore 125 Mc/sec per sec. The 2K50 (25,000-Mc/sec) oscillator time constant is about 1.6 sec; the band coverage is also 2000 Mc/sec; and band-center maximum follow rate is 625 Mc/sec per sec.

Two of the proposed circuits have been built sucessfully and are now in use in practical radar sets. They differ in that one system locks on one sideband only (see Figs. 3·6 and 3·21 and text, also below), rejecting the other "wrong" sideband automatically, whereas the other can lock *correctly* on either sideband. For the balance of this section, the first system will be referred to as the "single flip-flop" (SFF) control circuit, and the second system as the "double flip-flop" (DFF) circuit.

As will be shown later, the double-sideband locking property makes possible a *quick-relock* feature such that if the system should become unlocked because of momentary trigger failure or excessive follow-rate demand, it will automatically return to the correct locking position at least once for a second chance at locking without having to go through a complete searching cycle. Thus, with a 2K45, control may be restored in less than a second instead of having to wait for as much as 20 sec. This advantage is obtained only at the expense of considerable circuit complexity and extra tubes; when frequent unlocking is not expected, the simpler SFF system is to be preferred.

In addition to the on-off feature previously considered, both circuits utilize *push-pull* control. That is, both of the discriminator peaks are used to cause heat reversals. In the SFF circuit, positive pulses from the discriminator always result in the heat being turned on, if off, and have no effect if it is already on, whereas negative pulses have the converse effect. Examination of Fig. 3·32 shows that this results in locking on the low-frequency sideband only, locking on the other sideband being excluded. Of course, if the action of positive pulses were to turn the heat off or if the discriminator polarity were reversed, the system would lock on the high-frequency sideband.

In the DFF system, a similar one-to-one correspondence between heat reversals and pulse polarity exists as long as the system remains locked. However, the relation between heat reversals and pulse polarity is inverted at each end of every searching sweep, so that after each such search reversal, the relation is such as to allow locking on the nearest sideband.

Block diagrams of the systems are shown in Figs. 3·33 and 3·34. Complete circuit diagrams and detailed operational analyses are given elsewhere.[1]

In the SFF system, the heart of the circuit is the heat-reversing flip-flop, HR, or Eccles-Jordan trigger circuit. It has two stable conditions of equilibrium between which it may be switched as a result either of dis-

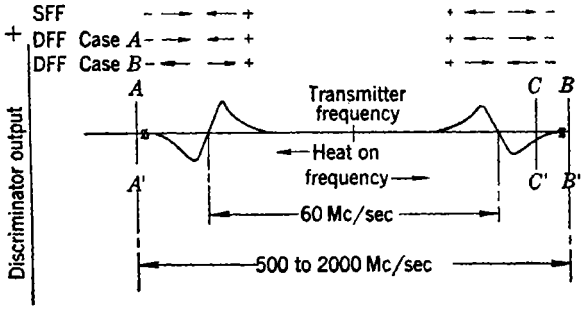


FIG. 3·32.—Locking and sideband rejection in thermal AFC.

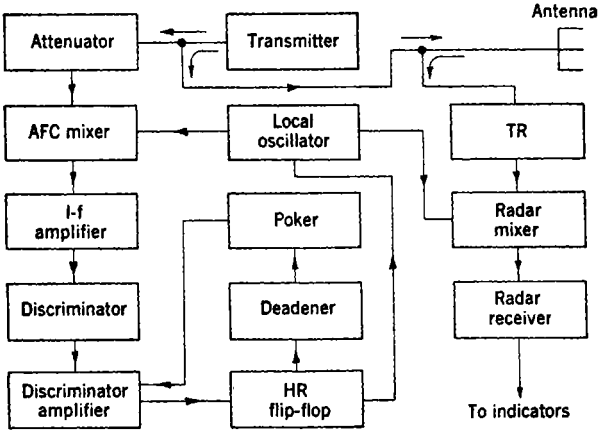


FIG. 3·33.—Complete block diagram of single flip-flop AFC system.

criminator pulses, in the manner discussed above, or of pulses from a "poker." The tuning triode power is either fully on or fully off according to the condition of the HR. A special coupling circuit ensures the required one-to-one correspondence between pulse polarity and HR condition.

The poker is a conventional multivibrator, one of whose control grids is tied directly to the suppressor (g_3) of the discriminator amplifier.

[1] Vol. 16, Secs. 7·15 and 7·16, Radiation Laboratory Series.

During half of the poker cycle, therefore, g_3 is close to ground, and the amplifier functions normally. During the other half, however, g_3 is well below ground, and the amplifier plate current is cut off, thus desensitizing the circuit for one direction of frequency sweep during hunting. Furthermore, the turning on and off of the amplifier plate current is so abrupt that the corresponding fall and rise of the plate voltage, differentiated, serve as trigger pulses to operate the HR flip-flop, causing a searching heat reversal at each reversal of the poker.

When the system is locked, similar but much more frequent reversals occur due to normal pulses from the discriminator. To prevent the poker from continuing to operate, thereby desensitizing the amplifier, the wave form at the HR plate is differentiated and applied to a diode detector, the "deadener," whose negative output is applied in turn to the grid of one of the triodes of the poker multivibrator. This triode is thus held in a nonconducting condition so that the completion of the multivibrator cycle is prevented.

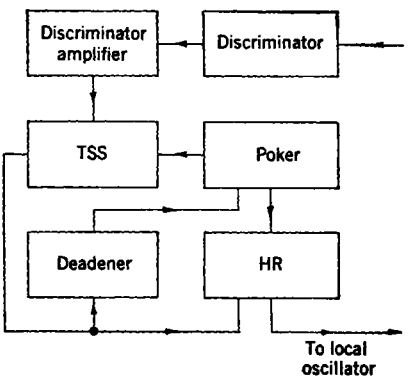


FIG. 3·34.—Block diagram of double flip-flop AFC control circuit.

In the DFF system, two flip-flops are used: a trigger sign selector (TSS) and a heat-reversal flip-flop (HR). The couplings from discriminator amplifier to the TSS and from HR to the tuner-triode grid are identical respectively with the input and output couplings of the HR of the SFF system. In addition, the two are coupled together in such a way that every reversal of the TSS causes a reversal of the HR. Thus, once the relative phases of the two flip-flops have been set, they act as one and provide locking and wrong-sideband rejection in a manner identical with that of the SFF system.

An entirely different poker mechanism is required. The device used is a screen-coupled phantastron oscillator, which has already been described in Sec. 3·13. The screen-voltage waveform (See Fig. 3·30*c*) is differentiated by an *RC* network, and the resultant positive pulses are used to operate the flip-flops. The phantastron period is long enough to allow the local oscillator to sweep over the entire band between successive pulses.

Assume that the transmitter is off, and consider the searching-sweep mechanism. Pulses from the poker are fed into the HR in such a way as to cause it to reverse at each pulse. Thus the tuner-triode power is

reversed at each pulse, and the oscillator sweeps back and forth over the band. Simultaneously, poker pulses are fed into the TSS, but the coupling is of such a nature that instead of always reversing it, they *preset* it to a definite condition. That is, after each pulse, the TSS will always be found ready to accept positive pulses from the discriminator. Examination of Fig. 3·31 shows that, with the TSS preset, coming in from *either* direction it will ignore the first (negative) pulses from the first sideband encountered and will accept the positive pulses that appear after the crossover frequency has been passed. If, therefore, the transmitter is turned on during a sweep, the TSS will reverse just after crossover has been passed. This reversal, transmitted through the HR to the tuner triode, causes a heat reversal. As soon as crossover is repassed, the TSS again reverses, this time from negative trigger pulses, and so on. If the LO frequency lies between the sidebands at the instant the transmitter is turned on, the positive pulse from the second sideband will cause a heat reversal, sending the oscillator back to the first sideband, where it will lock.

A "deadener" must also be used here, since one poke from the phantastron has an even chance of reversing the HR without reversing the TSS, thereby destroying the phase relation between the two and setting them up for locking on the other sideband. The deadener itself is similar to the one used in the SFF system, the negative bias from the diode being applied to the suppressor grid of the phantastron. If the system loses control, the deadener voltage will leak away in a time short compared with the total cycle, and a positive pulse will forthwith appear. This pulse will return the oscillator either to the original sideband or to the other sideband, according to which direction of drift existed at the time of loss of control. In either case, control will be reestablished in a short time.

ABSOLUTE-FREQUENCY AFC SYSTEMS

3·15. Crystal-controlled Frequency-multiplication Systems.—The use of a piezoelectric quartz crystal is standard procedure in commercial communications work at the lower frequencies. It is applicable whenever one or only a few fixed-frequency stations are to be received. It is difficult, however, to produce crystals whose fundamental frequency is much in excess of 10 Mc/sec and clearly impossible when the frequency is in the microwave region.

At lower frequencies, frequency multipliers are often used to obtain the advantages of crystal stability in the higher-frequency ranges. The method has been extended to the microwave region, but the equipment is somewhat complex to use in an ordinary receiver and is intended primarily as an accurate frequency standard. With such equipment, frequencies

have been established in the 3000- and 10,000-Mc/sec region with an absolute accuracy of one part in 10^6 or better.

Sperry has designed and put into production a system in which the output of a 5-Mc/sec crystal oscillator is successively doubled and tripled up to 270 Mc/sec, after which it is applied to a two-cavity klystron amplifier. The input cavity is tuned to 270 Mc/sec, and the output cavity to the eleventh or twelfth harmonic thereof. About 1 mw of r-f power in the 3000-Mc/sec region may thus be obtained.

The Radiation Laboratory Test Equipment Group has built a somewhat different device. The 5-Mc/sec signal is multiplied up to 50, 100, or 500 Mc/sec as desired and applied in considerable strength to an ordinary 1N23 silicon crystal. The nonlinearity of the crystal results in an output current that is rich in harmonics as high as 10,000 Mc/sec. Since no highly selective r-f circuits are used, an output signal is obtained that contains energy spaced at 50-, 100-, or 500-Mc/sec intervals. This device is very useful for generating pips, for calibration of spectrum analyzers over a wide band, etc.

The power output from the crystal harmonic generator is far too small to serve in place of a local oscillator for a receiver, since at least $\frac{1}{2}$ mw at the crystal is needed. It has been proposed to overcome this deficiency by using the signal from the standard as the reference signal for an AFC system controlling a normal local oscillator. That is, output from the standard and from the local oscillator would be applied to a crystal mixer, and the resultant i-f output would be amplified and fed into a conventional discriminator and control circuit, presumably of the d-c amplifier type. The LO output would then be controlled to a frequency differing from that of the standard by the amount of the intermediate frequency and of accuracy limited only by the accuracy of the AFC at the intermediate frequency, which could easily be made a small fraction of a megacycle per sec.

3·16. "Video Discriminator" or "Beacon" AFC.—Radar beacons offer a new and peculiar tuning problem. No sample of the transmitter to which the receiver is to be tuned is available except at the short moment during which the rotating antenna is pointing directly at the beacon to be received. Time is therefore not available for a controlled local oscillator to search for and lock the received signal. The problem of finding the signal manually is likewise difficult, since it is necessary to have the tuning correct at the instant when the antenna points to the beacon and, in general, the operator does not know from which direction the signal will come.

To solve this problem, the radar set is provided with a built-in frequency standard, set at the frequency at which the local oscillator should operate. All beacon transmitters are held accurately to a single fre-

quency, so all local oscillators operating with a given type of i-f amplifier also operate at a single frequency.

In the simplest case frequency is controlled manually, and the frequency standard is a precision cavity lightly coupled to the local oscillator with a crystal and crystal current meter attached to the far side. This device serves then as tuning indicator, the presence of maximum crystal current indicating correct tune.

If AFC is desired, the same r-f arrangement is used, but control circuits are provided that automatically set the local oscillator to the frequency at which crystal current is the maximum.

The most successful systems for both reflector and thermal AFC utilize reflector modulation and a gas-tetrode *coincidence tube*. A small, low-frequency (60 cps to 30 kc/sec) voltage is applied to the reflector, causing a frequency modulation with a deviation of about $\frac{1}{2}$ Mc/sec. If the oscillator center frequency coincides with the cavity resonant frequency, the crystal current will have no component at the modulation frequency; but if the oscillator is detuned slightly to one side or the other, a signal at modulation frequency will appear. On one side, the phase of this signal will be the same as that of the modulating voltage; on the other, the phases will differ by 180° because of the change in sign of the response curve slope. The output crystal current flows through an impedance, across which as much as 0.25 volt rms signal may be developed. This signal is amplified greatly (for example, a 50-to-1 stepup transformer with a pentode resistance-coupled amplifier) and applied to the control grid g_1 of a gas tetrode (2050 or 2D21). Both the control grid and the shield grid g_2 are biased about 10 volts negative with respect to the cathode. As long as g_2 remains at -10 volts, no voltage (up to the saturation limit of the amplifier) applied to g_1 can cause the tube to fire.

A second signal at the modulation frequency and of about 10 volts peak amplitude is applied to g_2. Again, the tube will not fire for any value of the shield-grid voltage as long as the control grid is at -10 volts. Consequently, the only condition under which the gas-tube can fire is that in which both grids are swung toward ground simultaneously by their respective applied signals. This condition will occur only if both signals are in the same phase.

Thus it is seen that if the LO frequency is on one side of the cavity peak, the gas-tube will never fire; but if the frequency swings to the other side, it will fire (regularly, at the modulation frequency). This is precisely the condition which is present in the control circuit of the gas-tube AFC of Sec. 3·12. All that is necessary here, to have an efficient AFC, is, therefore, to add the gas-triode search sweep generator (V_3 of Fig. 3·26), the coincidence tube serving as V_2 in this circuit.

Operation of a phantastron control circuit is likewise easy. A small

resistor (2000 ohms) is inserted in series with the gas-tetrode cathode. Since the gas-tube condenser (C_1 of Fig. 3·26, for example) is tied directly from the plate to ground, the initial surge that discharges it causes the cathode to rise sharply (after which it falls back) as the condenser is discharged through the tube and the small resistor. But this is a simple positive trigger which can readily be coupled into the shunt detector of Fig. 3·29, for instance, to develop the negative search-stopping voltage discussed in Sec. 3·13. Operation is then as described there.

The application of this circuit to thermally tuned tubes is not so simple. In essence, operation is as follows. The gas-tetrode circuit is operated between ground and a negative supply. The thermal-triode-tuner grid is tied directly to the gas-tube plate, so that in equilibrium (no firings), the tuner grid is at ground. When the gas tube fires, the

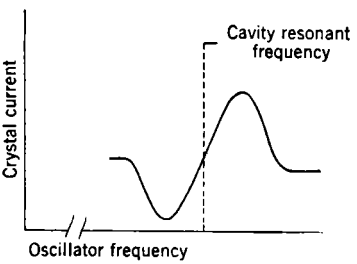


FIG. 3·35.—Microwave discriminator output.

plate is carried below ground, cutting off the tuner triode. The plate condenser is such that recovery requires 2 to 4 cycles at the modulation frequency. During this period, the LO frequency drifts away from the peak, after which it returns until another firing occurs.

Hunting is accomplished by making part of the gas-tube plate-load resistor the plate-load resistor of one section of a multivibrator. During part of its cycle, this section is nonconducting and has no influence on the circuit. During the other part it conducts and thereby effectively lowers the gas-tube plate-supply voltage to a point where the tuner triode grid will be cut off.

When the system is locked, a cathode-coupled diode detector similar to the one described in connection with the phantastron circuit is used to keep the grid of the nonconducting section of the multivibrator from coming close enough to its cathode to allow the cycle to be completed.[1]

3·17. The Microwave Discriminator.—One recent development that may have many future applications is a microwave discriminator whose

[1] Circuit diagrams and more complete descriptons of these circuits are found in Vol. 16, Secs. 7·17 to 7·19, Radiation Laboratory Series.

construction and theory of operation are described in detail elsewhere.[1] The microwave discriminator will be considered here as a "black box" containing a high-Q cavity (which may be tunable), a waveguide input to which an oscillator may be attached, and a low-frequency to d-c output lead. When the LO frequency is swept through the band of frequencies around the cavity resonant frequency, the output lead will trace out a typical discriminator response curve based about some zero level slightly above ground as shown in Fig. 3·35. That is, starting at a positive voltage, the output voltage will start to drop, then to rise, crossing the off-frequency value at crossover, continuing to a maximum positive value, and finally returning to the off-frequency condition.

This voltage may be applied to a high-gain d-c amplifier, whose output may then be used to control the LO-reflector voltage in the usual manner. The interesting property of this circuit is that with careful design enough follow rate may be achieved to compensate accurately for the inevitable frequency modulation of power-line frequency introduced by the a-c current in the heater circuit.

The stability achieved is remarkable. Two oscillators so controlled and operating at 10,000 Mc/sec can be beat together to produce a reasonably pure tone. The theoretical stability limit comes about from the inherent noise in the crystal. Since the circuit bandwidth need not be much greater than the power-line frequency, for which full compensation is required, the Johnson noise components are very small. There do appear to be extra low-frequency noise components, however, far greater than the Johnson noise. For best results extremely careful design of the d-c amplifier is necessary; push-pull circuits and much feedback should be used.

One immediate application would appear to be to make tunable receivers that can be set very accurately to a predetermined absolute frequency. The number of channels available within a given band becomes greater as frequency stability is improved. The possibility of using channels only 1 or 2 Mc/sec apart is apparent.

It may be noted that the previously discussed beacon AFC systems may likewise be used with tunable cavities. However, they all have present the rather large frequency modulation that must be applied to the reflector and are not suitable for precision work.

[1] Vol. 11, Chap. 2, Radiation Laboratory Series.

CHAPTER 4

I-F INPUT CIRCUITS

By Yardley Beers

4·1. Introduction.—In receivers for microwaves it is often the practice to feed the input signal from the antenna directly to a mixer having a conversion gain less than unity. To secure a good noise figure from such an arrangement considerable care must be used in the design of the coupling network between the mixer and the first i-f stage and of the first stage or stages of the i-f amplifier. This chapter will cover some of the considerations necessary in the design of this section of the receiver.

The noise figure of two networks in cascade may be written in either of two forms, as given in Chap. 1,

$$F_{12} = \frac{t_1 + F_2 - 1}{\mathcal{W}_1}, \tag{1·6}$$

or

$$F_{12} = F_1 + \frac{F_2 - 1}{\mathcal{W}_1}, \tag{1·3}$$

where F_{12} = noise figure of Networks 1 and 2 in cascade,
F_1 = noise figure of Network 1,
F_2 = noise figure of Network 2,
$\mathcal{W}_1$ = available power gain of Network 1,
t_1 = ratio of noise temperature of Network 1 to room temperature 300°K).

When dealing with crystal mixers it has become customary to use Eq. (1·6), with t_1 and $\mathcal{W}_1$ referring to the crystal. Since with present-day crystals t_1 is only slightly over unity, the noise figure of the receiver is almost directly proportional to F_2, which represents the noise figure of the i-f amplifier, thereby justifying the attention given to this factor. These considerations may be carried a step further by applying Eq. (1·3) to the i-f amplifier itself, letting Network 1 be the first stage and Network 2 be the rest of the amplifier. From this it is seen that if $\mathcal{W}_1$, the gain of the first stage, is high enough, the noise figure of the rest of the amplifier will have little effect on the over-all noise figure. If $\mathcal{W}_1$ is low, as may be the case in a very wide-band amplifier, the noise properties of the second or even the third stage become important.

4·2. Representation of Sources of Signal and Noise.—Before proceeding with a discussion of the ways of minimizing noise, some of the

methods commonly used for representing and characterizing sources of signal and noise will be reviewed. These follow directly from Thévenin's and Norton's theorems and the usual expressions for Johnson and shot noise and will be stated without proof.

1. Any two-terminal network containing linear impedances and generators may be represented as far as external effects are concerned by a single voltage generator E_a in series with an impedance Z_a or by a current generator I_a in shunt with an admittance Y_a. The maximum power will be obtained from the network if a load Z_a^* or Y_a^* (complex conjugate of Z_a or Y_a) is connected; this power is called the available power and is denoted by W_a. The network may be characterized by any of the four pairs of parameters: (*a*) E_a and Z_a, (*b*) I_a and Y_a, (*c*) W_a and Z_a, and (*d*) W_a and Y_a. These quantities are related by the following expressions:

$$I_a = \frac{E_a}{Z_a} = Y_a E_a, \tag{1}$$

and

$$W_a = \frac{E_a^2}{4R_a} = \frac{I_a^2}{4G_a}, \tag{2}$$

where R_a and G_a are the resistive and conductive parts of Z_a and Y_a, respectively.

2. A four-terminal network containing only reactive elements may be placed between the first network and the load without affecting the available power W_a. It will, in general, transform the other four quantities to new values: E_a', I_a', Z_a', and Y_a'. If the four-terminal network contains dissipative elements as well, W_a will be reduced.

3. Across the terminals of any two-terminal network containing no generator there appears a fluctuation voltage whose mean square value is

$$\overline{E^2} = 4kTR_aB, \tag{3}$$

where k = Boltzmann's constant = 1.38×10^{-23} joule/°K,
B = noise bandwidth (defined below), cps,
T = temperature of the network, °K,
R_a = resistive component of impedance between terminals.

If the terminals are short-circuited, a fluctuation current will flow whose mean square value is

$$\overline{I^2} = 4kTG_aB, \tag{4}$$

where G_a = conductive component of susceptance between terminals. From Eq. (2), the available noise power from the network is

$$W_n = kTB. \tag{5}$$

4. If a two-terminal network contains generators as well, the noise voltage, current, or power appearing at the terminals may be higher than the values given by Eqs. (3) to (5). A new value of temperature T_1, an effective noise temperature, may be used to describe this situation.

5. If a tube has in its plate circuit a noise current $\overline{I_n^2}$, such as that due to shot noise or partition noise, it may be replaced by a perfect tube with an equivalent voltage appearing in series with the grid

$$\overline{E_n^2} = \frac{\overline{I_n^2}}{g_m^2}. \tag{6}$$

This may, in turn, be replaced by a fictitious resistance, whose Johnson-noise voltage is equal to $\overline{E_n^2}$;

$$R_{eq} = \frac{\overline{I_n^2}}{4kTBg_m^2}. \tag{7}$$

If the grid circuit of a tube operated as a conventional grounded-cathode amplifier contains a signal source of internal impedance R, and if Johnson noise in the signal source and shot and partition noise in the tube are the only sources of noise, the noise figure of the combination is

$$F = 1 + \frac{R_{eq}}{R}. \tag{8}$$

The noise bandwidth B appearing in the equation is defined by the following expression

$$B = \frac{1}{\mathcal{W}_0} \int_0^\infty \mathcal{W}(f)\, df, \tag{9}$$

where $\mathcal{W}(f)$ = available power gain at frequency f,
$\mathcal{W}_0$ = maximum value of available power gain, usually occurring at band center.

The available power gain of a (four-terminal) network is the ratio of the available signal power at the output of the network to the available power from the signal source. The 3-db bandwidth $\mathcal{B}$ is more often used in characterizing the pass band of amplifiers and receivers. It is the width of the response curve between points where the response is 3 db below the maximum value, that is to say, where the power output for a constant input has dropped to half its maximum value. For a response curve of the form given by one single-tuned circuit,

$$B = 1.57\mathcal{B};$$

for two single-tuned circuits in cascade,

$$B = 1.22\mathcal{B};$$

whereas for most other pass bands normally met,

$$B \approx \mathcal{B}.$$

4·3. Properties of Amplifier Circuits.—The portion of the receiver under consideration here may be represented schematically as shown in Fig. 4·1. The mixer is represented as a current generator I_a shunted by

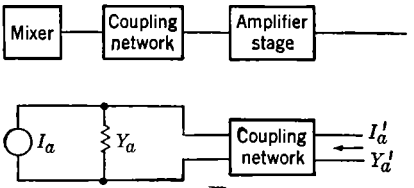


FIG. 4·1.—Equivalent circuit for mixer and coupling network.

an admittance Y_a. By the action of the coupling network these are transformed to new values I_a' and Y_a' as seen looking back from the first amplifier tube. It is to be expected intuitively that for a given amplifier tube and circuit there will be a value of Y_a' that gives the best ratio of signal to noise. It is the function of the design of the coupling network

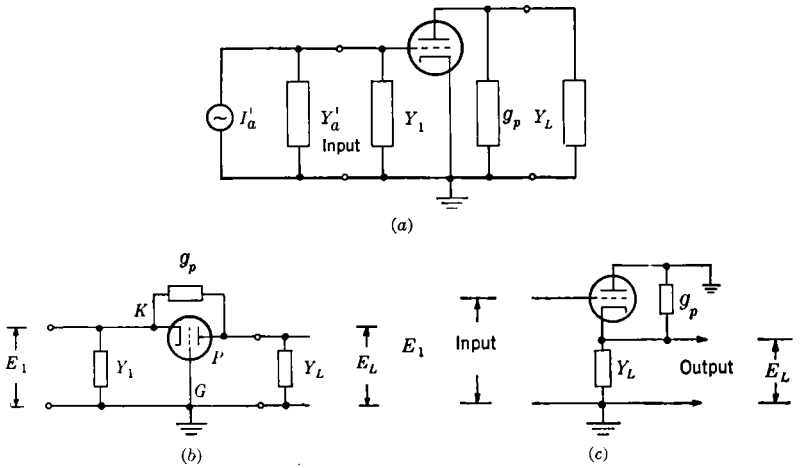


FIG. 4·2.—Amplifier circuits. (*a*) Grounded-cathode; (*b*) grounded-grid; (*c*) grounded-plate or cathode-follower.

to provide this value over the required pass band or, if the admittance value and pass band are incompatible, to achieve the best possible compromies. The properties of some of the possible amplifying circuits will be examined first to discover what values of Y_a' are required and what performance may be expected if the optimum transformed admittance is

attainable. Following this the problem of designing a coupling circuit will be taken up. For the first amplifier stages there is not only the choice between a triode and a tetrode or pentode but also the possibility of using a grounded-cathode, grounded-grid, or grounded-plate (cathode follower) circuit or even a combination of these in successive stages. As far as signal voltages are concerned, these three circuits may be represented as shown in Fig. 4·2*a*, *b*, and *c*. Sources of heater and plate power and screen or suppressor grids if present have been omitted.

The difference between the three circuits depends upon which of the three elements is common to the input and output. The admittance Y_t consists of the input capacity of the tube and the conductance that is due to any ohmic losses in parallel with the input. The conductive component of Y_1 is G_1. The dynamic plate conductance of the tube g_p is equal to the reciprocal of the dynamic plate resistance. The output capacity of the tube is considered as part of the load admittance Y_L. G_L is the conductive component of Y_L.

The properties of the three connections are summarized in Table 4·1. The equivalent noise resistance of the tube R_{eq} was defined in Eq. (7); g_m is the transconductance of the tube; and

$$\begin{aligned} g'_m &= g_m + g_p \\ &= \frac{(\mu + 1)g_m}{\mu}, \end{aligned} \tag{10}$$

where $\mu = g_m/g_p$ is the amplification factor of the tube. These formulas neglect transit time and other effects that become important at high frequencies. Some of the modifications that enter at higher frequencies will be considered in Chap. 5.

Input Admittance.—The input admittance of the grounded-cathode and cathode-follower amplifiers consists of Y_1 only, whereas that of the grounded-grid amplifier contains a second term, a so-called "electronic admittance," which is caused by feedback properties of this connection due to the flow of output current through the input circuit. In Sec. 4·5 it will be shown that such an electronic admittance can be produced by other types of feedback. In the case of the grounded-grid amplifier this electronic admittance almost always has a magnitude very much greater than $|Y_1|$.

Output Admittance.—In calculating the combined noise figure of two stages in cascade according to Eq. (1·3), the load resistor in the output of the first stage is included as part of the first stage. The noise figure of the second stage F_2 depends upon the admittance of the signal source, which in this case is the output admittance of the first stage, including this load resistor. Therefore, it is convenient to define output admittance in the following manner. Suppose that an auxiliary load admit-

TABLE 4·1.—PROPERTIES OF THREE AMPLIFIER CONNECTIONS

Property	Type of connection		
	Grounded cathode	Grounded grid	Cathode follower
Input admittance	Y_1	$Y_1 + \dfrac{g'_m Y_L}{g_p + Y_L} \approx Y_1 + g'_m$ *	Y_1
Output admittance	$g_p + Y_L$	$\dfrac{g_p(G'_a + G_1)}{g'_m + G'_a + G_1} + G_L$† $\approx \dfrac{(G'_a + G_1)}{\mu + 1} + G_L$‡	$g'_m + Y_L$
Voltage gain $\mathcal{G}$	$\dfrac{g_m}{(g_p + Y_L)}$	$\dfrac{g'_m}{(g_p + Y_L)}$	$\dfrac{g_m}{(g'_m + Y_L)}$
Available power gain $\mathcal{W}$	$\dfrac{g_m^2 G'_a}{(g_p + G_L)\lvert Y'_a + Y_1\rvert^2}$	$\dfrac{g_m'^2 G'_a}{[(G'_a + G_1)g_p + (g'_m + G'_a + G_1)G_L](g'_m + G'_a + G_1)} \approx \dfrac{G'_a}{G_L}$ ‡* †	$\dfrac{g_m^2 G'_a}{(g'_m + G_L)\lvert Y'_a + Y_1\rvert^2}$
Value of G'_a for maximum value of $\mathcal{W}$	G_1	$\sqrt{g'_m G_1 + G_1^2 + G_L \dfrac{g'_m(g'_m + G_1)}{g_p + G_L}}$ ‡ $\approx G_1 + g'_m$ *	G_1
Maximum value of $\mathcal{W}$, $\mathcal{W}_{max}$	$\dfrac{g_m^2}{4G_1(g_p + G_L)}$	$\dfrac{g_m'^2}{4(g'_m + G_1)G_L}$ *	$\dfrac{g_m^2}{4G_1(g'_m + G_L)}$
Noise figure F	$1 + \dfrac{G_1}{G'_a} + \dfrac{\lvert Y'_a + Y_1\rvert^2}{G'_a}\left(R_{eq} + \dfrac{G_L}{g_m^2}\right)$	$1 + \dfrac{G_1}{G'_a} + \dfrac{\lvert Y'_a + Y_1\rvert^2 g_m^2}{G'_a g_m'^2} R_{eq} + \dfrac{G_L}{G'_a}$ §	$1 + \dfrac{G_1}{G'_a} + \dfrac{\lvert Y'_a + Y_1\rvert^2}{G'_a}\left(R_{eq} + \dfrac{G_L}{g_m^2}\right)$
Value of G'_a for minimum noise figure	$\sqrt{G_1^2 + \dfrac{G_1}{R_{eq} + \dfrac{G_L}{g_m^2}}}$	$\sqrt{G_1^2 + \dfrac{(G_1 + G_L)g_m'^2}{R_{eq} g_m^2}}$ ‖	$\sqrt{G_1^2 + \dfrac{G_1}{R_{eq} + \dfrac{G_L}{g_m^2}}}$

* Approximate value for $|Y_L| \gg g_p$.
† With total input and output susceptances zero.
‡ Approximate value for $g_m' \gg (G_a' + G_1)$.
§ Fourth term (contribution of G_L) approximate. See * and ‡ above.
‖ Effect of load G_L is evaluated with approximations. See * and ‡ above.

tance Y'_L were connected in parallel with the output of the tube and the actual load Y_L and that Y'_L were varied until a maximum amount of power were dissipated in it. The output admittance is the complex conjugate of this value of Y'_L. The values given in Table 4·1 are obtained in this way.

It will be noted that the output admittance in each case is equal to the load admittance plus another term that represents the output admittance of the tube alone. In practice the amplification factor μ is large compared with unity, so that g_m and g'_m are large compared with g_p. Also g'_m is usually large compared with $(G'_a + G_1)$. Therefore the output admittance of the cathode follower g'_m is much larger than those of the grounded-cathode or the grounded-grid circuits. The cathode follower may be said to have a large electronic output admittance, just as the grounded-grid circuit has a large electronic input admittance.

Voltage Gain.—The amplification factor μ is almost always large compared with unity. Therefore g'_m is only a few per cent larger than g_m at most, and the voltage gain of a tube used as a grounded-grid amplifier is only slightly larger than that obtained with the same tube as a grounded-cathode amplifier. The voltage gain of a cathode follower can never exceed unity, because g'_m is always larger than g_m.

Available Power Gain.—In the definition of available power gain $\mathcal{W}_a$ an auxiliary load admittance Y'_L is employed as in the definition of output admittance. In this way $\mathcal{W}_a$ is defined as the maximum power that can be dissipated in Y'_L divided by the available power of the signal source. Under the conditions that G_L is large compared with g_p and that g'_m is large compared with $(G'_a + G_1)$, which are usually encountered in practice, the available power gain of the grounded-grid amplifier is very low, approximately equal to G'_a/G_L, which may be as small as unity. While the voltage gain of such a stage may be large, its available power gain is low because the input conductance is large compared with the output conductance of the tube. Although the voltage gain of a cathode follower is always less than unity, its available power gain may be considerably greater than unity because its output conductance is so much larger than its input conductance. However, as g'_m is considerably larger than g_p, the available power gain of the grounded-cathode amplifier is considerably larger than that of the cathode follower. Thus it will be seen that in reducing the effect of second-stage noise to a minimum by the use of a first stage having maximum available power gain, the grounded-cathode connection is the most suitable. The available power gain of the tube alone can be found by placing Y_L equal to zero.

Single-stage Noise Figure.—The expressions for noise figure contained in Table 4·1 have been based on the assumptions that transit time and other high-frequency effects can be neglected and that the first stage is

followed by other stages and an indicating device whose over-all pass band is less than the bandwidth of the input circuit. Also the effect of grid current has been neglected. In situations where the second assumption is not valid, these expressions can be considered as approximations, or, for exactness, the value of the noise figure averaged over the pass band with respect to the actual power gain can be calculated.

In each case, the expression for the noise figure contains four terms which represent, respectively, the contributions due to (1) the Johnson noise of the signal source, (2) the Johnson noise of the input conductance G_1, (3) the shot noise of the plate current, and (4) the Johnson noise of the load conductance. Except for this last term, the three types of connections have essentially the same noise figure. Generally G_L/g_m^2 is small compared with R_{eq} so that the contribution of the Johnson noise of G_L is virtually negligible in the case of the grounded cathode and cathode follower. However, the term G_L/G'_a, which gives an approximate value of the effect of the load G_L on the noise figure of the grounded-grid amplifier, may be appreciable. This effect is related to the low available power gain of the grounded-grid connection. In any case the contribution of the load resistor is usually somewhat smaller than the contribution due to noise in the second stage.

Several conclusions can be drawn from the expressions for noise figure. (1) The noise figure is a minimum when the susceptive component of $Y'_a + Y_1$ is zero, that is, when the total admittance in the input circuit is purely conductive. (2) The noise figure is lower when G_1 is made smaller. Therefore, the noise figure is never improved by increasing losses in the input circuit.

In the case of grounded cathode and cathode followers it can be seen that the minimum noise figure is obtained with a value of G'_a that is larger than G_1, the input conductance of the tube. When the contribution of G_L is small and when R_{eq} is small, the value of G'_a that gives the minimum noise figure is very different from G_1, and a considerable improvement in noise figure can be obtained by a proper impedance mismatch. However, it is obtained at the expense of gain. When R_{eq} becomes large, the optimum value of G'_a approaches G_1; that is, the optimum impedance approaches a perfect match. In other words, when R_{eq} is large, the noise originates almost entirely in the plate circuit, and therefore it is necessary to get the maximum gain at the input circuit to make the signal as large as possible compared with the plate noise.

With grounded-grid amplifiers, the value of G'_a that gives a minimum noise figure may be in theory either larger or smaller than the value that matches the input conductance and gives maximum gain. However, in practice, the value that gives maximum gain is very large and usually is greater than that which gives optimum noise figure. Therefore, the

direction of optimum mismatch is the reverse of that for grounded-cathode and cathode-follower amplifiers.

If the effect of the load resistor is neglected, if g_m/g'_m is replaced by its approximate value unity, and if the susceptive component of $Y'_a + Y_1$ is zero, the noise figure of all three connections is given by

$$F = 1 + \frac{G_1}{G'_a}\left[1 + G_1 R_{eq}\left(1 + \frac{G'_a}{G_1}\right)^2\right]. \tag{11}$$

The noise figure can then be expressed in terms of two parameters G'_a/G_1 and G_1R_{eq}. The loss in gain due to mismatching the signal source

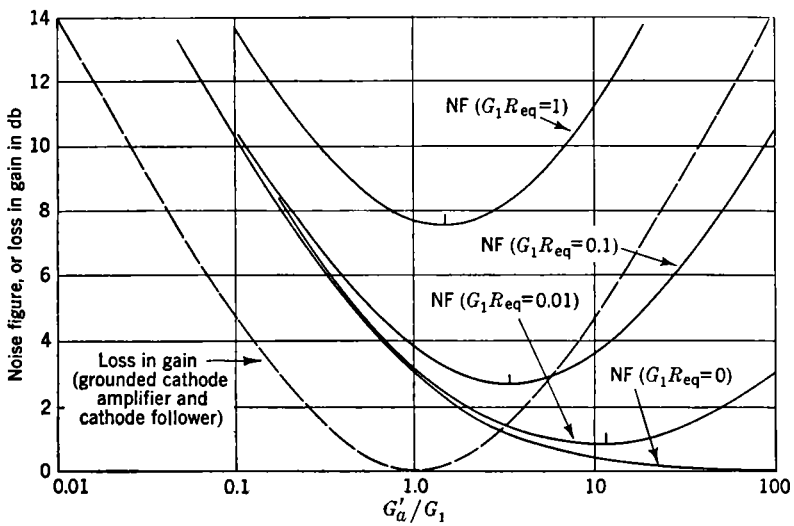


FIG. 4·3.—Noise figure and loss of gain as a function of mismatch.

to the input of a grounded-cathode or cathode-follower amplifier can also be expressed in terms of G'_a/G_1:

$$\frac{\mathcal{W}}{\mathcal{W}_{max}} = \frac{4\dfrac{G_1}{G'_a}}{\left(1 + \dfrac{G_1}{G'_a}\right)^2}. \tag{12}$$

Accordingly, both quantities are shown plotted against G'_a/G_1, in Fig. 4·3; different values of G_1R_{eq} giving a family of curves for the noise figure. It is to be noted that the optimum value of G'_a/G_1 approaches unity (a perfect match) as G_1R_{eq} increases. However, at any value of G'_a/G_1 the noise figure is always increased by an increase in G_1R_{eq}. The position of minimum noise figure has been indicated on each curve by a short vertical

line. For $G_1R_{eq} \ll 1$ the optimum value of G'_a/G_1 is approximately

$$\left(\frac{G'_a}{G_1}\right)_{opt} \approx \frac{1}{\sqrt{G_1R_{eq}}} \tag{13}$$

or

$$(G'_a)_{opt} = \sqrt{\frac{G_1}{R_{eq}}}. \tag{13a}$$

This is the value given in Table 4·1 for the transformed input conductance to produce minimum noise figure, with the additional approximation $G_L/g_m^2 \ll R_{eq}$ which permits neglect of the effect of the plate-load resistor.

In this treatment, wherein the noise in later stages is neglected and the bandwidth of these stages is assumed to be small compared with the bandwidth of the input circuit, it is sufficient in choosing a tube to find one in which the product of G_1 and R_{eq} is small, and either quantity may be large provided that the other is small. The question of whether a large value of R_{eq} with a small value G_1 is better or worse than a small value of R_{eq} and a correspondingly large value of G_1 depends on the desired bandwidth of the individual amplifier. In a wide-band amplifier, the admittance presented by the coupling network is usually large. Therefore, the input conductance of the tube itself has little relative effect, and in this case a tube that has a low value of R_{eq} but might have a relatively large input conductance is to be preferred to one whose value of R_{eq} is large and whose input conductance is small. With a narrow-band amplifier the reverse is true.

A situation wherein the combined bandwidth of the later circuits is wide compared with the bandwidth of the coupling circuit is to be avoided if the optimum noise figure is to be obtained. In such situations a reduction of the bandwidth of some circuit located in one of the later stages will reduce the contribution to the output of the shot noise generated in the plate of the first stage and by other sources further on in the amplifier without reducing the contribution of the signal or the Johnson noise of the signal source.

4·4. General Discussion of Feedback Effects.—The introduction of feedback produces changes in the bandwidth and gain of an amplifier but usually causes little or no change in the noise figure.[1] The change of bandwidth and gain can be represented as being due to change of input admittance and, to a less important extent, a change in output admittance of the stages containing feedback. One important consequence is that in situations where it is impossible to obtain sufficient input circuit

[1] The effect on the noise figure has been treated previously by W. A. Harris, "Fluctuations in Vacuum-tube Amplifiers and Input Systems," *RCA Rev.*, **V**, 505–524, April 1941, and **VI**, 115–124, July 1941. See also H. W. Bode, *Network Analysis and Feedback Amplifier Design*, Van Nostrand, New York, 1945, pp. 34–35.

bandwidth without the use of supplementary loading, much better noise figures can be obtained by producing this loading by controlled feedback than by the use of ohmic resistors. Another consequence is that when feedback is present owing to unavoidable distributed parameters within the tubes, the noise figure is generally not affected in any serious way.

The lack of effect on noise figure may be illustrated by the case shown in Fig. 4·4, in which B and F are amplifiers. Feedback can be introduced by closing the switch S so that some of the output of the amplifier B is transmitted through the network E back to the input. In the diagrams the input and output of the various units are denoted by single terminals; actually of course, they have pairs of terminals which may be connected at a junction of three units either in series or in parallel. The point J represents an independent source of noise within the amplifier B, and $\mathcal{G}_C$ and $\mathcal{G}_D$ denote the voltage gains between the input of B and J and

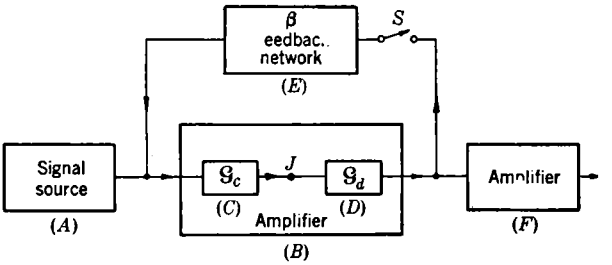


FIG. 4·4.—Block diagram of amplifier with feedback.

between J and the output of B, respectively; β is the voltage gain of the network E, which is generally less than unity. If the following assumptions are made: (1) the network E and the amplifier B transmit signals only in the directions denoted by the arrows, (2) at all times the amplifier B has sufficient gain to make sources of noise within amplifier F negligible, and (3) the network E contains no sources of noise, it can be shown that the introduction of feedback by closing S is the equivalent of modifying the emf of every independent noise (or signal) voltage contained within the loop, including those at the input and output, by the factor $1/(1 - \mathcal{G}\beta)$, where $\mathcal{G} = \mathcal{G}_C\mathcal{G}_D$ is independent of the location of the point J. Thus signal and all sources of noise are modified in the same way, resulting in no change in the noise figure or signal-to-noise ratio.

One situation that would not be covered by this argument would be that in which there is a source of noise within the second amplifier F that is coherent with a source in the first amplifier. By proper choice of the value of $\mathcal{G}\beta$ these sources could be made to nullify each other, resulting in an improvement of noise figure. Among the sources of noise that are of this type are those due to partition of the electron stream

between the screen grid and plate in a multigrid tube. At one instant a fluctuation in the screen-grid current is accompanied by a fluctuation in the plate current equal in magnitude but opposite in sign. By introducing feedback between the screen grid and the input of the tube, the fluctuation in the screen current can be made to cancel the effect of the fluctuation in the plate current.

Change of Impedance Due to Feedback.—Next, the change of input admittance will be calculated for one common situation shown in Fig. 4·5. One input and one output terminal of the amplifier are grounded, and feedback is produced by an admittance Y_{12} connected between the ungrounded terminals. The total output admittance Y_2 consists of the load Y_L and the output admittance of the last tube in parallel. The transformed internal admittance of the signal source is Y'_a and Y_1 repre-

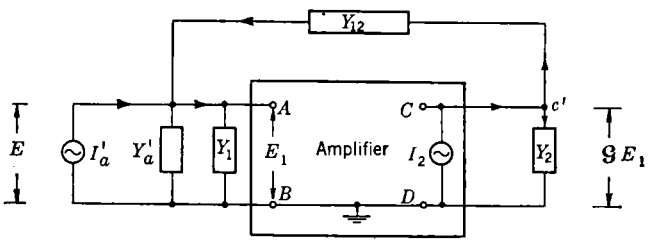


FIG. 4·5.—Amplifier with admittance between input and output circuit.

sents the rest of the admittance across the input (with Y_{12} disconnected). It will be assumed that Y_1 is independent of Y_2 (an assumption that is not valid if the amplifier consists of a grounded-grid amplifier whose load admittance Y_L is small compared with its dynamic plate conductance g_p). The voltage gain of the amplifier with feedback is denoted by $\mathcal{G}$, without feedback by $\mathcal{A}$. The amplifier may contain any number of stages, but it is assumed that at the center of the pass band the phase angle between the input and output voltages is either zero or some integral multiple of 180°. In the case of a single grounded-grid or cathode-follower stage, it is zero; and in the case of a single grounded-cathode amplifier, it is 180°. The total voltage across Y_{12} is then equal to $E_1(1 \pm \mathcal{G})$, where the upper sign pertains to a phase angle of 180° and the lower sign to a phase angle of zero. I_{12}, the current flowing through Y_{12}, is then

$$I_{12} = Y_{12}E_1(1 \pm \mathcal{G}). \tag{14}$$

The input current I_1 is the same as though an admittance Y_e were connected in parallel with the input and the feedback through Y_{12} were disconnected, where

$$Y_e = \frac{I_{12}}{E_1} = Y_{12}(1 \pm \mathcal{G}). \tag{15}$$

The relation between $\mathcal{G}$ and $\mathcal{A}$ can be found by writing the nodal equation for the node c' in each case:

$$I_2 = Y_2\mathcal{G}E_1 + (\mathcal{G} \pm 1)E_1Y_{12},$$

and

$$I_2 = Y_2\mathcal{A}E_1.$$

By equating these expressions,

$$\mathcal{G} = \frac{\mathcal{A} \mp \dfrac{Y_{12}}{Y_2}}{1 + \dfrac{Y_{12}}{Y_2}}. \tag{16}$$

The first term in the numerator is due to signal coming through the tube, and the second term is due to signal fed from input to output through Y_{12}. By substituting this value of $\mathcal{G}$ into Eq. (15),

$$Y_e = \frac{Y_{12}(1 \pm \mathcal{A})}{1 + \dfrac{Y_{12}}{Y_2}}. \tag{17}$$

If Y_{12} is small compared with Y_2, the voltage gain is not altered appreciably by connecting in the feedback admittance, or $\mathcal{G}$ is very nearly equal to $\mathcal{A}$.

Therefore it can be concluded that the output voltage is modified by the use of this type of feedback (1) by the change of loading on the signal source which changes the input voltage, a relatively large effect, and (2) by the change in voltage gain, usually a relatively small effect. This change of loading on the input circuit changes the pass band. If Y_e, averaged over the pass band, contains a positive conductance or a negative capacity, the bandwidth can be made to increase; whereas if it contains negative conductance or positive capacity, it will cause the bandwidth to decrease. If $Y_e = -(Y_a' + Y_1)$, the circuit will oscillate. The change in voltage gain can be represented as though it were due to a change in output admittance, and therefore it corresponds to a change in the pass band of the output circuit.

4·5. Miscellaneous Types of Feedback and Their Effects on Noise.—In this section various types of feedback and their direct and indirect effects on noise figure will be discussed. These topics will include the cathode-follower and grounded-grid amplifiers considered as grounded-cathode amplifiers with feedback, feedback through a capacity or a resistor connected from the output to the input, and feedback due to cathode-lead inductance.

Cathode Follower Considered as a Feedback Amplifier.—The cathode follower is essentially a grounded-cathode amplifier with feedback caused

by connecting the output voltage in series with the input. This type of feedback is of the general type covered by the discussion on noise figure in the previous section, and, in fact, all of the assumptions therein are completely satisfied. Therefore, according to the result of this discussion the noise figure of the cathode follower should be the same as for the same tube used as a grounded-cathode amplifier, a fact confirmed by direct calculation as indicated in Table 4·1.

The Grounded-grid Amplifier Considered as a Feedback Amplifier.—The grounded-grid amplifier can be considered as a grounded-cathode amplifier with feedback of such a nature that the output current flows through the input circuit. This type of feedback is not covered by the general discussion of Sec. 4·4, and therefore it is not surprising that the direct calculation of the noise figure gives a result (see Table 4·1) which is not completely identical with the expressions for the grounded-cathode and cathode-follower amplifiers. Nevertheless, except for the effect of the Johnson noise of the load, the expressions are essentially the same.

This type of feedback gives rise to an electronic input admittance that, as can be seen from Table 4·1, usually contains a very large conductive component. In fact, this amount of electronic loading is usually much more than is made necessary by bandwidth considerations, and it would be advantageous to trade some of the electronic loading for an increase in available power gain, a quantity in which (see Sec. 4·3) the grounded-grid amplifier is deficient. Unfortunately, this cannot be done with this kind of feedback; but by the resistor-feedback and cathode-lead-inductance feedback methods to be discussed later, it is possible to control the amount of feedback so that no more electronic loading and no larger reduction in gain are produced than is really necessary.

Feedback Produced by a Resistor between Input and Output.—The use of a resistor between the grid and plate of a grounded-cathode amplifier has been widely used as a means of increasing the over-all bandwidth of an amplifier. This subject has been discussed at length in Chap. 8 of Vol. 18. It can be seen by examination of Eq. (15) that when Y_{12} consists of a pure conductance, Y_e consists of a conductance at the center frequency of the output circuit, whereas at other frequencies it contains also positive and negative susceptances when the load contains positive and negative susceptive components, respectively. A. B. Macnee[1] has made an exact analysis of the noise figure of a single stage employing this type of feedback. In this analysis he has included the effects of the Johnson noise of the feedback resistor and the load resistor. His result for the noise figure at the center frequency (all admittances assumed to be pure conductances), assuming that the input loss conductance G_1 is

[1] RL Group Report No. 61–10/1/45.

negligible compared with the transformed signal source conductance G'_a, is as follows:

$$\left.\begin{aligned} F = 1 &+ R_{eq}G'_a\left(\frac{1+\dfrac{G_{12}}{G'_a}}{1-\dfrac{G_{12}}{g_m}}\right)^2 && \text{(Shot-noise term)} \\ &+ \frac{G_{12}}{G'_a}\left(\frac{1+\dfrac{G'_a}{g_m}}{1-\dfrac{G_{12}}{g_m}}\right)^2 && \text{(Term due to Johnson noise of } G_{12}\text{)} \\ &+ \frac{G_2G'_a}{g_m^2}\left(\frac{1+\dfrac{G_{12}}{G'_a}}{1-\dfrac{G_{12}}{g_m}}\right)^2 && \text{(Term due to Johnson noise of load resistor)} \end{aligned}\right\} \quad (18)$$

It can be seen by examining this equation that for small values of the feedback conductance G_{12}, the noise figure is not altered. However, when G_{12} approaches g_m, the noise figure approaches infinity. This corresponds to a value of G_{12} that gives zero gain, a situation that is not very practical.

Consider a practical case. Suppose the tube is a triode-connected 6AK5 (g_m equal to 5000 μmhos, g_p = 200 μmhos, and equivalent noise resistance R_{eq} = 500 ohms), that G'_a = 500 μmhos, and that the voltage gain without feedback is 10 (G_L = 300 μmhos). Then the noise figure without feedback is (assuming $G_1 = 0$)

$$F = 1 + 0.25 + 0.00 + 0.006 = 1.256 \text{ or } 1.0 \quad \text{db.}$$

Suppose G_{12} is selected to cause the input conductance of the tube to match the signal source conductance G'_a. According to Eq. (17) the proper value of G_{12} is 53 μmhos. The noise figure is then, from Eq. (18),

$$F = 1 + 0.309 + 0.123 + 0.007 = 1.44 \text{ or } 1.6 \quad \text{db.}$$

If the same amount of loading on the input circuit had been produced by connecting an ohmic resistor across the input of the tube, the noise figure, obtained by placing G_1 equal to G'_a in Table 4·1, is F = 3.02, or 4.8 db. Therefore, by producing the loading by feedback, the noise figure deteriorates only 0.6 db from the condition of no loading; but if the loading is produced by a resistor, the noise figure deteriorates by 3.8 db.

This value of G_{12} causes the voltage gain $\mathcal{G}$ to decrease from 10 to 9.0, and the available power gain is reduced from 20 to 16.2 db.

Feedback Produced by Cathode-lead Inductance.—It has been shown that feedback due to a cathode-lead inductance produces an input admit-

tance of a vacuum tube that contains a conductive component.[1] This type of feedback can also be used to load down the input circuit to widen its bandwidth. A. B. Macnee has calculated the noise figure for a grounded-cathode stage operated in this way[2] (see Fig. 4·6).

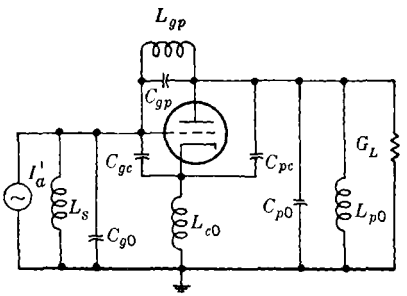


FIG. 4·6.—Amplifier with cathode-lead inductance.

When the assumptions stated below are fulfilled, it can be shown that the input conductance due to this type of feedback is

$$G_e = \frac{g_m C_{gc}}{L_{c0}\left[g_m^2 + \left(\omega_0 C_{gc} - \frac{1}{\omega_0 L_{c0}}\right)^2\right]}, \tag{19}$$

or when L_{c0} is small,

$$G_e = \omega_0^2 L_{c0} C_{gc} g_m \quad \text{approximately,} \tag{19a}$$

where g_m = the transconductance of the tube,
L_{c0} = the inductance between cathode and ground,
C_{gc} = the capacity between grid and cathode,
ω_0 = the angular frequency at the center of the band.

The assumptions are

1. The amplification factor of the tube is large compared with unity.
2. The load admittance connected from plate to ground is large compared with the admittance from plate to cathode.
3. The grid-plate capacity C_{gp} has been neutralized by an inductance L_{gp} in parallel with it, having such a value that the two are resonant at ω_0.
4. The inductance L_s between grid and ground satisfies the following relation
$$\omega_0^2 L_s(C_{gc} + C_{g0}) = 1,$$
where C_{gc} is the grid-cathode capacity and C_{g0} is the grid-ground capacity.
5. The inductance in parallel with the output L_{p0} satisfies the following relation
$$\omega_0^2 L_{p0}(C_{pc} + C_{p0}) = 1,$$
where C_{pc} and C_{p0} are the capacities from plate to cathode and ground, respectively.

[1] M. J. O. Strutt and A. Van der Ziel, *Proc. I.R.E.*, **26**, 1011–1032 (1936).

[2] A. B. Macnee, RL Group Report No. 61–5/25/1945. Equations (19) to (22) inclusive have been taken from this report.

From the exact expression, it can be seen that G_e has a maximum value when

$$L_{c0} = \frac{1}{\omega_0}\sqrt{g_m^2 + \omega_0^2 C_{gc}^2}. \tag{20}$$

Under these conditions, with the additional assumption that the ohmic-loss input conductance is negligible, Macnee has shown that the noise figure is

$$F = 1 + G_a' R_{eq}\left[(1 - \omega_0^2 C_{gc} L_{c0})^2 + \omega_0^6 C_{gc}^4 \frac{L_{c0}^2}{G_a'^2}\right] + \frac{G_2 G_a'}{g_m^2}\left[\left(1 + \omega_0^2 C_{gc} L_{c0}\frac{g_m}{G_a'}\right)^2 + g_m^2 \omega_0^2 L_{c0}^2\right] \tag{21}$$

where G_a' is the transformed signal source conductance and G_2 is the sum of G_L and g_p.

The available power gain is

$$\mathcal{W} = \frac{g_m^2}{G_a' G_2\left[\left(1 + g_m \omega_0^2 L_{c0}\frac{C_{gc}}{G_a'}\right)^2 + \omega_0^2 L_{c0}^2 g_m^2\right]}. \tag{22}$$

Consider the same numerical example contained in the discussion of feedback by a grid-plate resistor ($g_m = 5000$ μmhos, $g_p = 200$ μmhos, $G_L = 300$ μmhos, $R_{eq} = 500$ ohms, $G_a' = 500$ μmhos).[1] Suppose that at a frequency of 30 Mc/sec a value of cathode-lead inductance is chosen to produce an input conductance that matches this value of the transformed signal-source conductance G_a'. Equation (19) gives two values, 0.65 μh and 1.74 μh. The former gives a noise figure of 1.0 db and an available power gain of 13.2 db, and the latter gives a noise figure of 1.2 db and an available power gain of 7.1 db. These are to be compared with values of a noise figure of 1.0 db and an available power gain of 20 db with no feedback and a noise figure of 1.6 db and an available power gain of 16.2 db with resistor feedback.

This type of feedback gives a somewhat better noise figure than the resistor type, mainly because the resistor type introduces an additional source of noise, the Johnson noise of the resistor. However, it is generally not so well suited as a means of widening the pass band of the input circuit. The loading is not symmetrical around the center frequency, giving an asymmetrical pass band. On the other hand, a certain amount of this type of loading is inevitable at high frequencies, and this treatment indicates that the effect causes no great deterioration of noise figure.

[1] This value of G_a' equal to 500 μmhos is a little smaller than is usually encountered in practice at this frequency. Therefore the computed values of noise figure are slightly better than are usually realized.

The Effect of a Capacity between Input and Output.—In this case Y_{12} consists of a capacitance C_{12}. In the grounded-cathode amplifier, the upper (positive) sign is used in Eq. (15), indicating that at the center of the pass band of the plate circuit, the electronic input admittance Y_e consists of a pure capacitance. This effect, called the Miller effect, is well known. At other frequencies the phase between the input and output changes, yielding a conductive component to Y_e, and the capacitance decreases from a maximum at the center of the pass band of the plate circuit.

If the circuit in the plate has an inductive susceptance, which in the case of a parallel single-tuned circuit occurs at frequencies lower than the center frequency, it gives rise to a negative component of input conductance (which in extreme cases can produce oscillation); whereas, when the load contains a capacitive susceptance, a positive component of input conductance is produced. These effects produce an asymmetry in the pass band that is undesirable, and the increase in capacity tends to lower the gain-bandwidth product of the input circuit. With triode tubes the grid-plate capacity is relatively large, and these effects are serious unless the voltage gain is very small. Therefore, it is desirable to neutralize this capacity. A. B. Macnee, in making an exact analysis of the noise figure, has shown furthermore that even when the voltage gain is small, some improvement in noise figure can be obtained by neutralizing the grid-plate capacity, a conclusion that has been verified by experiment.

One method of neutralizing is to feed back to the input by an auxiliary network containing a small condenser a voltage that is equal in magnitude and opposite in phase to that which is fed back through the grid-plate capacity. The details of such networks have been known for many years and can be found in textbooks on radio engineering. A simpler method, which can be used when the center frequency of the amplifier is fixed, as is the case in an i-f amplifier, is to connect an inductance L_{12} in parallel with the grid-plate capacitance C_{12} such that L_{12} and C_{12} form a parallel-tuned circuit resonant at the center frequency of the amplifier. In this case Y_e is the admittance of a parallel-tuned circuit consisting of a capacitance $C_{12}(1 + \mathcal{G})$ and an inductance $L_{12}/(1 + \mathcal{G})$, and so is zero at the band center. It must, however, be allowed for in any calculations of the bandwidth of the input circuit. Thus it will be seen that this type of neutralization has not changed the bandwidth attainable in the input circuit, though it has reduced power transfer from output to input circuit.

In the case of the grounded-grid amplifier $\mathcal{G}$ is usually large, and in Eq. (15) unity can be neglected in comparison. The effects produced, which in this case is the capacity between cathode and plate, are essentially the same as with the grounded-cathode amplifier, but because the minus sign is used in Eq. (15), they are opposite in sign; Y_e con-

sists of a negative capacity plus a positive conductance with inductive loads and a negative conductance with capacitive loads. Usually, tubes of either triode or multigrid types can be found that have small values of C_{pc}, and with them these effects are small. In other cases it is necessary to neutralize as with the grounded-cathode amplifier.

With the cathode follower, $\mathcal{G}$ is always less than unity although frequently near it. Therefore, as the minus sign is used in Eq. (15), the increase of input admittance due to Y_{12} (which is connected from grid to cathode) is very small and, when $\mathcal{G}$ is nearly unity, is negligible. Consequently the cathode-follower connection affords a method of obtaining a small input admittance. A large portion of the input capacity of a tube, when used in the grounded-cathode or grounded-grid connections, is due to grid-cathode capacity. In the cathode-follower connection the effect of this grid-cathode capacity is nearly eliminated. It is interesting to note that with capacitive loads, the input admittance contains a negative conductance, relatively small in magnitude. Therefore it is theoretically possible to make the cathode follower oscillate in spite of the fact that its voltage gain is less than unity.

4·6. Selection of Tubes.—The selection of a tube for the input stage depends on several factors, the most important of which for most applications is the equivalent noise resistance R_{eq}, which is the measure of the noise produced by shot effect in the plate current. However, the available power gain also is of importance, of course, as the contributions of later stages can be made negligible if the input stage has sufficient gain.

In order to obtain a large available power gain the transconductance must be large and conductances connected across the input and output must be small. In order to have small conductances across the input and output in wide-band amplifiers, the input and output capacities must be small to maintain bandwidth. In the first stage a low input capacity is particularly desirable. Fortunately, as will become apparent from the discussion of equivalent noise resistances later in this section, generally speaking, a low value of the equivalent noise resistance is obtained when the transconductance is high. In special applications, of course, other properties may be of importance.

In triodes, the only source of random variations of plate current is the shot effect, which is much reduced by space charge at the cathode. In multielement tubes, there is an additional source, the random way in which the electron stream divides among the elements, giving rise to so-called "partition noise." Thus, if the sum of the currents to the other elements has a fluctuation at any instant, it is accompanied by fluctuation in plate current equal in magnitude but opposite in sign. When the current flowing to these other elements is large, the fluctuation in plate

current is especially large compared with that in a triode though the same value of plate current is flowing.

The following equations[1] for equivalent noise resistance of tubes are a useful practical guide:
for triode amplifiers,

$$R_{eq} = \frac{2.5}{g_m}; \tag{23}$$

for tetrode and pentode amplifiers,

$$R_{eq} = \frac{I_p}{I_p + I_{c2}}\left(\frac{2.5}{g_m} + \frac{20I_{c2}}{g_m^2}\right). \tag{24}$$

where R_{eq} = the equivalent noise resistance,
g_m = the transconductance,
I_p = the d-c plate current,
I_{c2} = the d-c screen current.

In Table 4·2 are shown values of equivalent noise resistances and transconductances of common receiving-type tubes. These values are calculated from the nominal values of the transconductance and plate and screen currents given in the *RCA Tube Handbook* or from the average of the minimum and maximum values in the JAN-1A specifications. Also average values of input and output capacitance are shown. The input capacitance does not include the effect of the grid-plate capacitance (Miller effect), which in the case of triodes is a large effect.

In Table 4·3 are shown data on noise figures of various amplifiers using various tube types. The measured noise figures are nearly in accord with the calculated equivalent resistances. However, Table 4·3 indicates that the triode-connected 6AK5 is superior to the 6J4 in contradiction to the data in Table 4·2.

If the noise figures are calculated from values of equivalent resistance and the values of circuit parameters, it is found that the experimental values of noise figures of amplifiers using 6AK5's agree with the calculated values whereas those of amplifiers using other tube types are somewhat worse. The reason for these discrepancies has not been explained. One possible explanation is the effect of grid current, which was not considered in these calculations. The 6AK5 has a gold-plated grid, causing it to have unusually low grid emission.

It can be seen from comparing triode and pentode tubes of the same g_m that the partition noise is a relatively large effect, which gives the triode tubes an appreciable advantage, a fact that is confirmed by measurements of noise figures as shown in Table 4·3.

Of course, there are wide variations in the equivalent noise resistances

[1] Equations (23) and (24) are taken from the paper by W. A. Harris, *RCA Rev.*, V, 505–524, April 1941.

of individual tubes of the same type. If a plot is made of cathode current vs. transconductance for a large number of tubes under fixed operating voltages, the points belonging to a majority of tubes lie close to a

TABLE 4-2.—EQUIVALENT NOISE RESISTANCES OF RECEIVING TUBES

Tube	g_m, μmhos	R_{eq}, ohms	C_{in}, $\mu\mu$f	C_{out}, $\mu\mu$f
		Triode amplifiers		
1LE3	855	2,920	....	....
3A5*	1,800	1,390	0.9	1.00
6AC7	11,250	220	11.0	....
6AK5	6,670	385	4.0	....
6C4	2,200	1,140	1.8	1.30
6F4	5,800	430	2.0	0.60
6J4	12,000	210	2.8	0.24
6J5	2,600	960	3.4	3.60
6J6*	5,300	470	2.2	0.40
6SC7*	1,325	1,890	2.2	3.00
6SL7*	1,600	1,560	3.2	3.60
6SN7*	2,600	960	2.9	1.00
7F8*	5,650	440	2.8	1.40
9002	2,200	1,140	1.2	1.10
		Sharp cutoff pentodes		
1L4	1,025	4,300	3.6	7.5
1LN5	800	3,450	3.4	8.0
6AC7	9,000	720	11.0	5.0
6AG5	5,000	1,640	6.5	1.8
6AJ5	2,750	2,650	4.1	2.0
6AK5	5,000	1,880	4.0	2.4
6AS6	3,500	4,170	4.0	3.0
6SH7	4,900	2,850	8.5	7.0
6SJ7	1,650	5,840	6.0	7.0
9001	1,400	6,600	3.6	3.0
		Remote cutoff pentodes		
1T4	750	20,000	3.5	7.3
6AB7	5,000	2,440	8.0	5.0
6SG7	4,700	4,000	8.5	7.0
6SK7	2,000	10,500	6.0	7.0
9003	1,800	13,000	3.4	3.0

* One unit of a dual-triode tube.

definite curve, indicating that high cathode current and high transconductance usually occur together. The equivalent noise resistance is found both theoretically and experimentally to decrease as the transconductance and cathode current increase along this curve. This

property of dependence of transconductance upon cathode current is utilized to reduce the variations in gain (and, as it turns out, in equivalent noise resistance) by a scheme known as "d-c degeneration"—the use of cathode bias resistors or large dropping resistors in the screen grid supplies. Thus the operating transconductance tends to be reduced by the less favorable operating voltages resulting for tubes with large current, whereas the reverse is true for tubes that draw smaller currents. A small

TABLE 4·3.—NOISE FIGURES OF AMPLIFIERS*

Circuit**	Tube type		Frequency, Mc/sec	Bandwidth, Mc/sec		Noise figure, db			C_{gp}, $\mu\mu$f
	1st	2nd		Input	Overall	Min	Ave	Max	
I	6AC7	6AC7	30	10.0	1.5	3.4	3.9	4.4	0.01
	6AK5	6AK5	30	12.0	6.0	2.6	3.3	4.5	0.05
	6AK5	6AK5	60	7.5	3.0	...	4.1	...	....
	6AK5	6AK5	60	....	16.0	...	6.5	...	....
II	6AK5	6AK5	30	15.0	1.5	1.3	1.5	2.1	1.2
	6AK5	6J6	60	25.0	12.0	...	3.5	...	....
	6AK5	6J4 / 6J6 / 6AK5	30	12.0	8.0	...	2.2	...	1.2
	6J6	6J4	30	6.0	2.0	...	2.0	...	3.6
	6J6	6J4	30	12.0	8.0	...	2.2	...	3.6
	6AS6	6J4	30	12.0	8.0	...	2.2	...	1.2
	½7F8	½7F8	30	6.0	2.0	...	3.0	...	....
	6J4	6J4	30	6.0	2.0	...	2.0	...	3.2
	6J4	6J4	180	....	....	...	5.5	...	....

* Minimum and maximum noise figures are given in case of amplifiers in production. In the determination of maximum noise figure, the worst 5 per cent have been neglected. C_{gp} is measured in the circuit and refers to the first tube.

** Circuit I consists of a grounded-cathode pentode followed by a grounded-cathode pentode. Circuit II consists of a grounded-cathode triode followed by a grounded-grid triode. See Fig. 4·12.

number of the tubes are found to draw appreciably more current than the value given by the curve, and an even smaller number are found that draw less current. The former are found to be particularly noisy. In fact, a tube with a medium transconductance and an abnormally large cathode current is generally worse, especially when d-c degeneration is employed, than a tube with a minimum transconductance and a normal current. One possible explanation of such behavior is the presence of gas. Another possible cause is a defect in the control-grid structure so that it does not cover the entire useful length of the element structure. The tubes with abnormally low currents, on the other hand, usually have very low equiva-

lent noise resistance. A possible explanation of their existence in the case of multigrid tubes is the accidental alignment of the wires of the various grids, which tends to reduce the screen current and hence the partition noise without reducing the transconductance.

The optimum operating voltages are found, of course, by experiment. In general, a tube should be operated to give as high a transconductance as possible, which is done at large plate and screen currents. As the input stage is not required to handle very strong signals, the first tube may be operated at moderately low plate and screen voltages, and a high transconductance can be obtained by the use of a low bias. However, if the bias is too small, grid current will flow and it will cause the noise figure to deteriorate, especially if the total conductance across the input of the first tube is very small.

4·7. Coupling Circuits.—Some of the features of the design of the input coupling circuit between the mixer and the first amplifier stage will now be examined. The i-f admittance of a typical microwave crystal mixer may be represented as that of a parallel circuit containing a capacitance that usually ranges from 5 to 30 $\mu\mu$f, depending on the mechanical design of the mixer assembly, and a resistance that may vary from 150 to 1000 ohms in extreme cases. The conductive component is a function of the amount of LO power applied to the mixer, of the r-f tuning of the system, and also varies from crystal to crystal. For usual operating conditions, 300 ohms is a good average value to assume, with variations due to interchange of crystals producing a range between 200 and 500 ohms.

For a conventional grounded-cathode amplifier the input admittance is determined largely by the capacitance from grid to ground, which for a tube such as the 6AC7 is of the order of 12 to 18 $\mu\mu$f depending on details of the wiring and for a miniature tube such as the 6AK5 may range from 5 to 10 $\mu\mu$f. The conductive component G_1 includes the effects of any leakage resistance over sockets and wiring, the effects of dissipation in the tuned circuits connected to the grid, and any transit-time loading that may occur. In this chapter transit-time effects are neglected. For a 6AC7 at 30 Mc/sec or a 6AK5 at 60 Mc/sec, this neglect is almost, but not quite, justified, a fact that has the effect of making the actual value of G_1 higher than would otherwise be expected.

The value of G_a' to produce minimum noise figure is given by

$$G_a' = \sqrt{\frac{G_1}{R_{eq}}}. \tag{13a}$$

Table 4·2 shows that the values of R_{eq} for different tubes range from about 300 ohms up to perhaps 3000 ohms for tubes useful in wide-band circuits. The value of $1/G_1$ may be expected to lie in the range of some tens of thousands of ohms, so that the optimum value of $1/G_a'$ will be of the order

of thousands of ohms. Thus the coupling circuit is called upon to produce an impedance transformation of the order of 10 to 1.

Two types of coupling network are particularly appropriate for use here. These are (1) the autotransformer or mutual-inductance coupled circuit with coefficient of coupling near unity and (2) the conventional

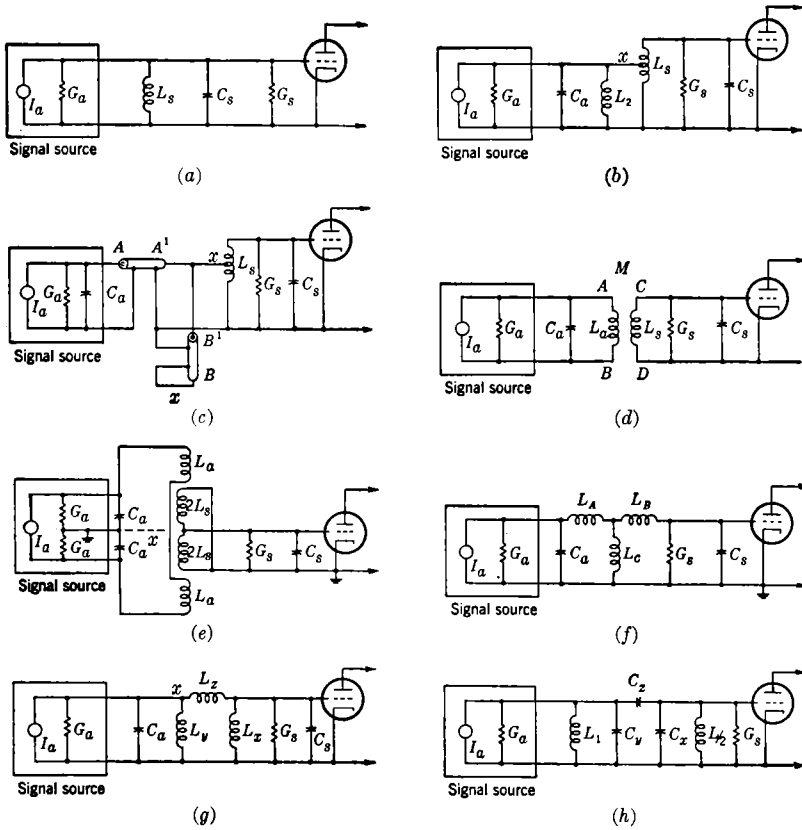


FIG. 4·7.—Coupling networks. (*a*) Single-tuned; (*b*) autotransformer; (*c*) autotransformer with transmission-line input; (*d*) mutual-inductance coupled; (*e*) mutual-inductance with push-pull input; (*f*) self-inductance coupled T; (*g*) self-inductance coupled π; (*h*) capacitance-coupled π.

double-tuned circuit or mutual-inductance coupled circuit with coupling near transitional. The T or π equivalents of the latter may also be used. One important difference between these two general types is found in the effect on G'_a of a change in G_a. In the circuits of Type 1 the changes are in the same direction. Thus if LO power is removed from the mixer, causing G_a to decrease, G'_a will also fall to a fairly low value. If the i-f

amplifier design is not good in regard to stability, this may cause oscillation. On the other hand, a coupling circuit of Type 2 will cause G_a' to increase under these conditions [see Eq. (37*a*)], which will usually result in no instability. The behavior of such a double-tuned circuit is in many ways analogous to that of a quarter-wave section of transmission line.

In Fig. 4·7 are shown various possible coupling circuits. The first three belong to Type 1, as given above, and the rest belong to Type 2. These will be discussed now in somewhat more detail.

4·8. Single-tuned or Autotransformer Circuits.—The simplest possible arrangement is that shown in Fig. 4·7*a*. Here an inductance L_s is used to form a resonant circuit with the capacitance associated with the mixer and the tube in parallel. It possesses no impedance-transforming properties, so that essentially $G_a' = G_a$. In view of the preceding discussion, it is seen that this is almost certainly not the appropriate value of G_a' for good noise figure, so that this circuit is ordinarily of little use. Since the mixer conductance G_a loads the circuit down heavily, the bandwidth is quite ample for most cases. If applications arise where the deterioration of noise figure is not prohibitive, this circuit has the advantage of extreme simplicity of construction and adjustment.

Impedance-transforming properties may be given to the circuit by tapping down on the coil as shown in Fig. 4·7*b*, thereby making it into an autotransformer. The value of L_s is such that it resonates with C_s, the input capacitance of the tube, at the center frequency, and L_2 resonates with C_a, the capacitance of the mixer. A quantity known as the current step-down ratio will be denoted by n, and L_a will represent the inductance measured between the tap and ground with L_s short-circuited. If a current source I_a is connected across L_a, then n is the ratio of I_a to the current that would flow through a short circuit connected across L_s. The value of n is given by

$$n = \frac{\sqrt{L_s}}{k\sqrt{L_a}} \tag{25}$$

where k is the coefficient of coupling.

If L_s consists of a solenoid very long compared with its diameter, n is equal to the ratio of the total number of turns to the number between the tap and ground. For a fixed value of n, L_a is smaller and has a larger susceptance as k becomes larger. If the coupling coefficient is near to unity n is approximately the voltage step-up ratio. The voltage at the input of the tube is the voltage that would be produced by a current of strength I_a' in a parallel single-tuned circuit consisting of a conductance G, an inductance L, and a capacity C where

$$I'_a = \frac{I_a}{n}, \tag{26}$$

$$G = \frac{G_a}{n^2} + G_s, \tag{27}$$

$$C = \frac{C_a}{n^2} + C_s, \tag{28}$$

and

$$\frac{1}{L} = \frac{1}{n^2 L_a} + \frac{1}{L_s}. \tag{29}$$

Also if

$$L_2 C_a = L_s C_s = \frac{1}{\omega_0^2},$$

then

$$LC = \frac{1}{\omega_0^2},$$

where ω_0 is the angular frequency at resonance. The transformed signal-source conductance is

$$G'_a = \frac{G_a}{n^2}. \tag{30}$$

At the center frequency the input voltage to the tube is

$$E_s = \frac{nI_a}{G_a + n^2 G_s}. \tag{31}$$

This has a maximum when

$$n^2 = \frac{G_a}{G_s}. \tag{32}$$

The bandwidth of the input circuit between the half-power points is

$$\mathfrak{B} = \frac{G_a + n^2 G_s}{2\pi(C_a + n^2 C_s)}. \tag{33}$$

It also can be shown that the figure of merit or gain-bandwidth product of such a network relative to a simple single-tuned circuit with a total capacity equal to $C_a + C_s$ is $n(C_a + C_s)/(C_a + n^2 C_s)$. When $n^2 = C_a/C_s$, this has a maximum value of $(C_a + C_s)/2\sqrt{C_a C_s}$, the ratio of the arithmetic mean to geometric mean, a quantity greater than unity.

The value of n that gives the minimum required bandwidth can be calculated from Eq. (33), and also the bandwidth for the value of n that gives optimum noise figure can be calculated. By comparison of these values, the correct value of n can be ascertained, and the tap on the coil can be made accordingly. This position of the tap should be checked by noise figure and bandwidth measurements.

A section of transmission line can also be used in place of L_2 as in Fig. 4·7*c*.[1] An inductive (negative) susceptance of any desired value can be produced by the use of the proper length of line (between zero and one-quarter wavelength long with a short circuit at the far end or between a quarter and a half wave long with an open circuit at the far end). Usually, lines with a short circuit at the far end are more convenient, as they afford a d-c return for the mixer-crystal current that can be isolated from the d-c return of the tube by a blocking condenser.

The alignment of the autotransformer input network presents no particular problem once the correct position of the tap has been ascertained. The value of L_2, or the length of the equivalent section of transmission line, is adjusted by impedance measurements, but the adjustment is not very critical. Then a signal generator is connected, and L_s is adjusted to give maximum response at the desired center frequency. Alternately, if L_2 and C_a are disconnected and the signal generator presents a pure conductance to the network, L_s is adjusted for maximum response, and then C_a and L_2 are connected and L_2 is adjusted for maximum response.

4·9. Double-tuned Input Circuits.—The circuit shown in Fig. 4·7*d* is typical of the form assumed by a double-tuned input circuit. The primary is loaded quite heavily by the mixer conductance G_a; but as shown previously, any unnecessary loading on the secondary should be avoided. Therefore the Q's of the primary and secondary will be quite a bit different. Furthermore, the ratio of bandwidth to center frequency is not small, with the result that a detailed mathematical treatment is both difficult and unprofitable.[2] For this reason a detailed development of the properties of the circuit will not be given at this point, but instead necessary formulas will be quoted without proof. Many of the parameters of the circuit are indicated on the diagram Fig. 4·7*d*. G_a is the conductance of the mixer plus the conductance due to losses in the primary coil and capacitance, which is usually negligible in comparison with the mixer conductance. The capacitance of the mixer assembly plus the distributed capacity of the primary circuit is denoted by C_a. If a *short* length of coaxial cable is used to connect the mixer to the main amplifier, it may be treated as a lumped capacitance and included in C_a. L_a and L_s are the inductances of the primary and secondary coils, respectively, measured with the other coil open-circuited. The mutual inductance between windings is M; G_s lumps all the conductances on the secondary

[1] This type of network has been used by the Bell Telephone Laboratories.

[2] In Chap. 6 there is a treatment of double-tuned circuits as interstage coupling circuits that summarizes the factors necessary for the design of such low-Q tightly coupled circuits. There is also a somewhat more detailed treatment of this subject in Vol. 18 of this series.

side, including losses in the coil, leakage resistance, and input conductance of the tube; and C_s is made up of the input capacitance of the tube plus stray wiring capacitance. The coefficient of coupling k is given by the usual expression

$$k = \frac{M}{\sqrt{L_a L_s}}. \tag{34}$$

Certain of the formulas are expressed more easily in terms of quantities L_{ac} and L_{sc} which are the inductances of the primary and secondary, respectively, measured with the opposite coil short-circuited. The following relations hold:

$$L_{ac} = L_a(1 - k^2),$$

and

$$L_{sc} = L_s(1 - k^2). \tag{35}$$

The action of the primary circuit upon the secondary may be described approximately in terms of an admittance Y'_a connected across the secondary side with the primary circuit removed, and conversely the action of the secondary upon the primary may be described by a similar admittance Y'_s. If Y_a and Y_s are the total admittances of the primary and secondary circuits, respectively, measured with the opposite circuit shorted, Y'_a and Y'_s are given by

$$\left.\begin{aligned} Y'_a &= \frac{k^2}{Y_s L_{ac} L_{sc} \omega^2}, \\ Y'_s &= \frac{k^2}{Y_a L_{ac} L_{sc} \omega^2}. \end{aligned}\right\} \tag{36}$$

If all of the Y's are written as $G + jB$, the following expressions hold:

$$G'_a = \frac{k^2 G_a}{L_{ac} L_{sc} \omega^2 (B_a^2 + G_a^2)}, \tag{37}$$

$$B'_a = -\frac{k^2 B_a}{L_{ac} L_{sc} \omega^2 (B_a^2 + G_a^2)}. \tag{38}$$

If $|B_a| \ll |G_a|$,

$$G'_a = \frac{k^2}{L_{ac} L_{sc} \omega^2} \frac{1}{G_a}, \tag{37a}$$

$$B'_a = -\left(\frac{k^2}{L_{ac} L_{sc} \omega^2 G_a^2}\right) B_a. \tag{38a}$$

This is the mathematical expression of the property of the double-tuned circuit mentioned previously.

The voltage developed across the secondary E_s is

$$E_s = \frac{\dfrac{k I_a}{j\omega \sqrt{L_{ac} L_{sc}}}}{Y_a Y_s + \dfrac{k^2}{\omega^2 L_{ac} L_{sc}}}. \tag{39}$$

This may be thought of as the result of a current I'_a flowing through an admittance $(Y_s + Y'_a)$ where

$$I'_a = \frac{I_a k}{j\omega Y_a \sqrt{L_{ac} L_{sc}}}. \tag{40}$$

The relation that determines the pass band of the circuit is the absolute magnitude of the transfer impedance E_s/I_a as a function of frequency. For the cases of practical importance this is too complicated to evaluate and is best determined experimentally.

Since the theoretical analysis of the circuits actually useful for coupling the mixer to the first amplifier stage is so complicated, it is a practical necessity to adjust the circuits empirically, measuring both noise figure and pass band, making small adjustments in the values of the various circuit constants until both noise figure and pass band are satisfactory.[1] A swept-frequency signal generator is almost indispensable for observing the pass band. The whole process of adjustment is easier if the initial values chosen for the circuit constants are approximately correct. For this reason a group of formulas will now be given that apply strictly only to the case where the bandwidth is small compared with the center frequency. They do, however, furnish a useful starting point for adjustment in the wide-band case.

The Q's of the primary and secondary are defined by

$$Q_a = \frac{\omega_0 C_a}{G_a}$$

and

$$Q_s = \frac{\omega_0 C_s}{G_s},$$

where ω_0 is the angular frequency at midband. For the value of k given by

$$k_{\text{crit}} = \frac{1}{\sqrt{Q_a Q_s}} \tag{41}$$

the coupling is known as *critical*. At this point

$$G'_a = G_s \tag{42}$$

corresponding to a condition of impedance match. If the coupling is increased above this value to

$$k_{\text{trans}} = \sqrt{\frac{1}{2}\left(\frac{1}{Q_a^2} + \frac{1}{Q_s^2}\right)}, \tag{43}$$

the coupling is known as *transitional*. For $k < k_{\text{trans}}$, the curve of transfer

[1] Some of the experimental techniques for measurement of noise figure and pass band are treated in Chap. 12.

impedance against frequency shows only a single maximum; for $k > k_{trans}$, there are two maxima. If $Q_a = Q_s$, $k_{crit} = k_{trans}$; but if $Q_a \neq Q_s$, $k_{trans} > k_{crit}$. For the case of a grounded-cathode amplifier, optimum noise figure is obtained with $G_a' \gg G_s$ and $Q_s \gg Q_a$. The degree of coupling is usually near to transitional and may be somewhat over or under this value.

In most cases Q_s is large enough so that k_{trans} may be written as

$$k_{trans} = \frac{1}{Q_a \sqrt{2}}. \tag{44}$$

The bandwidth at transitional coupling is approximately

$$2\pi\mathcal{B} = k_{trans}\omega_0 = \frac{G_a}{C_a \sqrt{2}}. \tag{45}$$

At the midfrequency the transfer impedance is approximately

$$Z_{12} = \frac{1}{G_a}\sqrt{\frac{2C_a}{C_s}} \tag{46}$$

and

$$G_a' = \frac{G_a C_s}{2C_a}. \tag{47}$$

It should be reiterated that these expressions are derived on the assumption that Q_a is moderately large compared with unity and so represent only a good starting point for experimental adjustment in the mixer–to–i-f–stage coupling network. Here Q_a is likely to be in the range 1 to 5.

In the first trial design of such a network the values of the mixer conductance G_a and of the primary and secondary capacitances will ordinarily be known. From these, the primary Q may be calculated, which, in turn, gives the coefficient of coupling for the transitional case. In Chap. 6 curves are given from which may be obtained the proper frequencies to which to tune the primary and secondary in the absence of the other circuit. With these values, the primary and secondary inductances may be computed, so completing the trial design of the transformer. The bandwidth and transformed signal-source conductance may then be computed.

The effect of the variation in G_a resulting from change of crystals and variation of other operating conditions may be estimated from the relations

$$\mathcal{B} \propto G_a, \tag{48}$$

$$G_a' \propto \frac{1}{G_a} \tag{49}$$

These are illustrated in Fig. 4·8 which shows experimentally obtained pass bands for a circuit with fixed coupling and G_a varied. The primary and

secondary capacitances are 25 and 7 $\mu\mu$f, respectively; the midfrequency is 30 Mc/sec, indicated by a marker pip on the pictures; and the secondary Q is about 75. The circuit is designed to be transitionally coupled for

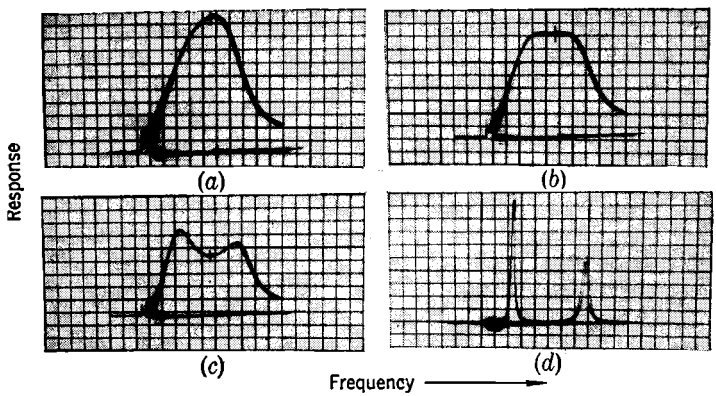


FIG. 4·8.—Experimentally obtained pass bands.

$1/G_a$ equal to 300 ohms, at which point Q_a is 1.7. The values of $1/G_a$ for the four pictures are as follows:

$$a.\ \frac{1}{G_a} = 200 \quad \text{ohms.}$$

$$b.\ \frac{1}{G_a} = 300 \quad \text{ohms.}$$

$$c.\ \frac{1}{G_a} = 500 \quad \text{ohms.}$$

$$d.\ \frac{1}{G_a} = 1800 \quad \text{ohms.}$$

In Fig. 4·9 the primary capacitance has been increased by 2 $\mu\mu$f for the condition $1/G_a = 300$ ohms. The peak of the response curve now occurs at 32.5 Mc/sec, although capacity has been added to the circuit. For $1/G_a$ equal to 200 ohms this amount of detuning would have very little effect on the response curve.

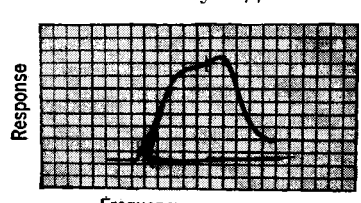


FIG. 4·9.—Pass band with primary detuned.

Mechanical Construction of a Double-tuned Transformer.—In the actual construction of the transformer, the coupling coefficient may be adjusted by impedance measurements by which the mutual inductance is measured. If the ratio of bandwidth to center frequency is large, the coupling coefficient must be large, which makes the design of the trans-

former difficult. In the treatment of the capacity-coupled π-network in Sec. 4·10 it will be seen that capacity coupling is undesirable in wide-band circuits. Thus it is necessary to achieve tight coupling between primary and secondary with a minimum of coupling through distributed capacities. Thus the terminals A and C in Fig. 4·7*d* should be isolated from each other as much as possible. Winding the primary on top of the secondary or interlacing the windings is to be avoided. Ideally, the primary and secondary should be wound as though they were portions of a single-layer solenoid, with the terminals in the following order: A, B, D, C. To get as large a coupling coefficient as possible, it is necessary to use a large-diameter form with close-spaced turns of fine wire. However, the fine wire tends to introduce losses, which in the secondary have an appreciable effect and cause the noise figure to deteriorate. Such a transformer is also difficult to adjust and to duplicate in quantity. The equivalent Π- or T-networks to be described later do not have these disadvantages. Theoretically, however, any coupling coefficient up to unity can be obtained with the transformer, but with the T- or Π-network, the largest that can be obtained is $\sqrt{C_s/C_a}$ or $\sqrt{C_a/C_s}$, whichever is smaller.

When moderate coupling coefficients are desired, the coupling due to stray capacities can be reduced to a negligible amount by the arrangement shown in Fig. 13·6. Here the primary has been split into two coils in series, with only the coil near ground being inductively coupled to the secondary. The effect of capacity between the windings is sufficiently reduced by this construction to permit winding the one section of the primary directly on the grounded end of the secondary. This arrangement has the additional advantage that it is possible to make adjustments in one of the quantities L_a, L_s, and k, which do not disturb the other two quantities so much as when the primary and secondary each consist of a single coil.

The arrangement shown in Fig. 4·7*e* was devised[1] for connecting a push-pull mixer to the input of a single-ended amplifier. The capacitive coupling is reduced to a minimum, and the same admittance to ground is presented to both terminals of the signal source. The secondary consists of two coils wound in opposite directions, adjacent to each other on the same form. The two near ends are connected together and form the ungrounded terminal of the secondary. The two far ends are connected together to form the grounded terminal. The secondary coils each have twice the desired inductance, so that when connected in parallel as indicated, they have the desired inductance. The primary coils are wound adjacent to the ends of the secondary. When the values

[1] By C. P. Gadsden at the Radiation Laboratory.

are defined as indicated in the diagram, all of the formulas for the conventional single-ended input transformer apply. There is no equivalent T- or Π-network which can be used to connect a push-pull signal source to a single-ended amplifier.

Sometimes a push-pull signal source is employed to balance out one source of incoming signal. An example of this is the elimination or reduction of LO noise in crystal mixers by the use of a push-pull or balanced mixer. Because the balance is usually not perfect, it is desirable to reduce the remaining portion of the undesired component further in any way possible. One thing that can be done is to connect a network between ground and the midpoint of the two primary coils that matches the impedance between these two points presented by the primary. This will absorb a large portion of the remaining part of the in-phase component. The value of this impedance Z, the impedance presented by the primary, is given by the following expression which assumes that the unbalance is small so that the interaction of the secondary can be neglected:

$$Z = \frac{1}{2}\frac{G_a - jC_a\omega}{G_a^2 + C_a^2\omega_0^2} + jL_a\omega. \tag{50}$$

4·10. T- and Π-networks.—There are two other networks, known as T- and Π-networks, which may often be used as coupling networks in place of a transformer.

T-network.—The method of selecting the correct parameters for the T-network shown in Fig. 4·7*f* is first to calculate the parameters L_a, L_s, M, and k of the transformer that would be required and then to employ the following equations:

$$L_a = L_A + L_C, \tag{51}$$

$$L_s = L_B + L_C, \tag{52}$$

$$M = L_C, \tag{53}$$

$$k = \frac{L_C}{\sqrt{(L_A + L_C)(L_B + L_C)}}. \tag{54}$$

It can be seen from Eq. (54) that when L_C is small, the coupling is small and, when L_C is large, the coupling is tight.

Π-*network.*—The values for the Π-network shown in Fig. 4·7*g* can also be calculated from the parameters of the equivalent transformer by the use of the following relations:

$$L_X = \frac{L_aL_s(1 - k^2)}{L_a - M}, \tag{55}$$

$$L_Y = \frac{L_aL_s(1 - k^2)}{L_s - M}, \tag{56}$$

$$L_Z = \frac{L_aL_s(1 - k^2)}{M}. \tag{57}$$

The relation between the values of the components of the Π-network and the coupling coefficient is

$$k = \frac{\frac{1}{L_z}}{\sqrt{\left(\frac{1}{L_x}+\frac{1}{L_z}\right)\left(\frac{1}{L_y}+\frac{1}{L_z}\right)}}. \tag{58}$$

It can be seen from Eq. (58) that a small value of L_z gives a large coupling coefficient.

Maximum Coupling Coefficient—The maximum value of the coupling coefficient for either T- or Π-networks that can be achieved without introducing negative inductances is

$$k_{max} = \sqrt{\frac{C_s}{C_a}}. \tag{59}$$

(If C_s is larger than C_a, the expression for k_{max} is the reciprocal of the above.) In this case the T- and Π-networks both degenerate into the same network containing two members, $L_C = L_y$ and $L_B = L_z$. Where coupling coefficients greater than k_{max} are desired, a transformer must be used. In situations where the T- or Π-networks can be used, they are usually to be preferred because of their ease of construction. Furthermore, as they consist of self-inductances only, they can be wound with heavy wire which avoids excessive losses. In cases where separate d-c returns for the signal source and the tube are desired, the Π-network is to be preferred to the T-, for these returns can be effected through L_y and L_z, respectively, by inserting a blocking condenser at X in Fig. 4·7*g*. To make a second d-c return with the T-network requires the addition of a shunt choke as well as a blocking condenser or the use of a bifilar winding for L_C. For coupling coefficients near to k_{max} in the Π-network one of the coils L_y or L_z becomes so large that it is difficult to construct without adding distributed capacity. On the other hand, L_A or L_B becomes small in the T-network, thus making the T-network easier to construct.

Capacity-coupled Π*-network*.—The capacity-coupled Π-network shown in Fig. 4·7*h* is the equivalent of a transformer or inductive T- or Π-network which would correspond to larger equivalent values of C_a and C_s. These larger capacities reduce the figure of merit of the network not only because they add to the total capacity but also because they tend to make the ratio of the capacities become nearer unity, causing the ratio of the geometrical mean of the two capacities to approach the arithmetic mean. Thus as the figure of merit (gain-bandwidth product) is smaller with such a network, it is not possible to achieve so small a value of the transformed signal source conductance with wide bandwidths, and therefore poorer noise figures are obtained with it.

Experimental data are available to confirm these statements. A certain type amplifier which was produced in quantity used a grounded-cathode 6AK5 pentode (7-$\mu\mu$f input capacity) in its first stage. It had a bandwidth of 7 Mc/sec centered at 30 Mc/sec, and the signal source admittance consisted of 300 ohms in parallel with 25 $\mu\mu$f. In the earlier models a capacitive Π-input with C_s equal to 10 $\mu\mu$f was used. Later this was replaced by an inductive Π of the same bandwidth. The average noise figure improved by 1.7 db, which is in agreement with direct observations of the gain-bandwidth product for the coupling circuit, which was found to increase by 22 per cent.

4·11. Alignment of Double-tuned-type Networks.—In the alignment of double-tuned circuits in wide-band amplifiers it is usually necessary to examine the whole pass band with more than one amount of loading. Thus a swept-signal generator is almost a necessity. In Fig. 4·8 are shown the pass bands of a correctly aligned double-tuned transformer with various amounts of loading. As the input conductance decreases, the pass band changes from a sharp-peaked curve to a flat-peaked curve and then to a double-peaked curve. A double-peaked curve can also be produced at loose coupling or with large loading by mistuning either the primary or secondary. Ultimately, with a large mistuning, one of the peaks can appear at such a remote frequency that it is beyond the range of observation. Without subsidiary information the observer has no way of knowing whether a single sharp-peaked curve indicates a correctly aligned network at loose coupling or is one peak of a very badly mistuned network or whether a double-peaked curve belongs to a correctly tuned network with overcoupling or to a slightly mistuned network with loose coupling. This subsidiary information can be supplied by observing the pass band with various amounts of loading. Methods for connecting the signal generator and networks that effectively change the conductance of the generator are described in Chap. 12. If the input conductance of a parallel-tuned network is made very small, a single-peaked curve belonging to a correctly aligned network changes to a double-peaked one, but that of the badly mistuned network does not. Therefore the first step is to observe the pass band under the two extremes of loading. Usually, if the network has been designed according to the calculations that have been described, it is found to be near enough in tune so that both peaks are detectable with small loading, and at large loading a single peak is observed in the vicinity of the desired center frequency. In the event that this is not true, it is necessary, of course, to retune whichever of the two circuits is badly mistuned. This can be ascertained by connecting a small condenser first across the signal source and then across the input of the tube. Usually at one position it does not affect the observable peak, but at the other it causes a shift in frequency. When it is in the position

where there is no shift, it is in parallel with the circuit that is badly mistuned.

Once the network is approximately in tune by this method, by small successive adjustments it can be given the desired properties of bandwidth and center frequency. Usually each adjustment alters all properties and requires changes in the other parameters, and therefore it is wise to make only small changes at one time. When the network is completely in tune, the center frequency should be independent of the conductance of the signal source, and at transitional coupling the maximum should be flat and the curve symmetrical. With large bandwidths, especially near transitional coupling and beyond it, the best measure of center of frequency is the geometrical average of the two frequencies at which the response has fallen by 3 db. At loose coupling or at high loading, the frequency of maximum response can be measured accurately because the peak is so sharp. This should agree with the geometrical average of the half-power frequencies.

In Fig. 4·8*b* is shown the actual response curve of a double-tuned circuit having very unequal Q's, near to transitional coupling. The marker is set at 30 Mc/sec near the maximum of the curve. Figure 4·9 shows the pass band of the same circuit except that the primary has been detuned by changing the capacity from 25 to 27 $\mu\mu$f. The marker is still at 30 Mc/sec, where the response is the same as before. The curve has become unsymmetrical with a maximum response at 32.5 Mc/sec. With a signal-source impedance of 200 ohms, very little effect was produced by an equal detuning. It is evident, in the first place, that the tuning in such a circuit near transitional coupling is very critical.

4·12. Tetrode and Pentode First Stages.—The conventional grounded-cathode amplifier has a distinct advantage over the cathode-follower and grounded-grid circuits in that it has a greater available power gain while having essentially the same noise figure as the other two. Therefore contributions of second-stage noise tend to be smaller with the grounded-cathode amplifier than with the other types. There is, however, one possible drawback to the grounded-cathode amplifiers, namely, the feedback due to the grid-plate capacity, which results in increased input capacity and an extra component of input conductance. At some frequencies this conductance is negative. If the input conductance is small and if the gain, grid-plate capacity, and frequency are high, instability arises. Of course, neutralization may be used; but if there is a strong tendency toward instability, the neutralizing adjustment is very critical. The grid-plate capacities of tetrode and pentode tubes (see Table 4·3)[1]

[1] The grid-plate capacities as given in Table 4·3 include the distributed capacity of the socket and wiring. The value of 0.05 $\mu\mu$f for the 6AK5 miniature pentode is considerably larger than the value quoted by the manufacturers for the tube itself,

are usually sufficiently low to permit their use in conventional grounded-cathode circuits without neutralization. However, with the 6AK5 pentode at 30 Mc/sec and voltage gains of about 10, these effects are sufficient to make an appreciable distortion of the pass band,[1] although not so serious that instability is imminent or neutralization is necessary. If either the gain or the frequency were increased by a factor of 5, this type of feedback would give rise to considerable difficulty.

It is useful to write down the expression for the combined noise figure of the first two stages, by substituting the value of the noise figure of the first stage F_1 and the value of the available power gain of the first stage from Table 4·1 into Eq. (1·3),

$$F_{12} = 1 + \frac{G_1}{G_a'} + \frac{|Y_a' + Y_1|^2}{G_a'} \left\{ R_{eq} + \frac{1}{g_m^2} [G_L + (F_2 - 1)(G_L + g_p)] \right\} \quad (60)$$

In Eq. (60) the load resistor is considered as part of the first stage, and the noise figure of the second stage is F_2 when it is connected to a signal source whose conductance is $G_L + g_p$.

With the approximations stated below, the noise figure at the center frequency of the input circuit becomes

$$F_{12} = 1 + \frac{G_1}{G_a'} + (G_a' + 2G_1)\left[R_{eq} + \frac{1}{\mathcal{G}^2}\left(\frac{1}{G_L} + R_{eq_2}\right)\right] \quad (60a)$$

where R_{eq_2} is the equivalent noise resistance of the second tube and $\mathcal{G}$ is the voltage gain of the first stage. It is assumed that (1) the interstage network is a simple parallel-tuned circuit at the same center frequency as the input circuit and all ohmic losses in parallel with the input of the second tube are considered part of G_L, (2) the dynamic plate conductance of the first stage g_p is negligible compared with the load conductance G_L, (3) sources of noise in the plate of the second stage and beyond are neglected, and (4) $G_1 \ll G_a'$. With tubes of high figure of merit such as the 6AK5, 6AC7, and 6AG5, the voltage gain $\mathcal{G}$ is approximately 5 with a single-stage bandwidth of 10 Mc/sec; therefore the last terms in the brackets are unimportant unless the bandwidth is somewhat greater than 10 Mc/sec.

It is theoretically possible to make the partition noise in the screen-grid circuit nullify that in the plate circuit by the proper use of feedback

0.01 to 0.02 $\mu\mu$f, and it is probable that it could be reduced by the use of supplementary shielding between the grid and plate terminals in the socket. The value of 0.01 $\mu\mu$f for the 6AC7 tube (octal base) is in agreement with values given by the manufacturers, indicating that there is little distributed capacity across the relatively large octal socket.

[1] Y. Beers, journal article in process of publication.

produced by a series reactance in the screen-grid circuit. In individual experimental amplifiers at 30 Mc/sec it has been possible to make improvements in the noise figure of the order of 1 db by selecting the proper value of screen bypass condenser or by connecting a small inductance in the form of an extra inch or two of lead wire between the screen-grid socket terminal and the bypass condenser. Such effects are very hard to control, and it has not been possible to manufacture amplifiers in quantity in which the noise figure was consistently improved by this kind of feedback. On the other hand, some of the variations in noise figure between amplifiers of the same type are probably due to the accidental existence of the correct screen-grid–circuit reactance in some of the amplifiers and not in others.

4·13. Triode Circuits.—Because of their large grid-plate capacities, triodes cannot be used in conventional grounded-cathode circuits unless neutralization is employed. Under the usual conditions, this neutralization is very critical in adjustment, and therefore such circuits are undesirable. If the gain of the first stage is reduced, the neutralization becomes uncritical in adjustment, but then second-stage noise makes an important contribution to the noise figure. Alternatively, a grounded-grid or cathode-follower circuit can be used as a first stage. Under conditions that are normally encountered these have little or no tendency toward instability. However, it was pointed out in Sec. 4·3 that it is not possible to obtain as large available power gains with these as with the grounded-cathode amplifier, so that second-stage noise is likely to make an appreciable contribution to the noise figure. Therefore, when triodes are used in the first stage, care must be given to the design of the second stage, as well as of the first, and to reduce noise in the second stage a triode should be used there as well.

In comparison with pentode circuits, the ideal triode input circuit should have the following features: (1) The improvement in input-stage noise figure which is theoretically possible should be fully realized in practice; (2) the contribution of later stages to the noise figure should be no greater than with pentode input stages of the same bandwidth; (3) the circuit should be no more critical in adjustment than pentode stages of the same bandwidth. In other words, it should be possible to use single-tuned circuits for all interstage coupling networks without appreciable deterioration of noise figure. There are nine possible combinations arising out of the fact that both the first and second stage may be connected as a grounded-cathode, grounded-grid, or cathode-follower stage. If the bandwidth is narrow or if critical adjustments are tolerated, all nine combinations give good noise figures. However, the only one that fully meets the requirements stated above is the grounded-cathode–grounded-grid combination. In it, the large electronic input admit-

tance of the second tube loads down the interstage network so that a large bandwidth is obtained without the need of an ohmic load resistor and permits the maximum possible available power gain for the first stage. At the same time, the output admittance of the first stage is small, so that the noise figure of the second stage is good. The first stage has a voltage gain of around unity, so that there is little tendency to instability.

Grounded-cathode Triode–Grounded-grid Triode.—The basic circuit is shown in Fig. 4·10.[1] The properties of this circuit can be calculated

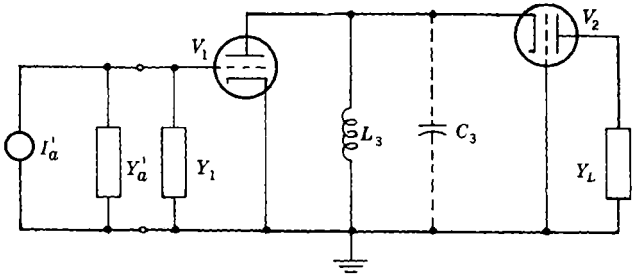


FIG. 4·10.—Basic grounded-cathode–grounded-grid circuit.

from the properties of the individual stages given in Table 4·1. The voltage gain is

$$\mathcal{G} = \frac{g_{m1}g'_{m2}}{(g'_{m2} + g_{p1})Y_L + g_{p1}g_{p2}}, \tag{61}$$

where the subscripts 1 and 2 pertain to the tubes V_1 and V_2, respectively. As Y_L is usually considerably larger than the dynamic plate conductances g_{p1} and g_{p2}, the voltage gain is nearly independent of the properties of V_2. If g_{m2} is increased, for example, the voltage gain of V_2 increases, but the input admittance of V_2 increases, reducing the voltage gain of V_1 by an amount that nearly compensates for the first effect. When the tubes have identical properties, this expression reduces to

$$\mathcal{G} = \frac{(\mu + 1)g_m}{(\mu + 2)Y_L + g_p}. \tag{61a}$$

The output conductance, including the load and assuming that the two tubes have identical properties, is

$$G_2 = G_L + \frac{g_p}{\mu + 2}. \tag{62}$$

It can be seen that the voltage gain is somewhat larger and the output

[1] The properties of this input circuit were discovered and its development was advanced at the Radiation Laboratory by H. Wallman, A. B. Macnee, and C. P. Gadsden.

conductance is smaller than the theoretical values pertaining to one such tube used as a grounded-cathode amplifier with the same load G_L. The available power gain is

$$\mathcal{W} = \frac{(\mu + 1)^2 g_m^2 G_a'}{(\mu + 2)[(\mu + 2)G_L + g_p]|Y_a' + Y_1|^2}. \tag{63}$$

As the amplification factor μ is large compared with unity, and as the dynamic plate conductance g_p is generally small compared with the load conductance G_L, Eq. (63) gives essentially the same value as would be obtained by using one such tube in a grounded-cathode amplifier with the load conductance G_L.

The noise figure, assuming that the tubes have identical properties, is given by

$$F = 1 + \frac{G_1}{G_a'} + \frac{|Y_a' + Y_1|^2 R_{eq}}{G_a'}\left[1 + \frac{(g_p + B_3)^2}{g_m'^2}\right], \tag{64}$$

where B_3 is the net susceptance of C_3 and L_3, which at the center of the band is zero. This expression neglects the Johnson noise of the load G_L. If a similar assumption is made in Table 4·1, it can be seen that Eq. (64), except for the last term which is due to the shot noise in V_2 which in practice is negligible, is identical with that for a simple grounded-cathode amplifier. Inasmuch as the available power gain is essentially the same, the effect of Johnson noise in G_L, as well as noise produced later in the amplifier, is essentially the same in the two cases. Although neutralization of the first tube is unnecessary for stability, it is usually desirable, as it produces a slight improvement in noise figure. However, the adjustment is not critical.

The noise figure does depend upon the careful selection of the tube type and operating conditions of the first tube, but the choice of the second tube and its operating conditions is relatively unimportant. Furthermore, the second tube can be operated with a moderately low plate current and transconductance with a saving in drain on the power supply and an increase in tube life. A tube with a low cathode-plate capacity simplifies adjustment procedures. The 6J4 or one half of a 6J6 (with the grid and plate of the other half connected to ground) is the best choice in this respect. In cases where the cathode-plate capacity is high, neutralization of the second tube may be necessary, but in such cases it is usually uncritical.

A Practical Grounded-cathode-triode–Grounded-grid-triode Amplifier.— The circuit diagram is shown in Fig. 4·11, and photographs of the unit are shown in Fig. 4·12*a*, *b*, and *c*. This unit was designed to replace the first three i-f stages of a radar receiver that employed a 6AC7 grounded-cathode-pentode input stage; consequently, the output and power connections are made through a built-in plug which plugs into the socket of

the third tube of the original receiver. The input cable which is attached to the unit goes to the output of a crystal mixer. Except for drilling a hole in the outer box of the original receiver for the feed-through fitting of this cable, and except for the realignment of the interstage network between V_4 and the fourth tube of the original receiver, no alterations in the original receiver were necessary to accommodate this unit. Data on the noise figure of the original stage can be found on the top line of Table 4·3, and data on the noise figure of this unit can be found in the fifth line of the same table. It can be seen that the use of this unit resulted in an improvement of noise figure of between 2.0 and 2.5 db.

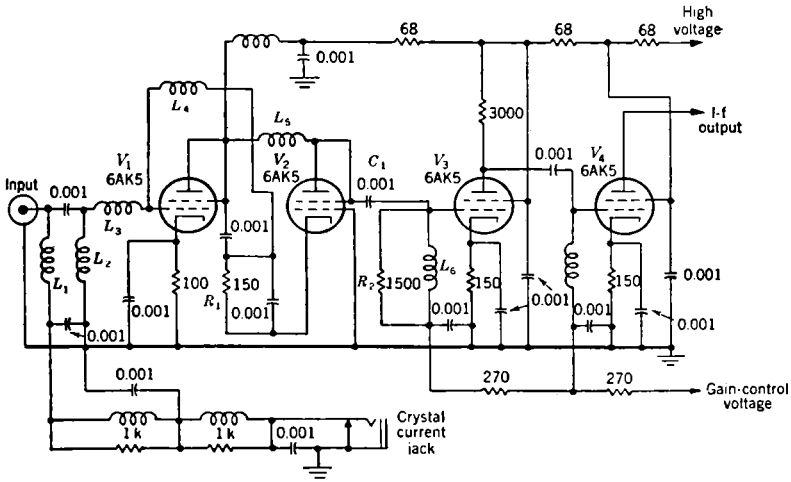
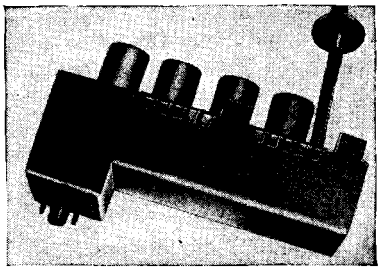
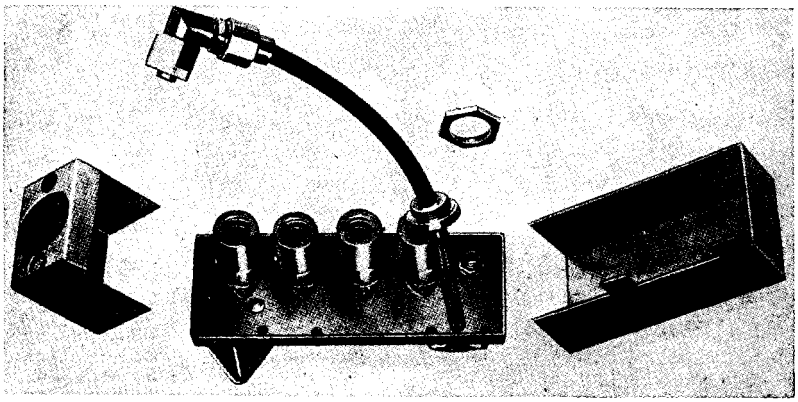


FIG. 4·11.—Circuit of amplifier with grounded-cathode–grounded-grid input.

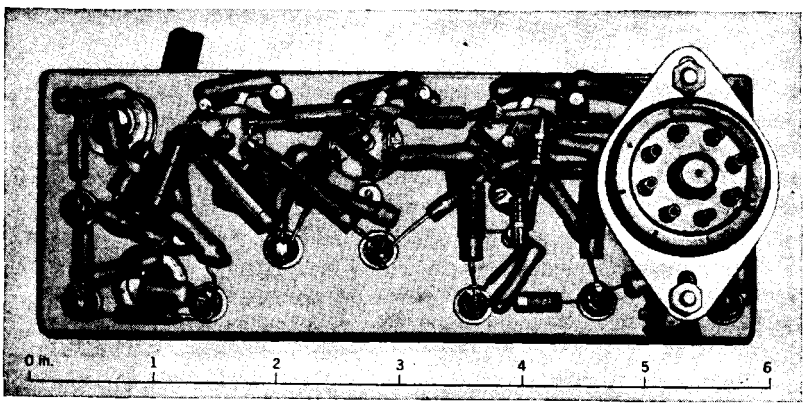
The design of this unit follows the principles described in this chapter. There are a few details that should be noted. The input circuit consisting of L_2 and L_3 is a degenerate T or Π operating at slightly less than transitional coupling. (L_1 is a choke which allows a separate d-c return for the crystal.) The first tube is neutralized by L_4, which is selected to resonate the grid-plate capacity of 1.2$\mu\mu$f. The use of neutralization of the first stage causes an improvement of noise figure of 0.25 db. The cathode-plate capacity of V_2, the grounded-grid tube, is 3.1 $\mu\mu$f, which is quite high, and it was found desirable to neutralize through the use of L_5, which resonates this capacity. As a matter of fact, the use of L_5 actually results in a net saving of one part. If L_5 were not used, the load resistance R_2 would have to be connected on the other side of C_1 in order to serve as a means of completing the d-c circuit to the plate. In that case two more parts, a bypass condenser and a decoupling resistor, would



(*a*)



(*b*)



(*c*)

FIG. 4·12.—I-f input unit. (*a*) side view; (*b*) top view, covers off; (*c*) bottom view, covers off.

have been required at one end of R_2. The d-c current flowing between the cathode of V_2 and ground follows a devious route: through R_1, which supplies grid bias, and then through L_4, L_1, and L_2 to ground. The 6AK5 was selected primarily because of its excellent performance with regard to noise figure as the first tube V_1 and secondarily for its high figure of merit as a conventional pentode amplifier as tubes V_3 and V_4. Its use as V_2 in the grounded-grid stage was purely a matter of convenience to minimize the number of tube types required. Had it not been for this consideration, the 6J4 or 6J6 would have been preferred because of their low cathode-plate capacity.

The adjustment of the neutralizing coils L_4 and L_5 may be done by connecting a signal generator to the input, placing a tube with one heater pin broken off in sockets V_1 and V_2 in turn and adjusting L_4 and L_5 so that no output is obtained at the center frequency.

The output admittance of a grounded-grid tube depends upon the admittance in parallel with the input. Therefore, if a signal generator with a large output conductance is connected across the input of V_2 for alignment of the inductance L_6, the tube will have a larger output conductance than with its normal input load, which is the output conductance of V_1. This effect is appreciable only when the dynamic plate conductance g_p of V_2 is comparable to that of the load R_2. In this case g_p is about one-third the conductance of R_2. Therefore, the bandwidth of the interstage circuit between V_2 and V_3 is approximately 1.33 times as large with the signal generator connected across the input of V_2 as it is in normal operation.

CHAPTER 5

VHF AMPLIFIERS, MIXERS, AND OSCILLATORS

By A. B. Macnee

This chapter will be concerned with the problems of designing the input stages or r-f heads of superheterodyne receivers. The r-f amplifiers that will necessarily be considered may equally well be used in TRF receivers. A block diagram of such an r-f head is shown in Fig. 5·1. The number of amplifiers required will vary according to the gain per stage,

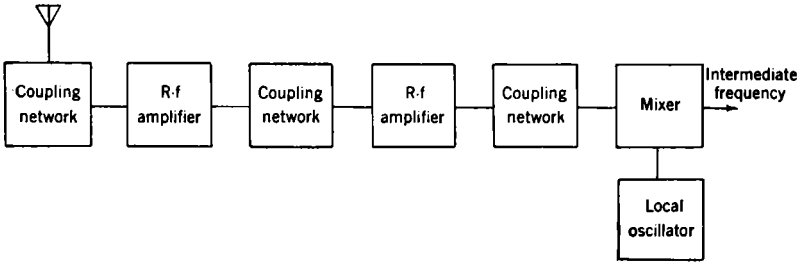


Fig. 5·1.—Block diagram of r-f head.

the bandwidth, and the noise figure required. Because of the ease in obtaining gain and selectivity in the i-f amplifier, the number of r-f amplifiers to be used is determined primarily by the maximum noise-to-signal ratio and image response allowable. The design of an r-f head can be broken down into the design of coupling circuits, amplifiers, a mixer, and a local oscillator.

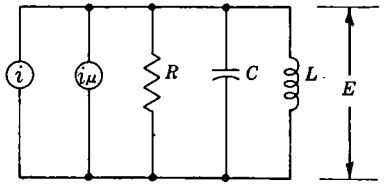


Fig. 5·2.—Parallel-tuned *RLC*-network.

COUPLING NETWORKS

5·1. Two-terminal Coupling Networks.—Whereas the basic types of coupling circuits used at these frequencies are the same as those used at lower frequencies, they assume somewhat different forms when each wire must also be considered as a transmission line. The simplest and most commonly used type of coupling network is the parallel-tuned *RLC*-network shown in Fig. 5·2. Its properties are discussed in Chap. 6. As frequencies approach 350 Mc/sec, it becomes difficult to

use solenoidal coils for the circuit of Fig. 5·2. For higher frequencies, sections of open- or short-circuited transmission lines may be used for the circuit elements. To reduce radiation and stray fields, a coaxial transmission line is usually employed.

A transmission line of length l and short-circuited at one end is equivalent to a parallel-lumped circuit, as indicated in Fig. 5·3, for frequencies in the neighborhood of the resonant frequency

$$f_0 = \frac{1}{2\pi\sqrt{L'C'}}.$$

In Fig. 5·3

$$\left.\begin{aligned} G' &= \frac{G_0}{2}\,l, \\ C' &= \frac{C_0}{2}\,l, \\ L' &= \frac{8}{\pi^2}\,L_0 l, \\ R' &= \frac{8}{\pi^2}\,R_0 l, \end{aligned}\right\} \quad (1)$$

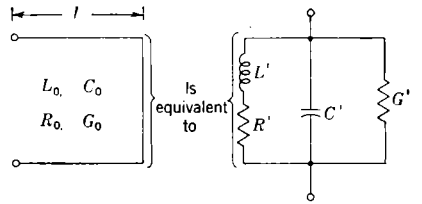


FIG. 5·3.—Transmission line used as tuned circuit.

where C_0 = the line capacity per unit length,
L_0 = the line inductance per unit length,
R_0 = the line series resistance per unit length,
G_0 = the line leakage conductance per unit length.

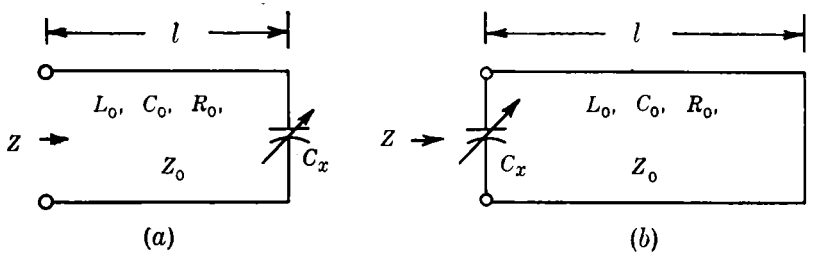


FIG. 5·4.—Transmission-line circuits with capacitance tuning.

$$l = \frac{1}{\omega_0\sqrt{L_0C_0}}\left[\frac{\pi}{2} + \cot^{-1}(Z_0\omega_0C_x)\right]$$

$$Q_0 = \frac{\omega_0L_0}{R_0}$$

$$Z_{max} = \frac{4}{\pi}\,Q_0Z_0m$$

$$m = \frac{\pi}{2\theta - \sin 2\theta}$$

$$\theta = \left[\frac{\pi}{2} + \cot^{-1}(Z_0\omega_0C_x)\right]$$

$$l = \frac{1}{\omega_0\sqrt{L_0C_0}}\cot^{-1}(Z_0\omega_0C_x)$$

$$Q_0 = \frac{\omega_0L_0}{R_0}$$

$$Z_{max} = \frac{4}{\pi}\,Q_0Z_0n$$

$$n = \frac{\pi\sin^2\theta}{2\theta + \sin 2\theta}$$

$$\theta = \cot^{-1}(Z_0\omega_0C_x)$$

From Eq. (1), it is apparent that the transmission-line circuit can be tuned by varying the line length. This is usually accomplished by short-circuiting the line with a metal plunger, the position of which may be varied. An example of this is shown in Fig. 5·27. A transmission line may also be tuned by shunt capacitance. Two possible ways of accomplishing this, with suitable design formulas, are shown in Fig. 5·4*a* and *b*. It is assumed that the leakage conductance in both the shunt condenser and the transmission line is negligible.

5·2. Four-terminal Coupling Networks.—The simple parallel circuit of Fig. 5·2 or its transmission-line analogues offer no impedance-transforming properties and therefore cannot improve the match between two amplifiers or between an antenna and the first amplifier. These circuits may be made into a transformer as well by tapping down on the coil with either the input or output terminal. The general manner in which this may be done with circuits using coils or transmission lines is indicated in Fig. 5·5.

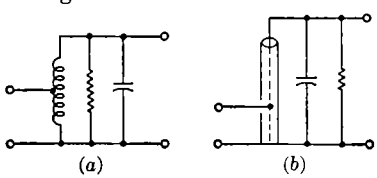


FIG. 5·5.—Impedance-transforming circuits using (*a*) coils, (*b*) transmsision lines.

For circuits in which the coil has a large number of turns and a rather uniform construction, the impedance stepup of this circuit is the square of the turns ratio as in a common autotransformer (see Chap. 4). For most vhf applications, however, the coils have shrunk to two or three turns or perhaps to a piece of transmission line, and any attempt to calculate the proper tapping point for a given impedance transformation becomes involved. The cut-and-try experimental technique is then a more satisfactory approach to the problem.

Two other four-terminal coupling networks that behave very much as a single-tuned circuit for small amounts of coupling are shown in Fig. 5·6*a* and *b*.

Although these circuits do have impedance-transforming properties, in both cases the figure of merit is a function of the maximum transfer impedance and approaches zero as the transfer impedance approaches infinity. This is to be compared with the parallel *RLC*-circuit where the figure of merit remains fixed as the transfer impedance is made to approach infinity. The parallel *RLC*-network with a tapped coil also has the good figure of merit of the simpler network as long as the tapped coil approximates an ideal transformer. Coaxial lines may be used for inductances in both coupling circuits but with more difficulty in the circuit of Fig. 5·6*a* because neither side of the coil is grounded. A still more complex coupling circuit, which is feasible for the lower range of fre-

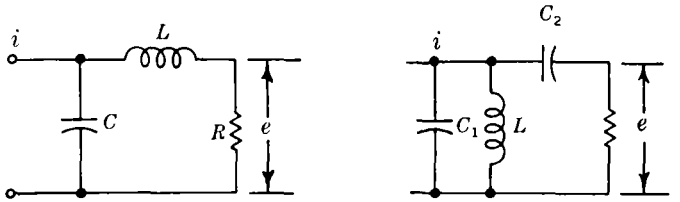


FIG. 5·6.—Four-terminal coupling circuits.

$$\omega_0 = \sqrt{\frac{1}{LC} - \frac{R^2}{2L^2}} \qquad \omega_0 \approx \frac{1}{\sqrt{L(C_1 + C_2)}}$$

$$(Z_{12})_{\omega=\omega_0} = \sqrt{\frac{L}{C}} \frac{1}{\sqrt{1 - \frac{R^2C}{4L}}} \qquad (Z_{12})_{\omega\approx\omega_0} \approx \frac{1}{\omega_0 C_2} = \frac{\sqrt{L(C_1 + C_2)}}{C}$$

$$\approx \sqrt{\frac{L}{C}} \approx \omega_0 L$$

$$\mathfrak{B} \approx \frac{R}{2\pi L} \qquad \mathfrak{B} = \frac{R}{2\pi L}\left(\frac{C_2}{C_1 + C_2}\right)^2$$

$$\left(FM = \frac{1}{2\pi C} R \sqrt{\frac{C}{L}}\right) \qquad \left[FM = \frac{1}{2\pi(C_1 + C_2)} R \sqrt{\frac{C_2^2}{L_1(C_1 + C_2)}}\right]$$

These expressions are valid for

$$R \ll \sqrt{\frac{L}{C}}$$

For these expressions ω_0 is the frequency for which Z_{12} is a maximum.

quencies at least, is the mutual-inductance-coupled double-tuned circuit, the various forms and properties of which are treated in Chaps. 4 and 6. Adapting this circuit to use distributed line parameters presents a problem, since either mutual inductance or a floating inductance (neither side grounded) is always needed. A double-tuned transformer designed for use at 200 Mc/sec is shown in Fig. 5·7.

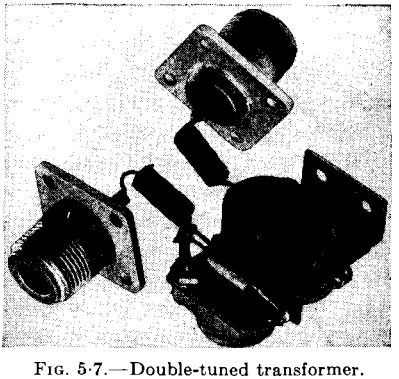


FIG. 5·7.—Double-tuned transformer.

Here the coils have shrunk to single loops of wire positioned by a grooved form, and the circuit is tuned by trimmer capacities. This particular circuit was designed as a band-pass antenna filter to improve the image response of a receiver in the 200-Mc/sec region.

AMPLIFIERS

The amplifiers to be discussed here are primarily r-f amplifiers. Sufficient gain is used to satisfy requirements of noise figure and image response, and the rest of the gain required is obtained at lower frequencies. Amplifiers have been built for i-f applications with a mid-frequency in the 200-Mc/sec region, but these differ somewhat in their basic philosophy and are described in Chap. 6. Here the trend in stability and noise figure will be considered as the frequency is increased from the normal i-f range to vhf.

GROUNDED-CATHODE CIRCUIT

The most common type of amplifier is represented in Fig. 5·8. The grounded-cathode amplifier shown may be considered as either a triode or a pentode in which the bypassed screen and suppressor grids have been omitted.

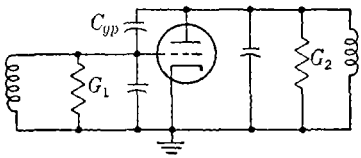


FIG. 5·8.—Essential circuit elements of grounded-cathode amplifier.

5·3. Stability.—The limiting condition for oscillation in this circuit, if the reactances are assumed to have the values most conducive to oscillation, is

$$C_{gp} \geqq \frac{2G_1G_2}{\omega g_m}. \tag{2}$$

For the special case of $G_1 = G_2 = G$, where the voltage gain g_m/G is written as $\mathcal{G}$, the maximum allowable voltage gain per stage is

$$\mathcal{G}^2 = \frac{g_m}{\pi f C_{gp}}. \tag{3}$$

Equations (2) and (3) give the oscillation conditions for a single stage as shown in Fig. 5·8. If a number of such stages are cascaded, the maximum C_{gp} allowable becomes smaller and approaches $2G_1G_2/3\omega g_m$ for four or more identical stages.

In actual amplifiers it is necessary that the point of normal operation be considerably removed from the oscillation points indicated by Eq. (2) or (3). In practice it is not desirable to allow C_{pg} to be larger than about 20 per cent of the value indicated by these equations. With this assumption, the maximum gain vs. frequency is plotted in Fig. 5·9 for a number of common pentode and triode tubes. It must be remembered that these curves indicate the maximum allowable gain for stability reasons and, because of other limitations, do not necessarily represent realizable gains with the tubes in question. For a given tube structure and stage gain, the stability of amplifiers is more or less inversely proportional to the frequency. This means that even in wide-band receivers at vhf, considerable care in shielding is required.

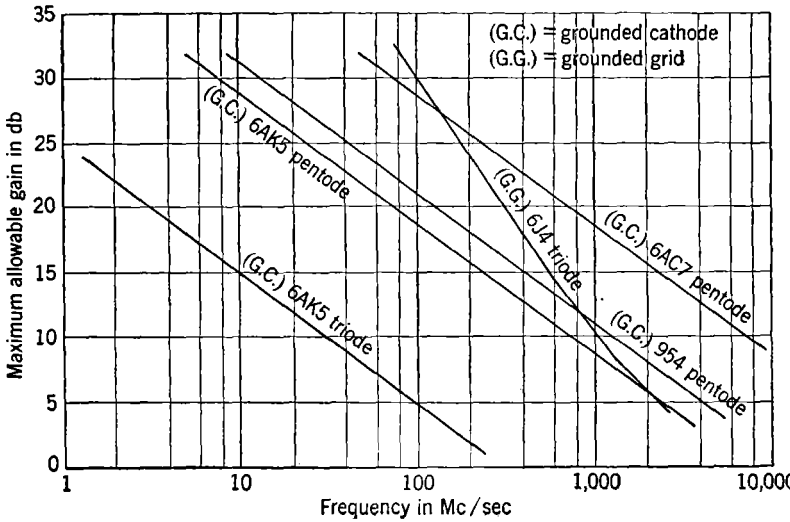


FIG. 5·9.—Maximum gain vs. frequency for common pentodes and triodes.

5·4. Noise Figure.—Another consideration which is usually of greatest importance is the question of amplifier noise figure. This has been treated in Chap. 4 for the low-frequency case where transit-time effects are neglected. At the frequencies considered here, this neglect is no longer permissible; hence the necessary extensions of the theory will be indicated. They will be found to affect principally the input conductance G_1; they make it larger than before and give it an effective temperature higher than room temperature. Calculation of the noise figure of a grounded-cathode amplifier may be done in terms of the equivalent circuit shown in Fig. 5·10.

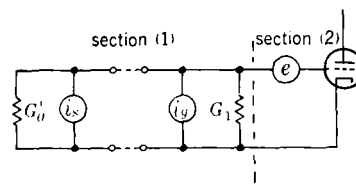


FIG. 5·10.—Equivalent input circuit for grounded-cathode stage.

Here G'_a = input loading due to signal source,

G_1 = input loading due to tube and coupling circuits,

$\overline{i_s^2} = 4kT_rBG'_a$,

$\overline{i_g^2} = 4kT_gBG_1$,

$\overline{e^2} = 4kT_rBR_{eq}$, (see Eq. 4·7)

T_r = absolute room temperature,

T_g = the effective temperature of the grid loading,

B = noise bandwidth,

k = Maxwell Boltzman constant.

It is assumed in this analysis that the signal-to-noise ratio at the center of the band is the only point of interest, and all reactances are considered to be tuned out at this frequency. A more detailed analysis might be made by calculating the noise figure as a function of frequency. The average noise figure of the amplifier could then be found by averaging this single-frequency noise figure over the usable amplifier pass band (usually between 3-db points). It will generally be found that the noise figure so calculated is only slightly larger than the band-center noise figure. Band-center calculations serve therefore to indicate the trends in noise figure and usually give a good approximation to the average noise figure over the band. The details of the calculation of the average noise figure are given in Vol. 18, Sec. 12·7. Considering then Section 1 of Fig. 5·10 the noise figure of this section may be written

$$F_1 = 1 + \frac{T_g}{T_r}\frac{G_1}{G_a'}.$$

The available power gain of this section is

$$\mathcal{W}_1 = \frac{G_a'}{G_1 + G_a'}.$$

The noise figure of Section 2 will be

$$F_2 = 1 + R_{eq}(G_1 + G_a').$$

Therefore, the over-all noise figure will be given by

$$\begin{aligned} F_{12} &= F_1 + \frac{F_2 - 1}{\mathcal{W}_1} \\ &= 1 + \frac{G_1}{G_a'}\frac{T_g}{T_r} + \frac{R_{eq}}{G_a'}(G_a' + G_1)^2. \end{aligned} \tag{4}$$

It is of interest to determine what value of G_a' will make this noise figure a minimum, since this is a quantity which can be adjusted through suitable input transformers. The optimum value of G_a' is found to be

$$G_a' \text{ (opt)} = \sqrt{\frac{T_g}{T_r}\frac{G_1}{R_{eq}} + G_1^2}. \tag{5}$$

5·5. Input Loading.—There are two primary sources of input loading in a grounded-cathode tube. One is due to the cathode-lead inductance and has been treated in Chap. 4. Equation (4·19*a*) gives the loading so produced:

$$G_{L_c} = \omega^2 L_c C_{gc} g_m. \tag{4·19a}$$

Fortunately since the equivalent temperature associated with this loading is considerably less than room temperature, its influence upon the

noise figure and optimum G_a' is not so great as might otherwise be expected. Loading of another type is caused by circuit losses of which the most important part is usually due to series-loss resistance in the circuit inductances. As the frequency of operation is increased, the equivalent shunt conductance will increase according to some power of the frequency. Alternative reasonable assumptions about circuit behavior give a dependence on either the first or the three-halves power of the frequency.

A last and most important type of loading that becomes evident at vhf is that due to electron inertia or transit time. This problem has been dealt with in considerable detail by Llewellyn,[1] and some of his results will be summarized here. The equivalent shunt input resistance for a pentode operating in Class A with its screen bypassed to the cathode is given by the expression

$$R_a = \frac{180 r_p}{\theta_c^2 \mu_0} \left[\frac{(y - h^3)^2}{y^2 A - yB + C} \right] \left[\frac{1 + \mu_0 + M}{\mu_0} \right], \tag{6}$$

where $A = 9 + 44h + 45h^2$,

$B = 51h^2 + 123h^3 + 55h^4 + 3h^5$,

$C = 45h^4 + 51h^5 + 24h^6 + 11h^7 + 3h^8$,

$M = \frac{4}{3}(y - h^3)(1 + h) - h^2(2 - h^2)$,

r_p = plate resistance,

μ_0 = amplification factor,

θ_c = cathode-grid transit angle,

and

$$y = \frac{x_{gs}}{x_{cg}},$$

$$h = \frac{t_{gs}}{t_{cg}},$$

x_{cg} = cathode-grid spacing,

x_{gs} = grid-screen spacing,

t_{cg} = cathode-grid transit time,

t_{gs} = grid-screen transit time.

For a given set of operating conditions, h is determined from the following expressions:

$$h = \frac{\frac{2}{3y}}{1 + \sqrt{\frac{V_p}{V_a}}},$$

$$V_a = \sqrt[3]{\left(\frac{10^6}{2.33} x_{cg}^2 I_c \right)^2} \quad \text{volts},$$

[1] F. B. Llewellyn, *Electron Inertia Effects*, Cambridge, London, 1941.

and θ_c is given by

$$\theta_c = -4.3 \times 10^{-9} f \left(\frac{x_{cg}}{I_c}\right)^{1/3} \quad \text{radians},$$

where f = frequency,

I_c = cathode current, amp per square centimeter.

In calculating Eq. (6) Llewellyn assumes that the tube has parallel-plane electrodes and that a condition of complete space charge exists between the grid and the cathode; that is, the electrons at the cathode have zero initial velocity and acceleration.

This loading has been observed to have an effective temperature of about five times room temperature.[1] As a result, it is very harmful to the noise figure and has a marked effect upon the optimum signal conductance as indicated by Eqs. (4) and (5).

5·6. Behavior of Noise Figure with Increasing Frequency.—To investigate in more detail the trend in noise figure as the frequency is increased, the signal conductance G_a' may be eliminated from Eq. (4) by substituting its value as given by Eq. (5). The resulting expression for optimum noise figure then becomes

$$F_{12} = 1 + 2R_{eq}G_1 + 2\sqrt{(R_{eq}G_1)^2 + \frac{T_g}{T_\tau}(R_{eq}G_1)}. \tag{4a}$$

It is apparent here that the optimum noise figure at band center is a function of only two things, the product of the equivalent noise resistance and the input loading due to the tube and the ratio of the effective temperature of that loading to the room temperature. If the majority of the input loading of the tube is due to transit time, the ratio of (T_g/T_τ) is found to be about 5. With this assumption, the optimum noise figure has been plotted in Fig. 5·11. Here it is seen that for values of $(R_{eq}G_1)$ less than 0.1, the expression for optimum noise figure may be written approximately

$$F_{12} \approx 1 + 2\sqrt{\frac{T_g}{T_\tau}(R_{eq}G_1)}. \tag{4b}$$

Since the tube input loading, whether due to lead inductances or transit time, is proportional to the square of the frequency, Eq. (4b) may be rewritten

$$F_{12} \approx 1 + 2f\sqrt{\frac{T_g}{T_\tau}(R_{eq}k)} \tag{4c}$$

where

$$G_1 = kf^2.$$

Thus

$$F_{12} - 1 = Kf. \tag{4d}$$

[1] D. O. North and W. R. Ferris, *Proc. I.R.E.*, p. 49, February 1941.

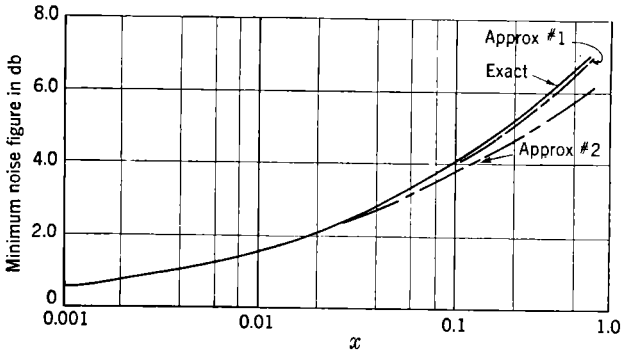


FIG. 5·11.—Universal optimum noise figure curve.

Exact: $F = 1 + 2x + 2\sqrt{x^2 + 5x}$ where $x = R_{eq}G_1$.

Approx. 1: $F = 1 + 2x + 2\sqrt{5x}$ $G_1 \approx k\omega^2$.

Approx. 2: $F = 1 + 2\sqrt{5x}$ $\frac{T_g}{T_r} = 5$.

The following values of the constants are approximately valid for the tubes indicated.

Tube	R_{eq}, ohms	k, μmhos per Mc/sec^2
6AK5(T)	300	0.0157
6AK5(P)	2500	0.0157
6J4	300	0.0281

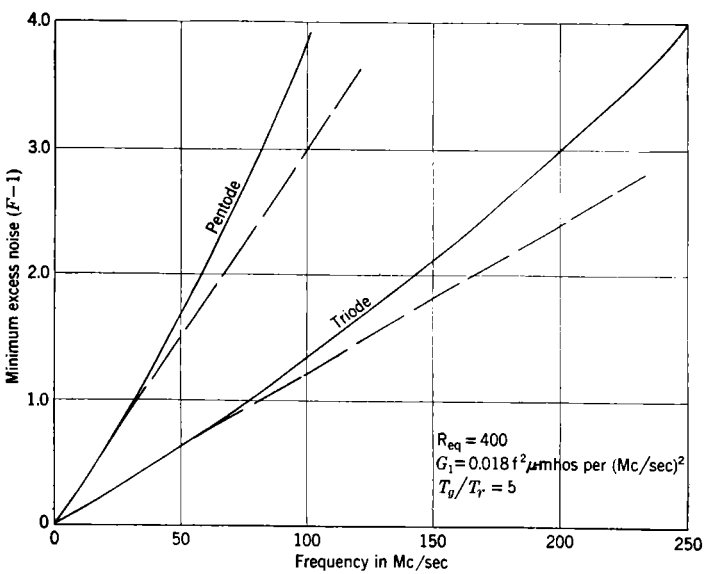


FIG. 5·12.—Minimum excess noise vs. frequency for 6AK5.

This last equation illustrates clearly the deterioration in noise figure that may be expected with increasing frequency for any given tube. Plots of Eq. (4*d*) for a 6AK5 connected as a triode or as a pentode are given in Fig. 5·12. For these plots the assumption was made that the equivalent noise resistance of the tubes does not vary with frequency, and the value of k used was calculated from Eq. (6).

The general trend as one goes to higher frequencies is for the minimum attainable noise figure and optimum signal conductance to become larger. These effects are reduced by tube designs that minimize cathode-lead inductance and electron transit time, but this is limited by the point where the tube elements become too small to manufacture and assemble.

5·7. Gain Attainable at Increasing Frequency.—The fact that both the transit-time loading and the feedback loading increase as the square of the operating frequency also affects the maximum gain available from

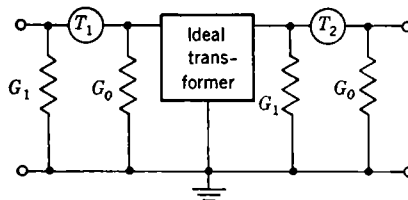


FIG. 5·13.—Cascaded amplifier stages.

a stage. Consider the case of cascaded grounded-cathode stages, each having an input conductance G_1 and an output conductance G_o and interconnected by some sort of a lossless transformer, as shown in Fig. 5·13 where T_1 and T_2 are the amplifier tubes. The input power to a stage is

$$P_i = e_g^2 G_1,$$

and the available output power of the stage is

$$P_o = \frac{(g_m e_g)^2}{4G_o}.$$

Therefore, the maximum power gain possible from the stage is

$$\mathcal{W}_{max} = \frac{g_m^2}{4G_o} \times \frac{1}{G_1}. \tag{7}$$

This varies inversely as G_1, which at high frequencies may be written by use of Eqs. (4·19*a*) and (6) as follows:

$$\begin{aligned} G_1 &= G_{Lc} + G_g = \omega^2 g_m (L_c C_{gc} + k^2) \\ &= K_1 \omega^2 g_m. \end{aligned} \tag{8}$$

With this value of G_1, Eq. (7) becomes

$$\mathcal{W}_{max} = \frac{g_m}{4K_1\omega^2 G_o}. \tag{7a}$$

Since the minimum attainable value of G_0 is the dynamic plate conductance g_p, this may be written

$$\mathcal{W}_{max} = \frac{\mu}{4K_1\omega^2}. \tag{7b}$$

The condition $\mathcal{W}_{max} = 1$ sets one upper limit to the frequency at which a given tube can be successfully used as an amplifier. The upper-frequency limit thus obtained will be the most optimistic one possible because it is determined on the assumption of ideal interstage transformers and does not consider any limitations on amplification other than input loading. Another upper-frequency limit for a tube is the frequency at which its cathode-lead inductance resonates the grid-cathode capacity of the tube. For a 6AC7 this resonant frequency is 254 Mc/sec, and for a 6AK5 it is about 500 Mc/sec.

Three limiting frequencies have now been given. In actual practice, the upper limit of operation will be considerably lower in frequency than any of these calculations would indicate. This is true partly because of the simplifying assumptions made in the calculations and partly because other considerations, such as the fact that the transconductance of a tube becomes complex at vhf, have been omitted. These calculations do, however, serve two very useful purposes. (1) They clearly indicate trends that will be encountered as amplification at higher and higher frequencies is contemplated, and (2) they afford a basis for comparison among the various available tubes. Some practical amplifiers will now be considered.

5·8. Typical Amplifier.—The circuit diagram of a typical pentode amplifier operating at about 200 Mc/sec is shown in Fig. 5·14. No matching transformer has been used in the interstage circuits of this amplifier, since the performance obtained with the much simpler single-tuned coupling circuit is sufficiently good to meet the requirements. With this interstage circuit, however, the maximum power gain per stage becomes

$$\mathcal{W}_{max} = \frac{g_m^2}{(G_0 + G_1)^2}, \tag{9}$$

which is somewhat smaller than the power gain possible with matching as indicated by Eq. (7). With pentodes such as a 6AK5, it should be possible to get some gain from an amplifier of this type at frequencies up to about 450 Mc/sec. The noise figure for a pentode deteriorates so rapidly with increasing frequency that it has not been found profitable to use such a circuit with 6AK5's at frequencies much above 200 Mc/sec.

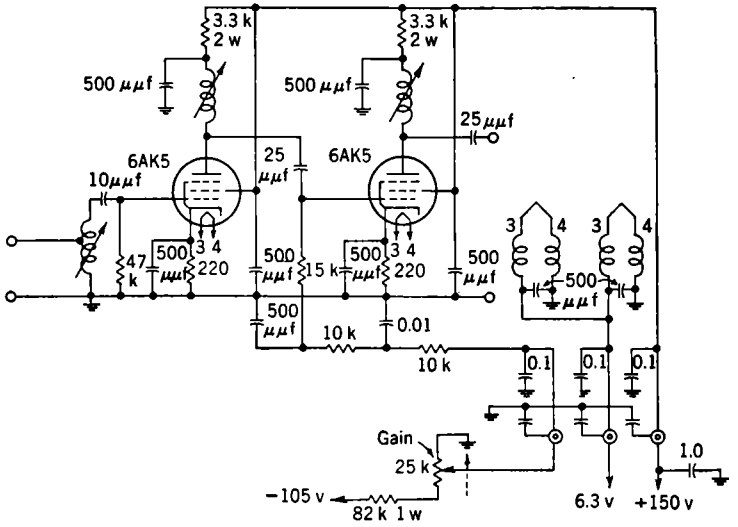


FIG. 5·14.—200-Mc/sec pentode amplifier.

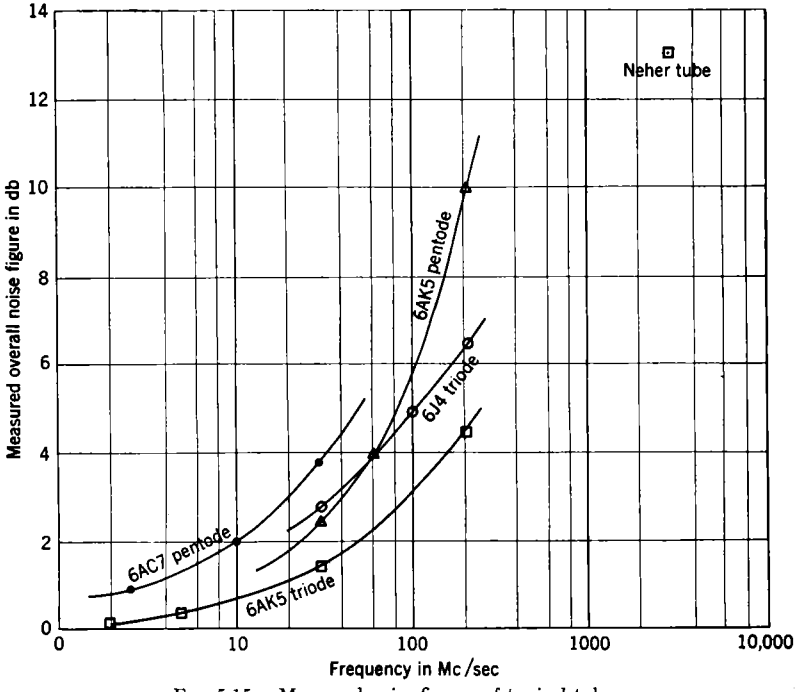


FIG. 5·15.—Measured noise figures of typical tubes.

The noise developed in a pentode amplifier is sufficiently confused even at low frequencies to defy quantitative treatment for other than the simplest cases. At vhf the noise problem is even more confusing; the best information possible, therefore, is that obtained by experimental investigations for each particular tube type. Figure 5·15 gives a plot of experimentally observed noise figures for various amplifier tubes, including a set obtained with 6AK5 pentode amplifiers having a bandwidth of approximately 2 Mc/sec and center frequencies from 5 to 200 Mc/sec.

At vhf the equivalent noise resistances of triodes or pentodes will in general become larger than the values given in Chap. 4. It is experimentally observed, however, that the noise figures of pentode amplifiers deteriorate at a considerably greater rate with frequency than do the noise figures of triode amplifiers. It seems intuitively reasonable that this is due to the more complex structure in a pentode which, at frequencies where mutual and self-inductances of even short leads are appreciable, results in many uncontrolled feedback paths. To illustrate this comparison, the observed noise figure for a two-stage grounded-grid triode amplifier is also plotted in Fig. 5·15.

GROUNDED-GRID CIRCUIT

5·9. Stability.—The elementary circuit and many of the properties of grounded-grid amplifier circuits have already been described in Chap. 4. The oscillation condition in this circuit is found to be

$$C_{pc} \geqq \frac{2(G_1 + g_m')G_2'}{\omega g_m'} - \frac{1}{\omega r_p}, \tag{10}$$

where

$$G_2' = G_2 + \frac{1}{r_p},$$

$$g_m' = g_m\left(\frac{\mu + 1}{\mu}\right).$$

For the special case

$$G_2 = G_1 = G \gg \frac{1}{r_p},$$

writing g_m'/G as $\mathcal{G}$, the maximum allowable voltage gain per stage is, for $\mathcal{G} \gg 1$,

$$\mathcal{G} \approx \frac{g_m'}{\pi f C_{pc}}. \tag{11}$$

Comparison of this expression with Eq. (3) shows that the maximum plate-cathode capacity that can be tolerated in a grounded-grid tube for a given stable gain is considerably greater than the plate-grid capacity that could be tolerated in a grounded-cathode stage having the same gain.

A single curve for a 6J4 grounded-grid triode has been plotted in Fig. 5·9. In this manner, it is possible to use triode amplifiers at frequencies all the way up to 1000 Mc/sec. A few tubes especially designed for grounded-grid operation at vhf are shown in Fig. 5·16.

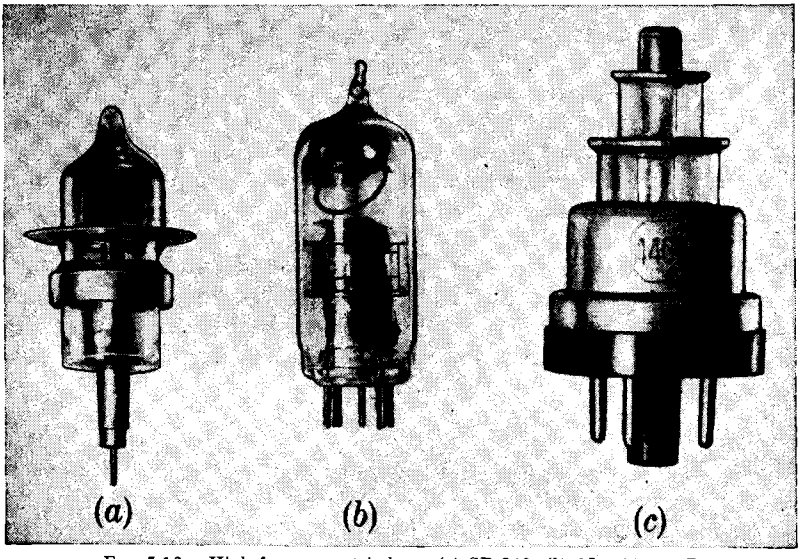


FIG. 5·16.—High-frequency triodes. (*a*) SB-846; (*b*) 6J4; (*c*) 446-B.

5·10. Gain.—For grounded-grid amplifiers operating with moderate transit angles (up to about 55°) Llewellyn's equivalent circuit shown in Fig. 5·17 may be used. The transit-time loading will be written as a shunt conductance proportional to the square of the frequency and having an equivalent temperature of about five times room temperature. The maximum available power gain for the grounded-grid circuit is

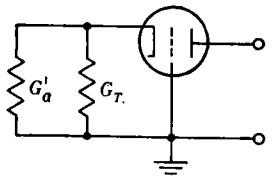


FIG. 5·17.—Equivalent circuit for grounded-grid amplifier.

$$\mathcal{W}_{max} = \left(\frac{\mu + 1}{M + 1}\right)^2 \frac{1}{r_p G_t} \qquad (12)$$

where μ = amplification factor,
r_p = plate resistance,
G_t = transit-time loading plus coil losses,
and

$$M = \sqrt{1 + \frac{g_m + \frac{1}{r_p}}{G_t}}.$$

Comparing this with Eq. (7) it is evident that the ratio of maximum available power gain for grounded-cathode and grounded-grid amplifiers is

$$\frac{\text{Max. grounded-cathode gain}}{\text{Max. grounded-grid gain}} = \left(\frac{\mu}{\mu + 1}\right)^2 \frac{(M + 1)^2}{4}. \tag{13}$$

Thus as the transit-time loading becomes comparable to the transconductance of the tube, the advantage of the grounded-cathode circuit over the grounded-grid circuit becomes smaller and smaller. Finally, for very large transit-time loadings, the maximum available power gain is the same with either circuit configuration.

5·11. Noise Figure.—The noise figure of the grounded-grid circuit with transit-time loading, if the noise from succeeding tubes is omitted, is given by

$$F = 1 + \frac{5G_1}{G_a'} \frac{2.5g_m}{\left(g_m + \frac{1}{r_p}\right)^2} \left(G_a' + 2G_1 + \frac{G_1^2}{G_a'}\right). \tag{14}$$

The value of G_a' that will make this optimum is

$$G_a' \text{ (opt)} = \sqrt{G_1^2 + \frac{5G_1}{R_{eq}}}. \tag{15}$$

This result is the same as was previously found for the grounded-cathode circuit and given in Eq. (5). Thus, as long as noise from succeeding stages is negligible, the grounded-grid amplifier compares favorably with the grounded-cathode amplifier. The importance of this second-stage noise will, in general, depend upon the over-all bandwidths being considered as well as upon the frequency of operation. An actual 200-Mc/sec two-stage amplifier using grounded-grid 6J4 tubes is described in some detail in Chap. 17. A single-stage grounded-grid amplifier for operation at 500 Mc/sec is shown in Fig. 5·18. The tuned circuits are brass straps. They are typical of what may be anticipated in the transition region between lumped- and distributed-circuit parameters.

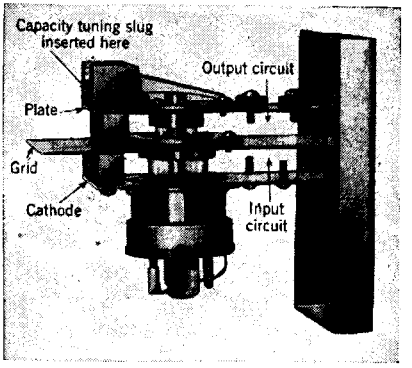


FIG. 5·18.—Grounded-grid amplifier for 500 Mc/sec.

GROUNDED-CATHODE GROUNDED-GRID CIRCUIT

In Chap. 4 a circuit has been described using two triode stages, the first a grounded-cathode and the second a grounded-grid stage. At 200 Mc/sec, this circuit has been found to give an improvement of 1 db in noise figure over an amplifier using two grounded-grid stages (the amplifier described in Chap. 17 and shown in Fig. 17·4). The tubes used are 6J4's in both cases, and the bandwidths are 4 Mc/sec. At higher frequencies or with narrower bands, the difference in noise figure should decrease.

MIXERS

Although the r-f amplifiers and coupling circuits described in previous sections are common to both TRF and superheterodyne receivers, the mixer is found exclusively in superheterodyne circuits and contributes considerably to the complexity of a circuit of this type. A superheterodyne converter may be broken down into two parts: a local oscillator which supplies a high-level voltage at a frequency differing from the signal frequency by the intermediate frequency and a mixer which combines the LO and signal voltages to give an i-f voltage. Although at lower frequencies the mixer and local oscillator are often combined into a single tube called a "converter," this technique is seldom feasible at vhf. The primary difficulty lies in obtaining satisfactory operation of the oscillator portion of converter tubes at such frequencies.

5·12. Conversion Transconductance.—Of primary interest among the characteristics of a mixer stage are its conversion gain, noise figure, and excitation requirements. The conversion gain is defined as the ratio of the output voltage, current, or power at the intermediate frequency to the input voltage, current, or power at the signal frequency. A quantity that is often of greater usefulness is the conversion transconductance, which is defined as the quotient of the i-f output current and the signal input voltage. In considering the design of a mixer stage, it is desirable to know the maximum value of conversion transconductance that may be expected from various types of tubes and the amounts of LO power or voltage necessary to obtain this optimum. If the mixer is to operate with little or no previous r-f amplification, it is also desirable to know its i-f noise output.

Conversion transconductance of any tube is determined by considering the modulation of the LO frequency by the signal inside the tube.[1] The general type of analysis that will be given here may be applied to any type of tube. If the signal voltages are assumed small and the LO

[1] E. W. Herold, "Superheterodyne Converter Considerations in Television Receivers," *RCA Rev.*, **4**, 324, January 1940; E. W. Herold and L. Malter, "Radio Reception at UHF," *Proc. I.R.E.*, **31**, 567, October 1943.

voltage large, the transconductance from the signal electrode to the output of the mixer tube will be a function of the LO voltage alone. This transconductance may, therefore, be considered as varying periodically at the LO frequency and can be written as a Fourier series:

$$g_m = a_0 + a_1 \cos \omega_0 t + a_2 \cos 2\omega_0 t + \cdots, \tag{16}$$

where ω_0 is the angular frequency of the local oscillator. If a small signal $e_s \sin \omega_s t$ is applied to the signal electrode, the resulting alternating plate current will be given by the expression

$$\begin{aligned} i_p &= g_m e_s \sin \omega_s t \\ &= a_0 e_s \sin \omega_s t + e_s \sum_{n=1}^{\infty} a_n \sin \omega_s t \cos n\omega_0 t \\ &= a_0 e_s \sin \omega_s t + \tfrac{1}{2} e_s \sum_{n=1}^{\infty} a_n \sin (\omega_s + n\omega_0) t \\ &\quad + \tfrac{1}{2} e_s \sum_{n=1}^{\infty} a_n \sin (\omega_s - n\omega_0) t, \end{aligned} \tag{17}$$

where only the first-order terms in e_s have been retained. If the intermediate frequency is to be lower than the signal frequency, the plate circuit of the mixer must be tuned to the frequency $(\omega_s - n\omega_0)$. As n may be any integer, the intermediate frequency may be chosen equal to the signal frequency minus any multiple of the LO frequency. This is true even if the LO signal is a pure sine wave, since the harmonics are introduced by the time variation of the signal transconductance. The conversion transconductance is given by

$$g_{cn} = \left.\frac{i_p}{e_s}\right]_{\omega = (\omega_s - n\omega_0)} = \frac{a_n}{2}, \tag{18}$$

where a_n is the coefficient of the Fourier series of Eq. (16). Therefore

$$g_{cn} = \frac{1}{2\pi} \int_0^{2\pi} g_m \cos n\omega_0 t \, d(\omega_0 t), \tag{19}$$

where n will be equal to 1, 2, or 3, depending upon whether the plate circuit of the mixer is tuned to $(\omega_s - \omega_0)$, $(\omega_s - 2\omega_0)$, or $(\omega_s - 3\omega_0)$. If a curve for the variation in g_m with oscillator electrode voltage is available, it is possible to assume an applied oscillator voltage and obtain a curve of signal transconductance vs. time which may be used to evaluate the integral of Eq. (19) either graphically or by approximate methods. One such

approximate method[1] which will give results of sufficient accuracy for many purposes is to assume a sine-wave oscillator voltage and carry through an analysis at 30° intervals by assuming straight-line connections between these points. This procedure is indicated for a typical transconductance curve in Fig. 5·19. If a diode is being analyzed, this curve

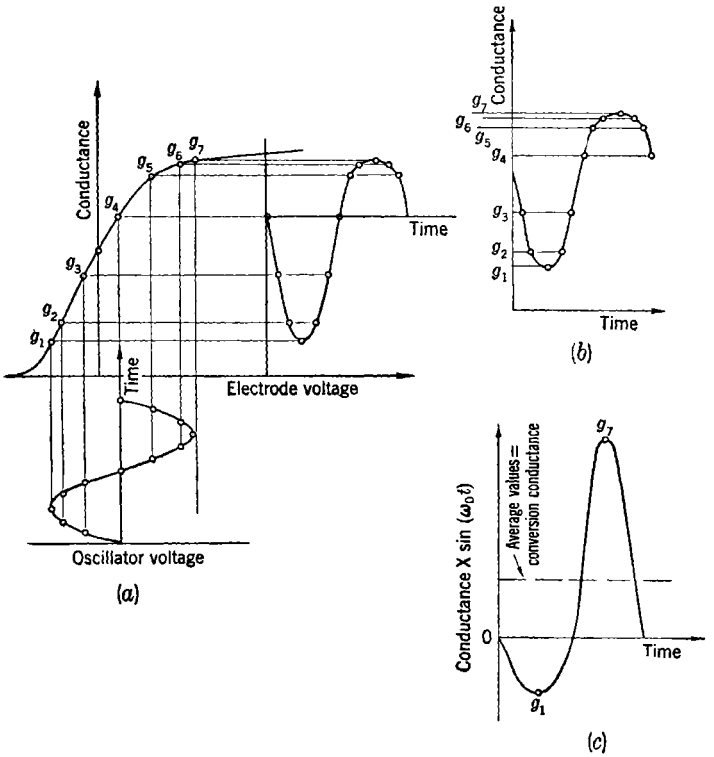


FIG. 5·19.—Graphical computation of conversion conductance.

would be the diode conductance vs. voltage; whereas if a triode or pentode is being considered, it would be the curve of grid-plate transconductance vs. grid voltage (assuming grid injection of the local oscillator).

From this method, the conversion transconductances for the first, second, and third harmonics of the local oscillator are

$$g_{c_1} = \tfrac{1}{12}[(g_7 - g_1) + (g_5 - g_3) + \sqrt{3}\,(g_6 - g_2)], \tag{20}$$

$$g_{c_2} = \tfrac{1}{12}[2g_4 + \tfrac{3}{4}(g_3 + g_5 - g_6 - g_2) - (g_7 + g_1)], \tag{21}$$

$$g_{c_3} = \tfrac{1}{12}[(g_7 - g_1) - 2(g_5 - g_3)]. \tag{22}$$

[1] Herold, *loc. cit.*

These expressions will be reasonably accurate for values of g_{c_1} but will be less reliable for g_{c_2} and g_{c_3}. To obtain accurate results for these higher harmonic conversion transconductances, a more careful integration would be required. It is apparent from Eq. (20) that for maximum conversion transconductance it would be desirable to have g_1, g_2, and g_3 all zero and g_5, g_6, and g_7 a maximum, or vice versa. From Fig. 5·19 it is apparent that this means that it is desirable to operate the mixer with the oscillator voltage swinging the transconductance into cutoff for an appreciable portion of the cycle and to have the slope of the transconductance curve a maximum in the region of operation.

5·13. Noise Output.—A second property that will be of interest in considering different types of mixer tubes is the fluctuation noise that they contribute to the circuit. From previous analyses,[1] the mean-squared noise current in the output of diodes, triodes, and pentodes can be determined as a function of transconductance and currents in the tube. These mean-squared output noise currents, denoted by $\overline{i_{pn}^2}$, must be averaged over the LO cycle to determine the resulting output noise of a mixer stage,

$$\overline{i_n^2} = \frac{1}{2\pi} \int_0^{2\pi} \overline{i_{pn}^2} d(\omega_0 t). \tag{23}$$

The equivalent noise resistance of a mixer tube will be

$$R_{eq} = \frac{\overline{i_n^2}}{4kTBg_{cn}^2}. \tag{24}$$

5·14. Diode Mixer.—A typical diode-mixer circuit is shown in Fig. 5·20.

This circuit may be divided into two equivalent circuits, one representing the circuit as seen from the input or LO terminals and the other representing the circuit as seen from the i-f terminals. These equivalent circuits are shown in Fig. 5·21*a* and *b* respectively.[2]

It will be assumed that the diode-conductance curve is the same for intermediate frequency as for vhf. This amounts to neglecting transit times in the diode. Then from Fig. 5·21*a* it is apparent that the input

[1] B. J. Thompson, D. O. North, and W. A. Harris, *RCA Rev.*, **5**, April 1941.

[2] This is really an oversimplification in the case of the diode mixer, because the input and output circuits interact to a greater degree than the analysis suggests. A complete treatment would have to include the remixing of the i-f output with the LO voltage to produce new voltages at both signal and image frequencies. Such a treatment may be found in Vol. 16, Radiation Laboratory Series, for the case of the crystal rectifiers used in the microwave region. Herold and Malter, *op. cit.*, have considered the interaction of the i-f circuit with the input but have not brought in explicitly the behavior of the circuits at the image frequency.

power to the diode is

$$P_i = e_d^2 \overline{g_d},$$

where $\overline{g_d}$ is the average value of the diode conductance over the LO cycle; that is,

$$\overline{g_d} = \frac{1}{2\pi} \int_0^{2\pi} g_d(\omega_0 t) d(\omega_0 t).$$

From Fig. 5·21*b*, the available output signal power of the mixer at

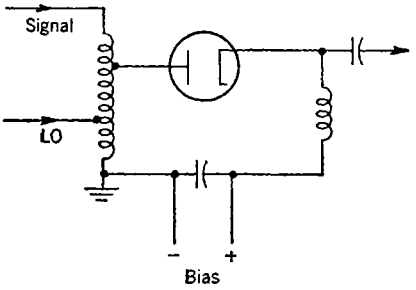


FIG. 5·20.—Typical diode-mixer circuit.

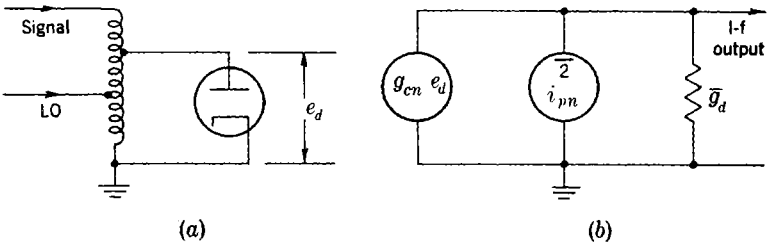


FIG. 5·21.—Equivalent circuits for diode mixer.

the intermediate frequency is

$$P_o = \frac{(g_{cn} e_d)^2}{4\overline{g_d}}.$$

Thus the maximum available conversion gain of the mixer is

$$\mathcal{W}_c = \frac{P_0}{P_i} = \left(\frac{g_c}{2\overline{g_d}}\right)^2, \tag{25}$$

since the conversion transconductance for the fundamental of the oscillator frequency is larger than that for any of its harmonics.

Using these same equivalent circuits, the noise figure of the mixer is calculated and found to be

$$F = 1 + \frac{\overline{i_n^2}}{\overline{i_s^2}} \left(\frac{G_a' + \overline{g_d}}{g_c}\right)^2, \tag{26}$$

where $\overline{i_s^2} = 4kTBG_a'$,

G_a' = the conductance of the signal source as seen from the diode terminals.

For a typical diode, with space charge, at any point in the oscillator cycle,[1]

$$\overline{i_{pn}^2} \approx 4kTB(2.2g_d).$$

Therefore,

$$\overline{i_n^2} = \frac{4kTB}{1}\frac{2.2}{2\pi}\int_0^{2\pi} g_d(\omega_0 t) = 4kTB(2.2\overline{g_d}). \tag{27}$$

Combining Eqs. (26) and (27),

$$F = 1 + \frac{2.2\overline{g_d}}{G_a'}\left(\frac{G_a' + \overline{g_d}}{g_c}\right)^2, \tag{28}$$

which has a minimum value for $G_a' = \overline{g_d}$ of

$$F_{\min} = 1 + 2.2\left(\frac{2\overline{g_d}}{g_c}\right). \tag{29}$$

The average diode conductance $\overline{g_d}$ may be obtained graphically from the diode characteristic. The same approximate analysis outlined for determining the conversion transconductance may be applied to find the conductance, which in the notation of Fig. 5·19 becomes

$$\overline{g_d} = \tfrac{1}{12}(g_1 + 2g_2 + 2g_3 + 2g_4 + 2g_5 + 2g_6 + g_7).$$

The ratio of $g_c/\overline{g_d}$ can be made to approach a maximum value of unity for the case where appreciable negative bias and large oscillator swing are applied to the diode. In this limiting case, the best noise figure for the thermionic diode is 9.9 db, and the maximum power gain is −6.0 db. From these figures, the optimum over-all noise figure that may be expected from a superheterodyne receiver using a diode mixer and no r-f amplification is

$$\begin{aligned} F_{12} &= F_1 + \frac{F_2 - 1}{G_1} \\ &= 9.8 + 4(F_2 - 1), \end{aligned} \tag{30}$$

where F_2 = noise figure of i-f amplifier.

Although the noise figures indicated by Eq. (30) are not so good as those obtained with other kinds of mixers or with r-f amplifiers, they are still sufficiently good for many purposes. Since the average impedance of a diode is rather low, the local oscillator must furnish a considerable amount of power. This is unfortunate, since it serves to aggravate the problems of LO stability.

[1] See Thompson, North, and Harris, *op. cit.*

5·15. Triode and Pentode Mixers.—A mixer of another type which is usable at vhf and which is capable of considerably better gain and noise figure than a diode is the triode mixer. A typical triode-mixer circuit is shown in Fig. 5·22.

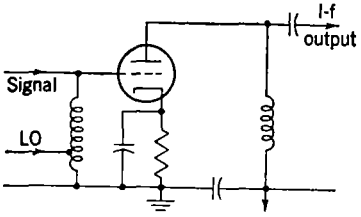


FIG. 5·22.—Triode-mixer circuit.

This circuit may also be divided into two equivalent circuits, one applying to the LO and signal frequencies and one applying to the intermediate frequencies. For the analysis of signal-to-noise ratio and gain, the local oscillator may be eliminated from the r-f circuit by assuming that the LO injection is always adjusted for optimum conversion transconductance, which may be shown to give both optimum gain and optimum signal-to-noise ratio. These two circuits are shown in Fig. 5·23*a* and *b*. In these circuits,

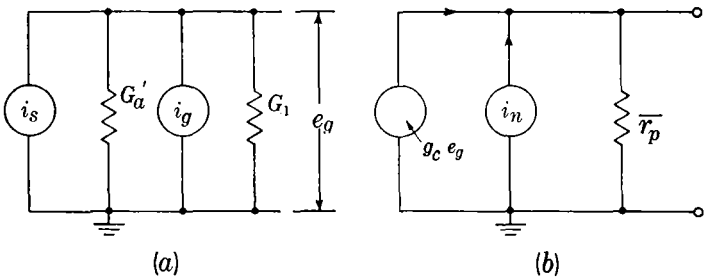


FIG. 5·23.—Equivalent circuits for triode mixer.

G_a' = input loading due to signal source,
G_1 = input loading due to tube and coupling circuit,
$\overline{i_a^2} = 4kT_rBG_a'$,
$\overline{i_g^2} = 4kT_gBG_1$,
T_r = absolute room temperature,
T_g = effective absolute temperature of grid loading,
B = noise bandwidth,
k = Maxwell-Boltzmann constant,

$$\overline{r_p} = \frac{1}{2\pi}\int_0^{2\pi} r_p d(\omega_0 t),$$

r_p = plate resistance for each grid bias.

By a procedure essentially the same as that used in Eq. (27), $\overline{i_n^2}$ may be written as

$$\overline{i_n^2} = 4kT_rB(2.5\bar{g}_m).$$

where

$$\bar{g}_m = \frac{1}{2\pi} \int_0^{2\pi} g_m(\omega_0 t) d(\omega_0 t)$$

and $g_m(\omega_0 t)$ is the grid-plate transconductance at any point in the LO cycle. This value for $\overline{i_n^2}$ may be substituted in Eq. (24) to give

$$R_{eq} = \frac{2.5\bar{g}_m}{g_c^2} \tag{31}$$

for the equivalent noise resistance of a triode mixer.

The noise figure may now be computed to be

$$F = 1 + \frac{T_g}{T_r} \frac{G_1}{G_a'} + \frac{2.5\bar{g}_m}{g_c^2} \frac{(G_a' + G_1)^2}{G_a'}, \tag{32}$$

which may also be written

$$F = 1 + \frac{T_g}{T_r} \frac{G_1}{G_a'} + \frac{R_{eq}}{G_a'} (G_a' + G_1)^2. \tag{32a}$$

This is seen to be identical with Eq. (4) for the noise figure of a triode or pentode amplifier; therefore Eq. (5) applies equally well to this mixer.

In general, it is necessary to evaluate both g_c and $\bar{g}_m$ from the tube characteristics as previously indicated. For most triodes with optimum LO injection, however, it is found approximately that

$$g_c = 0.28 g_0,$$
$$\bar{g}_m = 0.47 g_0,$$

where g_0 is the peak grid-plate transconductance in the LO cycle. Thus,

$$R_{eq} = \frac{15}{g_0}. \tag{31a}$$

From the equivalent circuits, it is also possible to calculate the available conversion gain of the triode mixer

$$\mathcal{W}_c = \frac{g_c^2}{(G_1 + G_a')^2} \frac{\mu G_a'}{\bar{g}_m}. \tag{33}$$

On substituting the values given above for g_c and $\bar{g}_m$ this becomes

$$\mathcal{W}_c = \frac{\mu}{6} \frac{G_a' g_0}{(G_a' + G_1)^2}. \tag{33a}$$

Comparing Eqs. (32a) and (33) with Eqs. (4) and (7) for amplifiers, it is observed that the only essential differences between the two lie in the magnitudes. Thus all commentaries made in the previous amplifier section on trends as the vhf region is approached apply equally well here. The important difference between the two cases is that the power gain

and noise figure which can be obtained from a given tube as a mixer can never be so good as the power gain and noise figure of the same tube operating as an amplifier. At the lower frequencies, the noise figures of a given tube either as a mixer or an amplifier can, in general, be made very good and nearly equal. In the vhf range, however, the amplifier is usually considerably better than the mixer. At still higher frequencies where amplification becomes very inefficient, the mixer may give better performance.[1]

The entire discussion of a triode mixer applies equally well to a pentode mixer except that the equivalent noise resistance of a pentode is larger than the equivalent noise resistance of the same tube as a triode. It is[2]

$$R_{eq} = \frac{15 + 21\dfrac{I_0}{g_0}}{g_0}, \tag{34}$$

where g_0 is the peak grid-plate transconductance, I_0 is the peak cathode current, and the LO excitation is adjusted to optimum. For the same peak transconductance from grid to plate, both the triode and pentode will have the same conversion transconductance; but if a single tube is compared for the pentode and triode connection, the triode will give a somewhat larger gain because its transconductance will be somewhat larger than that of the pentode for a given cathode current. Because of the higher output impedance of the pentode, however, the maximum available power gain will usually be greater for the pentode connection.

The discussions presented here for the triode or pentode mixer have all been based on calculations made at the center of the amplifier or mixer pass band, and the tube interelectrode capacities have all been assumed zero or tuned out. Consideration of these reactive elements will not change the general trends of noise figure and optimum input stepup noticeably. These band-center calculations may be taken as representing a lower limit to any results that may be obtained with actual circuits. If a more detailed knowledge of the circuit behavior is required, it will be necessary to calculate the noise figures and noise gains as a function of frequency and then average them over the pass band to determine the actual characteristics.[3]

Although these quantities are functions of frequency, their average values over narrow bandwidths will not differ noticeably from the band-center values.

[1] See E. W. Herold and L. Malter, *Proc. I.R.E.*, **31,** October 1943.

[2] E. W. Herold, *RCA Rev.*, **4,** 324, January 1940. The screen current has been assumed to be one-fourth as large as the plate current. For a ratio differing greatly from this a new expression for R_{eq} would be required.

[3] E. J. Schremp, RL Internal Report No. 61—5/18/1943.

Away from the band center the pentode has a considerable advantage over the triode because of its small grid-plate capacity. Because the impedance in the plate circuit of a mixer is usually very small at the signal frequency, there is seldom any danger of oscillation due to grid-plate feedback, but this feedback may cause undesirable interactions between the r-f and i-f circuits. Again, the feedback from the i-f into the r-f circuit may make it extremely difficult to obtain satisfactory tracking if the mixer is to be ganged with the local oscillator and one or more r-f stages. When a triode mixer is used, these feedback effects may be prohibitive for narrow-band circuits but are frequently quite tolerable in wide-band applications. They may be overcome in mixers, as they are in amplifiers, through neutralization or by using the triode in a grounded-grid circuit.

5·16. Local-oscillator Injection.—Throughout these comparisons, it has always been assumed that the LO injection is adjusted to optimum. It is worth while to consider briefly how this may be done. For all the mixers considered, it is customary to inject the local oscillator on the input-signal terminal. Screen- or suppressor-grid injection is possible with the pentode and may be used if LO reradiation is of extreme importance. However, the signal-to-noise ratio and the gain performance obtained in this manner will not be so good as would be possible with control-grid injection. In the triode or pentode mixer, it is also possible to inject the local oscillator in the cathode circuit, but the resulting input loading is usually undesirable.

With the diode mixer, optimum performance is obtained by biasing the tube considerably beyond cutoff and then using a rather large LO signal to swing the tube into the conducting region. This required bias for a diode mixer is most easily obtained through self-bias by means of a parallel *RC*-circuit in series with the diode.

For the triode and pentode mixers, the optimum LO voltage on the input grid is about seven-tenths as large as the cutoff bias. The operating bias for these mixers should usually be near the point of maximum curvature in the transconductance vs. grid-voltage curve. This bias may be obtained through a fixed voltage, a cathode resistor, or a high-resistance grid leak. Of the three possibilities, the use of the high-resistance grid leak is most desirable, provided there are no severe transient overload signals, because it tends to minimize changes in conversion transconductance with LO voltage; however, if for any reason the local oscillator fails entirely, the mixer current may become so high that the tube is damaged. This may be overcome in a pentode by the use of a screen dropping resistor which has such a value that the tube is protected for this condition. With a triode mixer, however, it is necessary to use either cathode-resistor bias alone or a combination of grid-leak and cathode-resistor

bias. Fixed bias is usually inconvenient to obtain and causes the conversion transconductance of the mixer to be markedly dependent upon LO voltage.

The problem of coupling the LO voltage into the grid circuit may be solved in various ways. Since most local oscillators will develop 60 volts or more of signal without trouble, and since triode or pentode mixers absorb almost no power, the problem is frequently to keep from injecting too much LO voltage on to the grid. Usually the oscillator voltage is attenuated by capacitive or resistive voltage dividers and then either tapped directly into the mixer grid coil or coupled in with a small condenser. With a diode mixer which requires considerably higher LO power, it may be necessary to match the local oscillator into the mixer through suitable transformers. This is an undesirable procedure, however, because of the resulting interaction between the local oscillator and the mixer.

5·17. Nonthermionic Mixers.—In addition to thermionic tubes, it is possible to use a number of other types of nonlinear devices as mixers. The most notable among these which may be used at vhf is the silicon-tungsten crystal. These crystals are the stock in trade of microwave receivers and were discussed in Chap. 2. At frequencies below about 1000 Mc/sec it has usually been found more profitable to use thermionic tubes, since the performance so obtained can be as good as or better than that to be had with the crystal and there is no danger of burnout under r-f overload.

LOCAL OSCILLATORS

The last major component of a superheterodyne r-f head is the local oscillator, which furnishes an output differing from the signal frequency by the intermediate frequency. At lower frequencies, many triode oscillator circuits employ mutual-inductance-coupling, for example, the Hartley circuit and its variations. At vhf the use of mutually coupled coils becomes either difficult or impossible, and the number of oscillator types is reduced to two.

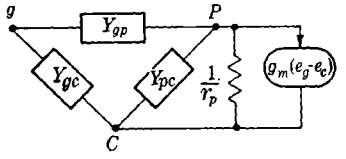


FIG. 5·24.—Oscillator circuit without mutual inductance.

5·18. General Circuit Usable at Vhf. The most general triode oscillator circuit without mutual inductance is shown in Fig. 5·24.

The general oscillation conditions in this circuit are established by making the determinant of this network equal to zero.

$$(Y_{gp} + Y_{gc})\left(Y_{pc} + \frac{1}{r_p}\right) + Y_{gc}Y_{gp} + Y_{gp}g_m = 0. \qquad (35)$$

Equation (35) is a complex expression; to make it zero, it must be broken up into its real and imaginary parts, and both parts must be made equal to zero. Before doing this it will be assumed that Y_{gc}, Y_{gp}, and Y_{pc} are all pure imaginaries. These elements are made up of the tube interelectrode capacities plus the external circuits including the coils,

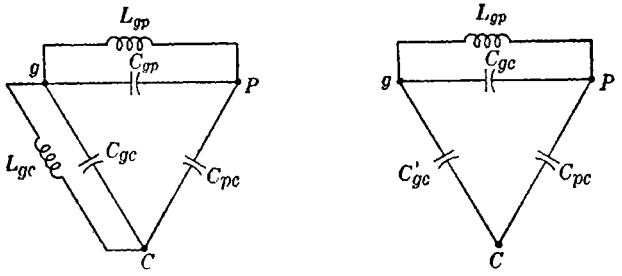


FIG. 5·25.—Colpitts circuit, Case I.

condensers, or lines used to tune the oscillators, which are assumed to be lossless. The losses in actual vhf components can usually be kept sufficiently low so that this assumption is essentially fulfilled. Thus, for

$$\left.\begin{aligned} Y_{gp} &= jB_{gp}, \\ Y_{gc} &= jB_{gc}, \\ Y_{pc} &= jB_{pc}, \end{aligned}\right\} \tag{36}$$

Eq. (35) can be broken down into its real and imaginary parts, and one

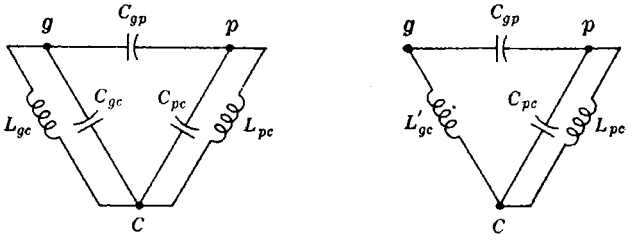


FIG. 5·26.—Tuned-plate tuned-grid circuit, Case II.

has as the conditions of oscillation

$$\left.\begin{aligned} B_{pc}(B_{gp} + B_{gc}) + B_{gc}B_{gp} &= 0, \\ \frac{1}{r_p}(B_{gp} + B_{gc}) + g_m B_{gp} &= 0. \end{aligned}\right\} \tag{37}$$

In order that these conditions be satisfied simultaneously, it is apparent that certain relations between the various B's must exist, namely,

1. B_{gp} and B_{gc} must have opposite signs.
2. B_{gc} must be greater than B_{gp}.
3. B_{pc} and B_{gc} must have the same sign.

There are only two possible combinations that will satisfy Eq. (37). These are indicated in the following table and have been called Case I and Case II. The corresponding circuit configurations in terms of coils and condensers are shown in Figs. 5·25 and 5·26.

TABLE 5·1.—COMBINATIONS NECESSARY FOR EQ. (37)

	B_{gc}	B_{gp}	B_{pc}
Case I	(+)	(−)	(+)
Case II	(−)	(+)	(−)

5·19. Colpitts Circuit.—The oscillator circuit of Case I is the familiar Colpitts circuit, and Case II is the equally common tuned-plate tuned-grid circuit. For each circuit, the actual configuration of parts has been drawn to the left, and the equivalent circuit at the frequency of oscillation is drawn to the right. At the frequency of oscillation the external inductance of the Colpitts oscillator L_{gp} must resonate with C_{gp} in parallel with some capacity less than C_{pc}, since B_{gc} is capacitive. For the tuned-plate tuned-grid oscillator, on the other hand, the external inductance L_{pc} must resonate C_{pc} in parallel with some capacity greater than C_{gp}, since B_{gc} must be inductive. Thus, for a given tube and a given external circuit, the Colpitts circuit can be made to oscillate at a higher frequency than the tuned-plate tuned-grid circuit. For the Colpitts circuit,

$$\left.\begin{aligned} B_{gp} &= \left(\omega C_{gp} - \frac{1}{\omega L_{gp}}\right), \\ B_{gc} &= \omega C_{gc}, \\ B_{pc} &= \omega C_{pc}. \end{aligned}\right\} \tag{38}$$

These values may be substituted, and Eqs. (37) solved for the oscillation frequency and the tube characteristics necessary to start the oscillation. These are

$$\omega_0 = \frac{1}{\sqrt{L_{gp}\left(C_{gp} + \dfrac{C_{pc}C_{gc}}{C_{pc} + C_{gc}}\right)}}, \tag{39}$$

$$\mu = \frac{C_{gc}}{C_{pc}}. \tag{40}$$

These conditions serve to describe only the circuit of Fig. 5·24 and not an actual oscillator. In particular, it was assumed that all elements in the oscillator behave as linear circuit elements, an assumption that is true only for the starting of the oscillation in an actual circuit.

5·20. Tuned-plate Tuned-grid Circuit.—In a similar manner the tuned-plate tuned-grid circuit may be analyzed with the assumption of

lossless external circuits. For this case

$$\left.\begin{aligned} B_{gp} &= \omega C_{gp}, \\ B_{cp} &= \omega C_{pc} = \frac{1}{\omega L_{pc}}, \\ B_{gc} &= \omega C_{gc} - \frac{1}{\omega L_{gc}}. \end{aligned}\right\} \tag{41}$$

Equations (37) give the frequency ω_0 at which the oscillation will begin and the value of tube amplification factor μ necessary for the oscillations to build up. For $\mu \gg 1$,

$$\omega_0 \approx \frac{1}{\sqrt{L_{cp}(C_{pc} + C_{gp})}}, \tag{42}$$

$$\mu \approx \frac{L_{cp}}{L_{gc}}\left(1 + \frac{C_{pc}}{C_{gc}}\right) - \frac{C_{gc}}{C_{gp}}. \tag{43}$$

In addition to operating at a lower frequency for a given tube and external circuit, the tuned-plate tuned-grid oscillator has been found to be rather unsatisfactory for vhf applications because of experimentally observed tendencies to instability and erratic operation. This circuit is, however, used to a considerable degree at lower frequencies.

5-21. Actual Oscillators.—At vhf the Colpitts circuit may take on a number of forms, the form used depending upon which terminal of the tube is to be at ground potential and whether the external circuits used are of lumped- or distributed-parameter types. For circuits using lumped coils, the grounded-grid configuration is frequently preferred, since it allows one side of both the tuned circuit and the cathode choke to be solidly grounded, and thus lowers the capacity effects on the tuning condenser if one is used. Two typical circuits of this type are shown in Fig. 5·27. The first of these circuits was designed for 200-Mc/sec operation and utilizes grid-leak bias. In this circuit, lumped parameters were used and the oscillator frequency was controlled by varying the grid-plate capacity. The second circuit utilizing a coaxial line as the tuned circuit has been used satisfactorily for frequencies up to 1000 Mc/sec. In this circuit, cathode bias was used

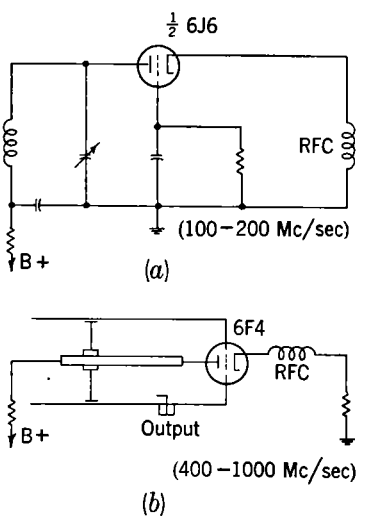


FIG. 5·27.—Oscillator circuits (a) 200 Mc/sec; (b) 1000 Mc/sec.

because of the difficulty of introducing grid-leak bias into the line construction.

A photograph of an oscillator of the type shown in Fig. 5·27*b* designed for 1000-Mc/sec operation is shown in Fig. 5·28. At even higher frequencies where it is desirable to used tuned lines for all the elements, all

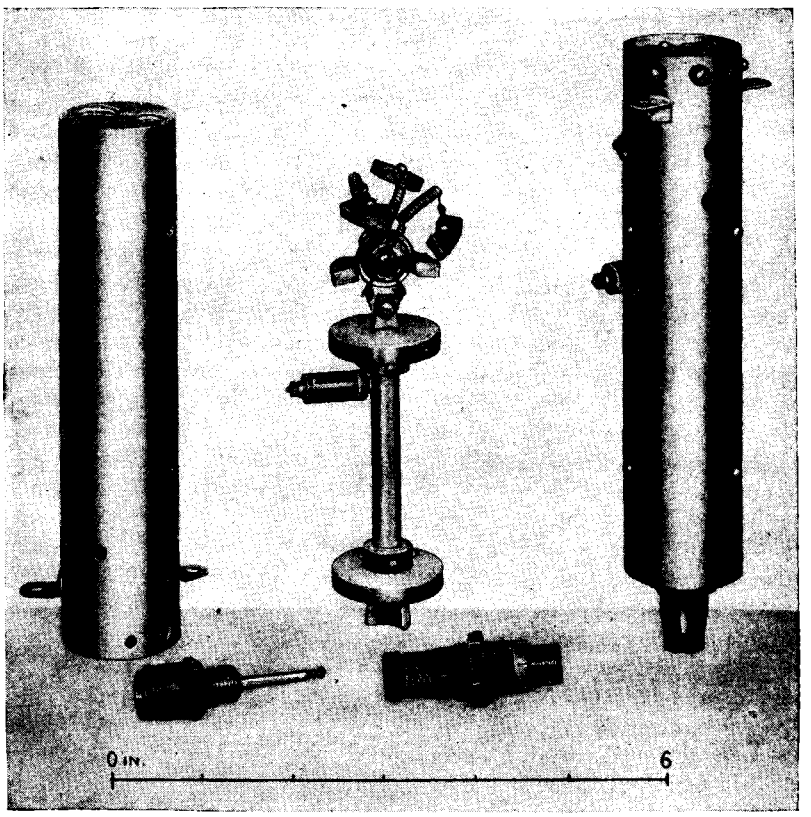


FIG. 5·28.—1000-Mc/sec oscillator.

three of the possible mechanical configurations shown in Fig. 5·29 have been successfully used, and tubes have been developed to suit each particular configuration. Electrically, all of these may be classed as grid-separation circuits. The operation of a high-frequency oscillator is frequently so inefficient that the limit of operation is determined largely by the dissipation of the grid or plate electrode. Depending upon which is the limiting element with regard to dissipation, the configuration of Fig. 5·29*a* or *b* may have a definite advantage over the form shown in

Fig. 5·29*c* because of the relative ease of cooling the hot electrode in these circuits. Figure 5·29*a* has the disadvantage of being somewhat more difficult to construct than Fig. 5·29*b*; changing tubes, however, is a much easier operation.

Tuning in each of these oscillator circuits is accomplished by moving the short-circuiting plug in the grid-plate line through some appropriate mechanical scheme such as a cam or lead screw. If extremely wide tuning range with uniform output is required, it will probably be found necessary to tune the grid-cathode line as well, although the tolerance for tuning this line is rather wide.

Frequency drift with temperature changes will always be present to some degree. As a rule, the distributed-line oscillators are considerably more stable with temperature changes than are lumped-element oscillators using solenoidal coils, since such coils change their inductance rather rapidly with temperature. In oscillators such as the one shown in Fig. 5·27*a*, however, the temperature drifts may be reduced to about one part in fifty thousand per degree centigrade through the use of suitable compensating condensers. The choice of the condenser size and the temperature coefficient needed will depend upon the individual circuit construction and must be determined separately for every case.

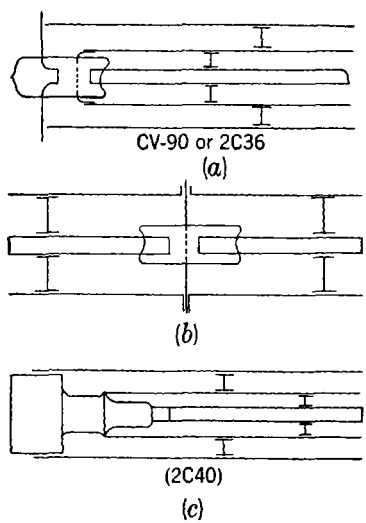


FIG. 5·29.—Mechanical configuration of grid-separation oscillator circuits.

The only other major problem in the design of a local oscillator is the choice of the LO frequency. For a given signal frequency ω_s and intermediate frequency ω_{IF}, the usable LO frequencies will be given by the expression

$$\omega_0 = \frac{\omega_s \pm \omega_{IF}}{n} \qquad (n = 1, 2, 3, \cdots), \tag{44}$$

where n equal to 1, 2, or 3 corresponds to operation on the first, second, or third harmonic of the local oscillator. It should be remembered that it is not necessary that the LO output itself contain any of these harmonic voltages, since the nonlinearity of the mixer will generate them even though a perfectly sinusoidal LO voltage is used. Equations (20) to (22) show, however, that the maximum conversion transconductance is obtained with the LO fundamental. The conversion transconductance

to be obtained at the second and third harmonics will be found to be respectively about one-half and one-third that obtained at the fundamental. Thus, use of an LO frequency lower than the fundamental results in a loss of gain and an increase in noise figure for the mixer stage. Furthermore, the number of possible image responses will be markedly increased by operation on an oscillator harmonic other than the fundamental. For a given oscillator frequency ω_0, there can be image responses at the frequencies

$$\omega_{\text{image}} = k\omega_0 \pm \omega_{\text{IF}} \qquad (k = 1, 2, 3, \cdots). \tag{45}$$

The only reason that might be advanced for operating the mixer on a harmonic of a local oscillator would be the inability to obtain satisfactory oscillator operation on the fundamental. This is a situation which will probably never be encountered at vhf. For the widest spacing of image responses and the highest conversion transconductance, it is desirable to operate the local oscillator at the frequency of $(\omega_s + \omega_{\text{IF}})$.

CHAPTER 6

INTERMEDIATE-FREQUENCY AMPLIFIERS

By H. Wallman

6·1. Introduction.—The amplifiers discussed in this chapter have been used as i-f amplifiers in superheterodyne pulse receivers. They have gains of about 100 db, bandwidths ranging from about 0.25 to 35 Mc/sec, and center frequencies from about 5 to 200 Mc/sec.

Among the desirable characteristics of pulse receivers are

1. High sensitivity, that is, the ability to detect weak signals.
2. Faithful pulse reproduction.
3. Rapid recovery.
4. Reasonable linearity between output and input signal levels.

The sensitivity of a pulse receiver for weak signals is influenced by many factors,[1] such as time and type of presentation, repetition rate, etc. The mportant properties of the receiver itself include low noise figure, adequate gain, and appropriate bandwidth. As shown in Chap. 4, the noise figure is determined almost entirely by the first one or two stages of the i-f amplifier. These stages will be assumed to be designed according to the principles discussed there.

6·2. Gain Requirements. *Combined I-f and Video Gain.*—In many applications the weakest usable signal has an amplitude about that of the residual noise[2] (thermal agitation, tube shot noise, etc.) in the receiving system. The gain required of the receiver is thus determined by the difference between the noise level presented to the input terminals of the first tube and the required output level.

An example will make this clear. Consider a microwave receiving system where in the present state of the art of r-f amplifiers, crystal mixers, and i-f amplifiers, the best over-all noise figure is attained by immediate conversion of the r-f signal to an i-f signal in a crystal mixer. For purposes of i-f noise level calculations, the crystal converter can be considered to be a resistor of about 300 ohms with a noise temperature close to 1. The crystal resistance is suitably transformed to a value yield-

[1] See Vol. 24, Radiation Laboratory Series.

[2] In certain applications, e.g., in cases requiring high fidelity of reproduction, the weakest usable signal is considerably above noise; in such cases the required gain is ess than described here.

ing optimum noise figure; the optimum value for the transformed resistance depends on center frequency, bandwidth, tube type, and tube configuration (see Chap. 4); for a grounded-cathode pentode at 30 Mc/sec it is about 2000 ohms.

The open-circuit thermal agitation noise voltage across this resistor is $\overline{e^2} = 4kTRB$; for $B = 1.6$ Mc/sec, $T = 292°$K, the rms voltage is $\bar{e} = 7$ μv. If the i-f noise figure is 3 db, the equivalent noise level at the grid of the first tube is $\bar{e}_{\text{equiv}} = 10$ μv.

Suppose now that the required output signal amplitude is 30 volts as might be the case for PPI presentation. Then the required over-all amplifier gain, i-f and video gains combined, from the grid of the first i-f amplifier tube to the grid of the cathode-ray tube, is 3×10^6, or 130 db.

This over-all amplifier gain varies, of course, with bandwidth, i-f noise figure, impedance level presented to the grid of the first tube, and required output level. It is further to be understood that the gain so calculated is a minimum gain and in the case of production amplifiers is to be secured even with low-g_m tubes.

I-f Amplifier Gain.—The question now arises as to the partition of the over-all gain between the i-f amplifier and the video amplifier. There are two reasons that favor assigning more gain to the video amplifier and less to the i-f amplifier;

1. For a given speed of pulse response, only half as much bandwidth is required in a video amplifier as in a centered i-f amplifier.
2. Higher gain-bandwidth products are practical in video stages than in i-f stages, because video circuits are less complicated than the equivalent i-f circuits.

It is nevertheless customary to apportion by far the larger part of the over-all gain to the i-f amplifier, for the following reasons:

1. High-gain video amplifiers are inferior to high-gain i-f amplifiers with regard to recovery time, transmission of long blocks of signals, and sensitivity to microphonics, hum, and power-supply variations.
2. The output vs. input curve for a second detector is linear above a certain level, say 1 volt, and square law below. For signals in the square-law range, 1 db of amplification preceding the second detector is equivalent to 2 db of amplification after the second detector; hence it is generally economical to provide enough gain in the i-f amplifier to bring the weakest desired signal up to the level where the second detector action is linear. In the example considered above this means an i-f gain of about 100 db.

In computing the gain of the i-f amplifier itself, account must be taken

of the usual second detector loss, which in the case of very wide-band video amplifiers may be as much as 10 db.

6·3. Bandwidth Considerations.—From the point of view of fidelity of pulse reproduction the greater the bandwidth the better. From the point of view of best signal-to-noise ratio there is, however, a certain optimum product of i-f bandwidth B and pulse duration τ,

$$B\tau \approx 1. \tag{1}$$

In this approximate result B may be taken as either the 3-db bandwidth or the noise bandwidth.

A detailed analysis of this point, on both theoretical and experimental grounds, is given in Vol. 24 of the Radiation Laboratory Series, but the following simple argument will show the plausibility of the assertion.

Imagine a system employing pulses of fixed duration and having an i-f amplifier whose bandwidth is allowed to increase from zero. The rms noise voltage is proportional to the square root of the bandwidth. At first, for small bandwidths, the peak amplitude of the reproduced signal increases linearly with the bandwidth (since the rise time of the leading edge of a step function is inversely proportional to bandwidth); hence, for small bandwidths, the ratio of peak signal to rms noise starts by increasing as the square root of the bandwidth. For large bandwidths, however, where the pulse response displays a flat top, further increase of bandwidth cannot increase the peak amplitude of the signal. Inasmuch as the rms noise level is still increasing as the square root of the bandwidth, the ratio of peak signal to rms noise then decreases as the square root of bandwidth.

Between the regions of increasing signal-to-noise ratio for small bandwidths and decreasing signal-to-noise ratio for large bandwidths, there is a certain optimum; as already stated[1] this occurs for $B\tau \approx 1$. Fortunately, this optimum bandwidth is extremely noncritical; thus a bandwidth either twice the optimum or half the optimum increases the minimum detectable signal by only about $\frac{1}{2}$ db.

Except for amplifiers of extremely large bandwidth, which are very difficult to obtain, it is common to make the i-f bandwidth closer to twice the reciprocal of the pulse length. This wider bandwidth costs little in minimum detectable signal and affords the advantages of greater pulse fidelity and reduced criticalness in LO tuning and automatic frequency

[1] A result of W. W. Hansen shows that under certain simplifying assumptions, for a given input signal wave form, the amplifier pass band maximizing the ratio of peak signal to rms noise is that which is identical with the frequency spectrum of the input signal waveform. For a 1 μsec rectangular pulse, for example, this means a $\sin x/x$ type of curve having its first zeros at ± 1 Mc/sec away from the band center.

control. Typical bandwidths are 0.5 Mc/sec for 5-μsec pulses, 1.7 Mc/sec for 1-μsec pulses, 35 Mc/sec for $\frac{1}{30}$-μsec pulses.

6·4. Choice of Intermediate Frequency.—The choice of intermediate frequency is an engineering compromise. In addition to the possibly compelling matters of availability of test equipment and existence of established standard frequencies, there are certain specific factors favoring a low intermediate frequency and others favoring a high intermediate frequency.

Before going into these questions, however, it is wise to dispose of two common, but incorrect, arguments. According to the first, which would favor high intermediate frequencies, it is easier to get a large bandwidth at a high frequency than at a low frequency. This is not the case, since the gain-bandwidth product for all the usual coupling networks is independent of frequency; that is, one gets no more gain in a 15-Mc/sec-wide amplifier at 60 Mc/sec than in an amplifier of the same bandwidth and type at 30 Mc/sec. It is true for double-tuned circuits, however, that small fractional bandwidths mean small coefficients of coupling and hence simpler coils, but the gain-bandwidth product is not affected unless the bandwidth is many times the center frequency.

The second incorrect argument, which would favor low intermediate frequencies, asserts that it is easier to get freedom from over-all regeneration at a low frequency. While it is true that one-stage feedback through grid-plate capacity is worse at high frequencies (see below under Miller effect), over-all regeneration is essentially independent of the center frequency and is dependent chiefly upon circuit decoupling and waveguide like feedback. It has proved possible to build very stable 200-Mc/sec amplifiers of high gain.

Arguments for a Low Intermediate Frequency.

1. Criticalness. For a given bandwidth the relative detuning caused by tube capacity variations is less for a low intermediate frequency. Hence amplifiers of a given bandwidth and type exhibit less variability because of manufacturing tolerances and tube replacement if they have a low center frequency.
2. Noise figure. The optimum noise figure attainable with a given tube type is lower at low frequencies. A working rule for optimum[1] noise figure F as a function of intermediate frequency f is $F - 1 \propto f$.

[1] That is, noise figure when the source resistance is so adjusted as to give lowest noise figure. Because the optimum source resistance decreases as frequency increases, the input circuit bandwidth of an amplifier adjusted for optimum noise figure is greater at the higher frequencies. Hence for bandwidths greater than that provided by the optimum source resistance, the relative noise figure advantage of a low intermediate frequency is not so great as it is for narrower bandwidths.

Thus noise figures of 1.06 (= 0.25 db) have been obtained with the two-triode combination of a grounded-cathode triode into a grounded-grid triode at 6 Mc/sec (optimum source resistance about 15,000 ohms, input circuit bandwidth about 2 Mc/sec, depending on circuit capacities and circuit type); 1.36 (= 1.35 db) at 30 Mc/sec (optimum source resistance about 3000 ohms, input circuit bandwidth about 12 Mc/sec); 3.4 (= 5.5 db) at 180 Mc/sec (optimum source resistance about 400 ohms, input circuit bandwidth about 40 Mc/sec).

3. Input loading and Miller effect. The input conductance of a tube due to cathode-lead inductance varies as the square of the frequency, and to the extent to which this loading is vexing there is an advantage in a low intermediate frequency. A similar advantage of a low intermediate frequency follows from the fact that the susceptance of the grid-plate capacitive feedback path increases with frequency.

Arguments for a High Intermediate Frequency.

1. Image rejection. The argument here is identical with that which has been well known in the design of communications receivers.
2. Small size. The higher the intermediate frequency the smaller are the tuning coils and the bypass condensers.
3. Separation of video frequencies from intermediate frequencies. For given video and i-f bandwidths, a high intermediate frequency reduces the possibility of trouble arising from transmission of i-f components of the rectified signal into the video amplifier. The importance of this point should not be exaggerated, however, for with a little care in decoupling it has proved not difficult to combine a 100-db i-f amplifier covering the range 20 to 35 Mc/sec between 3-db points with a video amplifier having its 3-db point at 15 Mc/sec and a gently sloping cutoff.
4. Reproduction of carrier-frequency waveform. In certain special applications, such as MTI (Chap. 21), the receiving system must reproduce the wave form of the modulation signals or pulses in a manner that depends as little as possible upon the chance phase relationship between the pulse and the intermediate frequency. This accuracy of reproduction is improved by the use of a high intermediate frequency.
5. Automatic-frequency control: local oscillator on wrong side of signal frequency. The majority of AFC schemes, although permitting the local oscillator to lock on either side of signal frequency, lead to large reduction in performance if the local oscillator is on the wrong side. In that case it may be possible to prevent wrong-

sideband locking by making the intermediate frequency so high that twice intermediate frequency exceeds the possible excursion in local oscillator electrical tuning. In the case of the 723A 10,000-Mc/sec local oscillator tube, for example, a 60-Mc/sec intermediate frequency makes wrong-sideband locking practically impossible, since very few tubes can be tuned over a range of 120 Mc/sec, while wrong-side locking was possible with a 30 Mc/sec intermediate frequency.

6. Automatic-frequency control: "harmonic" operation. When the local oscillator and signal differ in frequency by a half or a third, etc., of the intermediate frequency, there is always generated in the mixer of a superheterodyne a certain amount of i-f power; this harmonic power may be 20 or 25 db down from the power at the actual difference frequency; but if the AFC system has high gain, there is a danger of locking the local oscillator frequency to this spurious frequency, e.g., 15 Mc/sec away from signal frequency in a receiver having a 30-Mc/sec intermediate frequency. To a certain extent this possibility is less likely with a high intermediate frequency, because accidental slipping of LO frequency then has to be larger in magnitude before the local oscillator slips to a spurious frequency.

6·5. Synchronous Single-tuned Amplifiers.—The i-f amplifier types in most common use and the only types discussed in this chapter are the following:

1. Synchronous single-tuned (single-tuned stages tuned alike).
2. Staggered single-tuned.
3. Double-tuned.
4. Inverse-feedback.

Type 1 is characterized by a maximum of simplicity and noncriticalness but, because of its low efficiency, is not suited, with present tube types, to the construction of 100-db amplifiers of bandwidths larger than about 3 Mc/sec.

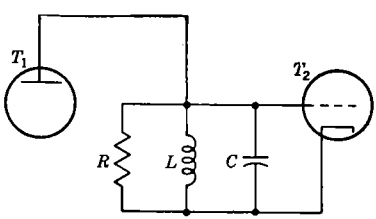


FIG. 6·1.—A-c diagram of single-tuned amplifier stage.

Types 2, 3, 4, which are roughly equivalent in efficiency, permit considerably larger bandwidths but require greater care in construction.

One Single-tuned Stage.—The a-c circuit diagram of one single-tuned stage is shown in Fig. 6·1, where C is the total interstage capacity ($= C_{out} + C_{in} + C_{stray}$), R is the damping resistor (= the plate-load

resistor in parallel with the plate resistance of Tube 1 and the input resistance of Tube 2), and L is the inductance resonating with C to the desired center frequency. The 3-db bandwidth[1] $\mathcal{B}$ is

$$\mathcal{B} = \frac{1}{2\pi RC}. \tag{2}$$

The voltage gain of the stage at midband is

$$\mathcal{G} = g_m R. \tag{3}$$

The product of gain and bandwidth[2] is thus $g_m/2\pi C$. For 6AK5 tubes with an assumed g_m of 4200 μmhos and C of 10.8 $\mu\mu$f, this gain-bandwidth product comes out to be 62 Mc/sec. Observe that bandwidth, gain, or gain-bandwidth product does not depend on the center frequency

$$f = \frac{1}{2\pi\sqrt{LC}}.$$

Cascaded Synchronous Single-tuned Stages.—As identical single-tuned stages are cascaded, the over-all bandwidth, measured between the 3-db points over all, goes down quite rapidly:

$$\text{Over-all bandwidth} = \text{bandwidth of one stage} \times \sqrt{2^{1/n} - 1}, \tag{4}$$

where n is the number of stages. Values of $\sqrt{2^{1/n} - 1}$ are given below:[3]

TABLE 6·1.—VALUES OF $\sqrt{2^{1/n} - 1}$

n	1	2	3	4	5	6	7	8	9
$\sqrt{2^{1/n} - 1}$	1	0.64	0.51	0.44	0.39	0.35	0.32	0.30	0.28

The product of gain *per stage* and bandwidth *over all* for cascaded synchronous single-tuned stages is thus $(g_m/2\pi C)\sqrt{2^{1/n} - 1}$; because this tends to zero as n increases, it turns out that for a given over-all gain and tube type the over-all bandwidth given by synchronous single-tuned stages first increases with the number of stages but later decreases as more stages are employed. For a 6AK5 amplifier of 100 db over-all gain, the maximum bandwidth attainable with synchronous single-tuned stages is about 6 Mc/sec. The number of stages needed for this bandwidth is 23. As a matter of fact it is impractical to build 100-db 6AK5 synchronous single-tuned amplifiers having bandwidths larger than about 3 Mc/sec.

[1] Derivation of this result and a complete mathematical treatment of single-tuned amplifiers will be found in Chap. 4 of Vol. 18 of the Radiation Laboratory Series.

[2] In the remainder of this chapter bandwidth always denotes bandwidth between the 3-db points.

[3] A useful approximation to $\sqrt{2^{1/n} - 1}$ is $1/1.2\sqrt{n}$.

Example. Design of 110-db I-f Amplifier about 2 Mc/Sec Wide at 30 Mc/Sec Employing 6AC7's Conservatively Rated.—To take account of tubes having g_m values at the low limit permitted by the JAN-1A specifications,[1] the g_m of the type 6AC7 is assumed to be 7000 μmhos. If the interstage capacity C is 25 $\mu\mu$f, the gain-bandwidth product $g_m/2\pi C$ for one stage is 44.5 Mc/sec.

By trial of several values it is concluded that six stages are needed. A gain of 110 db in six stages means 18.3 db per stage, or a voltage gain of 8.25 per stage. This yields a bandwidth per stage of $44.5/8.25 = 5.4$ Mc/sec. From Eq. (4) it follows that the over-all bandwidth is

$$5.4 \times 0.35 = 1.9 \text{ Mc/sec.}$$

The value of the damping resistor R is computed from the gain equation $g_m R = 8.25$; R turns out to be 1180 ohms. The design is completed by taking into account the shunting effect of the approximately 8000-ohm input resistance of 6AC7 tubes at 30 Mc/sec; this leads to a plate-resistor value of 1375 ohms. The next higher value in the preferred series, 1500 ohms, is chosen. The inductance of the tuning coils is, of course, selected to resonate with 25 $\mu\mu$f at 30 Mc/sec, namely, 1.13 μh. A complete circuit diagram of such an amplifier is shown in Fig. 13·6, and a photograph is shown in Fig. 13·7. In this amplifier the coils are tunable.

6·6. Stagger-tuned Amplifiers.—In stagger-tuned amplifiers the individual stages are single-tuned but are tuned to different frequencies and have different bandwidths.

The purpose of stagger-tuning is to reduce the shrinking of over-all bandwidth that occurs when cascaded single-tuned stages are tuned to the same frequency.

The selectivity curves achievable with stagger-tuning depend on the number of stages, their individual resonant frequencies, and their individual bandwidths. An especially simple family of selectivity curves is the "maximally flat" family of curves having selectivity functions of the form $1/\sqrt{1 + x^{2n}}$, where x is (for the case of small fractional bandwidth) proportional to frequency off resonance. Curves of this sort, corresponding to what are called flat-staggered n-uples are shown in the upper half of Fig. 6·2; corresponding step-function responses are displayed in the lower half of Fig. 6·2, which makes clear that[2]

[1] Under the JAN-1A test specifications the type 6AC7 is permitted to have a g_m varying between 7000 and 12,500 μmhos. These specifications include the current and hence g_m stabilizing effect of a 160-ohm cathode resistor; for the 50-ohm value of cathode resistor employed in the circuit of Fig. 13·6 the fractional variability in g_m is even larger.

[2] A discussion of transient response considerations will be found in Chap. 7 of Vol. 18 of the Radiation Laboratory Series.

1. Overshoot increases with squareness of pass band.
2. Speed of rise (between 10 and 90 per cent points) decreases somewhat with squareness of pass band.

The important feature of a flat-staggered n-uple is that its product of stage gain and over-all bandwidth is[1] $g_m/2\pi C$; in other words for a *stage*

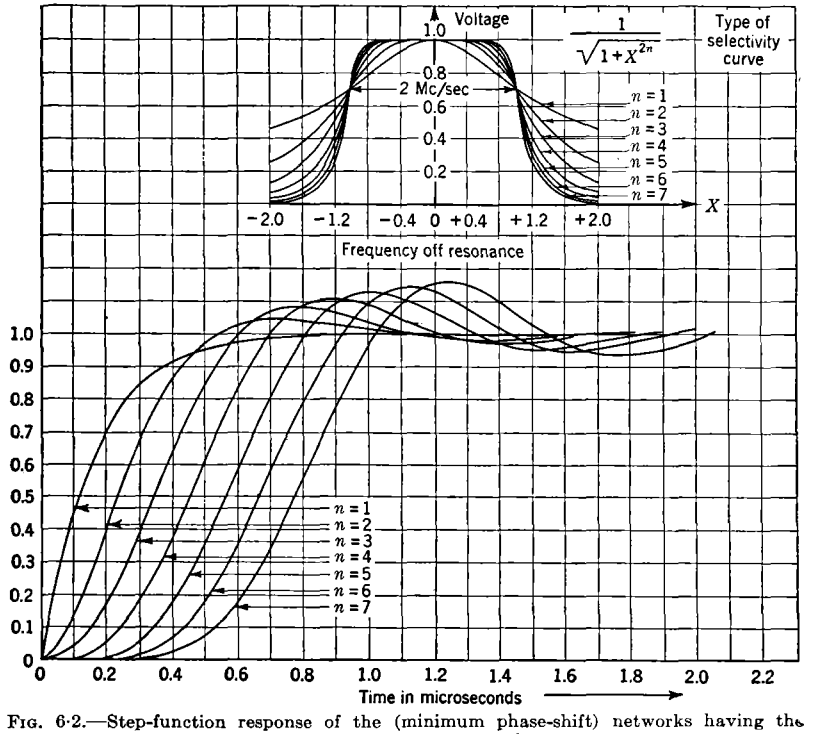


FIG. 6·2.—Step-function response of the (minimum phase-shift) networks having the absolute value curves shown above.

gain $\mathcal{G}$ the *over-all* bandwidth is the bandwidth of a single single-tuned stage of gain $\mathcal{G}$. This contrasts to the case of n synchronous single-tuned stages, where [see Eq. (4)] for a stage gain $\mathcal{G}$ the over-all bandwidth is $\sqrt{2^{1/n} - 1}$ times the bandwidth of a single single-tuned stage of gain $\mathcal{G}$. The elimination of the over-all bandwidth shrinking factor $\sqrt{2^{1/n} - 1}$ is the significant advantage of flat-staggered n-uples; the maximally flat character of the pass band is more likely to be a disadvantage than an advantage.

[1] The proof of this result and an extended treatment of stagger-tuning is contained in Chap. 4 of Vol. 18 of the Radiation Laboratory Series, *Vacuum Tube Amplifiers*.

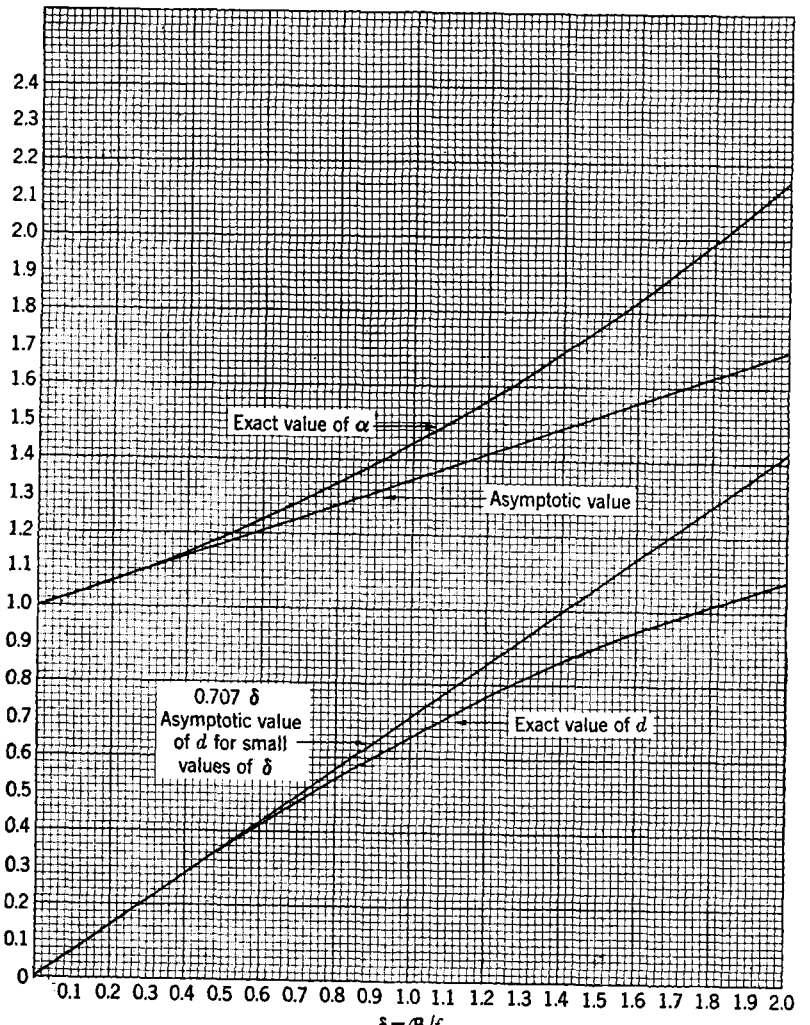


FIG. 6·3.—Exact flat-staggered pair. Note: (1) An exact flat-staggered pair of stage gain $\mathcal{G}$ has as great an over-all bandwidth $\mathcal{B}$ as does one single-tuned stage of gain $\mathcal{G}$, i.e., $\mathcal{G}\mathcal{B} = g_m/2\pi C$. (2) An exact flat-staggered pair of over-all bandwidth $\mathcal{B}$, geometrically centered at f_0, consists of two single-tuned stages staggered at $f_0\alpha$ and f_0/α of dissipation factor d, α being given in the upper graph and d in the lower graph as functions of $\delta = \mathcal{B}/f_0$. (3) An amplifier made up of n pairs has $1/1.1\sqrt[4]{n}$ the bandwidth of one pair.

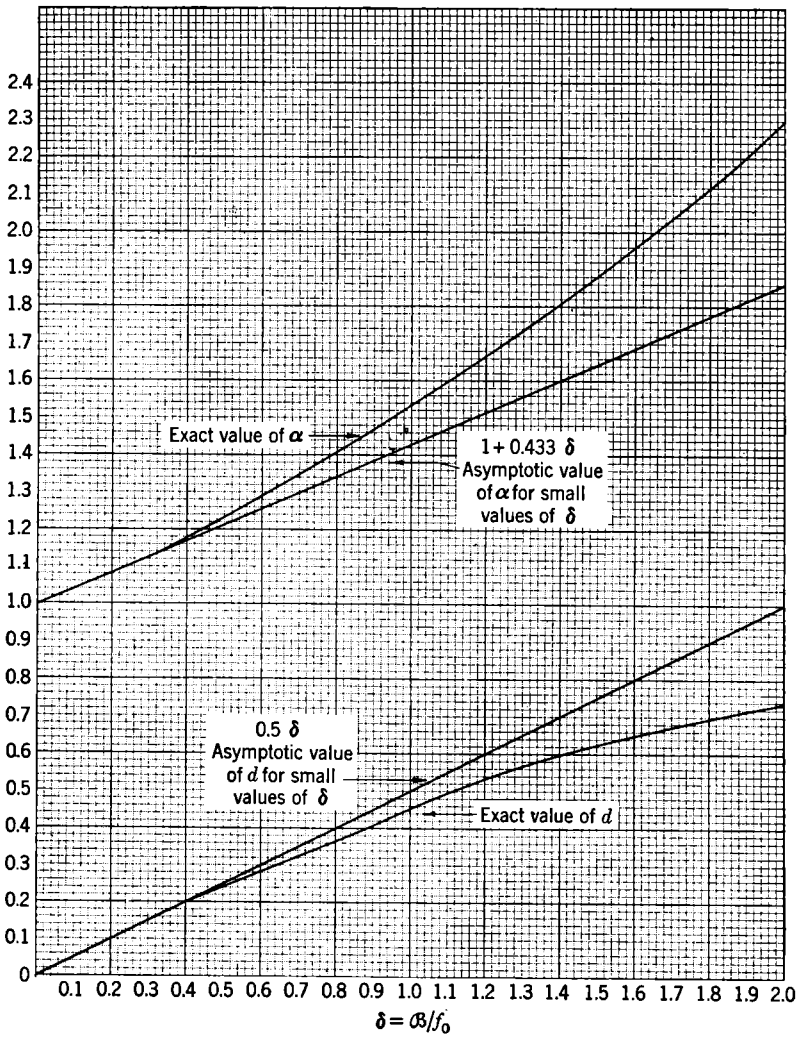


FIG. 6·4.—Exact flat-staggered triple. Note: (1) An exact flat-staggered triple of stage gain $\mathcal{G}$ has as great an over-all bandwidth $\mathfrak{B}$ as does one single-tuned stage of gain $\mathcal{G}$, i.e., $\mathcal{G}\mathfrak{B} = g_m/2\pi C$. (2) An exact flat-staggered triple of over-all bandwidth $\mathfrak{B}$, geometrically centered at f_0, consists of two single-tuned stages staggered at $f_0\alpha$ and f_0/α of dissipation factor d and one single-tuned stage centered at f_0 of bandwidth $\mathfrak{B}$, α being given in the upper graph and d in the lower graph as functions of $\delta = \mathfrak{B}/f_0$. (3) An amplifier made up of n triples has $1/1.06\sqrt[6]{n}$ the bandwidth of one triple.

A 100-db 35-Mc/sec-wide amplifier, at 80 Mc/sec, was constructed at the Radiation Laboratory consisting of three flat quintuples. This rather represented a tour de force, however, and for most purposes it is enough to use only staggered pairs or staggered triples.

The design data for flat-staggered pairs and flat-staggered triples are shown in the graphs of Fig. 6·3 and 6·4, in which the abscissa is the fractional bandwidth δ, or the ratio of the bandwidth to the center frequency for the pair or triple. These graphs are exact in the sense that they hold for fractional bandwidths of arbitrary magnitude, even larger than unity.

The use of these graphs is explained in the figure legends and is made clear by the following example. Suppose a flat-staggered pair of 8-Mc/sec bandwidth is to be designed with band center at 10 Mc/sec. Then $f_0 = 10$ Mc/sec, $\mathfrak{B} = 8$ Mc/sec, and $\delta = 0.8$ so that from Fig. 6·3 one finds that $\alpha = 1.33$ and $d = 0.535$. Therefore the pair is to be constructed of one stage centered at $10 \times 1.33 = 13.3$ Mc/sec, of dissipation factor[1] 0.535 and hence of bandwidth $0.535 \times 13.3 = 7.1$ Mc/sec, and one stage centered at $10/1.33 = 7.5$ Mc/sec, of dissipation factor 0.535 and hence of bandwidth $0.535 \times 7.5 = 4.0$ Mc/sec. If the $g_m/2\pi C$ ratio for a certain tube type were 48 Mc/sec, then for this flat-staggered pair of over-all bandwidth 8 Mc/sec, the stage voltage gain would be 6 ($= \frac{48}{8}$). For a pair of synchronous single-tuned stages having an over-all 8-Mc/sec bandwidth, on the other hand, the stage bandwidth would have to be 12.4 Mc/sec ($= 8/0.644$, see Table 6·1), and the stage gain would therefore be only 3.9 ($= 48/12.4$). An amplifier employing two flat-staggered triples is discussed in Sec. 16·8, and a circuit diagram is shown in Fig. 16.9.

Figure 6·5 shows a few stages of a 12-stage 100-db 6AK5 amplifier, 20 Mc/sec wide at 200 Mc/sec, consisting of four flat-staggered triples.[2]

Flat-staggered Pairs.—A flat-staggered pair has the same absolute value[3] and therefore[4] phase and transient response as a transitionally coupled double-tuned circuit.

The over-all bandwidth of m cascaded flat pairs is $\sqrt[4]{2^{1/m} - 1}$ times the bandwidth of one pair; the expression $\sqrt[4]{2^{1/m} - 1}$ is well approximated by $1/1.1\sqrt[4]{m}$.

The superiority in stage gain times over-all bandwidth product of an n-stage amplifier, n even, made up of $n/2$ flat-staggered pairs instead of

[1] Dissipation factor of a single-tuned circuit is the reciprocal of Q;

$$d = \frac{1}{\omega_0 RC} = \frac{\Delta f}{f_0}.$$

[2] This amplifier is discussed in detail in Sec. 5·11 of Vol. 18 of the Radiation Laboratory Series.

[3] At least for small fractional bandwidths.

[4] As a minimum phase-shift network.

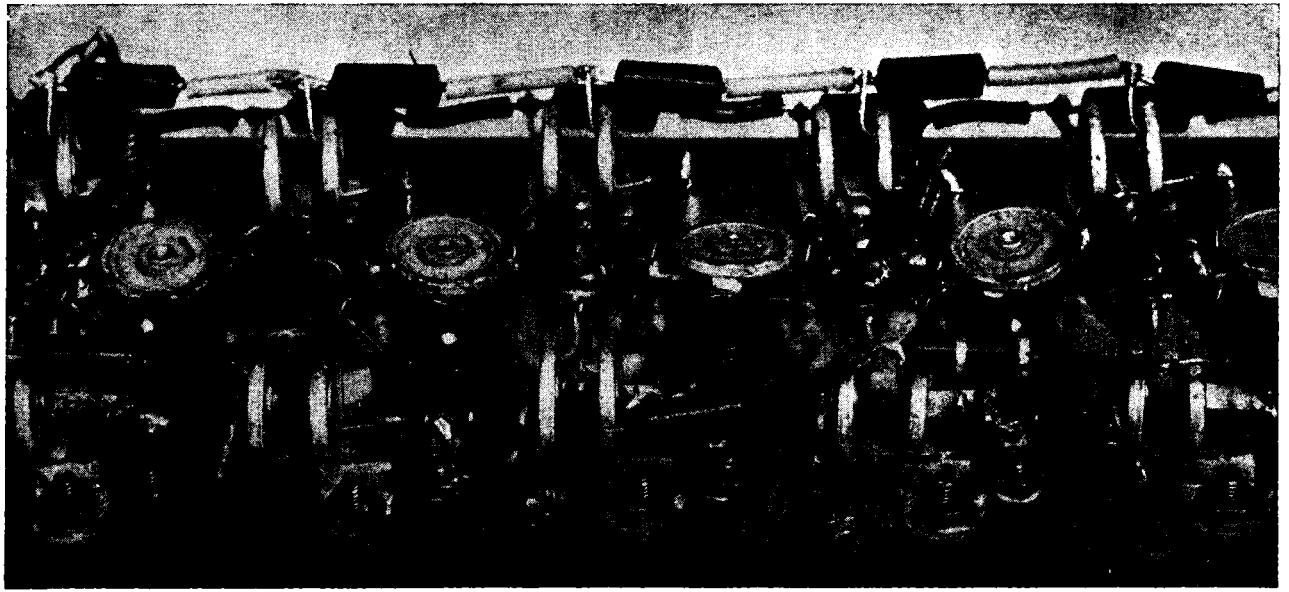


FIG. 6·5.—Bottom view of a few stages of a 12-stage 100-db 6AK5 amplifier made up of four flat-staggered triples, 20 Mc/sec wide at 200 Mc/sec.

n synchronous single-tuned stages is

$$\frac{1.2\sqrt{n}}{1.1\sqrt[4]{\frac{n}{2}}}. \tag{5}$$

Expression (5) has the value 2.0 for $n = 6$; that is, the stage gain times over-all bandwidth product for a six-stage amplifier made up of three flat-staggered pairs is twice that for a six-stage synchronous single-tuned amplifier.

Flat-staggered Pairs of Small Fractional Bandwidth.—For flat-staggered pairs of small fractional bandwidth, $\delta = \mathcal{B}/f_0$ less than 0.3, say, the pair consists quite accurately of two single-tuned stages staggered at $f_0(1 - 0.35\delta)$ and $f_0/(1 - 0.35\delta)$, of dissipation factor 0.71δ.

Flat-staggered Triples.—A flat-staggered triple has the same absolute value[1] and therefore phase[2] and transient response as a transitionally coupled triple-tuned circuit.

The over-all bandwidth of m cascaded flat-staggered triples is

$$\sqrt[6]{2^{1/m} - 1}$$

times the bandwidth of one triple; the expression $\sqrt[6]{2^{1/m} - 1}$ is well approximated by $1/1.06\sqrt[6]{m}$. The superiority in stage gain times over-all bandwidth of an n-stage amplifier, n divisible by 3, made up of $n/3$ flat-staggered triples instead of n synchronous single-tuned stages is

$$\frac{1.2\sqrt{n}}{1.06\sqrt[6]{\frac{n}{3}}} \tag{6}$$

Expression (6) has the value 2.5 for $n = 6$; that is, the stage gain times over-all bandwidth product for a six-stage amplifier made up of two flat-staggered triples is 2.5 times that for a six-stage synchronous single-tuned amplifier.

Flat-staggered Triples of Small Fractional Bandwidth.—For flat-staggered triples of small fractional bandwidth, $\delta = \mathcal{B}/f_0$ less than 0.3, say, the triple consists quite accurately of

1. Two single-tuned stages that are staggered at $f_0(1 - 0.43\delta)$ and $f_0/(1 - 0.43\delta)$ of dissipation factor 0.5δ.
2. One single-tuned stage centered at f_0 of bandwidth $\mathcal{B}$.

Gain Control.—The misapprehension sometimes exists that gain control can be applied to a stagger-tuned amplifier only if the high-

[1] At least for small fractional bandwidth.

[2] As a minimum phase-shift network.

frequency and low-frequency stages of the staggered pairs are controlled together. Although this would be the case if the selectivity curves of the individual stages were combined by addition, in stagger-tuning the selectivity curves are combined by multiplication, and it is therefore perfectly satisfactory to apply arbitrary gain control to any individual stage or combination of stages.[1]

6·7. Double-tuned Circuits.—The usual wide-band double-tuned circuit is shown in Fig. 6·6, where M denotes mutual inductance, and the

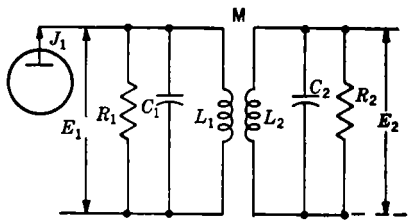


FIG. 6·6.—Inductance-coupled double-tuned circuit.

coefficient of coupling k is defined by $k = M/\sqrt{L_1L_2}$. The primary is assumed to be driven by a current source, and the voltage across the secondary terminals is the output signal. Capacity rather than inductance coupling between primary and secondary is inadvisable in the case of large fractional bandwidth, which is just the case where it would be most convenient, for the reason that capacity coupling seriously degrades the gain-bandwidth product by essentially adding to the primary a

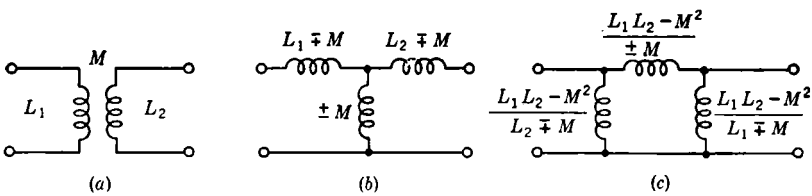


FIG. 6·7.—Transformer equivalent networks.

capacitance equal to the series combination of the coupling and primary capacitance and adding to the secondary a capacitance equal to the series combination of the coupling and secondary capacitances.

The transformer of Fig. 6.6 is sometimes replaced by the T or Π shown in Fig. 6·7; these T or Π self-inductance coupled networks are exactly equivalent to the transformer except for the d-c connection. If either $(L_1 - M)$ or $(L_2 - M)$ is negative in Fig. 6·6, however, neither network of Fig. 6·7 is physically achievable, in that negative inductances would be required.

[1] Subject, however, to the (generally unimportant) changes in input capacity and input loading attending grid-bias variation.

In the case in which the desired bandwidth is a moderately small fraction, less than $\frac{1}{5}$, say, of the center frequency f_0, the treatment of the double-tuned circuit is very much simplified by the assumption that

$$L_1C_1 = L_2C_2 = 1/4\pi^2 f_0^2,$$

that is, that the primary and secondary are tuned to the same frequency and that that frequency is band center. (As will be seen below, these approximations do not hold in the case of large fractional bandwidth.) This case will be discussed first.

Double-tuned Circuits of Small Fractional Bandwidth.—Denote by Q_1 the primary Q, namely $2\pi f_0 R_1 C_1$ and by Q_2 the secondary Q, namely $2\pi f_0 R_2 C_2$. If $Q_1 = Q_2$, the behavior of the double-tuned circuit as a function of increasing coefficient of coupling is well-known: The gain at midband increases, and the pass band becomes wider and squarer until the condition known as *critical* coupling is reached, where the midband gain is a maximum and the pass band is maximally flat, just failing to have humps. Larger coefficients of coupling lead to pass bands having humps, the height of the humps remaining constant, however, while the gain at the dip (band center) decreases.

The unequal-Q case is much less well known, however. A distinction must be made between *transitional* coupling, which means the coupling yielding a maximally flat pass band, and *critical* coupling, which means the coupling yielding maximum secondary voltage. (Critical coupling is also the condition of impedance match.) For a given Q_1/Q_2 ratio, $Q_1 \neq Q_2$, as the coefficient of coupling is increased from zero, the voltage gain at midband increases, and the pass band becomes wider and squarer, until maximum secondary voltage is reached at critical coupling:

$$k_{\text{crit}} = \text{geometric mean}\left(\frac{1}{Q_1}, \frac{1}{Q_2}\right) = \sqrt{\frac{1}{Q_1 Q_2}},$$

but the pass band is not flat. As the coefficient of coupling increases beyond critical, the midband secondary voltage starts to decrease and the pass band continues to become wider and flatter until the maximally flat pass band is reached at transitional coupling:

$$k_{\text{trans}} = \text{rms}\left(\frac{1}{Q_1}, \frac{1}{Q_2}\right) = \sqrt{\frac{1}{2}\left(\frac{1}{Q_1^2} + \frac{1}{Q_2^2}\right)}.$$

For coefficients of coupling larger than transitional the pass band shows humps whose heights do not stay constant with increasing k but decrease.[1]

[1] Graphs of this behavior, as well as a more extended treatment of double-tuned circuits in general, will be found in Chap. 5, Vol. 18, Radiation Laboratory Series.

The gain-bandwidth product at transitional coupling is larger than it is at critical coupling, although the gain itself is smaller at transitional coupling than it is at critical coupling.

For the sake of simplicity the only double-tuned circuit condition considered in the rest of this chapter is that of transitional coupling, despite the fact that from the point of view of transient response there is nothing the least bit special or noteworthy about transitional coupling.[1]

At transitional coupling the bandwidth $\mathcal{B}$ between 3-db points of a double-tuned stage is

$$\mathcal{B} = \frac{1}{2\pi}\frac{1}{\sqrt{2}}\left(\frac{1}{R_1C_1} + \frac{1}{R_2C_2}\right).$$

The voltage gain is

$$\mathcal{G} = g_m \frac{\sqrt{2}}{\sqrt{C_1C_2}} \frac{\sqrt{\left(\frac{1}{R_1^2C_1^2}\right) + \left(\frac{1}{R_2^2C_2^2}\right)}}{\left(\frac{1}{R_1C_1} + \frac{1}{R_2C_2}\right)^2}.$$

Special Cases.—In the equal-Q case

$$k = \frac{1}{Q_1} = \frac{1}{Q_2} = \frac{1}{\sqrt{2}}\frac{\mathcal{B}}{f_0},$$

$$\mathcal{B} = \frac{\sqrt{2}}{R_1C_1} = \frac{\sqrt{2}}{R_2C_2},$$

$$\text{Gain} = g_m \frac{R_1}{2}\sqrt{\frac{C_1}{C_2}}.$$

In the case in which $Q_2 = \infty$,

$$k = \frac{1}{Q_1\sqrt{2}} = \frac{\mathcal{B}}{f_0},$$

$$\mathcal{B} = \frac{1}{\sqrt{2}\,R_1C_1},$$

$$\text{Gain} = g_m\sqrt{2}\,R_1\sqrt{\frac{C_1}{C_2}}.$$

The case in which $Q_1 = \infty$ is completely analogous.

Comparison of Transitionally Coupled Double-tuned Circuit and a Single-tuned Circuit.—For a transitionally coupled double-tuned circuit with equal Q's the ratio of gain-bandwidth product to that of a single-tuned circuit is

$$\sqrt{2} \times \frac{\text{arithmetic mean } (C_1,C_2)}{\text{geometric mean } (C_1,C_2)},$$

[1] For graphs of the overshoot of a double-tuned circuit as a function of the coefficient of coupling see Chap. 7 of Vol. 18, Radiation Laboratory Series.

the interstage capacity C in the single-tuned case being given by

$$C = C_1 + C_2.$$

The corresponding ratio for a transitionally coupled double-tuned circuit loaded on one side only is

$$2 \times \frac{\text{arithmetic mean } (C_1,C_2)}{\text{geometric mean } (C_1,C_2)}.$$

With $C_1 = \frac{1}{2}C_2$, as is approximately the case for the types 6AK5 and 6AC7, the ratio

$$\frac{\text{Arithmetic mean } (C_1,C_2)}{\text{Geometric mean } (C_1,C_2)}$$

is 1.06, or 0.5 db, which is usually negligible. As a matter of fact the advantage due to the inequality of C_1 and C_2 can, if desired, also be obtained in single-tuned circuits by regarding the tuning coil as an autotransformer tuned with capacity $2C_1$ and stepped down in the impedance ratio C_2/C_1.

The main advantage in gain-bandwidth product of a transitionally coupled double-tuned circuit over a single-tuned circuit is the factor $\sqrt{2}$ (3db) in the equal-Q case and 2 (6db) in the one-side-loaded case. It is important to keep in mind that the 3-db further advantage of the one-side-loaded circuit over the equal-Q circuit comes at the expense of a very much greater criticalness to coefficient of coupling and capacity variability.

Cascaded Transitionally Coupled Double-tuned Circuits.—When n synchronous double-tuned transitionally coupled stages are cascaded, each of the same bandwidth (and without regard to the distribution of loading in each stage), the over-all bandwidth is $\sqrt[4]{2^{1/n} - 1}$ times the bandwidth of one stage. This is the same relation as that holding when m flat-staggered pairs are cascaded, for the reason that the absolute value curve of a transitionally coupled double-tuned circuit has the same shape[1] as that of a flat-staggered pair.

Double-tuned Circuits of Large Fractional Bandwidth.—The analysis required for this case is very complex and has been worked out only for the case of transitional coupling.[2] In order to get a pass band that is maximally flat at a frequency f_0, that is, whose first three derivatives at f_0 are zero, it turns out to be necessary to tune the primary and secondary windings to frequencies less than f_0; furthermore the primary and secondary frequencies are equal only if the primary and secondary Q's are equal.

[1] For small fractional bandwidth.

[2] See Chap. 5 of Vol. 18, Radiation Laboratory Series, for further references.

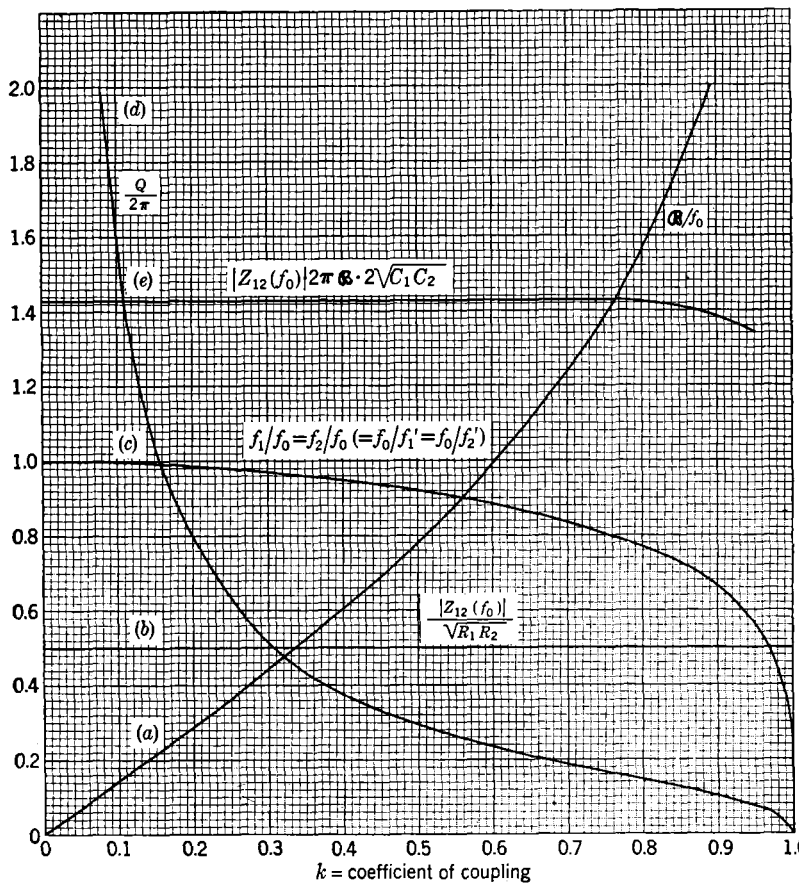


FIG. 6·8.—Low-Q transitionally coupled double-tuned circuit design data. k = coefficient of coupling; $Q_1 = Q_2 = Q$. (*a*) Fractional bandwidth: $\mathfrak{B}/f_0$; (*b*) normalized midband gain; (*c*) primary and secondary resonant frequencies; $f_1/f_0 = f_2/f_0 = f_0/f'_1 = f_0/f'_2$; (*d*) $Q/2\pi = f_0R_1C_1 = f_0R_2C_2$; (*e*) gain-bandwidth factor $|Z_{12}(f_0)|2\pi\mathfrak{B}2\sqrt{C_1C_2}$.

The design data for a transitionally coupled double-tuned circuit are shown[1] in Fig. 6·8 for the equal-Q and in Fig. 6·9 for the case $Q_1 = \infty$. In these curves the abscissa k is the coefficient of coupling, and the other symbols are as follows:

f_0 = band center, that is, the frequency at which the first three derivatives of the selectivity curve are zero,

[1] These graphs and the analysis upon which they are based are due to C. P. Gadsden.

$\mathfrak{B}$ = the 3-db bandwidth,[1]
f_1 = the primary frequency $1/(2\pi\sqrt{L_1C_1})$,
f_2 = the secondary frequency $1/(2\pi\sqrt{L_2C_2})$,
f_1' = the frequency of the primary with the secondary short-circuited,
f_2' = the frequency of the secondary with the primary short-circuited,
Q = the Q of either the primary or the secondary in Fig. 6·8, of the secondary in Fig. 6·9, but defined in either case at f_0; hence $Q/2\pi = f_0R_1C_1 = f_0R_2C_2$ in Fig. 6·8; $Q/2\pi = f_0R_2C_2$ in Fig. 6·9.

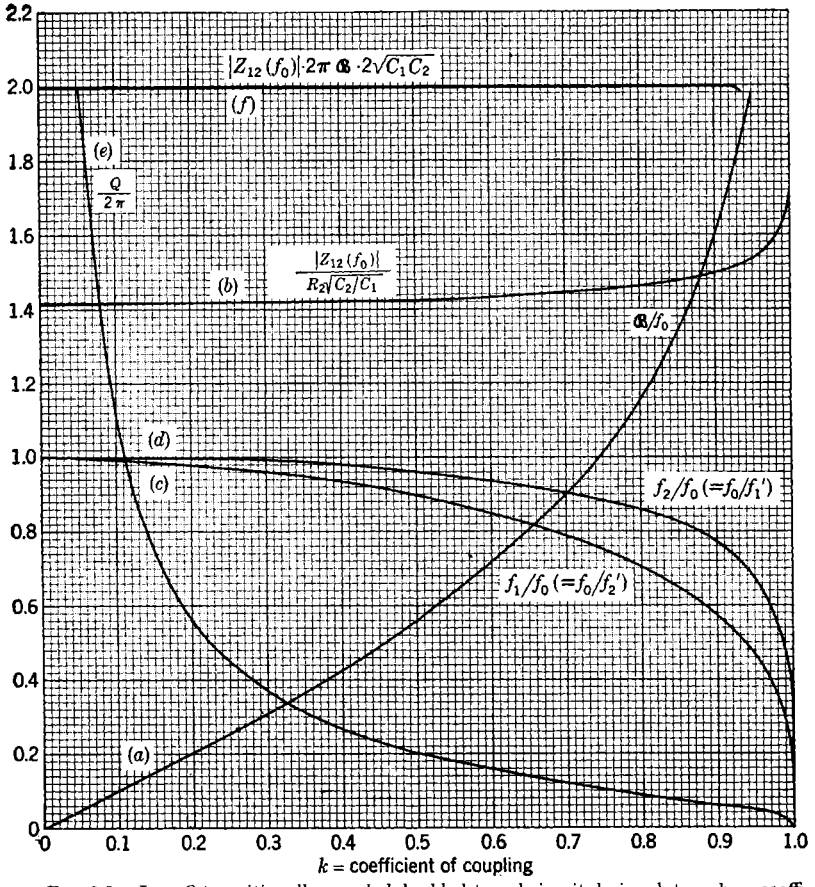


FIG. 6·9.—Low-Q transitionally coupled doubled-tuned circuit design data. k = coefficient of coupling; $Q_1 = \infty$, $Q_2 = Q$. (*a*) Fractional bandwidth: $\mathfrak{B}/f_0$; (*b*) normalized midband gain; (*c*) primary resonance frequency; (*d*) secondary resonance frequency; (*e*) $Q/2\pi = f_0R_2C_2$; (*f*) gain-bandwidth factor $|Z_{12}(f_0)|2\pi\mathfrak{B}2\sqrt{C_1C_2}$.

[1] As it turns out, f_0 is neither the geometric mean nor the arithmetic mean of the two 3-db points, but something in between.

As an example of the use of these graphs consider the design of a transitionally coupled double-tuned circuit loaded on the secondary only and having a 16-Mc/sec bandwidth centered at 30 Mc/sec. Because this case involves loading on one side only, the appropriate curves are those of Fig. 6·9. Here $\mathfrak{B} = 16$ Mc/sec, $f_0 = 30$ Mc/sec, so that $\mathfrak{B}/f_0 = 0.53$. From (*a*) of Fig. 6·9 it follows that the value of k is 0.48; this can be achieved by use of a powdered-iron core. The secondary $Q/2\pi$ is seen from (*e*) of Fig. 6·9 to be 0.21; from $Q/2\pi = f_0 R_2 C_2 = 0.21$ it follows that $R_2 = 1000$ ohms for $C_2 = 7.0$ $\mu\mu$f. Curve (*c*) shows that the open-circuit primary frequency is $30 \times 0.91 = 27.3$ Mc/sec, and Curve (*d*) shows that the open-circuit secondary frequency is $30 \times 0.97 = 29.1$ Mc/sec. From Curve (*d*) it follows that the primary frequency with the secondary short-circuited is $30/0.97 = 31$ Mc/sec, and from Curve (*c*) that the secondary frequency with the primary short-circuited is $30/0.91 = 33$ Mc/sec.

If it had been decided to leave the secondary unloaded, then it would have been the primary whose value of $Q/2\pi$ was 0.21; hence R would have had the value 1670 ohms for a C_1 of 4.2 $\mu\mu$f. The primary and secondary frequencies above would have been interchanged.

6·8. Inverse Feedback.—"Inverse feedback" as here applied to i-f amplifiers refers to a very special and simple feedback type only, namely, that effected by connecting a resistor between plate and grid of amplifying tubes that are coupled by two-terminal impedances. This is shown in Fig. 6·10.

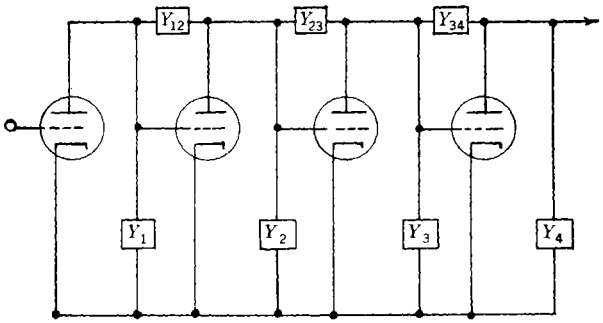


FIG. 6·10.—Feedback chain using two-terminal shunt impedances.

Although there is ideally an advantage of 2 in gain-bandwidth product over stagger-tuning to be achieved with feedback chains of the sort shown in Fig. 6·10, this advantage would require

1. Infinite stage gain.
2. An infinite filter terminating the feedback chain.
3. Infinite interstage impedance at midband.

When Conditions 1, 2, and 3 are not fulfilled, as is always the case, especially Point 1, in real amplifiers of fairly large bandwidth, the gain-bandwidth advantage over stagger-tuning rapidly decreases, and it becomes of little value to employ complicated terminating impedances. Not much is lost in performance by simplifying the feedback chain to one employing only single-tuned interstage circuits, and only this type is considered here.

It turns out[1] that the gain-bandwidth performance of feedback chains of any length employing single-tuned coupling elements can be completely duplicated by amplifiers employing feedback pairs, that is, pairs of stages of which only the second has feedback, as shown in Fig. 6·11.

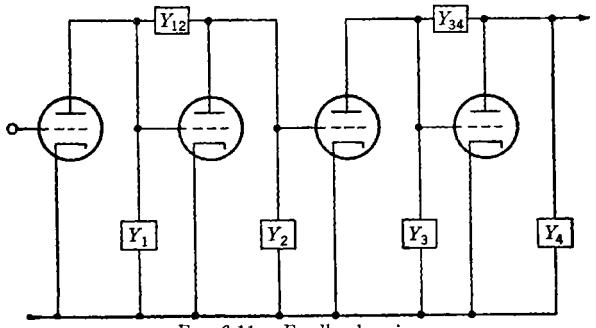


FIG. 6·11.—Feedback pairs.

An amplifier consisting of feedback pairs has the advantage that gain control can be applied to half the stages, namely, to the first stages of the pairs. A feedback chain, however, can be gain-controlled only in the first stage, because reducing the g_m of a stage around which feedback is connected has the effect of reducing bandwidth as well as gain. Consequently it seems wise to replace feedback chains employing single-tuned circuits by feedback pairs.[2]

The design of a feedback pair involves only the determination of the feedback resistance and the shunt resistances of the single-tuned interstage elements, since the stages are both tuned to the same frequency, namely, the desired midband frequency. There is a degree of freedom present, however, in regard to the distribution of the shunt loading

[1] See Chap. 6 of Vol. 18, Radiation Laboratory Series, for a general and extended discussion of plate-grid resistive feedback amplifiers.

[2] Sometimes feedback triples are employed; in that case feedback is connected around the second and third stages, so that one stage in three is available for gain control. The gain-bandwidth product of a feedback triple can be duplicated by the combination of a centered single-tuned stage and a suitable feedback pair (one with humps in its pass band).

between the two stages; but to make matters definite, it is here assumed that the two single-tuned circuits of the feedback pair are equally loaded.

Two sets of design data will now be given:

Flat Inverse-feedback Pair.—Let $\mathcal{B}$ be the bandwidth desired for the flat pair and g_m the transconductance of the tubes, assumed equal.[1] The feedback resistor R_{12} is

$$R_{12} = \frac{g_m}{(2\pi C\mathcal{B})^2},$$

and the resistor R loading each of the single-tuned circuits is determined by

$$\frac{1}{R} = \frac{2\pi C\mathcal{B}}{\sqrt{2}} - \frac{1}{R_{12}}.$$

Flat Triple Consisting of Single-tuned Stage and Feedback Pair.—Let $\mathcal{B}$ be the bandwidth desired for the flat triple and g_m the transconductance of the stages of the pair. Then the flat triple consists of a centered single-tuned stage of bandwidth $\mathcal{B}$ and a feedback pair having feedback resistor

$$R_{12} = \frac{4g_m}{3}\frac{1}{(2\pi C\mathcal{B})^2}$$

and resistor R loading each single-tuned circuit

$$\frac{1}{R} = \frac{2\pi C\mathcal{B}}{2} - \frac{1}{R_{12}}.$$

Example. Flat Feedback Pair of Bandwidth 10 *Mc/sec at* 30 *Mc/sec Employing* 6*AK*5*'s.*—Assuming $g_m = 5000$ μmhos, $C = 10.8$ $\mu\mu$f, the feedback resistor R_{12} between the plate of the second stage and the plate of the first stage or through a blocking condenser from the plate of the second stage to the grid of the second stage is 10,870 ohms, and the resistor loading the single-tuned circuits in the plates of the first and second stages is 2570 ohms.

Gain-bandwidth Product.—The gain-bandwidth product of an amplifier made up of feedback pairs is, except for the (generally negligible) reduction in effective transconductance of the second stage of the pairs, exactly equivalent to that of stagger-tuning.

[1] Actually the effective transconductance g'_{m2} of the second stage is less than g_{m2} by the conductance of the feedback resistor; the fractional reduction in effective transconductance is, however, less than the reciprocal of the over-all gain of the feedback pair. Taking into account the fact that the transconductance of the first stage is unaffected by the feedback, one sees that even if the stage gain is as low as 3 (corresponding to a bandwidth of 20 Mc/sec for a tube type having a $g_m/2\pi C$ of 60 Mc/sec) the gain of the pair is less than 1 db too high if one assumes $g'_{m2} = g_{m2}$, rather than the accurate value $g'_{m2} = g_{m2} - (1/R_{12})$.

Capacity across Feedback Resistor.—It is very important that the capacity across the feedback resistor be small, because its harmful effects are magnified, according to the Miller effect, by a factor equal to the voltage gain of the stage. The ordinary half-watt resistor has a capacity of about 0.3 to 0.5 $\mu\mu$f; this is much too large to be permitted except in very wide-band, low-gain cases. Although devices such as enclosing the feedback resistor in a grounded metal cylinder are of value, the only really satisfactory method is to use special resistors having extremely small end-to-end capacities. The IRC type MPN resistor, which has a long and extremely thin body, has an end-to-end capacity of about 0.02 $\mu\mu$f and has proved satisfactory.

6·9. Pros and Cons of Different Amplifier Schemes. *Gain-bandwidth Factor.*—Table 6·2 lists the bandwidths that can be attained with i-f amplifiers using 6AK5's for over-all gains of 80 and 100 db. In these calculations the type 6AK5 is fairly conservatively assumed to have a g_m of 4200 μmhos and an interstage capacity C of 10.8 $\mu\mu$f; this means a $g_m/(2\pi C)$ ratio of 62 Mc/sec. For other tube types or other modes of use of the type 6AK5, the bandwidths are multiplied by the ratio of the new $g_m/(2\pi C)$ to 62 Mc/sec.

The additional factor

$$\frac{\text{Arithmetic mean } (C_1,C_2)}{\text{Geometric mean } (C_1,C_2)}$$

favoring double-tuned circuits has been omitted from the double-tuned calculations for three reasons:

1. The factor is usually small: 1.06 for $C_1 = \frac{1}{2}C_2$, for example.
2. The factor is generally absorbed in the larger total interstage capacity occurring with double-tuned rather than single-tuned circuits; the larger total capacity is caused by the greater physical size of the double-tuned coil. In a 6AK5 amplifier, for example, the total interstage capacity is about 10 per cent larger in double-tuned stages than in single-tuned stages.
3. If desired, the advantage that it represents in gain-bandwidth product can be achieved with synchronous single-tuned or stagger-tuned amplifiers by regarding the tuning coils as unity-coupled autotransformers, in the way described in Sec. 6·7.

Table 6·2 also lists the overshoots accompanying the step-function responses of the various combinations.

Table 6·2 shows that synchronous single-tuned amplifiers are decidedly inferior in gain-bandwidth product to the other types. Flat-staggered pairs are slightly inferior in gain-bandwidth product to transitionally coupled equal-Q double-tuned circuits, and staggered triples are slightly

TABLE 6·2.—OVER-ALL 3-DB BANDWIDTHS OBTAINABLE WITH TUBES HAVING A $g_m/(2\pi C)$ RATIO OF 62 MC/SEC

Type of amplifier	Overshoot, %	Over-all gain = 80-db bandwidth, Mc/sec	Over-all gain = 100-db bandwidth, Mc/sec
Three stages:			
Three synchronous single-tuned	0	1.5	0.65
Flat-staggered triple	8.15	2.9	1.3
Three transitionally coupled equal-Q double-tuned	7.7	2.9	1.3
Three transitionally coupled one-side-loaded double-tuned	7.7	4.1	1.9
Four stages:			
Four synchronous single-tuned	0	2.7	1.5
Two flat-staggered pairs	6.25	5.0	2.8
Four transitionally coupled equal-Q double-tuned	8.4	5.8	3.3
Four transitionally coupled one-side-loaded double-tuned	8.4	8.1	4.7
Six stages:			
Six synchronous single-tuned	0	4.7	3.2
Three flat-staggered pairs	7.7	9.5	6.5
Two flat-staggered triples	11.2	11.5	7.8
Six transitionally coupled equal-Q double-tuned	10.0	11.2	7.7
Six transitionally coupled one-side-loaded double-tuned	10.0	15.8	10.8
Nine stages:			
Nine synchronous single-tuned	0	6.2	4.8
Three flat-staggered triples	13.0	17.8	13.8
Nine transitionally coupled equal-Q double-tuned	12.0	16.7	13.0
Nine transitionally coupled one-side-loaded double-tuned	12.0	23.6	18.4

superior. Transitionally coupled one-side-loaded double-tuned circuits are substantially superior, by a factor of $\sqrt{2}$, to transitionally coupled equal-Q double-tuned circuits.

Table 6·2 does not show inverse feedback pairs or triples for the reason that these are essentially identical in gain-bandwidth product with staggered pairs or triples (see Sec. 6·8).

Simplicity.—Undoubtedly the simplest type of amplifier is that employing synchronous single-tuned circuits. The tuning coils are as simple as possible and are identical from stage to stage.

Stagger-tuned and inverse-feedback amplifiers also employ very simple coils. Inverse-feedback amplifiers have the special requirement,

however, that for satisfactory results the feedback resistors must have very small end-to-end capacity, about ten times less than the usual half-watt resistor.

Double-tuned coils are considerably more complicated than single-tuned coils. Accurate control is required of the spacer controlling mutual inductance as well as the primary and secondary inductances. This greater coil complexity is in many cases the only significant disadvantage of double-tuned circuits.

Criticalness with Regard to Tube-capacity Variability.—About this very important matter adequate knowledge does not yet exist.[1] Many points have not been examined at all, while some points that have been considered are the subjects of opposing statements.

Some statements can be made, however:

1. For a given amplifier type the ill effects of capacity variability increase rather rapidly with increase in the ratio of fractional detuning[2] to fractional bandwidth; this ratio is the significant parameter.
2. Synchronous single-tuned amplifiers are very tolerant to tube-capacity variability. This fact together with the extreme simplicity and the good transient response of synchronous single-tuned amplifiers makes it wise to use that scheme[3] whenever it is not too uneconomical of tubes, even though its gain times band-width efficiency is poor.
3. Flat-staggered pairs are less critical than flat-staggered triples.
4. Flat inverse-feedback triples are less critical than flat-staggered triples.
5. One-side-loaded transitionally coupled double-tuned circuits are much more critical than equal-Q transitionally coupled double-tuned circuits. It is for this reason that use is not generally made of the $\sqrt{2}$ advantage in gain-bandwidth product of the one-side-loaded double-tuned circuit over the equal-Q double-tuned circuit.

Selectivity.—Synchronous single-tuned amplifiers have poor selectivity; flat-staggered or inverse feedback pairs have squarer pass bands; and flat-staggered or inverse-feedback triples are even squarer.

An amplifier of six stages made up of three flat-staggered pairs clearly has less remote-frequency rejection than a six-stage amplifier of the same

[1] See Sec. 7·12 of Vol. 18, Radiation Laboratory Series, for a somewhat fuller discussion.

[2] Fractional detuning is the square root of fractional capacity variability. For 6AK5's the JAN-1A specifications permit a capacity variability, input plus output, of ± 0.9 $\mu\mu$f. For a 10.8-$\mu\mu$f interstage capacity the fractional capacity variability is $\pm \frac{1}{12}$, and the fractional detuning is $\pm \frac{1}{24}$.

[3] Unless greater selectivity is needed.

over-all bandwidth made up of six transitionally coupled double-tuned stages.

Transient Response.—A useful principle to keep in mind is that in minimum-phase-shift networks the absolute value curve determines the phase curve, and hence the transient response is completely determined by the absolute value curve alone. All the networks discussed in this chapter are minimum-phase-shift networks; hence for all the i-f amplifiers described here the absolute value curve alone determines transient response.

Thus amplifiers having the same bandpass shape and bandwidth have the same transient response, whether they are made up of over-staggered pairs or overcoupled double-tuned circuits; flat-staggered pairs, flat inverse-feedback pairs, or equal-Q or one-side-loaded transitionally coupled double-tuned circuits; flat-staggered triples, transitionally coupled triple-tuned circuits, or combinations of an overcoupled double-tuned circuit and a single-tuned circuit.

Two important features of the transient response of i-f amplifiers are the rise time, between 10 and 90 per cent, of the envelope of the response to a step function of intermediate frequency, and the overshoot of the step-function response.

For all the usual types of i-f amplifiers there is the following approximate relation between rise time t as defined above and 3-db bandwidth $\mathcal{B}$,

$$t\mathcal{B} = 0.7 \text{ to } 0.85,$$

the values closer to 0.85 pertaining to amplifiers of very square pass band.

For synchronous single-tuned amplifiers

$$t\mathcal{B} = 0.7$$

very exactly, no matter how many stages are involved, and there is no overshoot.

In general the squarer the pass band the greater the overshoot. Hence larger overshoot is the price that must be paid for high selectivity.

The overshoot for m flat-staggered pairs (or m flat inverse-feedback pairs or m transitionally coupled double-tuned stages) is 4.3, 6.25, 7.7, 8.4, 10.0 per cent for $m = 1, 2, 3, 4, 6$ respectively.[1]

An amplifier of six stages made up of three flat-staggered pairs has smaller overshoot (7.7 per cent) than a six-stage amplifier made up of transitionally coupled double-tuned stages (10.0 per cent); the smaller overshoot is compensation for the smaller remote-frequency rejection.

[1] For graphs of these and other step-function responses, see Chap. 7 of Vol. 18, Radiation Laboratory Series.

The overshoot for m flat-staggered triples (or m flat inverse-feedback triples or m transitionally coupled triple-tuned stages) is 8.15, 11.2, 14.2 per cent for $m = 1, 2, 4$ respectively.

Gain Control and Gain Stability.—Except for inverse-feedback amplifiers, the individual stages of any of the amplifiers discussed here can be arbitrarily gain-controlled. Inverse-feedback amplifiers have the disadvantage that gain control cannot be applied to any stage around which there is inverse feedback, for the reason that reduction in g_m would also cause reduction in bandwidth. In feedback chains, therefore, only the first stage can be gain-controlled; for a long chain this may be a serious handicap. In amplifiers consisting of inverse-feedback pairs the matter is less serious, because every other stage can be gain-controlled. In amplifiers made up of triples consisting of an inverse-feedback pair and a centered stage, two out of three stages permit gain control.

Compensating somewhat for the smaller flexibility with regard to gain control is the greater gain stability of inverse-feedback amplifiers. The improvement in gain stability against g_m variations (due to tube replacement or heater-voltage variations, for example) is not very great and, although sometimes useful, is inadequate for use in calibrated instruments. The gain-stabilizing factor is always less than 2, even for long inverse-feedback chains,[1] and is less than $\frac{3}{2}$ for triples consisting of a feedback pair and a single-tuned stage. In order to achieve large gain-stabilizing factors, other means have to be employed.

6·10. Practical Considerations. *Safeguards against Regeneration. Over-all Feedback.*—Great attention to detail is needed to build a stable amplifier having 100-db or more gain and covering a wide band of frequencies. A positive feedback factor as small as 10^{-6}, anywhere in the band, is very serious. Every length of wire, no matter how small, must be regarded as both an inductance and a capacitance—a 1-in. length of No. 20 wire has a reactance of 8 ohms at 60 Mc/sec, for example, and an $\frac{1}{8}$-in. common coupling path in a high-gain amplifier may be more than enough to cause oscillation.

The important points to watch in guarding against over-all feedback are

1. Bad ground paths.
2. Waveguide feedback.
3. Inadequate bypassing, of heaters, $B+$, gain control, etc.

Point 1 is discussed in Sec. 10·7, and Point 2 in Sec. 10·6.

With regard to 3, it must be kept in mind that a bypass condenser is actually the series combination of its lead inductance and its capacitance.

[1] Sec. 6·5 of Vol. 18, Radiation Laboratory Series.

Because of poor layout or poor choice of type of condenser, it sometimes happens that a bypass condenser has a high reactance to ground and one that is rapidly varying over the band.

An especially effective means of bypassing involves the use of "series-resonant" bypass condensers, that is, condensers whose leads resonate with their capacitances at intermediate frequency. It is important to achieve this series resonance by means of a large C and a small L; otherwise the low impedance to ground will hold only over a small part of the band. With 2000-$\mu\mu$f condensers and a total lead length of only $\frac{1}{8}$ in. a bypass impedance to ground of less than $\frac{1}{2}$ ohm is maintained over the range 50 to 72 Mc/sec.

The Q of a bypass condenser is of little importance and may perfectly well be as low as 5 or 10.

A matter of practical importance is the necessity of grounding all coaxial leads connected to a high-gain amplifier by spreading the braid in circular fashion and grounding all around the perimeter.

Quite often regeneration is introduced into an amplifier by incorrect methods of attaching connections for measurements. The only safe way to make measurements is either to reduce the gain, preferably by removing tubes, or to make all connections very accurately coaxial. It pays very well to spend the time required to construct an adequate set of coaxial fittings, adaptors, and plugs.

One-stage Feedback.—This is of less common occurrence with wide-band amplifiers than over-all feedback but may occur at high intermediate frequencies if the stage gain is high, because of grid-plate capacity. A grid-to-plate voltage gain of about 20 is about as much as is comfortable with 6AK5's at 30 Mc/sec and less at 60 Mc/sec, according to the formula that oscillation can occur (if the plate circuit is suitably detuned from the grid circuit) for a stage gain greater than[1]

$$\sqrt{\frac{g_m}{\pi f C_{gp}}}.$$

Parasitic Oscillations.—Experience has shown danger of parasitic oscillations at about 500 or 600 Mc/sec in i-f amplifiers using 6AK5's when high-Q (silvered-mica button) bypass condensers are used. These parasitics are avoided by inserting 10-ohm carbon resistors in the paths between $B+$ and screen pins. Low-Q bypass condensers are also effective and are thus seen to be an advantage.

[1] With regard to the ratio g_m/C_{gp} the type 6AK5 is only about one-fourth as good as the type 6AC7; the g_m is only half, and the actual grid-plate capacity is about double, taking into account the larger capacity across the tube socket.

Means of Detecting Regeneration.—A variety of practical methods for determining and localizing regeneration are discussed elsewhere.[1]

Tube Type.—As is well known, for wide-band applications tubes of high g_m/C ratio are needed such as the type 6AC7 or type 6AK5. For the higher intermediate frequencies, above 30 Mc/sec, the type 6AK5 with its short cathode leads is preferable to the type 6AC7 because of its much smaller input conductance (about 2000 ohms for the type 6AC7 at 60 Mc/sec and 50,000 ohms for the type 6AK5). The type 6AK5 has a somewhat greater g_m/C ratio than the type 6AC7. Small-size tubes are, of course, needed for applications requiring great compactness.

The requirement for greater performance (greater bandwidths, for example) in i-f amplifiers will surely lead to even closer integration of the actions of the amplifying tube and the amplifying circuits, until an amplifier will be indistinguishably a tube and a circuit. A first step in this direction may be the development of "packaged" i-f amplifiers employing soldered-in baseless subminiature tubes together with circuits of high complexity and efficiency, adjusted at the factory to the particular tubes in use.

6·11. Overload Behavior of Amplifiers.—As a result of the wide dynamic range of input signals and the high gain of the i-f amplifier, the later stages of the i-f amplifier of a radar receiver are subjected to signals that are well above the level at which operation is linear. It therefore becomes important to study the behavior of an amplifier under such overload conditions. Effects of interest are found both during and following a strong signal and may be traced to factors involving circuit design as well as factors due to the internal behavior of the amplifier tubes.

Effects during the Signal.—If an adjustable c-w signal is applied to the grid circuit of a typical amplifier stage, the output of the stage will be found to increase linearly with the input as long as both are small. With larger signals, however, the output increases less rapidly and finally reaches a maximum. With still larger input, the output will usually decrease. For many applications the quantity of interest is the slope of the curve obtained by plotting input against output, since this *incremental gain* is a measure of the ability of the stage to reproduce a small variation in the input signal. A convenient method of measurement is to apply simultaneously a small pulse-modulated i-f signal of fixed magnitude and a variable c-w signal. Both input and output circuits of the stage under test are provided with diode detectors so arranged as to permit separation of the pulse and c-w components of the signals. If the pulse signal at the input circuit is maintained constant as the c-w input is varied, the magnitude of the output pulse will be a measure of the incremental gain, and the d-c output of the diode in the plate circuit will be a measure of

[1] Sec. 8·5 of Vol. 18, Radiation Laboratory Series.

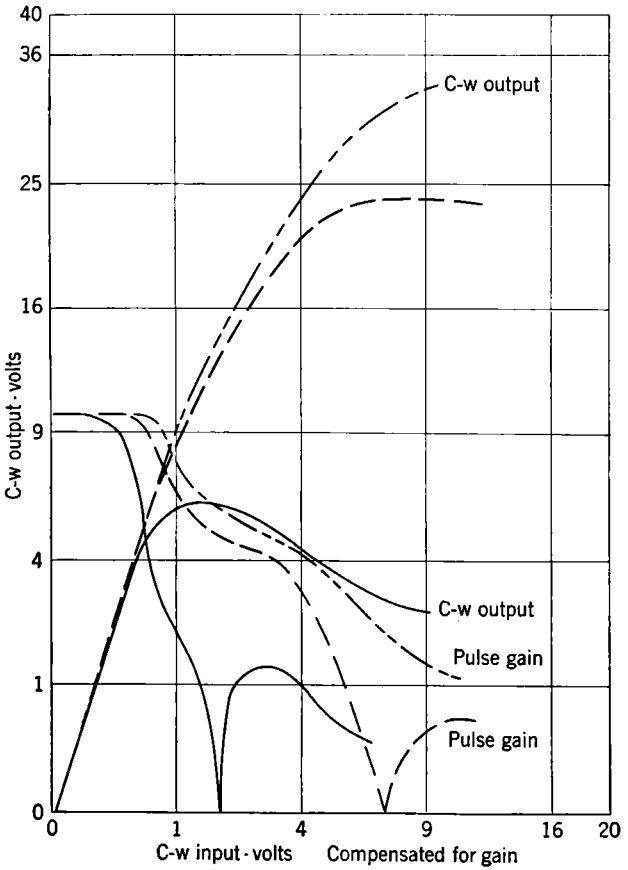


FIG. 6·12.—Overload characteristics of vacuum tubes.

	E_{sg}, volts	E_p, volts	Curve
6AC7	105	105	solid line
6AK5	105	105	dashed line
6AK5	150	150	dash-dot

the output of the stage. Results of such measurements are plotted[1] in Fig. 6·12. When the c-w input signal is greater than the value cor-

[1] In order to permit some compression of the larger values of ordinates and abscissas and at the same time to retain the zeros on both scales, the curves are plotted on coordinates in which the lengths of the ordinates and abscissas are proportional to the square root of the numerical magnitudes. Note that the geometrical slope of the input-vs.-output curve so plotted is no longer equal to the incremental gain.

responding to maximum output, the slope of the curve and hence the incremental gain become negative. The output pulse as viewed on an oscilloscope or synchroscope will be observed to become inverted in agreement with expectation. In Fig. 6·12, the absolute magnitude of pulse height or incremental gain has been plotted.

If the circuit is so arranged as to permit independent measurement of the d-c plate and screen-grid currents, it will be found that when the c-w signal reaches the point corresponding to zero incremental gain, the screen current rises sharply and the plate current may fall somewhat. This behavior is particularly marked in the case of tubes such as the 6AC7 or 6AG5 in which the plate is made up of two small rectangular sections instead of a cylinder completely enclosing the rest of the structure. The hypothesis may be advanced that this behavior is due to the action of the control-grid support rods.[1] As long as the control grid is negative, these rods act as an electron lens to focus the electrons in narrow beams that strike the plates. When the control grid becomes positive, however, the electron paths are such that many of them miss the plate and arrive instead at the screen. An increase in the plate voltage brings some of these stray electrons back to the plate and thereby increases the maximum output somewhat, as is observed experimentally. Another confirmation of this view may be found in the influence of a magnetic field. If a small permanent magnet is placed around a tube such as a 6AG5 in such a way that the magnetic lines are normal to the plates, the input signal for zero incremental gain may be increased by as much as 30 per cent. This action may be ascribed to the constraint on the electron paths causing more of them to be confined to the plates.

For the range of bandwidth or plate-load resistor of interest here, these overload effects set in at approximately a fixed plate current. The maximum output of a tube will therefore be approximately proportional to the load resistor or inversely proportional to the bandwidth. Tubes of different types having different plate-to-ground capacitances may be compared under conditions of equal bandwidth.

The performance of a multistage amplifier may be predicted by the usual graphical method in which the output of one stage is used as the input of the next. It is found that an important factor is the sign of the incremental gain at the point where the c-w output is equal to the c-w input. If the incremental gain is negative at this point, it is necessary to provide special protective circuits of the sort described in Chap. 9 to avoid undesirable overload effects. If, on the other hand, the incremental gain is still positive, as would be the case in an unusually wide-

[1] Chai Yeh, "The Effect of Grid-Support Wires on Focusing Cathode Emission," *Proc. IRE*, **34**, 444, July 1946.

band amplifier, the over-all performance without such protective circuits is relatively good.

Effects Following a Signal.—The usual manifestation of overload effects after a strong signal takes the form of a transient reduction of gain. The part that is due to the internal behavior of the tubes has sometimes been called the "Whippany effect" or "blackout effect." It appears to be closely analogous to the anode effect that has been studied by Copeland[1] and others. It is assumed that the surface of a metallic electrode such as the control grid of a tube may at times become partially covered with a very thin insulating layer. This layer might arise from traces of oxides, from material evaporated from the cathode, or from miscellaneous contamination. If an electron current is drawn to the electrode in question, some electrons may remain on the surface of the insulating layer for an appreciable time, of the order of microseconds. While these electrons are so trapped, they act to maintain the potential of the region near the electrode more negative than the normal value. Thus the tube is effectively biased off for a short time after the end of a signal strong enough to cause grid current to be drawn, and the gain is reduced accordingly. In the case of video amplifiers or plate detectors (See Sec. 7·7) the effect is much more serious, because the transient shift in apparent grid potential appears as a spurious signal instead of a change in gain. It is found that there is considerable variation from tube to tube, even in a single lot of supposedly identical manufacture. Otherwise good tubes that show this effect are, however, usually satisfactory in the early stages of the i-f amplifier where signals never reach a value sufficient to cause grid current.

The effects originating in the external circuit can usually be traced to transient shifts in electrode voltages produced by charges on bypass or coupling condensers. During the overload signal, the operation of the tube is nonlinear, and the average currents to the various electrodes, in particular the screen and control grids, may increase markedly. A somewhat more detailed consideration of the various circuits in which this effect may arise and of the effective time constants for its decay is given in Sec. 19·5. In i-f amplifiers, the ordinary way to avoid such difficulties is either to make the offending time constant so short that recovery is rapid enough to meet requirements or to make the time constant so long (in the radar case, long compared with the repetition rate) that a steady state is reached.

[1] P. L. Copeland, "Anode Effect as a Function of Temperature," *Phys. Rev.*, **57**, 625 (1940), and references there cited.

CHAPTER 7

SECOND DETECTORS

By W. H. Jordan

7·1. Introduction.—The output of the last i-f stage of a receiver is in the form of a modulated sine wave, the information being carried by the modulation frequencies. The sine wave may be either frequency modulated or amplitude modulated. Methods for removing the carrier frequency and recovering the modulating frequency in the case of amplitude modulation will be discussed in this chapter. Such a circuit is called a second detector.

The ideal second detector will reproduce faithfully the envelope of the carrier; that is, the rectified voltage at the terminals of the second detector will be directly proportional to the amplitude of the carrier. If the amplitude of the carrier changes greatly from cycle to cycle, it becomes difficult to define the envelope. Physically this means that the modulation frequencies are approaching the carrier frequency, so that a complete separation is impossible. This difficulty will not be troublesome so long as the bandwidth of the i-f amplifier is small compared with the intermediate frequency. In practice, the intermediate frequency must be chosen considerably higher than the highest modulation frequency desired to avoid imperfect reproduction of the modulation envelope.

In addition to reproducing the envelope of the i-f carrier, a practical detector should be reasonably efficient; that is, for a given carrier voltage from the last i-f stage, the rectified voltage should be as large as possible. Practical detectors may depart considerably from the ideal and thus give rise to various types of distortion of the signal. The amount of distortion tolerable and the requirements for the detector depend greatly on the application. In a broadcast receiver it is necessary that the detector be linear, for a square-law detector generates harmonics of the modulation frequency that are disagreeable to listen to. In a radar receiver considerable departure from linearity can be tolerated, but the transient response must be good. The emphasis in this chapter will be on detectors for high-frequency pulse reception.

7·2. Two-electrode Detectors. *Single Diode Tube.*—Perhaps the most widely used detector is the single diode tube. It is simple to use, is linear for signals above 1 or 2 volts, does not overload readily, and pro-

vides a convenient source of AGC voltage. It does not have the gain inherent in plate or grid detection, but this defect is partly offset by the small amount of power consumed and the small amount of space required. It also loads the i-f stage driving the detector fairly heavily, but this is not usually a disadvantage in wide-band receivers, for the stage must be loaded in any event.

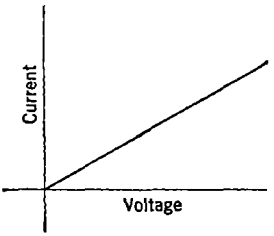


FIG. 7·1.—Ideal diode characteristics.

The ideal diode would have a voltage-current characteristic as shown in Fig. 7·1. There would be no conduction in the back direction, and the current would be a linear function of the applied voltage in the forward direction. The inverse of the slope of the curve (the forward resistance) should be small in order to achieve good efficiency with low load resistance.

Tube	Detector capacity		
	Input	Output	Anode cathode
6AC7	7.9	19.0	6.7
6AK5 (*A*)	4.2	7.9	7.1
6AK5 (*B*)	7.2	5.1	3.4
6J6	3.4	3.2	2.4
6H6	3.9	4.6	1.7
9006	2.4	1.8	1.6
6AG5 (*A*)	5.4	7.8	7.3
6AG5 (*B*)	8.5	7.0	4.0
CV140	2.1	2.9	1.8
6AL5 †	1.6	3.3	2.1
6AL5 ††	2.4	3.0	1.85

† Pins 1 & 7
†† Pins 2 & 5

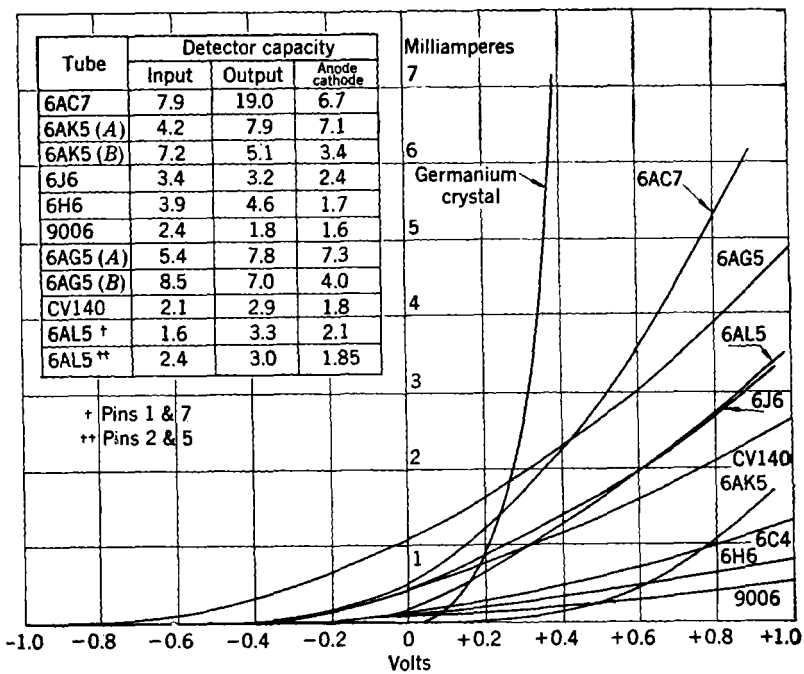


FIG. 7·2.—Low-level diode characteristics. Actual input capacity is equal to the value given in the table plus the anode-cathode capacity in series with the detector load capacity. Connections for 6AG5 and 6AK5 are as follows: (*a*) All positive electrodes used as anode; (*b*) grid used as anode, screen grounded through 0.15-megohm resistor band bypassed. Both connections have same d-c characteristics. For the 6J6, 6H6, and CV140, one half of the tube is used.

Characteristic curves for some commerically available diodes are shown in Fig. 7·2. They depart from the ideal curve in two respects. (1) The slope is not constant, and (2) the curves do not pass through the origin. By operating at i-f levels of several volts, the distortion is minimized and amounts to only a few per cent even for broadcast reception.

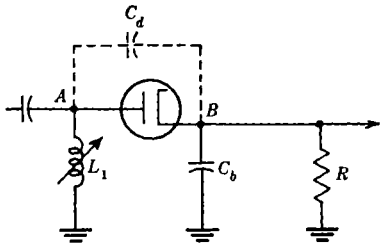


FIG. 7·3.—Diode-detector circuit.

The single diode detector circuit may take many forms, but the one shown in Fig. 7·3 is fairly typical. The inductance L_1 is tuned to resonate with the total capacity to ground at the intermediate frequency. This capacity is made up of the output capacity of the i-f stage, plus the input capacity of the diode, plus stray capacity due to sockets and wiring. The condenser C_b is an i-f bypass condenser. The signal voltage appearing across L_1 is divided across C_b and C_d, the plate-to-cathode capacity of the diode. To get good rectifying action, most of the signal voltage should be impressed across C_d, and therefore C_b should be much larger than C_d. The diode load resistance R then determines the video bandwidth.

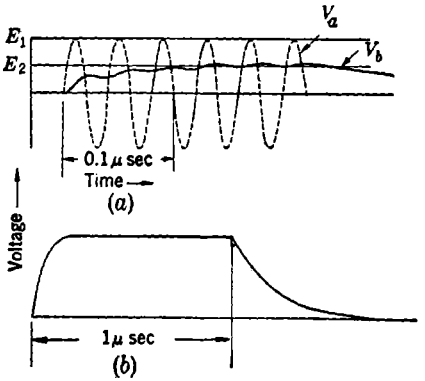


FIG. 7·4.—Second-detector action.

The operation of the detector can best be explained by reference to Fig. 7·4. The dashed line represents the voltage appearing across L_1, in this case a pulse packet of sine waves. Since C_b is assumed to be considerably larger than C_d, most of the voltage will be impressed across the diode. The solid line represents the voltage at Point B. When the voltage at Point A is greater than that at Point B, current flows through

the diode and condenser C_b is charged so that the potential at B is increased. The current is given at any instant by the difference in voltage between A and B and the diode resistance. As soon as the voltage at A becomes less than B, the diode ceases conducting and C_b discharges exponentially through R.

The time constant for discharge has arbitrarily been chosen as 6 cycles of the i-f sine wave. Then, if the intermediate frequency is 30 Mc/sec, the time constant would be 0.2 μsec. This corresponds to a video bandwidth of a little less than 1 Mc/sec.

The amount of charge conducted by the diode on succeeding cycles becomes gradually smaller for two reasons. (1) The voltage across the diode is less because of the increase in voltage at B. (2) Conduction takes place during a smaller fraction of the cycle as the voltage at B increases. As a result, the voltage at B approaches an equilibrium value E_2, which is less than the peak voltage E_1.

The detector efficiency η is defined as the ratio of E_2 to E_1; that is,

$$\eta = \frac{E_2}{E_1}. \tag{1}$$

By making the diode load resistance R large, the detector efficiency can be made to approach unity. When the detector efficiency is high, however, there is a third effect in addition to the two mentioned above; namely, the diode resistance becomes large because the peak voltage across it is small (see Fig. 7·2.) Efficiencies of 90 per cent may be obtained in broadcast receivers, whereas 50 per cent or less is common in radar receivers where R is usually less than 5000 ohms.

An expression for the efficiency[1] of a detector can be obtained if it is assumed that the current through the diode is a linear function of the voltage across it, that is, in the conducting direction it has a constant resistance R_d. First, consider the case where the condenser C_b (Fig. 7·3) is so large that there is only a small change in voltage across it during the cycle. Under these conditions it can be shown[2] that the efficiency η is given by

$$\frac{R}{R_d} = \frac{\pi\eta}{\sqrt{1-\eta^2} - \eta \cos^{-1}\eta}. \tag{2}$$

A graph of Eq. (2) showing η as a function of R/R_d is shown in Fig. 7·5. The more general case where the time constant RC_b is not large compared

[1] In defining efficiency it is postulated that the voltage applied to the detector shall be a pure sine wave and that steady-state conditions have been attained.

[2] C. E. Kilgour and J. M. Glessner, *Proc. IRE*, July 1933. For a thorough analysis of detectors see K. R. Sturley, *Radio Receiver Design*, Vol. 1, Wiley, New York, 1943.

with the reciprocal of the intermediate frequency is considerably more complex. When $C_b = 0$, the efficiency can easily be shown to be[1]

$$\eta = \frac{1}{\pi} \frac{R}{R_d + R}, \tag{3}$$

which has a maximum value of 28 per cent for $R_d \ll R$. As C_b is increased, the efficiency rises, approaching asymptotically the value given in Eq. (2). The efficiency will be considerably less than the asymptotic value as long as the reactance of C_b at the intermediate frequency is comparable to the resistance R, which is normally the case in a wide-band receiver.

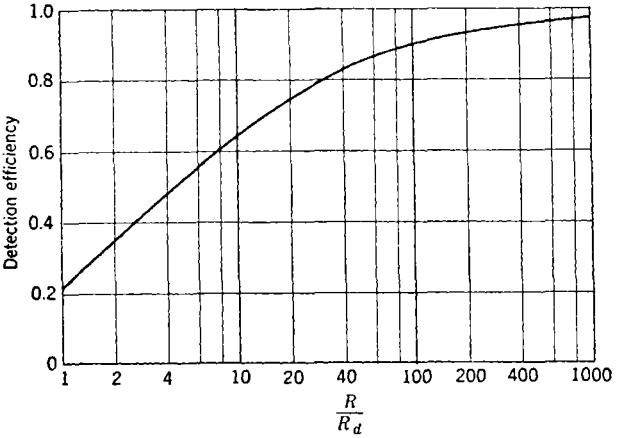


FIG. 7·5.—Diode detector efficiency as a function of the ratio of load resistance to diode resistance.

The effect of the diode load resistance on the pulse response may be seen in Fig. 7·4, where the voltage at B has risen to nearly the steady-state value in about 3 cycles. For a 30-Mc/sec intermediate frequency the time required would be 0.1 μsec, which would be reasonable for a 1-μsec pulse. The time for the pulse to fall from 90 to 10 per cent of the steady-state value is somewhat longer and can be determined exactly from a knowledge of the time[2] constant RC_b. As R is increased, the fall time becomes proportionately larger. The rise time also becomes somewhat greater but not in a direct proportion to R. As R is increased, the time required for V_b to reach a given value is slightly decreased because there is less discharge between cycles. On the other hand, since the equilibrium value is higher, more time will be required to reach 90 per cent

[1] Sturley, *op. cit.*

[2] The fall time for a single RC-circuit is 2.2 times the time constant, or 0.44 μsec in the present case.

of this value. Thus, in contrast to the usual video and i-f amplifier circuits, the rise time and fall time are relatively independent. This fact has sometimes led to confusion as to the choice of the diode load resistor. Actually, one must design conservatively and choose R on the basis of fall time, as is the case for any video stage (see Chap. 8).

The phenomenon of quick rise and slow fall has been made use of in the so-called "pulse-stretching" circuit, which is often used in radar sets in connection with beacon reception. The beacon signals are usually about 0.5 μsec long; and although they are much stronger than the noise, they may not stand out on the screen of the cathode-ray tube so clearly as might be desired, because they are not appreciably longer than the noise impulses and because the signals are limited in amplitude at a value no higher than that attained by many of the noise impulses. To make the signals stand out more clearly, a switch or relay is provided for increasing the diode load resistor by a factor of about 10. The beacon signals then come up to full brightness on the screen as quickly as ever and are extended in range over several microseconds. The ultimate sensitivity of the receiver to the beacon signals is not improved, but the signals can be seen much more readily.

In Fig. 7·4*a*, the envelope of the i-f pulse has not been faithfully reproduced by the detector because the detector time constant 0.2 μsec is much too long for the pulse length 0.2 μsec. Figure 7·4*b* shows the detector output voltage when a 1-μsec pulse of the same frequency and amplitude is applied. (The time scale has been compressed greatly.) The reproduction is now much better although still a bit slow. To be satisfactory for pulses as short as 0.2 μsec, the detector time constant would have to be approximately 0.02 μsec. With so short a time constant, the discharge between cycles would be a great deal more and the steady-state voltage would therefore be much less. This loss in efficiency can be overcome only by having a greater increase in the voltage V_b during the charging cycle or by increasing the time of charge by using a push-pull detector.

A greater increase in voltage during the charging portion of the cycle can be obtained only by increasing the current (less resistance) or by decreasing the capacity C_b (Fig. 7·3) and therefore correspondingly decreasing C_d. A diode, therefore, must be judged not only by its resistance but also by its capacity. Referring to Fig. 7·2, one can see that the 6AC7 has a low forward resistance but a large capacity. The 6H6 has very low capacity but a relatively high resistance. The 6AL5, a double diode in a miniature bulb, is reasonably good in both respects and has been widely used in recent receiver design. A possible figure of merit for a diode would be its forward conductance divided by the anode-cathode capacity. The conductance cannot, however, be reduced indefinitely,

even though the capacitance is correspondingly reduced, for the detector time constant depends on the total capacity across the diode load resistance. The detector must drive the grid of a following stage which has a fixed amount of capacity. Thus, the total capacity would approach not zero but rather the input capacity of the next stage. On the other hand, it would not be desirable to increase the conductance indefinitely with a corresponding increase in interelectrode capacitance, because the current through the diode depends not only on the diode conductance but also on the impedance and capacitance of the driving circuit.

What is needed in a diode, then, depends on the particular application. Its resistance should be low compared with the diode load resistance and the impedance of the i-f stage driving the detector. Its anode-cathode capacity should be considerably lower (by a factor of 10) than the input capacitance of the following stage.

Interaction between Detector and I-f Stage.—In the previous discussion it has been assumed that the i-f generator driving the detector is a constant-voltage, low-impedance generator that is not affected by the current drawn by the diode. Such is never the case in a practical receiver, and not only is the i-f amplifier loaded by the detector, but, conversely, the detector performance is dependent on the characteristics of the amplifier.

To get some idea of the amount of loading that the detector puts on the i-f stage, consider again the case of a zero-impedance voltage generator connected to Point A, Fig. 7·3. Assume the diode resistance to be zero and the bypass condenser C to be so large that there is no appreciable loss in voltage between cycles. These are the conditions necessary for 100 per cent rectification efficiency so that $E_2 = E_1$. Since the voltage E_2 appears across the diode load resistor R, the power dissipated in this resistor will be E_2^2/R. This power must be delivered by the generator. If the diode is replaced by a resistance R_e that will absorb the same amount of power from the generator, then

$$\frac{E_2^2}{2R_e} = \frac{E_2^2}{R},$$

or

$$R_e = \frac{R}{2}.$$

If the diode resistance is not zero, one can show that the equivalent resistance is given by the expression

$$R_e = \frac{\pi R}{\cos^{-1}\eta - \eta\sqrt{1-\eta^2}}. \tag{4}$$

In order to bracket the usual case where C is not very large, consider the other extreme of $C = 0$. By a somewhat similar argument it can be

shown that

$$R_e = 2R + R_d. \tag{5}$$

When C is small but not zero, the loading will be somewhere between the value given by Eqs. (4) and (5). A frequently used approximation is

$$R_e = \frac{R}{2\eta}.$$

The action of a second detector in a receiver can be approximated somewhat more closely by the circuit shown in Fig. 7·6*a*. The last i-f stage is represented by the sine-wave voltage generator V_1 of impedance R_1. C_1 is the total capacity to ground, that is, i-f stage output capacity plus diode input capacity plus stray capacities, all of which are resonated at the intermediate frequency by L_1. The voltage waveforms at Points A and B are shown in Fig. 7·6*b*. (Steady-state conditions have been reached.) The voltage appearing at A is no longer a sine wave, the departure being most noticeable at the peaks because of the clipping action of the diode. Energy is delivered to the tuned circuit by the generator over all of the cycle and extracted by the diode over a fraction of the cycle. When the voltage at A reaches a value E_5, the diode conducts and C_1 and C_2 are therefore in parallel. If the diode had not conducted, the energy stored in L_1 at the time t_1 would have been sufficient to charge C_1 from a voltage E_5 to E_3. (Some of the energy would have come from the generator.) Since this energy must be shared between C_1 and C_2, the voltage will rise only to a value E_4 such that

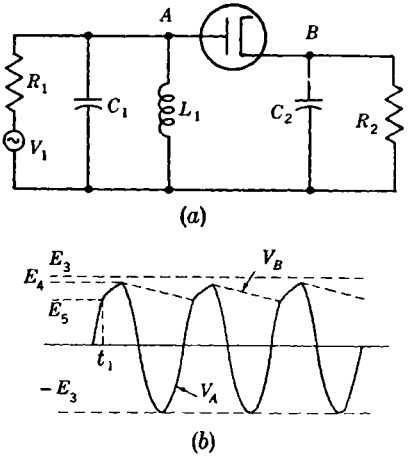


FIG. 7·6.—(*a*) Equivalent detector circuit; (*b*) detector waveforms.

$$(E_4^2 - E_5^2) = \frac{C_1}{C_1 + C_2}(E_3^2 - E_5^2).$$

Thus, even with a diode having zero resistance, the increase in voltage per cycle at the detector output is limited by the ratio of capacitances. The efficiency is affected and also the rise time.

This fact is of importance in the measurement of detector efficiency. Since efficiency is defined in terms of a sine-wave input, such distortion as is shown in Fig. 7·6*b* must be avoided. A high-Q circuit is usually

necessary to filter out generator harmonics; it is also necessary to make C_1 much larger than C_2.

7·3. Push-pull Detectors.—It has been shown that the time constant RC of the detector load must be short if the detector is to reproduce short pulses. Low efficiency results because the charging time is small and the condenser is rapidly discharged. It is possible to double the charging time by the use of the circuit shown in Fig. 7·7.

If the efficiency is defined as the ratio of the rectified voltage E_2 appearing at the output of the detector (Point B) to the peak voltage E_1 applied to each diode plate, then the efficiency of the push-pull circuit is higher than that of the single diode. On the other hand, the total voltage across the secondary must be twice that of the single coil used previously. In addition, there is the added complexity of the transformer and the additional diode. For these reasons the push-pull circuit is not commonly used. It does, however, have some additional advantages. The i-f return circuit is complete without any capacity at C_1, so that there is no loss in i-f voltage across the diode when C_1 is small. Also, the ripple frequency at the output of the detector is double the intermediate frequency, and the filtering problem is therefore greatly simplified (see Sec. 7·5).

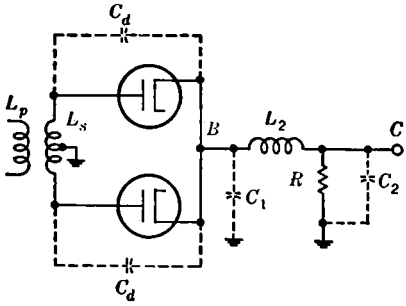


FIG. 7·7.—Push-pull detector.

Another advantage of a push-pull detector concerns the accuracy of reproduction of the pulse envelope. In an MTI system, echoes from stationary targets are eliminated by subtracting the video pulse received on a particular transmission from that received on the preceding transmission. How complete this cancellation is depends on how much alike the two video pulses are. The i-f pulses may be identical in every respect except for phase difference. As may be seen from Fig. 7·4, the exact timing and shape of the video-pulse rise depends upon the phase of the i-f pulse. There would be considerably less dependence on phase if a push-pull detector were used, because information as to the amplitude of the i-f sine wave would be received twice as often.

An alternative proposal for driving a push-pull detector has arisen in connection with the MTI-receiver development (see Chap. 21). It is shown in Fig. 7·8. A double triode is used, so that only a single tube envelope is required. In addition, it has been shown that triodes have better output capabilities than pentodes. The second triode section,

operating as a grounded-grid stage, needs no neutralization; the first section is neutralized by connecting a condenser from the plate of the second section to the grid of the first, the capacity being equal to the plate-to-grid capacity of the first triode section. Each plate is connected to one of the diodes of the push-pull detector through a single-tuned coupling network, thus avoiding the use of a coupling transformer and the associated difficulties of maintaining balance in spite of stray coupling.

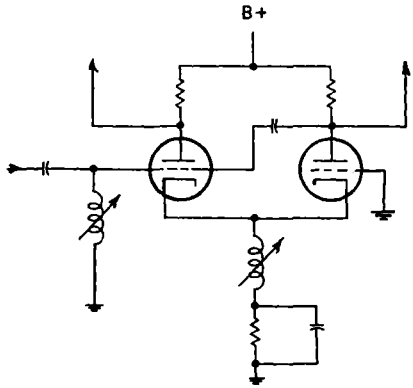


FIG. 7·8.—Double-triode i-f stage for driving a push-pull detector.

7·4. Crystal Detectors.—Crystals have been used as low-level detectors for many years. Only recently, however, have crystals been developed with properties that make them suitable for use as second detectors where the voltages are relatively high. The development of the high-voltage germanium-crystal rectifier has made this possible.[1] For comparison with the more usual vacuum-tube diode, the characteristic curve for a typical crystal is shown in Fig. 7·2. The conductance in the forward direction is considerably higher than that of any tube, the forward resistance being usually less than 100 ohms at 1 volt. This value would indicate that the crystal would be particularly useful in wide-band receivers where the diode load resistance is low.

In addition to the property of high forward conductance, the crystal has other inherent advantages over the tube diode. Its interelectrode capacity and capacity to ground are very low (around 0.5 $\mu\mu$f), and this property is helpful in both the i-f and video circuit. It requires no heater power and can contribute no hum. It requires little space, since the size of the unit is about that of a half-watt resistor. It is equipped with pigtail leads and can be soldered into the circuit, thus obviating the need

[1] For a more complete treatment of the subject of germanium crystals see Vol. 15 of this series.

for sockets. Since the crystal also has a very small and stable contact potential, the current-voltage curve passes through the origin. This, coupled with the further advantage of reaching the linear region at a low voltage, makes the crystal detector more efficient at small input voltages.

Crystals are inferior to tube diodes in two respects. (1) They do conduct in the reverse direction, and (2) the reverse voltage that may be applied is limited. It is in these two respects that germanium crystals are superior to crystals previously developed. How good they are can be seen from the specifications placed on two types of germanium rectifier units manufactured by Western Electric, the D171561 second-detector unit and the D172925 d-c restorer unit. The former type is required to withstand an inverse peak voltage of at least 50 volts, and the latter is required to have a back resistance greater than 60,000 ohms at -50 volts and greater than 0.25 megohm at -5 volts.

The finite value of back resistance and inverse peak voltage limits the usefulness as a second detector in two ways. First, the crystal loads the i-f stage even though the diode load resistance is very high. The amount of loading varies greatly from crystal to crystal and becomes heavier as the i-f voltage increases. The loading also increases with frequency so that it cannot be predicted from an inspection of the d-c characteristic of the crystal. A minimum equivalent resistance of 2000 ohms at 30 Mc/sec is specified for the D171561 unit. Although this much loading is tolerable in wide-band receivers, the variation in loading is inconvenient. In narrow-band receivers a tube diode will give better over-all performance. The second limitation is on the maximum voltage out of the detector. With a good crystal this may be nearly 20 volts, with a poor crystal only 8 or 10 volts.[1] This voltage may, however, be adequate for many applications. Even in radar receivers where a large ratio of maximum voltage to limit level is desired, there is always the possibility of limiting after some video amplification. This procedure is especially indicated when crystals are used, because crystals are linear for smaller i-f voltages than tube diodes.

Recent receiver developments have included receivers with logarithmic response to signals and receivers with d-c feedback loops around individual i-f stages to prevent overloading. Such circuits require detectors at the plates of several of the i-f stages. If tube detectors are used, the added capacity reduces the gain-bandwidth product, thus making wide-band receiver design very difficult. In such cases crystal detectors have proved valuable.

[1] These results were obtained with a wide-band detector circuit in which the diode load was 1000 ohms and the i-f bypass condenser was 10 $\mu\mu$f.

At present it appears that germanium crystal rectifiers can be used in place of tube diodes in detectors for wide-band receivers. Developments in the art of making germanium crystals have shown that considerably higher peak voltage and back resistance are possible. If such crystals can be produced in large numbers at reasonable cost, their usefulness as second detectors will be greatly extended.

7·5. Low-pass Filter between Detector and Video Amplifier.—It has been shown that the voltage at the output of the detector is not steady but varies during the i-f cycle as the condenser is charged and discharged. In very wide-band circuits this variation or ripple may be as large as the d-c component. There are several reasons why it is undesirable for these fluctuations to be transmitted to the video amplifier. (1) The design of the amplifier must provide for larger grid swings if the ripple is superimposed on the video signal. (2) In a half-wave detector the fundamental frequency of the ripple is the same as the intermediate frequency. Since the level at the detector output is high, only a very small amount of coupling to the input of the i-f amplifier is required to produce regeneration. The video amplifier frequently provides this feedback path, and it is much simpler to attenuate the i-f signal by a filter between the detector and video amplifier than by screening and filtering the leads to the video amplifier or improving the i-f amplifier shielding. This is usually the compelling reason for including the filter in radar receivers. Receivers with push-pull detectors do not suffer from this trouble because the fundamental ripple frequency is double the intermediate frequency. (3) In some receivers, such as television receivers, low-order harmonics of the ripple frequency may fall within the r-f pass band. If these harmonics are allowed to get into the video amplifier, they may be reradiated and picked up on the antenna, causing annoying interference effects.

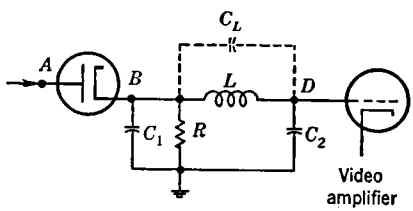


FIG. 7·9.—Low-pass filter.

In the event that a low-pass filter is required between the detector and the video amplifier, it frequently takes the form shown in Fig. 7·9. The inductance L should present a high impedance to the ripple frequencies but a low impedance to the video frequencies. "High" and "low" are relative to the impedance of C_2 at the same frequencies. In a receiver with a 30-Mc/sec intermediate frequency and a video amplifier 2 Mc/sec wide, the choice of L is noncritical. Frequently L is chosen so as to be self-resonant at the intermediate frequency. A 10-μh pie-wound choke, for example, has a distributed capacity of about 2.5 $\mu\mu$f, so that it will be self-resonant in the region of 30 Mc/sec and will thus present a

high impedance at this frequency. At frequencies much higher than the resonant frequency, the coil acts like a condenser and the filter becomes a capacity divider. Since this capacity can never be higher than the distributed capacity, the coil may still be a fairly effective filter. If, for example, the input capacity C_2 of the video amplifier were 15 $\mu\mu$f, the coil would attenuate high frequencies by a factor of at least 15/2.5 or 6. Thus, higher harmonics are considerably attenuated. If the self-resonant frequency of the coil is considerably below the intermediate frequency, it will still be a good attenuator at the intermediate frequency as long as the distributed capacity is low. This property accounts for the observed noncriticalness of the coil.

Although the inclusion of the inductance L does not reduce the video bandwidth and may be chosen to increase it, it does adversely affect the efficiency. For this reason some designers omit it whenever possible. It has previously been pointed out that when the voltage at Point B (Figs. 7·3 and 7·6) falls rapidly between i-f cycles, the efficiency is low. If the filter is effective, the voltage at Point D (Fig. 7·9) will be constant under steady-state conditions; the capacity C_2 is, therefore, unimportant. The voltage at B, however, will fall off exponentially between cycles with a time constant of RC_1. Therefore, even with diodes having low plate-to-cathode capacity, C_1 must be fairly large to obtain reasonable efficiency. For example, if R were 1000 ohms and C_1 were as much as 10 $\mu\mu$f, the efficiency at 30 Mc/sec would be only 40 per cent, even with a perfect diode. Thus, the efficiency depends upon the time constant RC_1, whereas the video bandwidth depends on the time constant $R(C_1 + C_2)$. If there is no filter, the two time constants are the same and the capacity is kept as low as possible. With the filter, a capacity C_1 must be added and R reduced. One attempts to find the value of C_1 and R that will give the most gain and yet maintain the required video bandwidth. The gain is measured under steady-state conditions in terms of the d-c voltage out of the detector and the signal voltage at the grid of the last i-f stage.

Whether R is placed at the input or the output terminals of the filter has been shown experimentally to make little difference in detection efficiency. In the case of a perfect diode with $C_1 = 0$ and L infinite, the resistance should be at the diode side of the filter, for it has already been shown that in this case the efficiency is $1/\pi$. If the resistor were on the output side of the filter, the current through the inductance would have to be constant. Since the capacity C_1 is zero, the current through the diode would be the same constant and hence the voltage at Point B would be the same as the voltage at A. But since the average value of the voltage at A is zero, the average value at B must also be zero. Hence, the voltage at D is zero, and the detector efficiency is zero.

If the inductance L is chosen to series-peak the video, the location of R is determined by the ratio of the capacities C_1 and C_2.

7·6. Video Peaking of Detector Circuits.—It is possible to obtain a considerable improvement in the gain-bandwidth product of the detector by the use of video-peaking circuits. This improvement becomes necessary in wide-band receivers where it is important to maintain a large ratio of maximum detector output to video limit level. The limitation in advantage to be gained by the use of peaking circuits is set by the amount of overshoot on the signal that can be tolerated. Ordinarily detector-

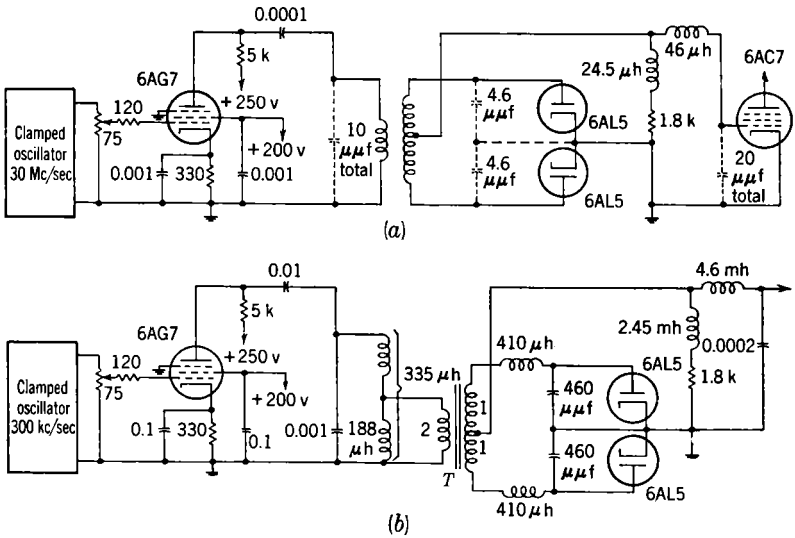


FIG. 7·10.—Push-pull detector with video peaking. (*a*) 30-Mc/sec detector; (*b*) analogue of circuit at 300 kc/sec. Notes: T is a pulse transformer, type OA-15, manufactured by Utah Radio Products Company.

peaking circuits are designed with the same overshoot criteria used for video peaking, in which case the design formulas given in the next chapter can be applied. Recent investigations,[1] however, have shown that the situation is not the same in a detector circuit as in a video circuit and that more peaking can be used if necessary in the former. The remainder of this section will be devoted to a summary of these investigations.

Figure 7·10*a* shows a particular 30-Mc/sec detector circuit, and Fig. 7·10*b* shows the analogue of this circuit at 300 kc/sec. Here the condensers, even the stray ones, have been measured in the 30-Mc/sec circuit and reproduced one hundred times larger in the 300-kc/sec circuit, and

[1] Unpublished work done by P. R. Bell and Warren Henry, Radiation Laboratory.

the double-tuned i-f circuit between the i-f stage and the detector has been replaced by a suitable equivalent. A 300-kc/sec pulse about 15 cycles long is applied to this circuit, and the results photographed. These are the results which could be expected of the 30-Mc/sec circuit, with the time scale divided by 100, that is, a 0.5-μsec pulse. The use of this analogue allows the work to be done readily, since the observing instru-

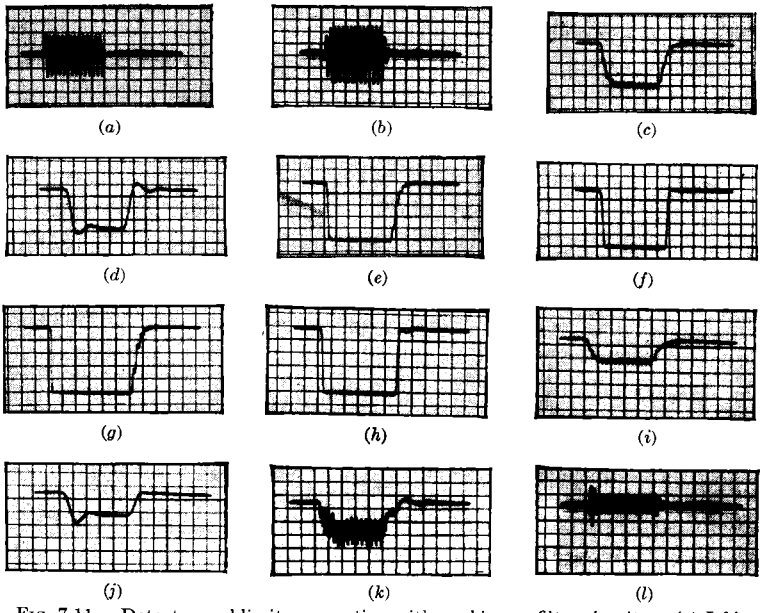


FIG. 7·11.—Detector and limiter operation with peaking or filter circuits. (*a*) I-f input pulse; (*b*) i-f pulse at diode plate; (*c*) video output pulse with small input signal, no filter; (*d*) video output pulse with small input signal, with filter; (*e*) video output pulse with input signal somewhat above limit level and no filter; (*f*) video output pulse with input signal somewhat above limit level and with filter; (*g*) video output pulse with very large input signal, no filter; (*h*) video output pulse with very large input signal, with filter; (*i*) voltage at grid of limiter for conditions used in obtaining (*g*); (*j*) voltage at grid of limiter for conditions used in obtaining (*g*) with filter; (*k*) pulse appearing at center tap of i-f coil secondary under conditions that produced (*d*); (*l*) pulse showing effect upon receiver circuits when signal is detuned from midfrequency.

ment (an A/R-scope) with its input capacity of 25 $\mu\mu$f and resistance of 1 megohm can be attached anywhere in the circuit with negligible effect.

Figure 7·11*a* shows the input i-f pulse. Figure 7·11*b* shows the i-f pulse applied to one of the diode plates with the diode removed and a resistor shunted from plate to plate to give the same effect on the average as the diode does.

Figure 7·11*c* shows the video output pulse with a small i-f signal input and no video filter at all except the input capacity C of the 6AC7 video

stage and the 1800-ohm load resistor. Figure 7·11*d* shows about the same small signal with the video-peaking circuit in place. There is actually no observed change in amplitude of output. Note that the signal is now delayed slightly more and falls more quickly with a considerable overshoot.

Figure 7·11*e* and *f* shows the output pulse without and with the filter for a signal considerably above limit level. The signal in the video-peaked case is now considerably shorter than in the no-peaking case, and no overshoot appears. This is the effect of the diode, there being, of course, an overshoot at the input of the video filter that is acted upon by the diode.

Figure 7·11*g* and *h* shows the output pulse without and with the filter for a very large input signal. The improvement for the video-peaking filter is still obvious. An overshoot does appear but it is very small. Figure 7·11*i* and *j* shows the output pulse of the detector at the 6AC7 grid, but as it would appear if no limiting took place. There is an attenuation factor of 30 used between these pictures and Fig. 7·11*g* and *h*. Note that the percentage overshoot is quite small for the peaked case. Figure 7·11*k* shows the pulse appearing at the center tap of the i-f coil secondary under the conditions that produced Fig. 7·11*d*. Figure 7·11*l* is an interesting one included to show the effect upon receiver circuits when the signal is detuned from midfrequency. This picture was taken under the same conditions as Fig. 7·11*b* except that the signal is detuned with no change of amplitude or gain. Notice the beating between the transient excited at the beginning of the pulse and the driven oscillations that are at the pulse frequency. The transient at the end can also be seen to be of a different frequency. There is no beating here because the driven oscillation has ceased. This i-f signal results in a video output pulse with ears or horns preceding and following. This effect is often observed in wide-band receivers, and the beating phenomenon is observed in receivers having i-f bandwidth switching in only one or two stages when in the narrow-band position with a detuned signal.

The following points should also be noted in connection with these results.

When the signal is present, the diode output impedance is comparatively low; but when the signal ceases, the diode becomes an open circuit as long as the video voltage at its output is larger than the falling i-f voltage in the i-f coupling circuit. Also, if the signal out of the detector load is equal to or smaller than the video limit level, a considerable overshoot transient (perhaps 5 to 10 per cent) on top and after the signal can be tolerated. When the signal is larger than the video limit level, the size of overshoot on top of the signal is immaterial, since it is beyond limit level and not reproduced. The overshoot after the signal, however,

cannot be allowed to increase much in actual voltage and must decrease, therefore, in percentage as the signal increases or it will become a sizable part of the video limit level and may appear as a second signal or will at least lengthen a large signal. The diode helps very much with this overshoot and may remove it entirely (as seen at the output of the video-coupling network), since the diode will always be conducting when the video output changes sign.

7·7. Multielectrode Detector.—The chief disadvantages of the diode detector described in Sec. 7·2 are the low gain and the loading on the i-f amplifier. Other types of detector such as the grid detector, the plate detector, and the infinite-impedance detector have been used in radio receivers for many years. Each of these types of multielectrode detector has certain advantages over the diode that for some applications make it more desirable. On the other hand, each has some disadvantages, and in the case of radar receivers these have been so serious that the diode detector is used in the majority of cases. Grid detectors have not found application in radar receivers and will not be discussed. The plate detector and the infinite-impedance detector have been used considerably and will be described briefly in this section.

If an r-f signal is applied to the grid of an amplifier tube that is biased at or near cutoff, the average plate current increases, the amount of current depending on the amplitude of the signal. Thus, the tube functions as a detector; and if a resistance is placed in the plate circuit, a rectified voltage will be developed that is proportional to the average plate current. In the ideal detector the plate-current grid-voltage characteristic would be linear, and the tube would therefore function as a linear detector when biased to cutoff. The rectified voltage E_2 developed across the load resistor R_L would be

$$E_2 = \frac{1}{\pi} g_m R_L E_1, \tag{5}$$

where E_1 is peak i-f voltage and g_m is the mutual conductance of the tube. It is assumed that the tube is a pentode, and its plate resistance is neglected. Thus, the voltage E_2 is the same as if the tube were used as an amplifier following a detector of efficiency $1/\pi$. If the transconductance were assumed to be 5000 μmhos and the load resistor 2000 ohms, the gain E_2/E_1 would be 3.2. In practice, the efficiency is always less than this because the cutoff is not sharp and g_m is not constant; in many cases it is still further reduced by the use of an unbypassed cathode resistor.

The plate detector may be operated with either fixed bias or self-bias. If fixed bias is used, the situation is fairly simple, and what will happen under different circumstances can be readily predicted. With small signals, the efficiency will be low because of the curved characteristic.

As the signal is increased, the detector becomes fairly linear until the signal gets large enough to draw grid current. Unfortunately, the range of signal amplitude between the start of linear operation and grid current is not large. This sets a serious limitation in the use of the plate detector. As the signal is still further increased, the rectified voltage increases at first, but for very strong signals it begins to decrease because of the current drawn by the grid and the screen. How serious this effect is depends on how large the rectified voltage is compared with the voltage required to produce limiting in the video amplifier.

Before the reversal of output occurs, two other troubles may be encountered. One is the "blackout" effect previously described (Sec. 6·11). As the grid draws current, it collects a negative charge which may not be dissipated for several microseconds after the signal is removed, no matter how low the resistance external to the tube may be. Thus, after a strong pulse, the current through the tube is reduced. In an i-f amplifier, a reduction in gain results; in a detector the effect is more serious, for it results in a video overshoot that is amplified and appears in the output signal. This effect has been exhibited by a fairly large percentage of 6AC7's and 6AK5's.

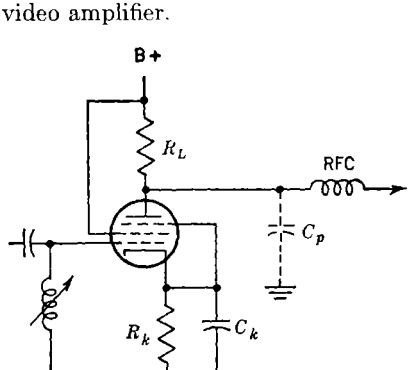


FIG. 7·12.—Plate-detector circuit.

The same sort of trouble occurs when the resistance in the grid circuit exceeds a few hundred ohms. Obtaining a fixed bias of 4 or 5 volts across such a low resistance requires considerable current from the negative supply. For this reason, plate detectors are usually operated with self bias by means of a resistor in the cathode circuit as shown in Fig. 7·12. The load resistor R_L is chosen to give the required video bandwidth, and C_p is usually just the output capacity of the tube. Sometimes additional capacity is added to reduce the impedance of the plate circuit at the intermediate frequency. The efficiency is not affected by C_p when a pentode is used. The resistor R_k must be large enough to bias the tube well into the nonlinear region; a value of 600 ohms has been found fairly satisfactory with a 6AK5 operating with 105 volts on the screen. The best value for C_k will be discussed in some detail.

If the condenser C_k is very large, say 50 μf, the operation and gain will be the same as for fixed bias as long as the duty ratio is low. When the duty ratio is high, the cathode becomes more positive and the tube is

biased beyond cutoff so that the bottom part of the signals and noise is clipped. Although this defect could be taken care of by a lower cathode resistor if a fixed duty ratio were maintained, such conditions seldom exist in a radar set. The duty ratio changes as the antenna scans, and the amount of noise is not always kept the same. Also, it has been found that the blackout effect is much worse when a large condenser is used, probably because of the larger amount of grid current and the higher gain as a video amplifier. These factors, plus the inconvenience of mounting so large a condenser, have made this scheme unpopular.

The next possibility is to make $C_k = 0$. The efficiency for this case may be calculated as follows. If an i-f signal of peak voltage E_1 is applied to the grid, the peak plate current will be

$$I_p = \frac{g_m E_1 R_{eq}}{R_k},$$

where R_{eq} is given by

$$\frac{1}{R_{eq}} = \frac{1}{R_k} + g_m + \frac{1}{r_p}$$

and r_p is the plate resistance when operated as a triode. The average voltage across R_L due to half sine waves of peak value I_p will then be

$$E_2 = \frac{1}{\pi} g_m E_1 R_L \frac{R_{eq}}{R_k}. \tag{6}$$

Comparison with Eq. (5) shows that the gain is reduced by a factor of R_{eq}/R_k. With a g_m of 5000 μmhos and a cathode resistor of 600 ohms, R_{eq}/R_k is approximately $\frac{1}{4}$ (neglecting $1/r_p$) and the gain E_2/E_1 will therefore be 0.8, which is not a great deal more than could be expected from a diode with a 2000-ohm load resistor. This result has been confirmed by experiment.

The third possible choice for C_k is to make it large enough to have a fairly low impedance compared with R_k at the intermediate frequency and yet small enough to have the time constant R_kC_k small compared with the pulse length. A value of 75 $\mu\mu$f has been used in one case. Under these circumstances the cathode reaches its equilibrium potential E_3 during the pulse. As will be shown later when infinite-impedance detectors are discussed, the magnitude of E_3 can be calculated as in a diode detector and can be expressed by the relation

$$E_3 = \eta E_1. \tag{7}$$

Neglecting the current drawn by the screen,

$$E_2 = E_3 \frac{R_L}{R_k},$$

therefore,

$$E_2 = \eta E_1 \frac{R_L}{R_k}. \tag{8}$$

If η is assumed to be 0.4, the gain is

$$\frac{E_2}{E_1} = 1.3,$$

which is approximately 2.5 times as much as could be expected from a diode detector.

Against this rather considerable gain of a plate detector must be weighed its disadvantages. It suffers from overload troubles and requires plate power. The rectified voltage is generated in the plate circuit and is

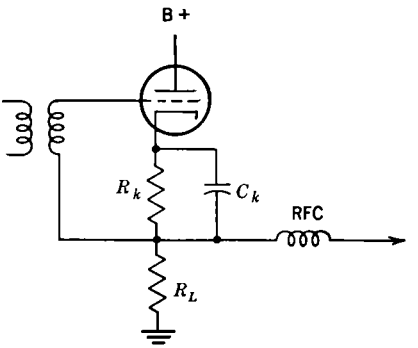


FIG. 7·13.—Plate detector with load in cathode circuit.

not at a convenient level for use in an AGC circuit. Troubles with power-supply hum and feedback are much more likely.

The last-mentioned difficulties can be avoided to a large extent if the load resistance is placed in the cathode circuit as shown in Fig. 7·13. The i-f voltage is applied by means of a transformer between the grid and cathode. The rectified voltage is taken across R_L. Since this voltage is positive and is not zero in the absence of signal, it is not readily adapted for use in AGC circuits. Also, since most video limiter circuits require a negative signal, an inverter is required between the detector and the limiter.

Another detector that has been used in radar receivers is the infinite-impedance detector shown in Fig. 7·14. As the grid goes positive, the tube conducts, thus charging C_L to a positive potential. As C_L becomes more positive, the grid conducts over a smaller fraction of the i-f cycle until an equilibrium value is reached. The action is similar to the diode detector and can best be analyzed in terms of the equivalent circuit shown

in Fig. 7·14*b*. The internal impedance of the generator R_0 is the usual value for the output impedance of a cathode follower; namely,

$$\frac{1}{R_0} = \frac{1}{r_p} + g_m.$$

As before, if a pentode is used, r_p is the value pertaining to the triode connection. The detection efficiency is determined in the same manner as for a diode and depends on the ratio of R_L to R_0 and on the ratio of the time constant R_LC_L to the period of an i-f cycle.

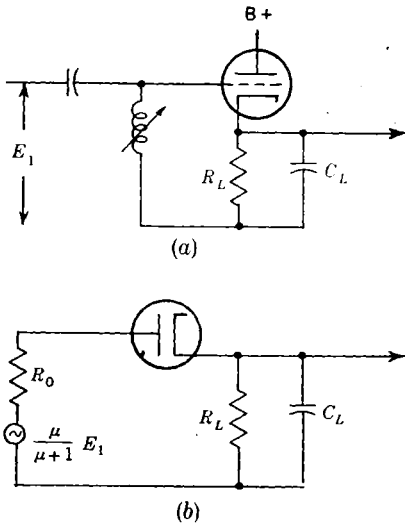


FIG. 7·14.—(*a*) Infinite impedance detector; (*b*) equivalent circuit.

The chief advantage of the infinite-impedance detector is that it does not load the i-f amplifier. On the other hand, it is subject to overloading, though to a much less extent than is the plate detector. Also it has low gain and positive output.

7·8. Testing of Second Detectors.—The tests that should be performed on a detector are largely determined by the requirements placed upon it in a given type of service. For pulse reception, the detector should have high gain, be reasonably linear, have good transient response, and be capable of handling large signals. These factors are by no means independent or complete. If tests show that the detector measures up to requirements in each of these respects, however, it will probably perform well in the receiver. Unfortunately, because detector performance cannot be completely separated from that of the last i-f stage, in many cases it is necessary to include this stage in the tests.

It is usually easier to make c-w measurements of gain and linearity than to make pulse measurements. This procedure is permissible for detectors whose operation is independent of duty ratio, as is the case for all of the detectors described thus far except the self-biased plate detector with a large cathode bypass condenser. When measuring gain by the c-w method a calibrated signal generator is connected to the grid of the last i-f stage and tuned to the intermediate frequency. The rectified voltage is measured with a d-c voltmeter connected across the detector output terminals. This connection is preferably made through a fairly large series resistor so that the detector capacities are not altered. If peak i-f voltage is plotted against rectified d-c voltage, a curve similar to that shown in Fig. 7·15 will be obtained. This curve shows the range over which the detector is linear and the gain at any given level. The leveling-off of the characteristic at high voltage is due to overloading in either the detector or the i-f amplifier. If the detector is not linear, the law of the detector can be determined by making the plot on logarithmic paper.

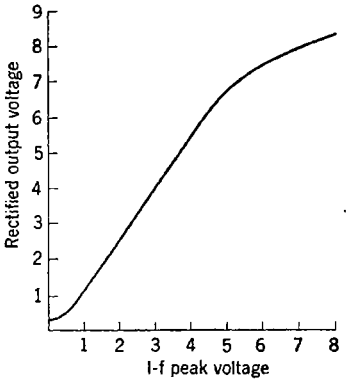


FIG. 7·15.—Over-all detector characteristics.

Although the measurement of the combined gain of the last i-f stage and detector is fairly simple, the measurement of detector efficiency is considerably more difficult. For this measurement it is necessary to drive the detector with a pure sine-wave voltage of known amplitude. The sine-wave generator must have a low impedance compared with the diode resistance. It is not easy to make a high-frequency signal generator that has a low percentage of harmonics, an internal resistance of 50 ohms, and an output of about 10 volts. Consequently, the alternative of driving a high-Q resonant circuit from a high-impedance (pentode) generator is usually adopted. It must be remembered, however, that the Q must be high (at least 10) even with the diode loading. Since

$$Q = \omega RC,$$

a fairly large value of C is necessary; a value of 100 $\mu\mu$f has been found satisfactory at 30 Mc/sec.

Testing transient response requires somewhat special equipment and techniques. First, there must be a means for viewing and measuring pulses with rise and fall times of a fraction of a microsecond. An A/R-scope or a Model 5 synchroscope has been used for this purpose. It is

not advisable to connect the synchroscope directly to the detector, because the capacity would be changed. It is usually connected to the plate of the first video amplifier; the load resistor is shunted with a low resistance (about 100 ohms) to prevent distortion of the pulse coming from the detector. A broad-band video amplifier must be used with the synchroscope.

There are several possibilities for connecting the pulse signal generator to the detector. If the performance of the last i-f stage plus detector is to be tested, the signal generator is connected to the grid of the last i-f stage. For wide-band amplifier testing, pulse signal generators have been built that produce very nearly perfect pulse packets at 30 and 60 Mc/sec; that is, the amplitude of the first oscillation is as large as the amplitude of the following oscillations. At the end of pulse the oscillations are damped within a quarter of a cycle.

If the combined response of the i-f amplifier and detector is not tested, the effect of the former can be made negligible by loading the i-f stage with a resistance of about 100 ohms. Alternatively the signal generator can be applied directly to the detector.

In the event that a pulsed i-f signal generator is not available, a video-pulse generator can be used. The rise and fall time of the video pulse must be much less than the detector time constant. The pulse generator is applied to the detector through a series resistor[1] connected to the top of the detector load resistor. The pulses are then observed on the synchroscope in the same manner as described above. The trailing edge of the observed pulse will be the same shape as that obtained with an i-f pulse generator, but the rising edge will generally not be the same. For this reason the method described earlier is preferred.

The high intermediate frequencies and low capacities involved in pulse receivers make investigation of detector performance particularly difficult. These difficulties can be largely avoided by using the low-frequency analogue of the detector circuit, as was done in obtaining Fig. 7·11. Efficiency, gain, and overload characteristics are the same in the low-frequency circuit as in the high-frequency circuit; transient response times have been increased by the ratio of the frequencies, so that ordinary pulse generators and oscilloscopes are adequate. It is now easy to observe what is happening during the charge and discharge cycle.

7·9. Effect of Noise on Signal.—In order to realize the maximum possible range for a given radar set, it is necessary to detect signals that are weaker than the noise. It has been shown[2] that when signals are

[1] The series resistor should be at least ten times as large as the load resistor if the time constant is not to be changed by more than 10 per cent.

[2] S. A. Goudsmit, "The Comparison between Signal and Noise," RL Report No. 43-21, Jan. 29, 1943; W. H. Jordan, "Action of Linear Detector on Signals in the

comparable to noise, the detector modifies the signal-to-noise ratio of the i-f amplifier. Since an ideal detector is one that reproduces the i-f envelope exactly, this would seem to be an anomalous situation. It must be remembered, however, that the detector reproduces only the envelope, and therefore phase relationships are lost. The discussion to follow will be limited to a perfect linear detector; noise will be thermal agitation and shot noise. At the output terminals of the i-f amplifier, the noise will appear as sine waves of varying amplitude and frequency; the rms value of these waves will be represented by n. During the pulse or signal there will be superposed sine wave of rms value s. When the i-f waves are detected, a voltage A proportional[1] to the amplitude of the waves will be produced. With noise alone, the voltage will fluctuate about some mean value $\bar{A}_1$, the rms value of the fluctuation being the video noise N. When a signal s is added to the noise, the mean value of A is increased to a value $\bar{A}_2$. The difference between $\bar{A}_2$ and $\bar{A}_1$ is the video signal S.

When the signal is much larger than the noise, $S = \sqrt{2}\, s$, as would be expected. (The factor $\sqrt{2}$ is introduced in changing from rms to peak value of S.) The video noise N increases from $N = 0.65n$ for no signals to $N = n$ for large signals. The expression for S is a rather complicated hypergeometric function; but for $s \leqq n$, it can be approximately expressed by the relation

$$S = \frac{1}{2}\sqrt{\frac{\pi}{2}}\frac{s^2}{n}. \tag{9}$$

Hence, even for a linear detector, the video signal falls off as the square of the i-f or r-f signal when $s/n \leqq 1$. The situation is much the same for a square-law detector. Since the video signal is considerably smaller than it would be if noise were not present, this is sometimes referred to as "suppression" of the signal by the noise. A signal equal to noise is suppressed 8 db; a signal $\frac{1}{10}$ noise is suppressed 27 db.

The effect of suppression can be easily demonstrated. To the input terminals of a low-gain i-f amplifier are connected a signal generator and a thermal noise source. A narrow-band output meter, such as the General Radio 760A sound analyzer, is connected to the output of the detector. When a modulated c-w signal is fed into the receiver and the sound analyzer tuned to the modulation frequency, a reading will be obtained on the analyzer. Then, as noise is introduced into the amplifier, the reading will decrease. If it were not for the suppression of the signal,

Presence of Noise," RL Report No. 61-23, July 6, 1943; R. E. Burgess, "The Rectification of Signal and Noise by Linear and Square Law Detectors," National Physical Laboratory (British) Report RBB/C.93.

[1] For simplicity A will be taken as equal to the i-f amplitude.

the reading would be expected to increase. In one such experiment with a signal slightly less than noise, a decrease in signal of 13 db was observed.

In a radar indicator one is usually interested in the size of the video signal as compared with the noise N on either side of the signal. Since $N = 0.65n$, Eq. (9) becomes

$$\frac{S}{N} = \frac{1}{1.3}\sqrt{\frac{\pi}{2}}\frac{s^2}{n^2}. \tag{10}$$

This equation shows that the video signal-to-noise ratio falls off as the square of the i-f signal-to-noise ratio for weak signals. This is particularly discouraging when one tries to improve the radar by designing a more sensitive indicator for the signals. If the indicator sensitivity is improved by 2 db, the over-all sensitivity is improved by only 1 db.

CHAPTER 8

VIDEO AMPLIFIERS

By F. M. Ashbrook

8·1. Introduction.—Although it is possible to provide all the amplification needed in a receiver in the form of i-f gain, it is much more common to use additional amplification after the second detector. These circuits must transmit the appropriate band of frequencies, often from a fairly low frequency of a few cycles per second up to several megacycles per second. Circuits of this type, which first found wide application in television practice, have come to be known as "video" circuits.[1]

Radar receivers pose some special problems because of the wide range of signal level at the second detector. The maximum output of the second detector may be as much as 40 to 60 db above the smallest useful signal. On the other hand the cathode-ray tube to which the signal is ultimately applied can tolerate a range of less than 20 db for efficient operation. Limiting or some form of level compression is therefore required somewhere in the chain. If it is applied early, all the video stages following may be designed for the reduced dynamic range and simplifications in design thereby made possible. It has become customary, accordingly, to combine the functions of video amplifier and limiter in the first stage following the second detector. This chapter will be concerned largely with the design of such combined amplifier-limiter stages and of line-feeding stages, these being the two types usually incorporated into the receiver package.

8·2. Video Bandwidth.—The bandwidth required of a video amplifier in a receiver is dependent on the specific requirements of the equipment. In general, however, it has been the practice in radar reception to make each stage of the video amplifier have a bandwidth of just one-half of the apparent bandwidth[2] of a single i-f stage. When this condition is attained, each stage, i-f and video alike, has about the same effect on the over-all pulse response. If the receiver is to be designed with the

[1] A treatment of the fundamental principles and practical features of the design of such amplifiers will be found in Vol. 18, Chap. 3, Radiation Laboratory Series.

[2] In synchronous single-tuned i-f stages, the apparent bandwidth of an i-f stage is the bandwidth of the single stage. In double-tuned and stagger-tuned amplifiers, etc., the apparent bandwidth is taken as the bandwidth of each stage of a synchronous single-tuned amplifier having the same number of i-f stages and the same over-all bandwidth.

best possible performance against interference, the over-all video bandwidth should be at least twice as wide as the whole i-f bandwidth. This conclusion has been borne out both experimentally and theoretically.

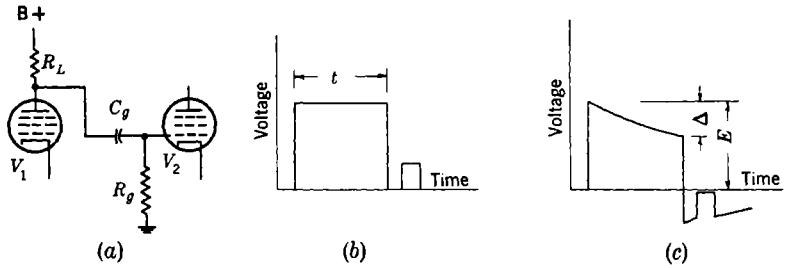


FIG. 8·1.—The effect of imperfect low-frequency response. (*a*) Amplifier stage; (*b*) signal at plate of V_1; (*c*) signal at grid of V_2. Note that the small pulse is now entirely below the baseline and may be lost altogether with some indicating methods. $\Delta \approx E(t/R_0C_0)$.

For the video-amplifier circuit it is also necessary to consider the low-frequency response. If it is assumed that the system is to reproduce pulses as in radar, it is fairly simple to establish the basic low-frequency requirements. If a square pulse is applied to a video-coupling circuit, the failure of the low-frequency response will show up as decay in the voltage received at the grid of the next stage, as shown in Fig. 8·1. The over-all effect will be, of course, the aggregate effect of the individual stages.

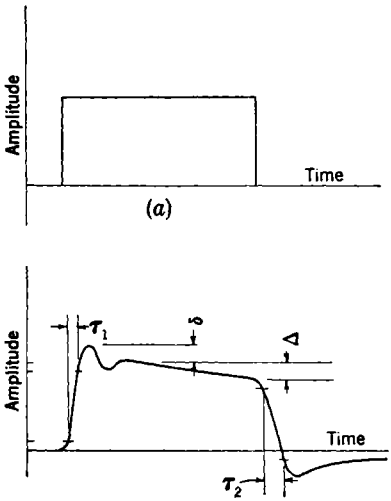


FIG. 8·2.—Specification of performance of a video amplifier. (*a*) Input signal; (*b*) output signal.

It has been common practice to describe amplifiers in terms of their steady-state frequency response, an easy characteristic to measure. For those circuits used with pulses, particularly such pulse signals as are met in radar application, a much more significant description can be given in terms of the response to a square wave or pulse. To measure such characteristics, a square wave of appropriate duration is applied to the circuit, and the output signal is viewed on an oscilloscope. There are four features of this output wave that are of importance. As shown in Fig. 8·2, which represents a hypothetical output wave, these are

1. The rise time τ_1.
2. The fall time τ_2.
3. The percentage of overshoot δ.
4. The percentage of droop Δ.

Rise and fall times are here defined as being measured between the points corresponding to 10 and 90 per cent of the final amplitude. For a video circuit whose high-frequency cutoff is at frequency f, the rise and fall times are approximately

$$\tau_1 \approx \tau_2 \approx \frac{0.35}{f}.$$

The presence of nonlinear elements in the circuit may cause the rise and fall times to differ. The test signal should, of course, have a rise and fall time considerably shorter than does the circuit under test.

The relation between bandwidth and rise time may be misleading if there is a considerable amount of overshoot on the output signal. At one time a video amplifier was built that had a flat frequency response to above 70 Mc/sec. This amplifier should have reproduced pulses as short as $\frac{1}{50}$ μsec with little distortion. It failed completely, however, when it was used with pulses as long as 2 μsec because of an irregularity in the response at fairly low frequencies. It must be remembered that when one is dealing with pulses, the transient behavior of the amplifier is of prime importance and the frequency response tells only the steady-state characteristics. If both the frequency and phase characteristics are available, it is possible to convert these characteristics into the pulse response. Since the mathematics involved for more than a few special cases is extremely difficult, if not impossible, the pulse technique is preferred for testing amplifiers designed for pulse reception.

8·3. Low-frequency Response.—This has to do principally with the response of the receiver to long pulses. In deciding just what response is necessary, two questions must be answered: (1) What are the longest pulses that are to be expected, and (2) what are the characteristics of the indicating device, that is, how much "droop" can be tolerated without the loss of information either during or after the pulse? It is convenient to describe the phenomena in terms of the picture seen on an ordinary oscilloscope in which a linear time sweep is the horizontal deflection and the signals give the vertical deflection. For such a presentation considerable droop of the baseline after a pulse can be tolerated without serious effect. When intensity-modulated tubes are used, rather serious signal loss can occur if the depressed area is too deep. These tubes are normally operated with the baseline at about the cutoff voltage on the tube. Consequently, any signal more negative than that will drive the tube beyond cutoff. While the tube is in this cutoff condition, a fairly

large signal could fail to bring the potential up to the normal bias level and thus would fail to show. Figure 8·1 shows how this might occur.

It has generally been considered good practice in radar work to make the time constants such that the over-all signal depression on a long sustained signal block would be less than 10 per cent. Some of the higher-power radar sets use a 600-μsec signal for a test of acceptability. The normal receiver practice has been to make the time constants of all the circuits at least 10,000 μsec. Low-frequency compensation is sometimes used where it is felt that this value should be extended or where conditions make it difficult to use condensers and resistors that will give a sufficiently long time constant. Such low-frequency compensation schemes should be used with caution, and the compensation should be

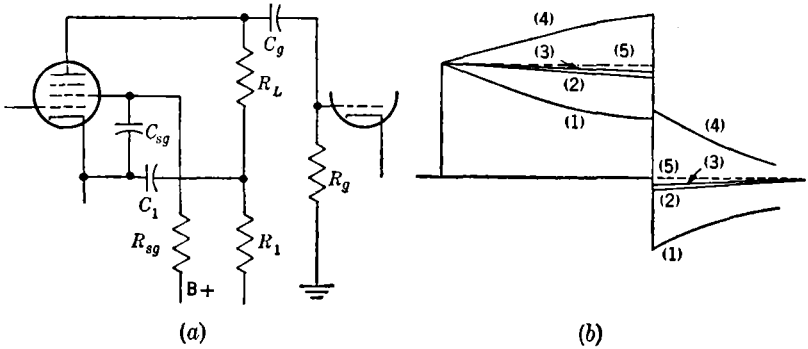


FIG. 8·3.—Low-frequency compensation. (*a*) Circuit; (*b*) the effects of low-frequency compensation on a step function: (1) Undercompensated (R_0C_g); (2) compensated to flat frequency response $R_0C_1 = R_gC_g$; (3) compensation beginning with zero slope $R_lC_1 = R_gC_g$; (4) overcompensated low-frequency circuit; (5) ideal response.

checked by introducing a long pulse into the receiver and observing the results. The commonly used low-frequency peaking circuit is an RC-network in series with the plate-load resistor, as shown in Fig. 8·3.

There are two possible methods of choosing the constants for low-frequency compensation. In the first, the constants are chosen so that the steady-state frequency response is maintained flat. The second case is designed so that the initial slope following a step function is zero. In the circuit of Fig. 8·3 a flat frequency response is obtained when

$$R_0C_1 = R_gC_g, \tag{1}$$

where

$$R_0 = \frac{R_1R_L}{R_1 + R_L}.$$

If a pulse response that has an initial zero slope is desired,

$$R_LC_1 = R_gC_g. \tag{2}$$

For practical reasons, R_1 is usually about five times as large as R_L; thus the difference between these two formulas is small. In fact the difference between the two will usually be less than the tolerance of the components used. Although Eq. (2) does not involve R_1, the amount of compensation obtained is a function of the resistance R_1. For $R_1 \approx 0$, the low-frequency compensation will be negligible. When $R_1 = 5R_L$, the time required for the pulse to fall to 90 per cent of its original amplitude is about five times as long as in the case of the uncompensated circuit.

Until now the discussion has assumed that the screen voltage is fixed. If it is necessary to use a dropping resistor for obtaining the screen voltage, the screen time constant must also be taken into account. If i'_p is the change in plate current due to the grid signal, then

$$i'_p = g_{m_1} e'_g,$$

where g_{m_1} is the transconductance from control grid to plate and e'_g is the grid signal. The change in plate current i''_p due to a change in screen voltage will be

$$i''_p = g_{m_2} e''_g,$$

where g_{m_2} is the screen-grid-to-plate transconductance and e''_g is the change in screen voltage. For a signal that is not too long, the change in screen voltage is approximately

$$e''_g \approx \frac{g_{m_1} e'_g}{k} \frac{t}{C_{sg}},$$

where k is the ratio of plate current to screen current,[1] t is the length of the pulse, and a linear discharge of the condenser C_{sg} is assumed. The fractional change in output signal caused by a change in the screen-grid voltage will be[2]

$$\Delta = \frac{i''_p}{i'_p} \approx \frac{g_{m_2}}{k} \left(\frac{t}{C_{sg}} \right). \tag{3}$$

[1] This ratio is approximately equal to 4 in 6AC7 and 6AG7 tubes, which are the most commonly used video tubes.

[2] The full expression for the fractional change in output signal for an input pulse of any length is

$$\Delta = \frac{g_{m_2}}{k} R_{sg} (1 - e^{-\frac{t}{R_{sg} C_{sg}}}). \tag{i}$$

In this expression it has been implicitly assumed that the change in screen-grid voltage does not produce a change in control-grid transconductance. For usual values of R_{sg} and for pulses long enough to show any difference in the value of Δ given by the two expressions, this assumption will not be justified. Accordingly, the simpler expression in the text is adequate.

If a maximum allowable value is assigned to Δ, C_{sg} will be given by

$$C_{sg} = \frac{g_{m_2} t}{k\Delta}. \tag{4}$$

Since the screen transconductance g_{m_2} in 6AC7's and 6AG7's is between 200 and 500 μmhos, a signal change less than 1 per cent in 500 μsec would require a screen bypass condenser of about 6 μf. In many of the amplifiers that have been built in the laboratory, 1-μf condensers have proved satisfactory.

8·4. High-frequency Response.—The requirement for substantially constant gain over a wide frequency range with uniform time delay over the same range of frequencies is best satisfied by a resistance-coupled amplifier. As in the low-frequency case, it is possible to improve the response somewhat by the use of peaking or compensating circuits. A parameter of importance is the frequency at which the response of the normal resistance-coupled circuit has dropped 3 db. This is

$$f_c = \frac{1}{2\pi R_L C'}, \tag{5}$$

where R_L is the load resistor[1] and C' is the total capacity associated with the stage. This total capacity consists of the output capacity of the stage plus the input capacity of the next stage and the shunt capacity to ground of the circuit components and wiring. The value of C' should be held as low as possible by careful wiring and placement of parts.[2]

Figure 8·18*b* shows the careful construction of one wide-band video circuit. It should be noted how all the components are carefully mounted to keep capacities low and lead lengths short. Because this circuit was designed for an over-all bandwidth of 15 Mc/sec, more than usual care was taken in placing the components. This particular amplifier circuit contains all the video amplification up to the cathode-ray tube (a 5JP1 with push-pull deflection). The unusually wide bandwidth together with the requirement of over 100 volts of output makes the large output tube necessary.

The usual methods of improving the response of an amplifier at frequencies around f_c involve the addition of one or more inductances or of

[1] The shunting effects of the grid-leak resistor and the plate resistance have been neglected, since they are both generally large compared with the plate-load resistor.

[2] As an example of the effect of poor wiring, the following instance may be cited. One video circuit was designed to have a bandwidth of 4 Mc/sec but, when measured, actually had a bandwidth of less than 1 Mc/sec. The components for the amplifier had been located on a resistor board some distance away from the tubes, and all the wiring had been carefully cabled; thus the stray capacity had been increased from a few micromicrofarads to well over 50 $\mu\mu$f.

both inductance and capacitance to the circuit. Circuits may also be built using various types of inverse feedback, but these will not be discussed here. Although the circuits are frequently catalogued according to the increase in high-frequency cutoff (frequency at which response is down 3 db), it must be remembered that the transient response is the ultimate criterion and that a good steady-state response for high frequencies is of no use if it is obtained at the expense of bad overshoots.

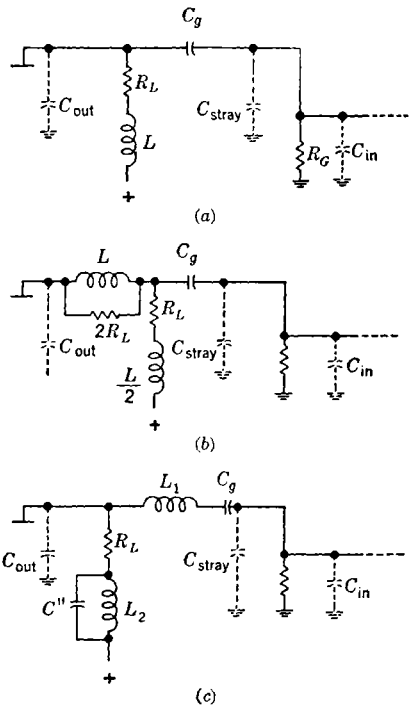


FIG. 8·4.—Video peaking circuits. (*a*) Shunt peaking: $L = 0.25\ R_L^2C'$; $C' = C_{out} + C_{in} + C_{stray}$. (*b*) shunt-series peaked: $L = R_L^2C_{in}$; (*c*) linear phase network. The capacitor C'' should resonate with L_2 at $3f_c$, where f_c is the unpeaked cutoff frequency.

The simplest peaking circuit is the so-called "shunt-peaked" circuit where an inductance is added in series with the load resistance (see Fig. 8·4*a*). This circuit is capable of increasing the video bandwidth by a factor of $\sqrt{2}$ without introducing any overshoot on the signal. The inductance L (Fig. 8·4*a*) is equal to

$$L = 0.25R_LC'. \tag{6}$$

In practice, the peaking can be easily adjusted for optimum by starting with a coil having a little too much inductance. A *square* pulse is then

introduced into the circuit, and a few turns at a time taken off until the overshoot has completely disappeared. The peaking coil is usually placed between the load resistance and the $B+$ supply rather than between R_L and the plate because the coil often has more capacity to ground than the resistance, and in this combination it exhibits fewer ill effects on the circuit. The coil form for the peaking inductance should be one of low capacity like those used in Fig. 8·19. Slug-tuned coils that mount on the chassis are not recommended because of their high capacity to ground.

The second circuit is often called the "series-shunt-peaked" circuit and has a gain in bandwidth over the unpeaked case of 2.25. However, there is a 7 per cent overshoot introduced by this circuit if the output capacity of the previous amplifier is one-half the input capacity of the next tube. If the output and input capacities are equal, the overshoot is only about 4 per cent, but the rise time is not so fast. This circuit, illustrated in Fig. 8·4*b*, has an inductance of

$$L = R_L^2 C_{in}. \tag{7}$$

The same constructional details as in the shunt-peaked circuit apply here. There is, however, no simple experimental method of adjusting the inductances. Probably the best procedure is to determine the total capacity associated with the stage and then calculate the proper value for each coil. Adjustment to the correct inductance may then be made with an inductance bridge or Q-meter. The capacity may be determined either by direct measurement or by measurement of the frequency characteristics of the uncompensated stage (both coils short-circuited). The frequency at which the gain has dropped 3 db is f_c in Eq. (5), from which the capacity may be computed.

Figure 8·4*c* is called a "linear phase network." This peaking circuit increases the bandwidth of the normal circuit by a factor of 2.4 and introduces an overshoot of only 0.3 per cent. This circuit is good only for the case where the output capacity is one-half of the input capacity of the next tube. The same construction techniques mentioned above apply in this case. It has been common practice in the laboratory to wind the shunt coil in such a way that its distributed capacity furnishes the proper capacity across the coil. If the ratio of C_{out} to C_{in} is not exactly $\frac{1}{2}$, a resistance is required across the series inductance to make the damping of the two inductances equal. The bandwidth advantage in this case will not be so great as when the ideal capacity ratio is obtained.

As the gain in the bandwidth increases in these various peaking circuits, the tolerance of the components decreases; consequently, if the more powerful peaking circuits are used, the inductances must be more

uniform and the resistor tolerance small if the overshoots are to be kept at a minimum.

8·5. Bandwidth Requirements Preceding a Limiter.—Any video circuit preceding the point at which limiting occurs requires special attention with regard to its frequency response if proper transient behavior is to be maintained. In radar receivers in which the first video stage is a combined amplifier limiter, the coupling circuit between the second detector and the grid of the video stage falls in this category.

The simplified circuit shown in Fig. 8·5*a* may be used to demonstrate the effects produced by restricted high-frequency response. The limitation of response is represented by means of the *RC*-circuit whose high-frequency cutoff is

$$f_c = \frac{1}{2\pi RC}.$$

Suppose that there is applied to the circuit a large square negative current pulse 1 μsec long, as shown in Fig. 8·5*b*. The voltage across the circuit will then have the form shown in Fig. 8·5*c*, which shows the voltage pulse for two values of f_c, 0.5 and 1.0 Mc/sec. Suppose now that this negative voltage pulse is applied to the grid of a tube that requires a voltage for complete cutoff of plate current represented by the line marked "limit level" in Fig. 8·5*c*. The plate current of this tube will then have the forms shown in Fig. 8·5*d* for the two assumed values of f_c. Note that the form of the voltage pulse in Fig. 8·5*c*, even for $f_c = 0.5$ Mc/sec, does not represent an excessive distortion of the input signal but that the plate-current pulse in Fig. 8·5*d* is nearly twice as long as the input signal. Thus, the limitation of high-frequency response ahead of limiting may produce very serious lengthening of pulse signals.

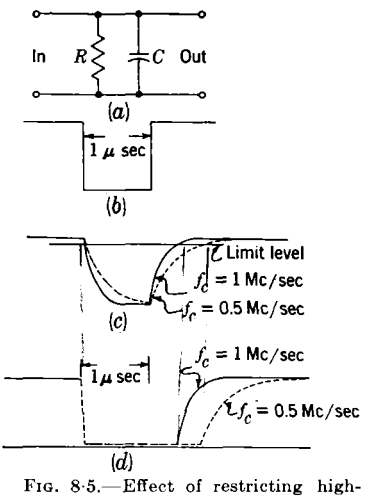


FIG. 8·5.—Effect of restricting high-frequency response before limiting. (*a*) Circuit; (*b*) input current; (*c*) output voltage; (*d*) plate current.

Next, consider the effect of imperfect low-frequency response, as represented by the circuit of Fig. 8·6*a*. Again, a large negative current pulse is applied to the input, and the voltage across the output terminals is determined. The input and output are shown in Fig. 8·6*b* and *c*. As before, the output voltage is applied to the grid of a tube whose limit

level is indicated on Fig. 8·6c. The plate current of this tube will then have somewhat the form shown in Fig. 8·6d. Although the overshoot in the voltage pulse of Fig. 8·6c is not particularly great, the output signal has an overshoot comparable in size to the signal itself. In effect, the fractional size of the overshoot has been multiplied by the ratio of maximum voltage pulse to limit level. In radar receivers it is necessary to make this ratio fairly large, approximately 10, to secure good performance in the presence of certain types of interference. (The maximum voltage in this case is taken as the greatest output of which the detector is capable when driven by the i-f amplifier at its overload point.) Thus it is necessary to be much more careful with both low- and high-frequency response in the detector-video coupling circuit than might otherwise seem to be the case. Similar caution will be necessary in any other video circuit preceding a limiter if the maximum input signal is large compared with the limit level.

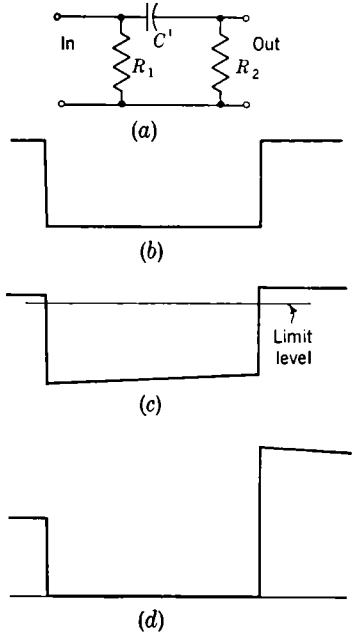


FIG. 8·6.—Effect of poor low-frequency response before limiting. (*a*) circuit; (*b*) input current; (*c*) output voltage; (*d*) plate current.

8·6. Practical Coupling Circuits.—Figure 8·7 shows some of the forms that these coupling circuits may take in radar receivers. Probably the one most commonly used in simple receivers is that shown in Fig. 8·7*a*, in which a moderately large coupling condenser is used between the detector and the video tube. The high-frequency response is assured by a fairly low value of detector load resistor and by the use of some peaking or compensation as described in Chap. 7. The low-frequency response is obtained by the use of a large enough value for the coupling condenser and grid leak. It might seem preferable to avoid all possibility of trouble from poor low-frequency response by the use of d-c coupling as shown in Fig. 8·7*b*. If, however, any c-w interference is present, it may produce enough d-c output from the detector to keep the video stage biased beyond cutoff at all times and thereby completely eliminate all signals. With the condenser coupling, such a d-c voltage rapidly dies out on the grid of the video tube; thus this tube may again respond to changes in

signal level. A second possible cure is found in the use of a special protective circuit for the detector, which is discussed in Chap. 9 (detector back bias, DBB). The receiver described in Chap. 14 incorporates this scheme.

In radar sets in which interference modulated at a moderate frequency, either frequency modulation or amplitude modulation, is expected, it may be desirable to give up completely the possibility of reproducing long pulse signals. If an *RC*-coupling network with a short time constant is used, a large overshoot may be produced in the manner shown in Fig. 8·6. It will, however, decay at a rate determined by this

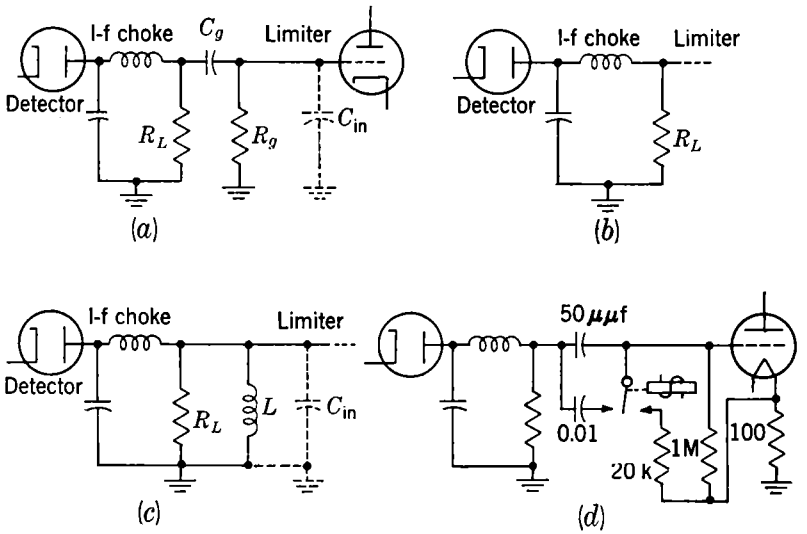


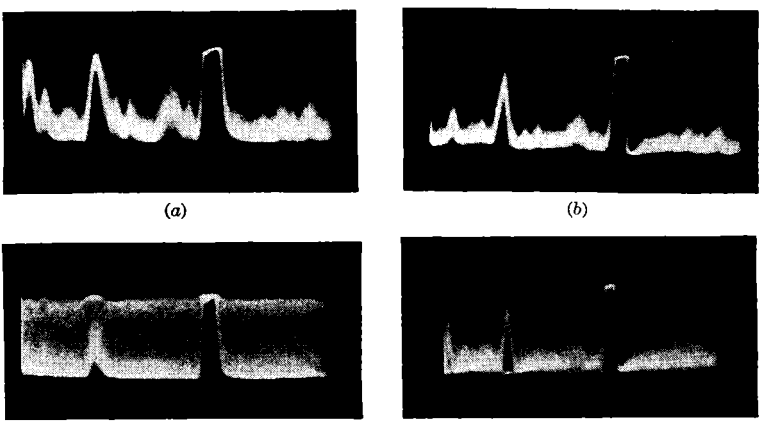
FIG. 8·7.—Coupling circuits between second detector and video amplifier. (*a*) *RC*-coupling circuit; (*b*) direct-coupling circuit; (*c*) inductance-coupling circuit; (*d*) switching circuit for FTC.

short time constant and hence may be tolerable. The constants of the circuit are usually chosen so that the product R_gC_g (see Fig. 8·7*a*) is about equal to the length of the transmitted pulse of the radar system. Since C_g and C_{in}, the input capacity of the tube, form a voltage divider, C_g should be at least twice C_{in}. In many cases the amplifier tube is run at zero bias, so that the current drawn by the grid, if it tends to go positive as a result of the overshoot in the circuit, will act to suppress the overshoot. This action is facilitated if C_g is not too large. Accordingly, the compromise value $C_g = 2C_{in}$ has been found generally satisfactory for time constants from 0.25 to 5 μsec. This circuit has come to be known as a "fast time constant" (FTC) circuit and has found use not only in furnishing some protection against interference but also in increas-

ing the discernibility of single radar echoes in the presence of ground clutter or sea return (see Chap. 9, also Vol. 24, Chap. 11, of this series).

Another circuit whose performance is essentially equivalent to the FTC circuit is shown in Fig. 8·7*c*. In it the quantity L/R_L plays much the same role as the time constant R_gC_g of the FTC. The coupling from the detector may pass direct current, since the added inductance provides a low-resistance path that prevents any biasing off of the amplifier.

Whenever FTC or inductance differentiation circuits are provided, it is usual to arrange for their removal to permit normal response of the receiver for testing, etc. An arrangement that has often been used with FTC is shown in Fig. 8·7*d*. The relay used should have low capacity to ground in order to avoid undue narrowing of the video bandwidth.



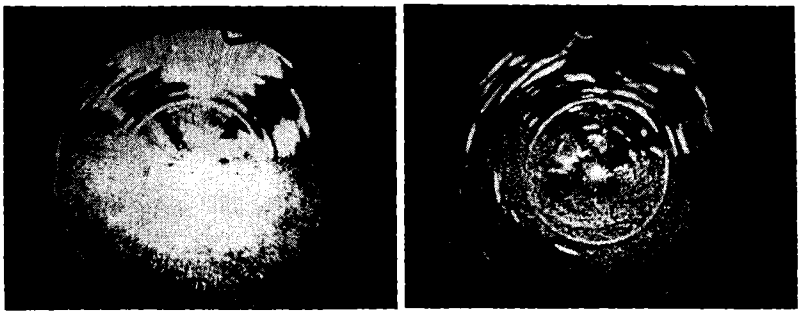
(*a*) (*b*) (*c*) (*d*)

FIG. 8·8.—The action of the FTC circuit on signals alone and in the presence of interference. (*a*) Normal signals; (*b*) FTC introduced; (*c*) signals plus interference (c-w modulated at a low frequency); (*d*) signals plus interference with FTC.

A considerably more elaborate coupling arrangement is shown in Fig. 8·21. In it a multisection high-pass filter is used together with an additional amplifier stage to make up for the insertion loss of the filter. This stage is followed by a full-wave detector. The general action of the circuit is fundamentally the same as that of the FTC or inductance differentiation, differing only in the degree of refinement.

Figure 8·8 shows some oscilloscope photographs that indicate the general action of these coupling circuits. Figure 8·8*a* shows the echo signals on a radar system under normal conditions. The length of the sweep is slightly under 20 μsec (corresponding to slightly less than 4000 yd in range). It has been delayed after the transmitted pulse by an arbitrary amount that is the same in all four pictures; therefore, the actual range of the objects producing the echoes is considerably greater

than 4000 yd. In Fig. 8·8*b*, the FTC circuit has been added. It is seen to have little apparent effect on the signal presentation. In Fig. 8·8*c* and *d* some modulated interference has been added, the FTC being out of the circuit in the first case and in the circuit in the second. Although many of the signals can be seen in Fig. 8·8*c* with this particular type of presentation, they would not be visible if an intensity-modulated CRT such as a PPI were used. However, the conditions shown in Fig. 8·8*d* would give essentially normal appearance on the PPI. Figure 8·9 shows the effect of the FTC circuit against clutter arising from sea return or ground echoes with a PPI presentation. The FTC is out and in respectively in Fig.



(*a*) (*b*)

FIG. 8·9.—The action of the FTC circuit on signals in the presence of ground clutter and sea return. (*a*) Normal coupling circuit; (*b*) FTC introduced. These are photographs of the presentation on a PPI. The radar set operates in the region of 3000 Mc/sec with a pulse length of 1 μsec. The sweep length is 5 miles. The set is located directly on the coast, looking out to the south over the ocean. The upper half of each picture is due to land signals. The ring visible in (*b*) about halfway out is a pulse signal from a signal generator, which is used to get an estimate of the intensity of signals. Note that at about 9 o'clock the sea return has overridden the marker. The individual signals outside the marker between 7 and 9 o'clock are ships that are completely hidden in (*a*).

8·9*a* and *b*. In the latter picture, many individual echoes that are completely hidden under the usual conditions may easily be seen.

8·7. Video Limiters.—The limit level is established in almost every case by an amplifier that is driven to cutoff. Although grid current has been used in some instances to provide limiting action, it is in general less satisfactory because it is not a hard limiter (that is, the larger signals limit at a higher level than the smaller signals). In addition, the charge that collects on the coupling condenser during the limiting action either has to be removed by the use of a diode or causes a shift in the operating characteristics of the limiter and thus a change in the limit level for succeeding signals.

These limiters perform best when the tube used is a sharp cutoff pentode such as a 6AC7. The signal level at which limiting occurs is controlled by the screen voltage of the tube. The limiter characteristics

are in a large measure determined by the type of screen-voltage supply used and must be chosen to fit the particular situation. All the limiters discussed here will be of the type where a tube is driven to cutoff by the signal and will first be classified according to the means used to supply the screen voltage.

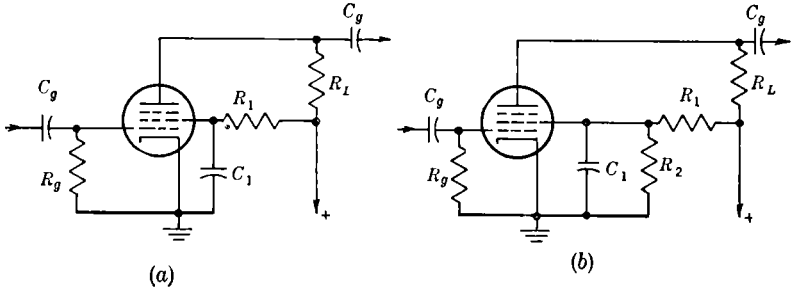


FIG. 8·10.—Limiter circuits. (*a*) Series screen resistor; (*b*) screen voltage supplied from voltage divider.

The simplest arrangement is to supply the screen voltage from a series dropping resistor and a bypass condenser, as shown in Fig. 8·10*a*. If signals with a very small duty ratio are applied to such a circuit, the average screen current will be almost unchanged and so will the drop in the series resistor and hence the voltage at the screen. Thus the plate current in the absence of signal voltage will be nearly constant. The magnitude of the limited output of such a stage is the product of the load resistor R_1 and the plate current in the absence of signal. Therefore anything that changes plate current in the absence of signal changes the size of the limited output, an effect that is usually undesirable. In the circuit under consideration, increase of duty ratio of the signal will have such an effect, since it will decrease the average screen current and hence increase the screen voltage and thereby increase the plate current during the intervals between signals.

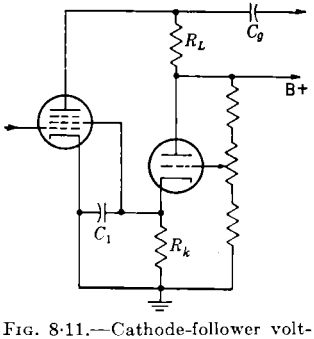


FIG. 8·11.—Cathode-follower voltage-stabilized limiter.

This effect may be reduced somewhat by the addition of a second resistor R_2 as shown in Fig. 8·10*b*. If the current taken by R_1 and R_2 in series is large compared with the screen current, there will be little shift of screen voltage with duty ratio and consequently little change in size of output.

A further step in the reduction of this effect may be made as shown in Fig. 8·11 in which a cathode follower is used to supply the screen voltage.

In this way the internal impedance of the screen supply may be reduced to a few hundred ohms. To obtain this impedance with a straight resistance voltage divider would require much more power. The screen voltage is easily varied by changing the grid voltage of the cathode follower. This grid voltage should be well filtered to avoid hum and pickup that would be transmitted through the main video channel. The condenser C_1 from screen to ground may have a value of 0.001 μf.

The Influence of the Input Coupling Circuit on Limiter Performance.—A second factor that enters in the change of size of limited output with duty ratio is the type of input coupling circuit from the detector to the limiter. As illustrated in Fig. 8·12, the important point is whether there is d-c coupling or condenser coupling. Suppose that the input signal is a series of pulses (Fig. 8·12*a*) in which the spacing between pulses is not large compared with their width and the duty ratio is therefore appreciable. With d-c coupling the voltage excursion on the grid of the video tube will be between the same points, whatever the duty ratio. With condenser coupling, however, the *average* voltage at the grid of the limiter tube will remain fixed (assuming that the train of pulses is long compared with the time constant of the coupling circuit). The situation will then be as shown in Fig. 8·12*a*, *b*, and *c*. Note that there is a break in the time scale; the first pulse on the left is assumed to be the initial one of a long train of pulses that has been preceded by a long interval of no signal. The rest of the pulses shown come well after the beginning of the train, and new equilibrium conditions have been reached. Note that the grid voltage and thus the plate current in the intervals between pulses are now greater than before and the limited output is therefore greater.

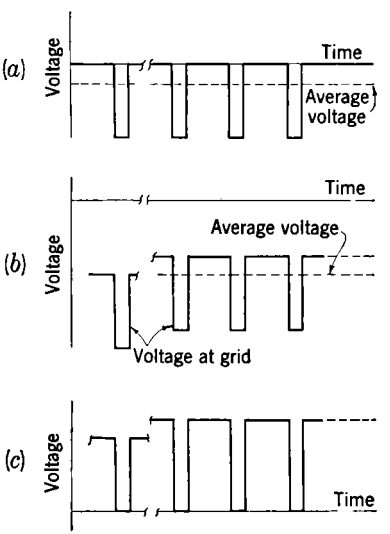


FIG. 8·12.—Effect of condenser coupling on limit level. (*a*) Input signals; (*b*) signal after condenser coupling; (*c*) plate current of limiter tube.

If the limiter tube is run at zero bias instead of having some negative bias as indicated in Fig. 8·12*b*, grid current will be drawn in the intervals between pulses and will tend to prevent the shift in potential. A separate diode, often known as a "d-c restorer," may also be used for this purpose (see Sec. 8·9).

Special Limiters.—It is sometimes possible to combine in one stage the functions of limiter and output tubes. The circuit shown in Fig. 8·13 provides the limiting action and at the same time is capable of driving a terminated 75-ohm cable with an 0.8-volt negative signal. The grid input could also be direct-coupled circuit, and the screen could obtain its voltage from the other sources previously mentioned. Since the 75-ohm termination to the cable provides degeneration in this circuit, the signal voltage required for cutoff is somewhat higher than would be the case with no degeneration. Thus it becomes somewhat harder to obtain a high ratio of maximum detector output to cutoff voltage. As mentioned previously, a high value of this ratio is desirable in the presence of certain types of interference. However, this circuit saves one tube, about 4 watts of heater power, and 4.5 watts of high-voltage power as compared with a circuit using a separate output tube capable of 2 volts output into the 75-ohm line.

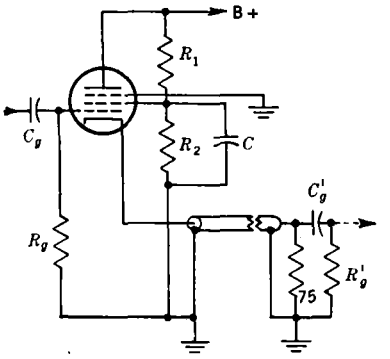


FIG. 8·13.—Combined limiter and output circuit.

Another special function that is sometimes required is a means of electronically changing the magnitude of the limited output signals.

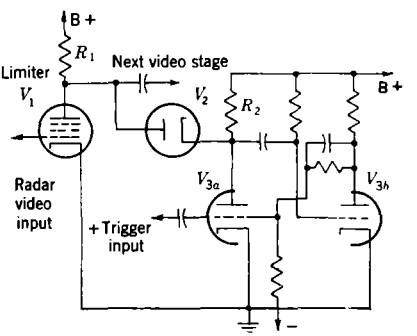


FIG. 8·14.—Circuit for changing limit level electronically.

Manual variation of this quantity may be obtained by merely changing the plate-load resistor of the limiter tube. The circuit shown in Fig. 8·14 effectively provides a means of shunting an extra resistance across the load resistor by electronic means. It was originally applied to a radar receiver in which it was desired to have the limited signals alternately large and small after successive transmitted pulses. When

V_{3a} is nonconducting, V_1 acts as a normal limiter with load resistor R_1. The circuit constants are so chosen that when the multivibrator consisting of V_{3a} and V_{3b} flips to the condition in which V_{3a} is conducting, the plate potential of V_{3a} comes to nearly the same value as that of V_1 in the absence of video signals. Therefore, R_1 is effectively shunted by the parallel combination of R_2 and the plate resistance of V_{3a} (assuming the diode to be perfect).[1] In the receiver to which this circuit was applied, simultaneous variation of i-f gain was produced.

8·8. Video Output Circuits.—If the signals furnished by the receiver are to be used at a distance of more than 3 or 4 ft from it, it is almost imperative to transmit them by way of a terminated line. This may be an ordinary coaxial cable, in which case the characteristic impedance will ordinarily not be more than about 110 ohms, or in some special cases it may be a continuously loaded coaxial line[2] with a characteristic impedance of 1000 ohms. Since no generally useful impedance-transforming systems are available for the whole video band, the characteristic impedance of the cable determines the load that is presented to the output tube. The value of the load is thus much lower than would ordinarily be chosen on grounds of bandwidth or power-output capability. Under these conditions the current that the output tube can furnish determines its performance. The size of the signals that can be furnished is the product of the characteristic impedance of the cable and the current capability of the tube. Thus a requirement for large output signals becomes very expensive in terms of power, since the large current must be furnished at the full B+ voltage.

The output cable may be connected in either plate or cathode circuit of the output tube. In the first case the B+ voltage appears on the center conductor, since a prohibitively large condenser would be required to maintain low-frequency response if capacity-coupling into the low impedance presented by the cable were attempted. For this reason the output is ordinarily taken from the cathode circuit as a cathode follower. The voltage gain of this arrangement is less than 1 and is given approximately by the expression

$$\mathcal{G} = \frac{g_m R_c}{1 + g_m R_c},$$

[1] It would not do for this particular application to produce variations of limited output by changing the screen voltage, since this would produce a condition that would move the bottom of the limited signals and leave the top fixed, whereas the reverse behavior is necessary.

[2] This cable was standardized under the type number RG-65/U (see Vol. 17, Chap. 1, of this series). It is satisfactory mechanically, but its fairly high attenuation makes runs of more than about 50 ft impractical. In addition the attenuation to frequencies above 2 or 3 Mc/sec rises fairly rapidly.

where R_C is the characteristic impedance of the cable. The output tube may be either a triode or tetrode or a tetrode or pentode used in triode connection. Sometimes it is possible to make savings in the power supply by the use of a tetrode, since the plate voltage may be supplied from a source with comparatively little filtering, whereas the screen supply, furnishing much less current, is given full filtering. Types 6AG7 and 6L6 are frequently used for this purpose.

The video signals found in radar or other pulse applications are usually not symmetrical, and it is therefore possible to make worth-while savings in power consumption by the proper choice of polarity for the output signals. The output tube is arranged to run at comparatively low current in the absence of signal and is turned on by the signal, thereby giving a positive output signal in the usual cathode-follower circuit. The use of a d-c restorer (see Sec. 8·9) permits use of a lower value of quiescent current without having the tube driven to cutoff in the intervals between signals of high duty ratio. It is not always possible, however, to use a d-c restorer in radar receivers because of its effect on performance in the presence of interference.

If the cable run from the receiver to the point where signals are used is short enough to permit the use of an unterminated line, it is still desirable to use coaxial cable to provide proper shielding. Such short lengths of cable act as a shunt capacitance equal to the capacitance per foot times the length in feet. The effect of this capacitance on high-frequency response may be calculated by the usual expression

$$f_c = \frac{1}{2\pi R_L C},$$

where R_L is the load impedance. If such an unterminated line is used with a cathode follower, the effective value of R_L may be considerably different during positive-going and negative-going parts of the signal. When the grid of the tube is being driven positive, the output impedance of the cathode follower is approximately $1/g_m$. This impedance appears in parallel with whatever resistor is present. On the other hand, when the grid is driven negative, the output tube may be nearly cut off. Its output impedance is then high, and the effective load resistance is determined largely by the resistor in the circuit. If a square positive pulse is applied to such a circuit, this may cause the output wave to rise much faster than it falls.[1]

8·9. D-c Restorer.—The d-c restorer, as used in the video amplifiers under discussion here, acts to maintain the signal baseline, or the bottom of the signals, at a fixed potential as the duty ratio changes. It is applicable to either positive or negative signals.

[1] See, for example, H. Goldberg, *Proc. IRE*, **33**, 778–782 (1945).

Figure 8·15*a* shows a positive signal with a duty ratio of about 10 per cent being applied to an amplifier whose standing current is about one-third the maximum signal current. When the duty ratio of the signal is increased to 50 per cent and the restorer is not present, the signal will appear as in Fig. 8·15*b*. Figure 8·15*c* shows the effects of the d-c

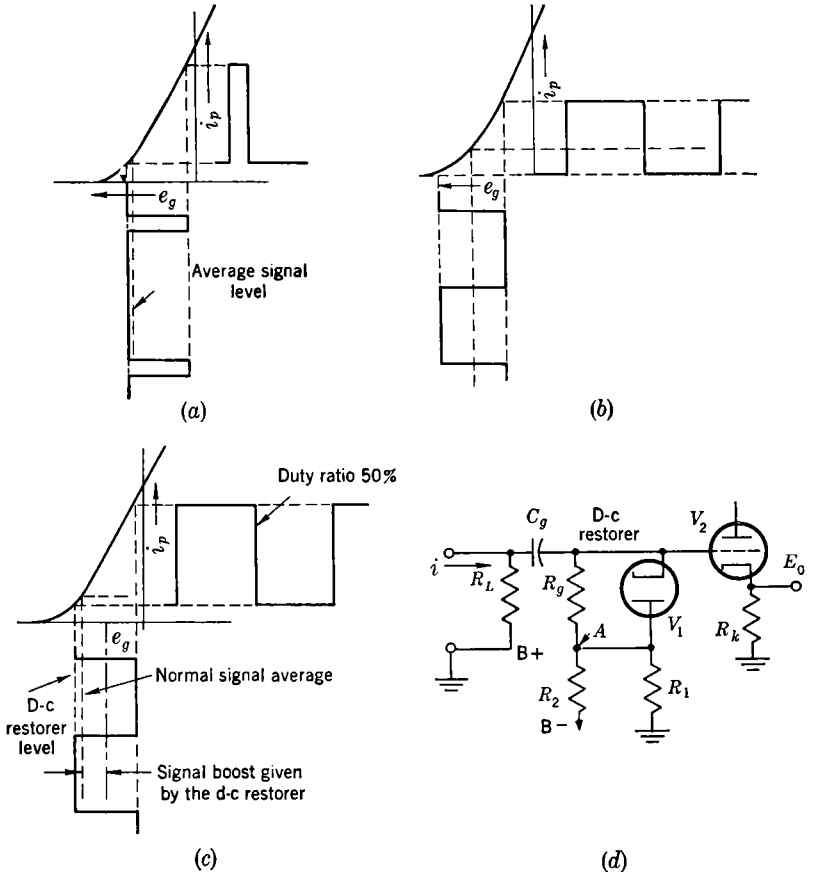


FIG. 8·15.—Action of a d-c restorer. (*a*) Input signal with duty ratio of 10 per cent; no d-c restorer; (*b*) input signal with duty ratio of 50 per cent; no d-c restorer; (*c*) input signal with duty ratio of 50 per cent; d-c restorer added; (*d*) circuit of d-c restorer.

restorer in preventing the signal from backing off the characteristics of the amplifier tube. A typical circuit is shown in Fig. 8·15*d*. Resistors R_1 and R_2 make up a voltage divider for furnishing the desired bias voltage for V_2; in the absence of signal the grid of V_2 is at the same potential as Point A in the diagram. If a positive signal is now applied, the grid of V_2 will be driven more positive than Point A. Since this is in

the nonconducting direction for the diode V_1, this tube will have no effect on the circuit. If a negative signal is applied, however, the diode will become conducting and charge will flow on to the coupling condenser C_g. If the diode were perfect, that is, if it had no internal impedance, and if the impedance from Point A to ground were zero, the grid of V_2 could never go more negative than the potential of Point A. In the actual circuit it does go somewhat more negative. The extra charge that has appeared on C_g acts to shift the average potential of the grid of V_2 enough more positive to cause only the tips of the negative-going signals to go more negative than Point A. Thus the base of the signal remains approximately fixed as the duty ratio increases.

Caution should be used in applying a d-c restorer to a circuit in which low-frequency compensation is used, because the compensation depends on the balance between the time constant made up of the coupling condenser and the grid resistor and the compensating time constant. A nonlinear element in the grid circuit changes that time constant and thus affects the compensation. A d-c restorer may be unsatisfactory in a radar receiver under conditions where a c-w carrier is introduced into the receiver along with the desired signals. This causes the signals in the video amplifier to become two-sided, that is, to extend both above and below the baseline. At the detector proper the signal beating with the c-w carrier produces an output that goes above and below the d-c output due to the carrier; but when the signal passes through the condenser coupling into the video, the d-c level becomes the baseline and thus the signal appears to go both positive and negative. The d-c restorer sets the level so that only the negative-going peaks so produced extend down to the proper baseline. This action interferes with the proper functioning of the radar indicators.

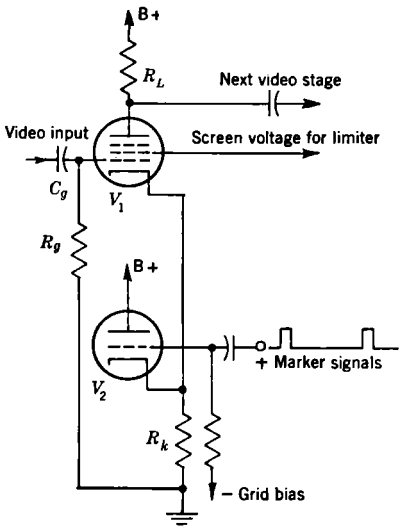


FIG. 8·16.—A mixing circuit in which range marks, heading markers, etc., are mixed with video signals so that all limit at the same level.

8·10. Introduction of Special Signals.—It is often desirable to be able to mix some other signals with the normal video signals. In the case of radar systems such signals might be range marks, heading markers, angle

markers, signals from other systems, etc. Since all the receivers described in this volume incorporate a limiter circuit in the video amplifier, it is necessary to decide whether the output due to the special signals is to be limited at the same value as the regular signals or is required to be larger. In this latter case the special signals must be introduced after the limiter.

Suppose that the limiter in the first video stage is the only limiter in the video system and that the special signals are to be limited at the same time as the regular signals. The circuit shown in Fig. 8·16 is then applicable. The regular video signals from the detector are applied to the grid in the normal way, and the marker signals are introduced through the cathode follower V_2. The common cathode resistor is generally of

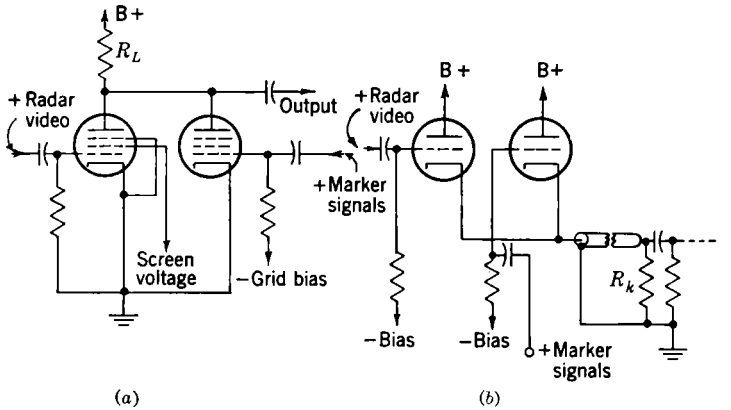


FIG. 8·17.—Marker mixing. (*a*) On common plate resistor; (*b*) on common cathode resistor of cathode-follower circuit.

low resistance (usually 100 ohms). The cathode follower V_2 is biased so that with no signal it does not conduct any current. The positive marker signals on the grid must be sufficiently large to develop enough voltage across the common cathode resistor to cut off the limiter tube. The marker signals should be provided with a control so that their intensity can be controlled; a switch should also be present to permit removal of the markers if desired.

Figure 8·17 shows two circuits that have been used for introducing the markers after the limiter. The circuit in Fig. 8·17*a* mixes the signals in a common plate resistor. Since both tubes are pentodes and can be considered constant-current generators, the output voltage for small input signals is

$$E_0 = (E_1 g_{m_1} + E_2 g_{m_2}) R_L,$$

where E_1 and E_2 are the input video signal and marker signal respectively and g_{m_1} and g_{m_2} are the transconductances of the two tubes. The tube

introducing the markers is normally biased to cutoff and conducts only when the marker pulse is on its grid.

In Fig. 8·17*b* the signals are mixed in the common cathode resistor, and again the tube that mixes the markers is normally biased to cutoff, conducting only when the markers are applied. If it is assumed that the

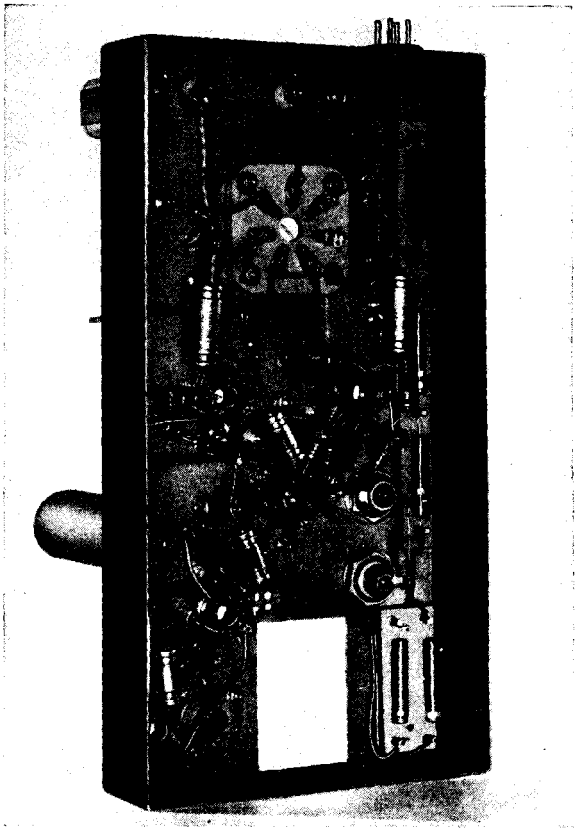


FIG. 8·18*a*.—Photograph of a wide-band amplifier.

μ of the two tubes is the same, then it can be shown that the output voltage E_0 is

$$E_0 = (E_1 + E_2) \frac{g_m R_K}{1 + 2g_m R_K \left(\frac{1 + \mu}{\mu}\right)},$$

where E_1 and E_2 are the video voltage and marker voltage respectively. This equation is true only when E_1 and E_2 are small signals.

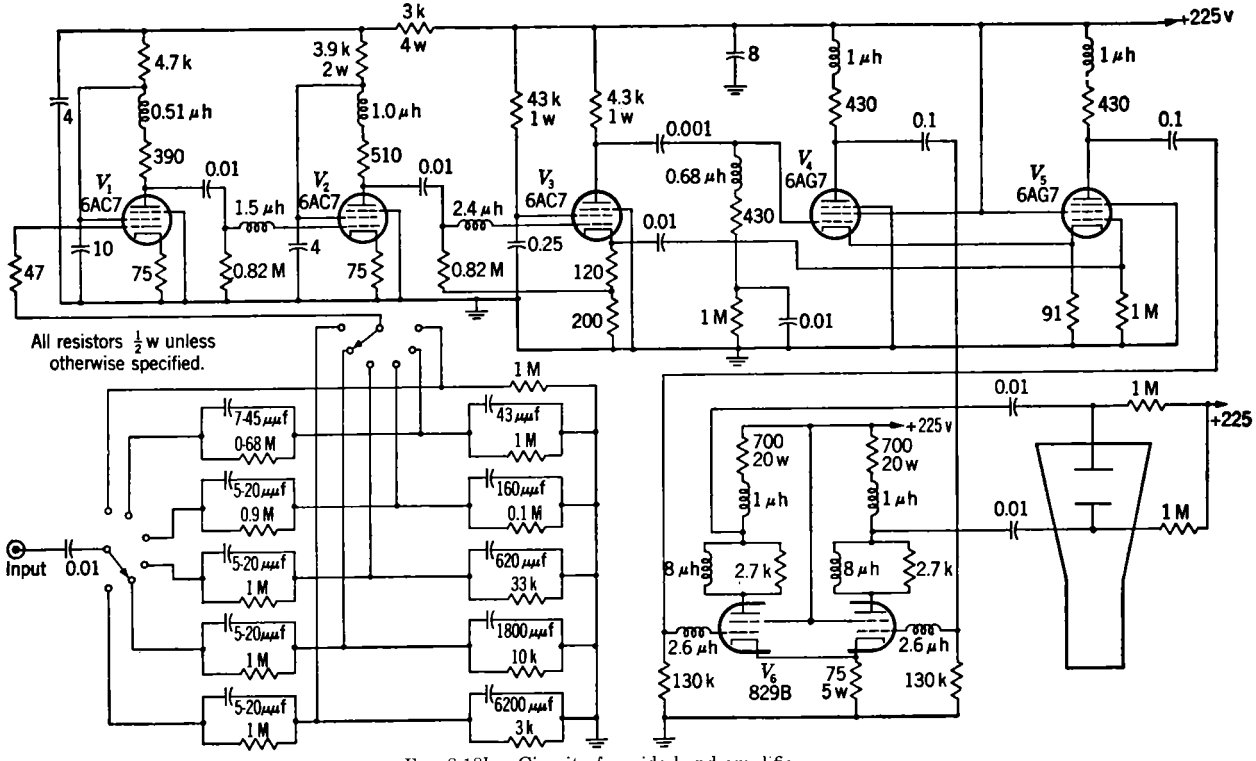


FIG. 8·18*b*.—Circuit of a wide-band amplifier.

8·11. Typical Video Amplifiers.—Figure 8·18*a* shows a photograph of an amplifier of unusually wide band, which is used as part of a test oscilloscope. The frequency response is down 3 db at 18 Mc/sec and at 11 cps. Exceedingly careful placement of parts and reduction of wiring

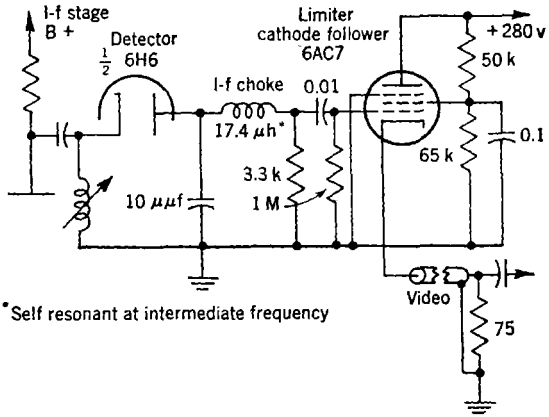


FIG. 8·19.—Detector and video circuit of a medium-power microwave radar.

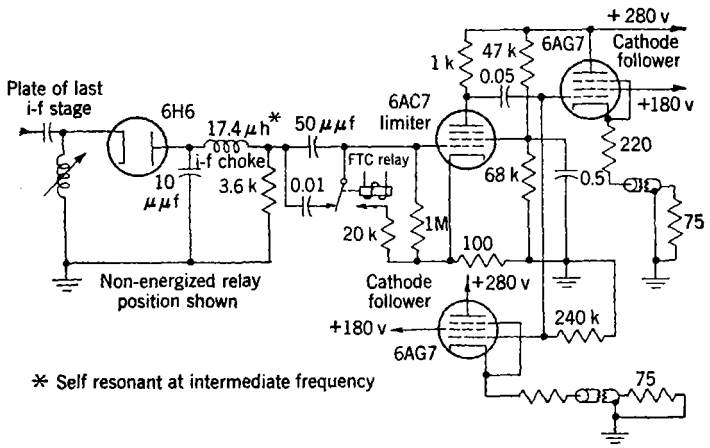


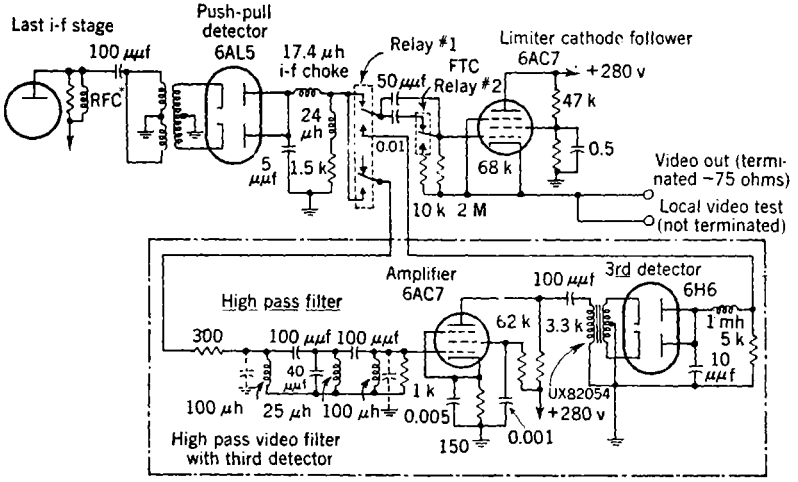
FIG. 8·20.—Two-stage video amplifier in a receiver.

to an absolute minimum are required in order to obtain this wide response. The circuit diagram for this amplifier is shown in Fig. 8·18*b*.

The circuit illustrated in Fig. 8·19 is the video amplifier from the receiver described in Chap. 13. This circuit combines the limiter and cathode-follower operations as described earlier in this chapter. The set is designed to operate on a medium-power radar system where 10

to 15 miles of continuous signal is the maximum expected. A video signal of about −3.0 volts at the grid of the output tube is required to cause the tube to limit, and a signal of about −0.9 volt is transmitted through the line.

Figure 8·20 shows a conventional two-stage video amplifier in a receiver. The cathode followers deliver +2.0-volt signals to the terminated cables. A short- as well as a long-time-constant coupling circuit is provided, and a relay provides the means of switching circuits.



*Self resonant at intermediate frequency

FIG. 8·21.—Detector and video circuit for special receiver.

The circuit illustrated in Fig. 8·21 provides the long-, short-, and high-pass video filter with a third detector for the coupling circuit. The relay connections provide a selection of any of the three circuits as well as a combination of the high-pass filter and FTC circuits. The numerous relay contacts and component parts connected with this circuit make it necessary to use extreme care in the construction and selection of components, particularly the relays, in order to maintain the video bandwidth required.

CHAPTER 9

GAIN-CONTROL CIRCUITS

By S. N. Van Voorhis

9·1. Introduction.—Of the characteristics of an amplifier subject to control during operation, gain and bandwidth are most frequently varied. Means for accomplishing gain control will be discussed in this chapter. This discussion will be limited to circuits applicable to i-f amplifiers, video circuits being treated separately in Chap. 8. There are at least four ways in which it may be desired to have gain vary:

1. As a function of a knob setting, ordinary manual gain control.
2. As a function of the strength of some particular received signal, automatic gain control, AGC.
3. As a prescribed function of time, sensitivity time control, STC.
4. As a function of the *immediately* preceding signal strength, instantaneous automatic gain control (IAGC).

9·2. Manual Gain Control.—This is usually accomplished by varying the potential applied to one or more electrodes of the amplifier tubes in one or more stages. The following questions must be answered in the course of the design:

1. To which electrode or electrodes should the control be applied?
2. To how many stages should it be applied?
3. If only to part of the whole amplifier, to which stages?

Answers to these will be found from considerations of the total range of gain reduction required, of the overload characteristics necessary, of linearity for small and large signals, and of the characteristics of the tubes being controlled.

The amplifiers under discussion, of necessity, use high-transconductance pentodes such as 6AC7, 6AK5 etc., with load impedances small compared with the plate resistance of the tube. Thus, the gain of a stage is proportional to the transconductance of the tube, which is in turn at least roughly proportional to the plate current. Thus anything that changes plate current is at least potentially useful for control of gain. This shows at once that variation of plate-supply voltage alone will not be effective but that either screen-voltage or grid-bias variation will produce the desired result. It is immaterial at this point whether

increased grid bias is obtained by applying a negative voltage to the grid or a positive voltage to the cathode, though later considerations will show that the first is usually preferable.

If the transconductance of the tube is reduced to zero, the signal transmission will not vanish altogether, because of various modes of spurious coupling between the input and output circuits, such as grid-plate capacity within the tube, capacity across the socket, coupling of circuit elements, and wiring, etc. These last are, of course, greatly dependent on details of construction and layout but will be minimized in a well-designed amplifier. In an ideal design the internal plate-grid capacity alone is effective. For a 6AK5 tube this has a maximum value of 0.02 $\mu\mu$f, corresponding to a reactance of 265,000 ohms at a frequency of 30 Mc/sec. Thus if the plate-load impedance is around 2000 ohms, the ratio of input to output voltage will be about 100, corresponding to a loss of 40 db for the stage. Since it would have a gain of around 20 db under normal operating conditions, this is equivalent to a range of gain of 60 db. Very similar results would be obtained if the calculations were made for a 6AC7 or any other tube suitable for a wideband amplifier. In actual practice, a figure of 40 db for the total available range of control per stage is probably more conservative. Thus it will be necessary to apply control voltage to at least two stages in most cases. Linearity considerations may call for control on a still larger number of stages. If a very large signal is being applied to the input of the amplifier, the input to the first stage being controlled may be fairly large, of the order of some volts. To reduce the gain by such large factors it will then be necessary to bias the tube beyond cutoff (assuming for the moment the use of grid-bias control) by an amount comparable to the peak positive signal swing. Then a signal somewhat smaller may not reach the conducting region of the tube at all and so will be discriminated against to a considerable degree. In a great many pulse applications this is of no importance; where it is, the solution is to apply control voltage to enough stages so that no one need be biased as far as cutoff. If screen-voltage control is used, a still more serious difficulty arises from the decrease in the signal capable of producing overload in the controlled stage. This may have the result that two fairly strong signals of different input levels may both overload the controlled stage if its gain is reduced enough. Then both will appear in the output having the same strength, and no manipulation of the gain control will separate them. This effect is so serious in most pulse applications as to rule out almost completely the use of screen-voltage gain control.

In deciding which stages to control, the guiding principle is again the avoidance of overload ahead of the point at which control is applied. In a receiver for amplitude modulation, this is of paramount importance;

in others it is highly desirable. This calls for application of gain control early in the receiver. In the case of a microwave superheterodyne having an i-f amplifier working out of a crystal mixer with no r-f amplification, it has been found desirable to begin with the second i-f stage and control as many of the following stages as are needed to meet other requirements. The crystal mixer will overload itself before its output becomes sufficient to cause overload in the first i-f stage. Furthermore, as shown previously (Sec. 4·1) the ultimate sensitivity of such a receiver is governed very largely by the first i-f stage; it is therefore convenient to set up operating conditions for this tube for best signal-to-noise performance and to leave them undisturbed. For i-f bandwidths of the order of a few megacycles per second the gain of the first stage will be low enough so that the maximum signal which the mixer will give is insufficient, when amplified by the first stage at full gain, to drive the grid of the second stage positive, particularly when control by means of grid bias is applied to this second tube.

In amplifiers having still wider bands, the stage gain may be reduced to the point where this maximum signal will be insufficient to produce overload at the grid of the third tube, with the first two stages operating at full gain. Also both the first and second stages may make significant contributions to the noise output. In this case it may be found desirable to begin gain control at the third stage. If a tube mixer is used in a microwave receiver, two changes are made: (1) The maximum output of the mixer is considerably increased, and (2) the amount of noise contributed by the mixer is increased to the point where the contribution of the first i-f tube becomes somewhat less important. These together point to application of control to the first i-f stage. At frequencies where r-f amplification is used, it will usually be desirable to apply control to these stages, particularly if more than one is used. Arguments on noise and maximum received signal similar to those above may preclude control of the first stage.

In amplifiers that are subjected to wide ranges of signal strength, such as those encountered in many pulse applications, considerable attention must be paid to transient effects on gain following very strong signals. These arise primarily from the nonlinearity of various parts of the amplifier at high signal levels. This nonlinearity produces a shift in average current or voltage, which may appear on various capacitances in the circuit. After the strong signal is removed, the voltages return to normal at a rate governed by the time constant of the circuit involved. If this time constant is short compared with the reciprocal of the i-f or video bandwidth, currents and voltages in the amplifier will be back to normal by the time the output of the receiver has fallen to normal and no ill effects will be observed. It may also be possible to make such time

constants quite long and reach another tolerable condition. For an amplifier a few megacycles per second wide it then becomes advisable to avoid time constants in the neighborhood of a microsecond. This has various effects on gain-control circuits. For one thing, the type of gain control frequently used in narrower-band amplifiers, consisting of a variable resistance in the cathode circuit of one or more stages, sometimes with a bleeder resistor from B^+ in addition, now becomes less desirable because of the difficulty of avoiding the wrong RC values. For example, if the cathode bypass condensers are 0.001 μf, resistance values from about 500 to 20,000 ohms should be avoided, thus making it less straightforward to secure considerable control range per stage. A second less onerous restriction arises when control is secured by grid bias. Here again it is likely that a bypass condenser of 0.001 μf will be used for the control voltage. A potentially harmful time constant will then be formed by this bypass condenser and the internal impedance of the control voltage supply. It will usually be found that if this impedance is kept below 1000 or 2000 ohms, no trouble will result.

One additional means of gain control in a pentode amplifier stage is the application of a negative voltage to the suppressor grid. This reduces plate current by the mechanism of turning electrons around and sending them to the screen, thereby increasing the screen current by very nearly the amount by which the plate current is reduced. Screen dissipation is likely to be increased above the danger point unless the normal operating currents are held down to an undesirably low value. Furthermore, this means of control is obviously impossible in the case of a tube such as the 6AK5 in which the suppressor is not brought out separately.

For the case of a moderately wide i-f amplifier such as is used in a microwave radar system, the following answers to the original questions have been reached:

1. Control shall be obtained by applying a negative bias to the grid.
2. This will occur in at least two stages, possible three.
3. These stages will be the second and third and possibly the fourth.

A typical example of such a receiver will be found in Chap. 13, which shows how these answers are incorporated in an actual design.

9·3. AGC Circuits.—The useful output dynamic range of a receiver of ordinary design seldom exceeds 20 db. The limits are set on the one hand by overloading in the i-f amplifier and on the other hand by poor functioning of the second detector for signals small enough to cause it to act as a square-law device. Therefore, if information is to be extracted from the amplitude of a signal whose input dynamic range is greater than 20 db, some control of the gain of the amplifier must be provided. If rapid fluctuations of the input level are expected, or if it is desired to hold the

average output level accurately constant, it may be desirable to provide such gain regulation automatically by electronic means. Such circuits are known as "automatic gain control" (AGC) circuits and have been used for many years in receivers of various types.[1] Attention here will be focused on the special problems introduced by the application to pulse signals, by the necessity of maintaining good transient response, and by the high accuracy of control required in some radar applications.

An AGC circuit is an example of a feedback loop and is accordingly subject to the limitations imposed by stability requirements. It should be noted that the loop is closed by way of the input signal to the receiver or by noise if the gain of the amplifier in the absence of control is sufficiently high. A block diagram of the elements of an AGC circuit is shown in Fig. 9·1. The circuit is a conventional superheterodyne with

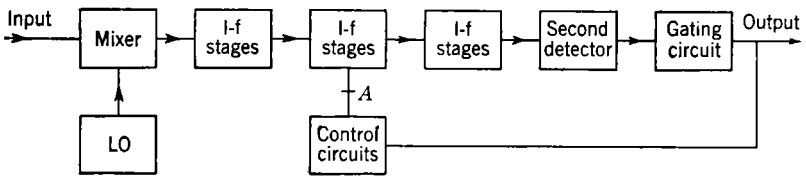


FIG. 9·1.—Block diagram of receiver with AGC.

the addition of the gating and control circuits. The gating circuits are made necessary by what may be called the "time-sharing" property of pulse systems. The desired signals may be present for only a small fraction of the total time (given by the duty ratio); other signals occupying the same radio frequency may be present in the intervening time. It is desired to have the gain of the system controlled entirely by the amplitude of the wanted signals even though the unwanted ones may be much larger. The function of the gating circuit is to select the desired signals so that only they are applied to the control circuit. This action may be represented as shown in Fig. 9·2. Various signals are shown appearing at the input terminal of the receiver, the desired ones being indicated by arrows. The gating circuit is arranged to have zero transmission except during a small interval of time containing the desired signals, thereby singling them out to pass on to the output and to the control circuit. An important prerequisite of such a procedure is accurate knowledge of the time of arrival of the desired signal. In radar systems, the signal of interest is the echo from a particular object. This echo follows the transmitted pulse by a time determined by the range of the object. Even for a rapidly moving object, the change in range or

[1] See for example: F. E. Terman, *Radio Engineers' Handbook*, McGraw-Hill, New York, 1943, p. 639.

time from one pulse to the next is small enough to permit such gating action. Another circumstance in which gating may be applied is that in which the desired signals occur at an accurately known repetition rate.

The presence of the gating circuit has a fundamental effect on the consideration of the stability of the system. The feedback loop is now closed only during the time that the gating circuit transmits, in contrast to the usual case in which it is closed continuously. There have been two mathematical approaches to this problem, one by MacColl, the other

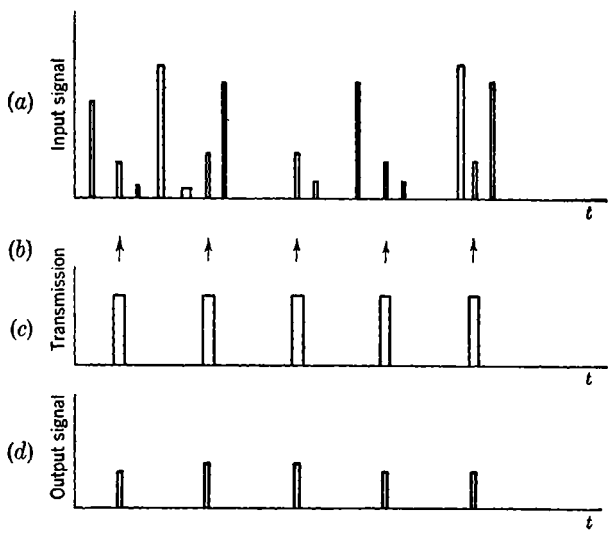


FIG. 9·2.—Action of gating circuit. (*a*) Input signals; (*b*) desired signals; (*c*) transmission of gating circuit; (*d*) output signals.

by Hurewicz.[1] It will be recalled that in the conventional treatment of a continuous feedback circuit, stability is investigated by means of a plot of the complex transmission function of the system on the complex frequency plane. The condition of stability is found to be the absence of roots in the half plane to the right of the line whose real part is equal to -1. To establish this condition, the transmission function along a contour given by the real frequency axis may be plotted on the impedance plane. If the curve so obtained encloses the point $(-1,0)$, the system is unstable (Nyquist criterion.)

For the case of the intermittently closed feedback loop, which he terms a "sampling servomechanism," MacColl derives a modified trans-

[1] L. A. MacColl, *Fundamental Theory of Servomechanisms*, Van Nostrand, 1945, Chap. X; Vol. 25, Radiation Laboratory Series.

mission function, subject to the same limitations as the location of roots to which Nyquist's criterion may be applied in its usual form. Hurewicz, on the other hand, deals with the ordinary transmission function and finds a new contour on the complex frequency plane enclosing a region outside which there must be no roots. A plot on the impedance plane of the transmission function along this contour furnishes a criterion of stability completely analogous to the usual one.

Another difficulty in the study of the stability of an AGC system arises from the circumstance that the feedback loop is closed by way of the input signal. This signal will almost certainly vary over a wide range, and with it the gain around the AGC loop may vary. It is, therefore, not enough to establish stability at a single level of input signal. A procedure by which a system such as that shown in Fig. 9·1 might be studied is the following. Suppose that the loop is broken at point A and a convenient steady voltage is applied to the i-f amplifier terminal so exposed. Now apply to the receiver an input signal whose level is so adjusted that the output of the control circuit is just equal to the voltage applied to the amplifier, in other words so that the connection at A could be remade without altering conditions. If now a small test signal of variable frequency is applied to the i-f amplifier at A in addition to the steady voltage, the output of the control circuit will contain a component of the same frequency, differing in general in both amplitude and phase from the input signal. The transmission function is the quotient of these two signals. The process may then be repeated for a different value of steady voltage or input signal level.

For usual values of circuit constants it will be found that the part of the feedback loop involving the i-f amplifier contributes practically nothing to the *change* of gain or phase with frequency in the range of interest. Its effect is restricted to change in gain with input signal level. This circumstance makes possible a considerable simplification in the process of investigating stability, for it is necessary only to make a complete study of the system as a function of frequency at a single input level. From this it is possible to ascertain the frequency at which a phase shift of 180° has occurred, as well as the change in loop gain required to produce instability. A separate investigation of the gain as a function of input level may then be made at any convenient test frequency.

A still further simplification may be made by relating the effective gain of the i-f amplifier portion of the feedback loop to the characteristics of the i-f tubes that are controlled. Suppose for the sake of definiteness that gain control is accomplished by means of control-grid bias on one or more i-f stages. If now the feedback loop is broken at A and a steady input signal applied to the amplifier together with the appropriate bias voltage, a straightforward calculation of the gain may be made.

Let E_i = input signal,
E_o = output signal from second detector,
E_B = bias voltage applied to the amplifier,
E_C = output voltage from control circuit,
$\mathcal{G}$ = gain of i-f amplifier and second detector.
Then

$$E_o = \mathcal{G}E_i. \tag{1}$$

The total gain $\mathcal{G}$ may be written as the product of individual stage gains

$$\mathcal{G} = \mathcal{G}_0 \prod_a g_m R^a, \tag{2}$$

where $\mathcal{G}_0$ represents the gain of the stages not controlled. Suppose now that an increment ΔE_B is applied, producing a change in the transconductance Δg_m. The change ΔE_o will then be

$$\Delta E_o = \mathcal{G}E_i \sum_a \frac{\Delta g_m^a}{g_m^a} = E_o \sum_a \frac{\Delta g_m^a}{g_m^a}. \tag{3}$$

It is ordinarily desirable that E_o, the output of the receiver, should be moderately large and at the same time should not vary appreciably with change in input signal level. This last requirement calls for high gain around the feedback loop, which would in turn limit E_o to a small value if applied directly. To permit simultaneous fulfillment of these conflicting requirements, a *delay voltage* is usually applied in the control circuit in such a way that feedback is effective only when E_o exceeds a fixed value E_o^0. The relation between E_C and E_o may then be expressed as follows:

$$\begin{aligned} E_C &= k(E_o - E_o^0) && E_o \geq E_o^0 \\ &= 0 && E_o < E_o^0. \end{aligned} \tag{4}$$

With the help of this relation Eq. (9·3) becomes

$$\Delta E_C = kE_o \sum_a \frac{\Delta g_m^a}{g_m^a}. \tag{5}$$

If now the loop gain of the feedback network is denoted by $\mathcal{G}_L$, Eq. (5) may be written

$$\mathcal{G}_L = kE_o \sum_a \frac{\Delta g_m^a}{g_m^a \, \Delta E_B} \tag{6}$$

or, on passing to the limit $\Delta E_B \to 0$ and identifying ΔE_B with the change in grid voltage Δe_g of the controlled stages

$$\mathcal{G}_L = kE_o \sum \frac{d(\ln g_m)}{de_g} \tag{7}$$

Thus if the logarithm of the transconductance of a tube is plotted as a function of grid voltage, the slope of the curve so obtained will be a measure of the loop gain for that tube in the AGC system. Figure 9·3*a* shows such a plot for three 6AK5 tubes selected from a lot of 27, and Fig. 9·3*b* shows a plot of the slope of the curves against grid voltage. Figure

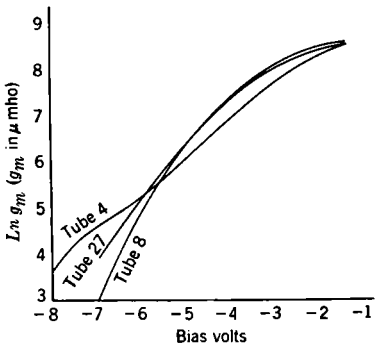


FIG. 9·3*a*.—Typical curves for single tubes, type 6AK5.

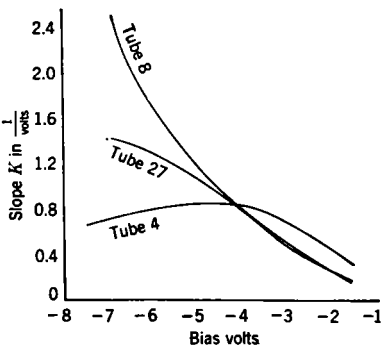


FIG. 9·3*b*.—Typical curves for single tubes, type 6AK5.

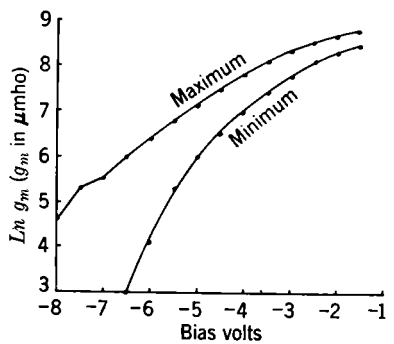


FIG. 9·4*a*.—Results of measurements of 27 tubes, type 6AK5.

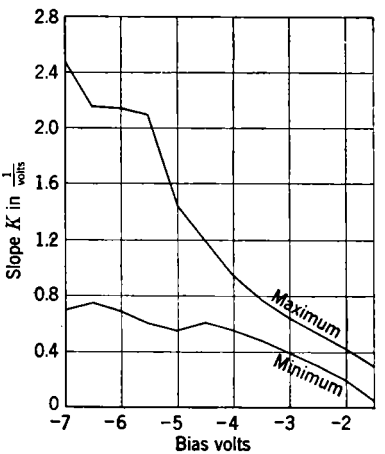


FIG. 9·4*b*.—Results of measurements of 27 tubes, type 6AK5.

9·4*a* and *b* shows the limiting values for both sets of curves in the whole set of 27 tubes. This set of data was obtained in the course of the development of the receiver described in Chap. 15, to which reference should be made for further details.

It is apparent from these curves that if the bias of a controlled stage is subject to wide excursion, the change in gain around the feedback loop

will be considerable. These excursions may be reduced by applying the control to a greater number of stages, thereby reducing the safety factor that must be provided to ensure stability under all conditions.

It was stated previously that the i-f amplifier usually contributes little to the frequency characteristics of the feedback loop. This may be made clear by noting that the AGC voltage is applied as a modulation to the input signal of the receiver, thereby producing sidebands on either side of the input carrier. The effect of the i-f amplifier on these sidebands may be represented as approximately equivalent to the effect on the original AGC signals of a low-pass circuit whose frequency characteristic about zero frequency is the same as that of the i-f amplifier about its midfrequency. The rest of the AGC loop will ordinarily be much narrower than this. If, however, an attempt should be made to provide a fast acting AGC system, the situation is changed.

Suppose, for example, that an attempt is made to secure sufficiently rapid AGC to permit control of the gain of the receiver to be completed before the end of the pulse that initiates the controlling action. To do this, the entire feedback loop must transmit a band of frequencies extending to a substantial fraction of the reciprocal of the pulse length τ. If the receiver bandwidth is approximately matched to the length of the pulse signals, each stage of the receiver has an i-f bandwidth of approximately $3/\tau$ or is equivalent to a low-pass circuit having a bandwidth of approximately $1.5/\tau$ as far as effects on the feedback loop are concerned. The i-f circuits will therefore be an important influence on the over-all frequency and phase characteristics of the feedback loop. Detailed consideration shows that the achievement of stability in these circumstances calls either for application of feedback around only a single i-f stage, with attendant practical difficulties in securing sufficient loop gain, or, if feedback extends over several stages, for unusably small loop gain. The first alternative is approached in the IAGC circuits to be discussed in Sec. 9·4.

It may be remarked that if AGC action of this rapidity were attainable, mathematical consideration of the circuits could proceed on the usual steady-state basis without regard to the pulse nature of the information. It is apparent, however, that the only method of securing such rapid action is to use a receiver whose bandwidth is considerably greater than would otherwise be chosen.

There are considerable similarities between the problems of AGC and those of AFC, previously mentioned. In particular, the difficulty in securing control of gain within the length of a pulse is completely analogous to the difficulty in controlling frequency within the same time. There is, however, one fundamental difference between AGC and AFC in regard to the requirements placed on circuit action in the intervals between

pulses. Proper AFC action almost always requires that the frequency be maintained constant in the intervals between pulses, in order that the receiver may be ready to accept any signals which may appear. On the other hand, no requirement is placed on the gain in the intervals between pulses.[1] In some receivers, provisions have been incorporated to allow the i-f gain to be at its maximum except for an interval around the selected signal during which time the gain is under the control of this signal. This procedure permits simultaneous use of the receiver at maximum sensitivity in searching for other signals. If the interval in which gain is controlled is longer than the gate already present in the feedback loop (see Fig. 9·1), no change in the action of the circuit will be produced, since the procedure is equivalent to breaking the feedback loop in two places rather than one in the intervals between signals.

If an AGC system is used in connection with signals carrying intelligence in the form of amplitude modulation, it is essential that there be proper correlation between the speed of action of the AGC and the frequency spectrum of the intelligence. A sufficiently low modulating frequency will be entirely removed by the AGC action from the normal output of the receiver. It will, however, appear in the control voltage that is fed back and may, by proper circuit design, be recovered and utilized at that point. In the receiver described in Chap. 15 this procedure is followed for reasons that are given there.

9·4. Instantaneous Automatic Gain Control (IAGC) Circuits.—A situation that arises often enough in radar systems and possibly also in other pulse applications to demand special consideration is that in which interference in the form of a c-w carrier or an amplitude-modulated wave is present. If the amplitude of the interference is small compared with the desired pulse signal, it is ordinarily sufficient to provide adequate gain control as described in Sec. 9·2. Where the relative amplitudes are reversed, however, additional design features may be useful. These features act in such a way as to preserve the incremental gain of the receiver for the desired pulse, while greatly reducing the response to the interfering signal.

Suppose that the desired signal is made up of a carrier frequency f_0, amplitude-modulated by a rectangular pulse. It is convenient to focus attention on the i-f amplifier, and therefore f_0 may be assumed to be in the neighborhood of the intermediate frequency, say 30 Mc/sec. For ordinary pulse lengths, such as 1 μsec, the signal then consists of approximately 30 oscillations. Suppose also that the interfering signal is a c-w carrier of frequency f_1. If $|f_1 - f_0| > 1$ Mc/sec, the envelope of the combination—signal plus interference—will exhibit at least one complete

[1] Unless, of course, one is dealing with a situation such as a television signal in which the pulses serve as a pilot control for the intervening signals.

cycle of the difference frequency $(f_1 - f_0)$. If, however, $|f_1 - f_0| \ll 1$ Mc/sec, the envelope of the combination will be essentially the steady value characteristic of the c-w carrier with a rectangular pulse superimposed on it. According to the relative phase of signal and interference, the resultant may be either larger or smaller than the c-w interference; therefore the envelope may appear to have either a positive or a negative pulse of amplitude equal to that of the signal, or there may be any intermediate condition. If repeated pulses are viewed on an oscilloscope in the presence of such interference, all values of the relative phase between signal and interference are run through, and the pattern appears to be filled in as shown in Fig. 9·5. It is apparent therefore that for part of the time at least, there is a change in the amplitude of the envelope of signal plus interference, which is comparable in magnitude to the amplitude of the signal alone. If this change in envelope can be preserved, the signal may be recovered from the interference.

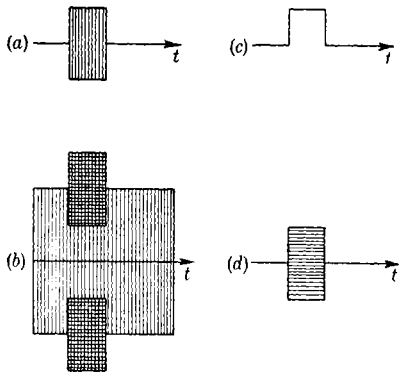


FIG. 9·5.—Waveforms of signal plus interference. (*a*) I-f envelope of signal alone; (*b*) i-f envelope of signal plus interference; (*c*) second detector output, signal alone; (*d*) second detector output, signal plus interference (after *RC*-coupling circuit).

The phenomenon of i-f overloading has already been described in Sec. 6·11. It will be recalled that if the input vs. output characteristic of a single i-f stage is measured, a level of input signal is easily found such that increase in input produces no increase in output (see Fig. 6·12). If this stage is so placed in the receiver that the level of the interfering carrier at its grid is in this region of no change of output, it is plain that the signal pulse superimposed on the carrier envelope will be entirely wiped off and the signal will be lost. Two procedures might be adopted to combat this signal loss. The first, reduction of gain ahead of the point at which overloading occurs, is effective but has the disadvantage of reducing the size of the desired signal at the receiver output. The second procedure is the provision of *back bias* in some form. An additional bias voltage is added between grid and cathode. Ideally, the magnitude of this bias should be comparable to the peak voltage, or envelope, of the interfering carrier at the grid of the tube in question. In this case, the peaks of the waves of the interference come to approximately the normal quiescent point of the stage, and changes in the size of the peaks are preserved.

Probably the simplest circuit that may be considered for this procedure is obtained by increasing the size of the normal cathode resistor to the point where the tube acts as an infinite impedance detector (see Sec. 7·7). To prevent excessive loss of gain, the resistor is bypassed for the intermediate frequency. Normal operating current in the tube may be obtained by returning the cathode resistor to a negative voltage or the grid circuit to a positive voltage.

This simple circuit suffers from two disadvantages. First, it is difficult to secure enough bias voltage. In order that the bias be equal to the peak voltage of the interfering carrier, it would be necessary to achieve unity detection efficiency. A large value of cathode resistor

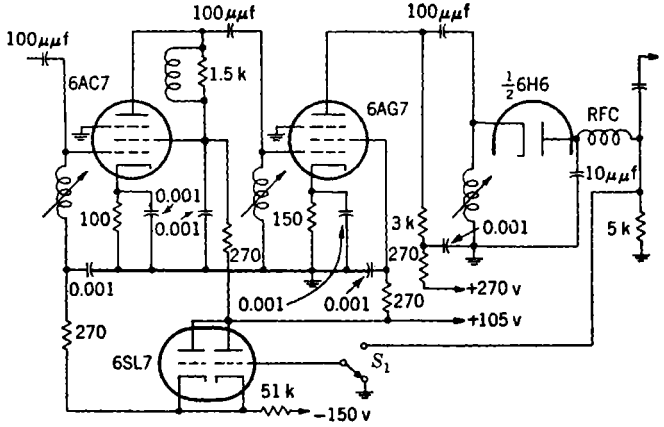


FIG. 9·6.—Back-bias circuit.

and bypass condenser would then be required. To obtain normal gain from the stage in the absence of interference, normal values of cathode current must be used, thereby setting a limit to the size of the cathode resistor. The second difficulty arises from the undesirable time constant of the cathode resistor and bypass condenser. If a strong signal, such as a block of normal echoes, ends suddenly, the cathode bypass condenser will be left charged to a positive voltage from which it can discharge only through the cathode resistor, since the tube will be cut off. This discharge may take several microseconds, during which time the receiver is insensitive to weak signals.

A somewhat more versatile circuit may be obtained by applying the bias voltage to the grid, as shown in Fig. 9·6. A cathode follower is used to provide a low-impedance source for the bias voltage. A diode detector with negative output supplies the voltage for the grid of the cathode follower. The best relation between the bias voltage and the amplitude

of the interfering carrier is usually obtained if the diode detector is connected to the plate of the stage following the one to which the bias voltage is applied. In other words, the feedback extends over two i-f stages. If the regular second detector of the receiver is a diode with negative output, its output voltage may be applied to the grid of the next-to-last i-f stage by way of a cathode follower, thereby adding only one tube to the usual complement. The principal disadvantages of this circuit are difficulty in assuring stability and lack of flexibility in adjusting the ratio of bias voltage to interference amplitude. The fact that the feedback extends over two stages is one of the important reasons for stability difficulties.

A circuit of still higher performance is used in the receiver described in Chap. 14 and is illustrated in Fig. 14·15. Reference should be made to that chapter for a detailed description of the circuit. It will be noted that feedback extends over only one stage and that a video amplifier of adjustable gain has been incorporated into the feedback loop. This circuit may therefore be applied to any i-f stage in which overloading is troublesome, without regard to the following stages. In the receiver described in Chap. 14, the last three i-f stages are so protected. This protection, however, involves the addition of loading due to the diode detector across each i-f interstage coupling circuit. In a receiver with a wide i-f pass band (several megacycles per second), this added loading may cause serious reduction in gain. The capacitive loading may be reduced by the use of germanium crystals in place of the diode (see Sec. 7·4). The resistive loading presented by the germanium crystal, although seldom heavier than the amount required to secure the necessary bandwidth, does vary with signal level, and the pass band of the amplifier accordingly varies.

The receiver of Chap. 14 incorporates an additional circuit (Fig. 14·16) known as "detector-balance-bias," DBB, whose action has many points of similarity to the IAGC circuits discussed above. It differs in two important respects: (1) It is not a degenerative or feedback circuit, since the input for the circuit comes from a point in the receiver ahead of the point at which the output is applied, and (2) bias is applied to a diode detector instead of an i-f amplifier. A delay line is included in the circuit to prevent any biasing action until a time comparable to the pulse length of a normal signal has elapsed. Such a delay line would impose severe problems in maintaining stability in a degenerative circuit and therefore could not be incorporated directly into the IAGC circuits. Reference should be made to Chap. 14 for a detailed discussion of the action of the DBB circuit.

9·5. Sensitivity-time-control (STC) Circuits.—In addition to the IAGC circuits, a system of time variation of gain, which has come to be known as "sensitivity time control," is often used as an aid to radar

operation in the presence of land or sea clutter. The inverse-fourth-power dependence of echo signal power on range leads to extreme variation in the intensity of clutter signals. In many cases it is necessary to reduce gain by as much as 40 db in order to make near-by objects visible in the presence of severe clutter. Although the IAGC circuits are capable of producing this much reduction in gain, their operation is somewhat improved if they are not called upon to do so. It is accordingly advantageous to produce a smoothly varying change of gain with range (or time following the transmitted pulse), which is nearly sufficient to permit visibility of small objects in the areas of least clutter. The IAGC circuits then need produce only the rest of the required decrease in gain. The STC circuits may function in this way. They may also be used on receivers not provided with IAGC to give improved visibility in the face of clutter.

Circuits used for STC have usually been designed to apply to the normal gain-control terminals of the i-f amplifier an additional voltage pulse of the proper shape and magnitude to produce the desired variations of gain. It has been found that a comparatively simple waveform with two or three adjustable parameters offers a sufficiently close approximation. The TR tube and to a lesser extent other receiver components do not recover completely for a few microseconds after the transmitted pulse. Accordingly it has been found desirable to have the i-f gain remain constant during this time and then return to normal. Typical voltage waveforms as well as a circuit for producing and inserting them are shown in Fig. 14·18.

It is, of course, possible to apply voltage waveforms of other sorts to the gain-control terminals. For example, a square wave may be used to give alternate high- and low-gain conditions after successive transmitted pulses. Such a system is used in conjunction with the circuit for changing video limit level which is described in Sec. 8·7. If the attempt is made to apply a waveform that is too steep, frequency components may be present that fall within the i-f pass band. These appear in the video output as "spikes." This phenomenon will ordinarily be eliminated by the i-f decoupling in the gain-control lead if enough decoupling is provided to ensure stability of the amplifier. If relatively steep waveforms are to be inserted, excessive decoupling should of course be avoided.

CHAPTER 10

MECHANICAL CONSTRUCTION OF RECEIVERS

By F. M. Ashbrook

10·1. Introduction.—The mechanical construction of receivers used for microwave reception must be considered primarily with regard to the effect that it has on the functioning of the electrical circuit. Microwave receivers often use intermediate frequencies of 30 Mc/sec or higher. At such frequencies a chassis can no longer be considered a universal and fixed ground, because its dimensions may be an appreciable fraction of a wavelength. Instead, it must be regarded as a conducting sheet on which there are currents and potential differences. The coupling of these ground currents produces exactly the same effect, good or bad, as any other form of coupling between circuit elements.

It is just as important to prevent external fields from generating ground currents on the chassis as it is to prevent the intercoupling of the internally generated ground currents. The interconnecting of the various component parts by means of cables must also be considered from the standpoint of stray signals that may be introduced into the circuits. The unit must be constructed with mechanical strength and durability, and adequate space must be provided so that the component parts will be readily accessible for servicing. The necessity for close correlation between the mechanical design and the electrical properties of the receiver has already been stressed in Chap. 6.

10·2. The Construction of the Receiver Box.—The box that houses the receiver has two general functions. (1) It must protect the receiver from externally generated fields that could interfere with the functioning of the receiver. (2) It provides mechanical protection for the receiver so that the component parts are not physically damaged. The box can serve also as a guiding element to aid the circulation of air for cooling.

Both the basic design and the details of construction of the box depend a great deal on the receiver and the system as a whole. For example, the receiver used for a fixed ground installation obviously does not require the same construction as that which is applicable in a low-power lightweight airborne installation. As a first subdivision, the receivers can be divided into two classes: those using the regular or large tubes such as 6AC7's, etc., and those using miniature and subminiature tubes. These two general groups of receivers may also be classified

according to the method of packaging. A first group has the complete receiver, i-f amplifier, local oscillator, and power supply in one package. A second group has the complete i-f amplifier, with or without local oscillator, in one unit, with the power supply in a remote unit. A third group has the i-f amplifier split into two sections, one part located at the mixer, the rest in the main package. The last group, which almost always uses miniature tubes, has a unit containing an i-f amplifier mounted as part of a larger unit which supplies mechanical protection.

Although each type of receiver presents an individual packaging problem, there are fundamental requirements that must be met by all receivers. These are ease of assembly of the component parts in the unit, ease and simplicity of making repairs, and completeness and durability of the electrical shielding.

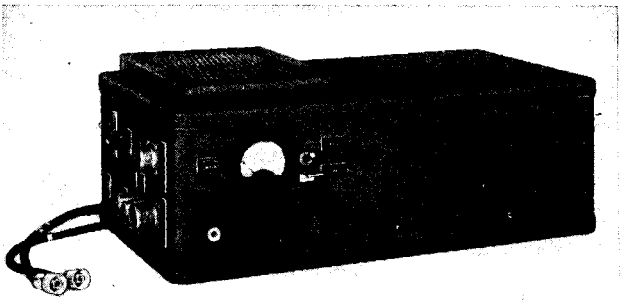


FIG. 10·1.—Typical large-tube receiver in one box. Note construction with covers top and bottom, an air filter on the top cover, operating controls on front and side.

The Exterior Box.—One type of exterior box which houses the complete receiver consists of three pieces. The outer shell, or "wrap-around," is an open box into which the receiver chassis is fastened. On the top and bottom of this box are covers that may be removed to provide access to the component parts. Figure 10·1 shows a view of a receiver that has both covers in place. The controls for adjustment and operation of the receiver are on the front. The two cables extending out in front are the cables that connect the i-f amplifier and the mixer. The bead rolls on the box just below the top and bottom edges have the purpose of strengthening the box and thus ensuring a better contact with the covers. The chassis is a tight-fitting shelf with a lip just large enough to provide a means of fastening the chassis securely to the box. Figure 10·2 shows part of a chassis assembled in a box. The two parts are held securely together by means of screws at regular intervals. This means of assembly provides maximum accessibility of the component parts during manufacture; and because the chassis is tight-fitting, the screws can provide the necessary electrical bond between the box and chassis.

The only practical way to exclude externally generated interference from the receiver is to have an electrically tight box. The first step is to provide adequate means for maintaining a good electrical bond between the receiver box and the covers. There have been several types of

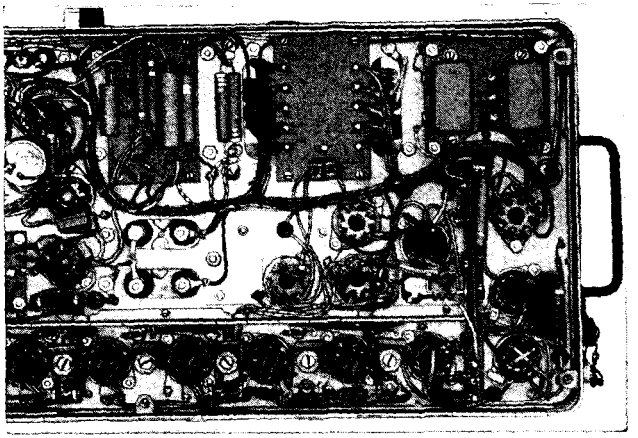


FIG. 10·2.—Bottom view of a receiver. The chassis edges are bent up and fit tightly against the walls of the box.

device used, but on the whole the most satisfactory, particularly from the standpoint of durability and economy, is the "antirattle clip"[1] (see Fig. 10·23*n*). This clip was developed for use in automobile radios and has been widely used where it is desirable to maintain a good electrical connection between a box and cover. It is common practice to roll around the edge of the box a groove that engages the two spring fingers on the clip and prevents it from coming off the box when the cover is removed. Figure 10·23*o* shows a section of the box with the groove and an antirattle clip in place.

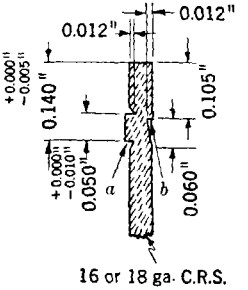


FIG. 10·3.—Rolled section for antirattle clips.

The cross section of the groove or rolled section is shown in Fig. 10·3. This section was designed with the idea of obtaining a flow of material rather than of having two opposing cutting rollers which would weaken the box. It is necessary that the edges marked *a* and *b* be maintained sharp enough to retain the antirattle clips. It has been found that an allowance of $\frac{1}{32}$ in. on each side is desirable between the box and

[1] The antirattle clip is manufactured by the United-Carr Fastener Co., Cambridge, Mass.

the cover for the antirattle clips. The edge of the cover should extend vertically downward at least $\frac{1}{2}$ in. for best results.

The smaller antirattle clip (Fig. 10·23*m*) was developed for use in the miniature-tube amplifiers where the large clip was too bulky to use. Figure 10·23*k* is a spring-finger clip, similar to antirattle clips, which was also developed for use in miniature-tube amplifier strips.

10·3. Introduction of Cables and Leads.—Secure bonding of the covers prevents signals from leaking into the receiver by way of the covers, but it is also necessary to remove any other possibilities of leaks. The most careful attention should be paid to all of the cables entering or leaving the unit to be sure that they will not conduct externally gener-

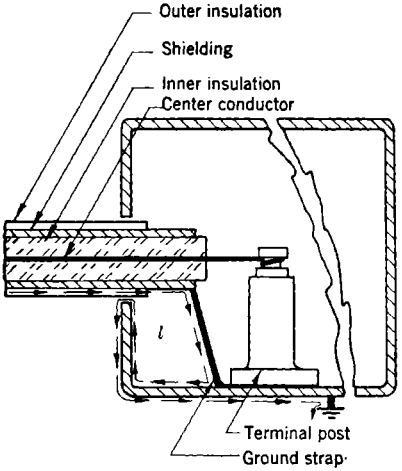


FIG. 10·4.—Incorrect method of introducing shielded cable.

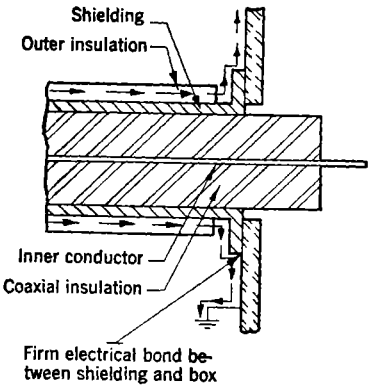


FIG. 10·5.—Correct method of introducing shielded cable.

ated interference into the receiver. One source of trouble is the introduction of a shielded cable into a completely sealed box, as shown in Fig. 10·4. The shielding on the cable does not make contact with the box where it is introduced but is grounded internally. The arrows in the diagram show the path that any signal or interference which is picked up on the *outer* surface of the shielding must follow in order to return to ground. The currents around the loop will generate a field in the enclosed box in just the same way as do the coupling loops used with resonant cavities. Figure 10·5 illustrates the correct method of introducing shielded cables into a box where shielding must be maintained. If external interference is to be kept from the receiver proper, then the shielding on the interconnecting coaxial cables must be sufficiently heavy to prevent voltage differences generated on the outside

of the shielding from penetrating to the inner surface of the shielding. Only the heavy double-shielded cables have been found good enough to give adequate protection from interference such as might arise from a near-by radio transmitter. Figure 10·6 is an example of the application of this shielding in the receiver discussed in Chap. 14. The two input circuits in this illustration are for the signal channel and for the AFC circuit. This is a particularly good example, because the shielded cable enters not only the receiver box but also the i-f subchassis box. The mechanical construction is as follows: a short lead-through tube is welded securely to the outer box and extends through to the inner box with the threaded section shown; the i-f cable stripped of its outer covering is

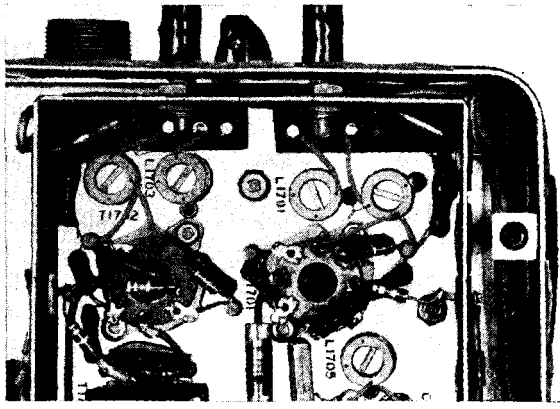


FIG. 10·6.—Method of introducing i-f cables into receiver.

forced through the tubing until enough of the cable extends through the tubing to complete the electrical connection; the cable shield is soldered to the lead-through tubing; and the two nuts on the tubing provide the means of securing the inner box and making certain of good electrical connection.

Any other cables that enter or leave the unit must be treated in a similar manner. Particular emphasis should be placed on the treatment of the power and control circuits. If the receiver and power supply are in one unit, the only connections into the receiver will be the main power leads together with any external control circuits. The a-c power should be handled carefully to prevent pickup of stray signals. If possible, it, too, should be in a shielded cable, and once inside the receiver it should go directly to the power-supply transformer and *should not be cabled with other wiring.* The control circuits require even more care, because they generally are high-impedance circuits. It is usually neces-

sary to provide some sort of filtering for these leads. If this is not done just as soon as the circuits enter the receiver box, most of the effectiveness of the filtering is lost, because that part of the lead between the entrance to the box and the filter serves as a means of feeding the interference into the box.

With high-impedance circuits it is also necessary to be careful of the grounding. For instance, if a tuning control for a local oscillator is grounded in a control unit located at a distance from the receiver, then any voltage drop in the ground circuit between these two units adds a voltage in series with the control potential for the local oscillator. If the current in the ground circuit fluctuates so as to cause this voltage drop to change, then the tuning also fluctuates. Large metal structures with much electrical machinery are likely to have considerable voltage differences in the structure. *It is very desirable, therefore, that the ground circuit of all voltage-sensitive devices be returned to the unit that is being controlled and that the controlled unit furnish the only ground for that circuit.*

10·4. Holes for Ventilation and Adjustment.—The last consideration is the question of shielding the holes that are necessary for ventilation and adjustments. Many of the adjustment holes can be protected most of the time by a tight-fitting snap plug such as those shown in Fig. 10·1. The plug must make good electrical connection with the wall of the box, or it will not be effective in preventing external fields from being coupled into the unit. There are certain other types of hole, such as ventilation holes, that cannot easily be plugged. For these it is necessary to provide a suitable grating to attenuate the signal sufficiently to prevent trouble.

The signal leakage through a hole that has a diameter equal to one or more wavelengths can be found by using the principles of physical optics. The case where the hole is about the same size as the wavelength has not been calculated; but for a hole less than $\frac{1}{10}$ wavelength in diameter, the leakage through the hole is proportional to[1]

$$S_{\text{tot}} \approx \frac{A^6}{\lambda^4},$$

where S_{tot} is the Poynting vector, A is the radius of the hole, and λ the wavelength of the signal. This statement assumes that the thickness of the material in which the hole is punched is negligible. If the material has thickness t, then there is an additional attenuation of

$$\alpha = 32\,\frac{t}{r} \qquad \text{db},$$

[1] H. A. Bethe, RL Report No. V-15S, Jan. 23, 1942; *Phys. Rev.*, **66**, p. 163 (1944).

where r is the radius of the hole. Since the leakage increases so rapidly with hole size, it is immediately obvious that any grating, if it is to be effective, must have the smallest holes possible.

These calculations of attenuation are based on round holes. If slots are used, the leakage is largely governed by the long dimension. Consequently, a slot $\frac{1}{4}$ by 2 in. produces about the same attenuation as a hole 2 in. in diameter for the worst orientation of the plane of polarization of the incident interference.

To obtain some estimate of the absolute magnitude of the transmission through such a small hole, one may consider the coupling holes used in waveguides, for which the attenuation has been measured. For a hole 0.040 in. in diameter in the narrow side of $\frac{1}{2}$- by 1-in. waveguide at a frequency of 10,000 Mc/sec the attenuation is approximately 20 db.

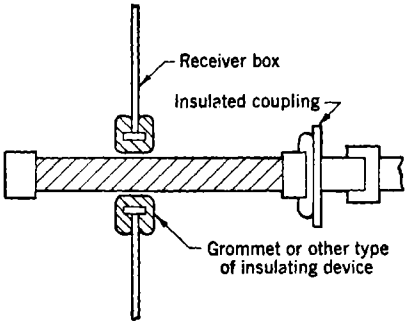


FIG. 10·7.—A mechanical coupling link acting as a dual dipole.

From this datum the attenuation of a hole $\frac{1}{4}$ in. in diameter for a frequency of 30 Mc/sec may be estimated to be about 70 db. Inasmuch as the gain of a typical 30-Mc/sec i-f amplifier may be as much as 120 db, it can be seen that such a $\frac{1}{4}$-in. hole might cause difficulty in some cases. If external interfering fields are expected, it is wise to test the unit by generating a field of the expected magnitude and operating the unit in that field.

Any hole that is located in the receiver box and used to permit operation of a mechanical adjustment inside the receiver requires special attention. A typical example is a shaft connected to a variable condenser, LO tuning strut, or the like. If the shaft is metal, it may act as an antenna on both sides of the box (see Fig. 10·7). Whether or not the part of the shaft extending out of the box is grounded, it is still easy for the right frequency to excite the shaft. When this happens, the shaft will reradiate on the inside of the box. If the shaft were to go through a metal tube without making particularly good contact, the energy transfer might still be good because the shaft in the tubing would

act just like a coaxial conductor. To eliminate this transfer of energy into the box, it is necessary to do one of two things: either to make a good electrical ground to the shaft at the point of entry to the box or to make the shaft of an insulating material.

It is necessary to shield a receiver box for both i-f and r-f signals. If a local oscillator is present in the box, the requirements for r-f shielding are obvious. In units where only the i-f amplifier is present, however, the shielding is still very important, because it is often found that such r-f signals, even at 10 cm, are rectified at the grid of the i-f amplifier tubes and thus produce a signal in the i-f channel. If an AFC circuit is present, the shielding must be very carefully done for the reasons outlined in Chap. 3.

If the receiver is part of a larger unit that is enclosed and provides adequate shielding against externally generated interference, it is possible to take some liberties with the construction of the receiver box. Most of the miniature-tube strips used as illustrations in this chapter were designed to be used in larger units which provided the external shielding requirements. Therefore, it is necessary to provide only enough shielding to ensure the stability of those units plus a moderate amount of shielding from externally caused interference.

10·5. Construction Materials.—The chassis material plays an important role in determining the serviceability of an amplifier. Silver, copper, and aluminum are the three best materials from the standpoint of electrical conductivity, but the material selected must have durability and resistance to corrosion as well. It must also be adapted to establishing and maintaining a good electrical connection. In this last respect, aluminum is a poor construction material for use with any techniques so far developed because it is difficult to make a soldered connection to aluminum that is durable. Therefore all the grounds must rely on pressure joints. These joints soon begin to give trouble from high resistance as a result of the very tough oxide coating that forms on aluminum. The chief possibilities for improvement would seem to be the use of plated aluminum to which solder connections can be made or the use of welded connections for grounds.

Electrolytic corrosion is another source of trouble in the use of aluminum and similar materials. Most of the basic components and the wire used in the construction of receivers are composed largely of copper. These two metals, aluminum and copper, are widely separated in the electromotive series. Therefore at the junction of these two metals there will be a strong tendency toward electrolysis which furthers the corrosion at the junction, with consequent increase in resistance. The material selected for the chassis and box should therefore be a metal that is free from corrosion and placed in the electromotive series near to

copper and tin, which are prominent in component parts. Copper, iron, tin, nickel, and cadmium are all close together and grouped around hydrogen, the zero point in the table. Iron and nickel are not satisfactory materials, both because of oxidation (rusting) and because of difficulty of soldering. Copper (or brass) and cadmium are therefore left as the most acceptable materials. Copper and brass are excellent materials, easy to work with and to solder, but in damp climates are susceptible to corrosion. Cadmium is never used as a base metal by itself, but rather as a protective plating for other base metals. Cadmium-plated steel and brass are very durable materials. One objection to cadmium plate is that it tarnishes wherever it is touched. However, a "water-dip lacquer" treatment at the time of plating protects the surface from finger marks without damaging the electrical characteristics.

10·6. The I-f Amplifier.—With i-f amplifiers that have gains of about 120 db and bandwidths of 1.5 Mc/sec and more, the problem of maintaining the stability of the amplifiers, particularly those made in production quantities, is a serious one. The mechanical layout of the component parts must be done with meticulous care, and each part must be placed with due regard to possible coupling with other parts and coupling through ground currents.

Each i-f amplifier stage has its own set of ground currents. If the amplifier as a whole is to be stable, then the coupling of these ground currents must be a minimum. The simplest and most direct approach to this problem is to build the amplifier so that all of the i-f circuits are in a straight line, one following the other. Any stage that "goes around the corner" will have its ground currents flowing in such a way that coupling between it and the previous stage is more likely to occur, as illustrated in Fig. 10·8. In actual practice, when amplifiers are built on a large chassis and bent around a corner, it has been found necessary to provide more careful shielding as well as to keep a tight cover over the bent stages. With an in-line amplifier strip it is possible to test the amplifier with no covers on, but with the bent amplifier this is usually not true.

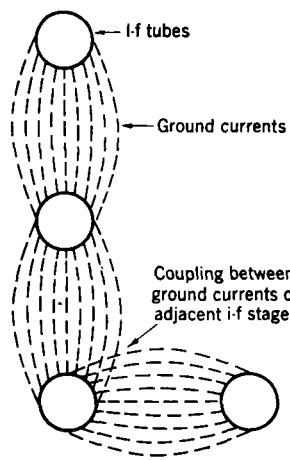


FIG. 10·8.—The coupling between ground currents that occurs when i-f amplifier tubes are not in a straight line.

Instead of considering the currents that flow in the chassis, attention may be focused on the electric and magnetic fields inside the enclosure, inasmuch as the fields and the current in the boundary are directly re-

lated by Maxwell's equations. By adopting the field point of view, it is possible to introduce many of the concepts and semiempirical results used in waveguide circuitry. The box in which the amplifier is mounted plays the part of the waveguide, and the coils or other circuit elements serve to set up fields in the enclosure. It is well known that a waveguide of a given size will transmit freely only waves of a frequency above a limiting or cutoff frequency. Frequencies lower than this are attenuated at a rate that is governed almost entirely by the dimensions of the waveguide. This behavior is made use of in the waveguide attenuator (see Chap. 11). It may be shown that the cutoff frequency is related to the width of a rectangular waveguide as follows:

$$f_c = \frac{c}{2w}.$$

It may also be shown that a frequency $f \ll f_c$ will be attenuated 27.3 db in going a distance equal to the width of the waveguide. Both these



FIG. 10·9.—Detail of bottom of six-stage i-f amplifier.

results assume that the field configuration belongs to the so-called "lowest mode" in which the electric field is parallel to the shortest dimension or height of the guide. Other modes have a higher cutoff frequency and are attenuated more rapidly.

For a frequency of 30 Mc/sec it would be necessary to have a waveguide 5 m wide to transmit signals freely. Accordingly, any box that is likely to be encountered may be considered as a waveguide below cutoff. Consider now an amplifier of many stages laid out in line and mounted in a rectangular box. As a signal goes through the amplifier, it is continually being amplified. At any given point, for example, one of the inductances in a particular stage, there may be coupling between the amplifier circuit and the waveguidelike field. This may produce a wave traveling back toward the input stage of the amplifier. This wave will be attenuated at the rate of 27.3 db/width. If, however, it arrives at the input end of the amplifier sufficiently strong to produce a signal voltage greater than that originally introduced, oscillation may result. This situation may be avoided by so spacing the amplifier stages that the gain in a length equal to the width of the box is less than 27.3 db. Alternatively, the effective

width of the box may be decreased by such an expedient as the use of a post making *good* electrical connections with the top and bottom of the box. Figure 10·9 illustrates the use of two blocks that act as waveguide filters in an i-f strip.

Where an i-f strip is part of a larger box, as in Figs. 10·10 and 10·11, it is necessary to provide extra attenuation between input and output by

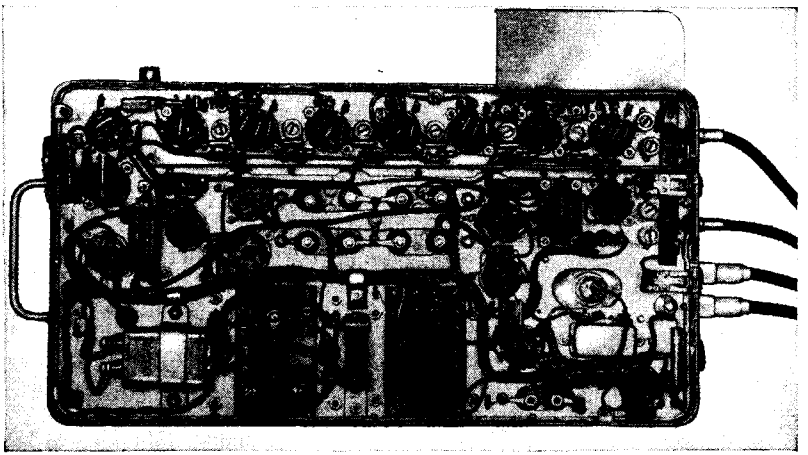


FIG. 10·10.—A complete receiver in one box with i-f strip along top and cover for first two i-f stages.

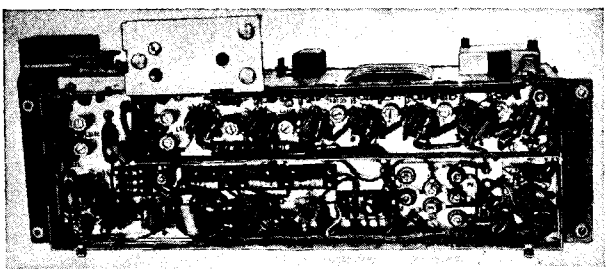
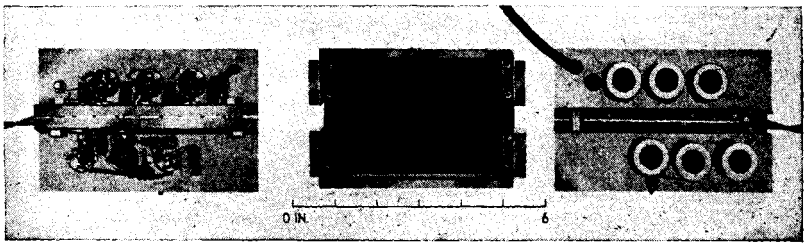


FIG. 10·11.—A 3-cm receiver. I-f strip is in boxed-off section. Small cover fits over first two i-f stages.

some means, because the full width of the box is effective as a waveguide and thereby greatly reduces the attenuation from output to input. Without the extra attenuation provided by the small cover that fits over the input circuit in Fig. 10·10, the attenuation from the detector coil to the input coil, if unity coupling between coil and guide is assumed, would be only about 33 db. In Fig. 10·11 the attenuation without the extra cover is about 57 db. In both cases this was insufficient attenuation to prevent

regeneration. Regeneration caused by feedback of this type may show up only when the outer cover is on and connecting all the way around.

The two sets illustrated in Figs. 10·10 and 10·11 are perfectly stable when the outer cover is off. This feature is one of the requisites of good design because it allows repairs or adjustments to be made to all components of the receiver with the set in operation. If the outer cover is put on without the inner cover, the sets are regenerative. If, however, the outer cover is placed on in such a way that it contacts one of the sides and *does not touch the other side*, the set is again stable if the regeneration is caused by this waveguide feedback. The addition of the small covers over the first two stages of the i-f amplifier provides adequate attenuation to insure stability. If these covers are to function satisfactorily, it is necessary to ensure a good electrical bond between the walls and the



(*a*) (*b*) (*c*)

FIG. 10·12.—Special-purpose four-stage miniature-tube i-f amplifier. (*a*) Bottom view, cover off; (*b*) cover; (*c*) top view.

covers. Such a bond can be accomplished by the use of antirattle clips on the walls as shown. This shielding takes care of the electromagnetic coupling in the amplifier. To improve the electrostatic shielding, the partition that supports the cover is extended the full length of the amplifier. This partition does not make contact with the outer cover.

These conditions are not in themselves sufficient to ensure the stability of a receiver; but when coupled with proper design and layout of single-stage circuits, they will produce a receiver that can be made in quantity without individual tailoring for stability.

It sometimes becomes necessary to bend an i-f strip in order to fit it into the allotted space. Generally, this procedure is very uneconomical because of the complications that arise in the construction of a bent amplifier that is sufficiently stable to permit quantity production. The amplifier in Fig. 10·12 is perhaps one of the exceptions. This particular amplifier has four i-f stages with a total gain of only 80 db. The other two stages are detector and video stages. The only wiring that is not complete in this unit is the video output cable connection and the connection between the third and fourth i-f stages, which crosses between

the two channels. The tuning inductances are mounted on the wall of the channel, and the load resistors are mounted on the tube sockets. The bypass condensers are button mica condensers as in Fig. 10·22*e*, and the voltage supply and decoupling resistors are in the interior part of the channel. The mechanical construction of the unit is such that if major repairs are required, the coupling connection between the two channels can be removed and the two parts easily separated. The stability of this particular amplifier is good because the gain is low, there is a minimum amount of wiring to pick up stray signals, and all ground connections are extremely short. This set operates without regeneration at its normal gain with or without the cover.

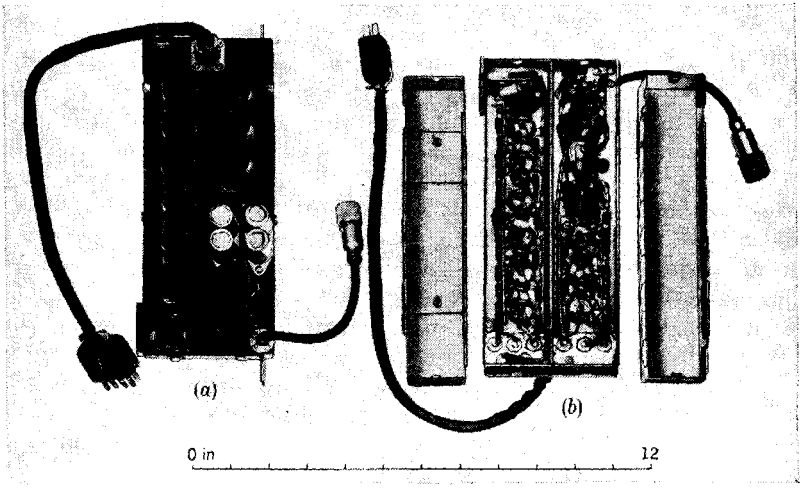


FIG. 10·13.—Wide-band 60 Mc/sec bent i-f strip. (*a*) Top view; (*b*) bottom view, cover off.

If all six stages were i-f stages with gain of about 20 db per stage, this construction would probably be troublesome from the standpoint of stability. The cover would have to make firm contact with the central channel partition, and it would be impossible to operate the unit without a cover on at least one of the channels. If the front stud that holds the two units together (near the input cable) were of insulating material instead of a conductor, then the chance of stability might be increased. As long as the metallic conductor exists between the input and output stages, there is a short path between the ground currents of the input and output stages. Without this connection between the two chassis, the ground currents would have to flow back along the one chassis to the connection point, then over to the other chassis, and along that chassis to the input circuit. The path is thus considerably lengthened, and the ground

currents are the same as they would be if the i-f amplifier were in a straight line.

The amplifier in Fig. 10·13 is a wide-band 60-Mc/sec i-f amplifier with a gain of about 115 db. This receiver is a good example for demonstrating the problems that arise in bent amplifiers. In order that this amplifier may not regenerate or oscillate, it is necessary that one of the two covers be fastened in place. This is a handicap when experimental work or servicing is being done. The two springs that project downward in the cover contact the vertical strip on the chassis in which the bypass condensers are mounted. These two spring fingers contacting this upright have the effect of decreasing the width of the waveguide and so ensure stability in the amplifier.

The input circuit in this receiver is not in place, since it attaches directly to the mixer output probe. The raised platform effect on the back of the receiver where the power cable enters contains filter sections for all of the incoming power and control leads. The antirattle clips ensure complete connection of the covers at all times. The covers are retained in place by six screws in the side. In this way a very compact construction is achieved. The circuits are all fixed-tuned, and the inductances are wound on the coil forms illustrated in Fig. 10·22*b*.

In general, then, if it is necessary to build an i-f amplifier so that it makes a right-angle bend or doubles back on itself, the construction should be such that the ground currents set up in the chassis of the amplifier flow in a direction that makes the amplifier appear to be of straight-line construction. Either this can be done by careful design of covers, which will restrict the entry of currents to the proper place in the amplifier, or it can be done by having the chassis of the amplifier split into two separate pieces joined only at the point where the i-f amplifier proper bridges the two chassis. If this latter practice is followed, it must be remembered that the two chassis can have only *one* connection and that any supporting brackets or mounting devices, etc., *must not connect the two chassis*. The cover must be of two separate pieces or constructed in such a way that it makes definite contact with the center partition and, therefore, is of two pieces electrically.

10·7. The Ground Circuits for I-f Stages.—The individual stages rather than the general structure of the i-f amplifier will now be considered. At frequencies of 30 Mc/sec and higher the chassis can no longer be considered an equipotential ground plane of zero inductance, and the inductance of short lengths of wire begins to have an appreciable effect. Consequently, the arrangement of the component parts and the grounding points for individual stages become extremely important.

A typical i-f stage is shown in Fig. 10·14. The resonant circuit consists of the inductance L in parallel with C_{in} (input capacitance of V_2),

C_{out} (output capacitance of V_1), and any stray capacitance in the wiring of the circuit. Both the cathode bypass condenser C_k and the screen bypass condenser C_s carry i-f signal currents to ground. These circuits that carry i-f currents should be carefully grounded with as short lead lengths as are practical.

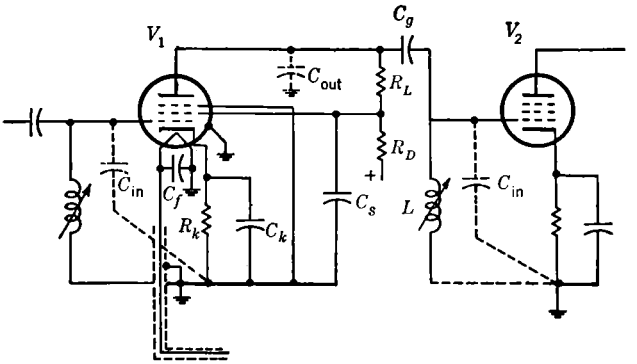


FIG. 10·14.—Complete circuit for i-f stage showing the various ground circuits.

Those components which are not part of the actual i-f circuits, such as the shell of the tube and the heater circuit and its bypass condensers, should be grounded with leads as short as possible to a separate point from the i-f components, because any inductance in a common ground lead couples i-f signals in or out of these non-i-f components. The sup-

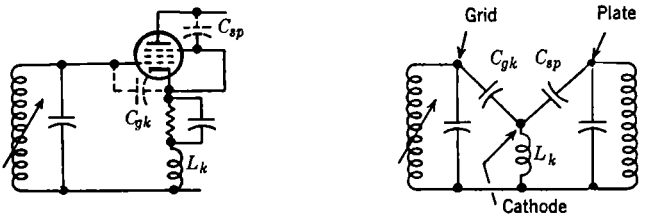


FIG. 10·15.—Feedback in a stage having the suppressor tied to the cathode with inductance in the common ground lead. (*a*) Complete circuit; (*b*) equivalent circuit.

pressor grid has been shown grounded, as is common practice for all high-frequency amplifier tubes. If the suppressor is connected to the cathode and if there is some inductance in the cathode-to-ground circuit, then feedback can occur from the plate to the grid of the stage, as shown in Fig. 10·15*a*. The equivalent circuit is that shown in Fig. 10.15*b*.

The grounding of the shells of 6AC7's should be done with care. The connection between the tube pin and the shell is extra long, because this

connection is used in flashing the getter in the tubes. Thus, even though the socket pin is firmly grounded, there may be an appreciable impedance between the shell and ground and consequently a substantial signal voltage may be built up on the shell. This can cause undesirable coupling into and out of the circuits. The receiver described in Chap. 13, as first built, was in difficulties from this source and was modified slightly as described there.

Good construction practice then indicates that for the most reliable operation, each i-f stage should have all of its circuit components grounded in one place and with short and low-inductance connections. Because the output capacity is inherent in one stage and the input capacity is inherent in the next stage, the two stages should be as close together as possible. All of the stages should be arranged in a straight line or in such a way that the ground currents of the stages are kept isolated. The inductances should be oriented so that currents induced in the chassis by their fields are a minimum. It is also necessary to locate the components so that there is no undue electrostatic coupling. The ground circuits for elements not part of the i-f circuit should be separate from the circuit ground. The heater circuit should be carefully isolated so that feedback from input to output terminal by way of it is a minimum. The wiring and grounding of the heater circuit should be so done that the wiring and chassis current do not form a large loop that couples in with the signal currents of the amplifier.

10·8. Typical Construction with Large Tubes.—When large tubes are used (6AC7's), it is generally most convenient and practical to mount most of the components on the tube sockets as shown in Fig. 10·16*a*. (Figure 10·16*b* gives the circuit diagram for the portion shown here.) In this way, the lead lengths of the components will be at a minimum. Pin 1 on the socket is the shell and is grounded by itself. Pin 2, the heater, is grounded, and the heater bypass condenser is returned to this pin. The heater wire across the socket goes into a shield that is grounded at the same ground point. This construction keeps the grounding leads short and keeps the loop formed by the heater circuit and the bypass condenser small. It also keeps the chassis loop small, since the heater and shield are both grounded at the same place. The suppressor, Pin 3, is used as the circuit ground. The grid coil, the cathode and screen bypass condensers, as well as the plate bypass condenser, if there is one, are all returned to Pin 3 on the tube socket. This is not a difficult construction for production receivers because the tube sockets can be made up as subassemblies. The mechanical strength of the mounting is good—vibration and shock tests have failed to dislodge the components—and, most important, all of the elements that carry i-f currents are confined to a small space and short lengths of wire.

The heater circuits are wired in pairs. The wire goes from the first tube into the shield and along the chassis to the second stage; from the second heater it returns to the shield at the heater ground point and then

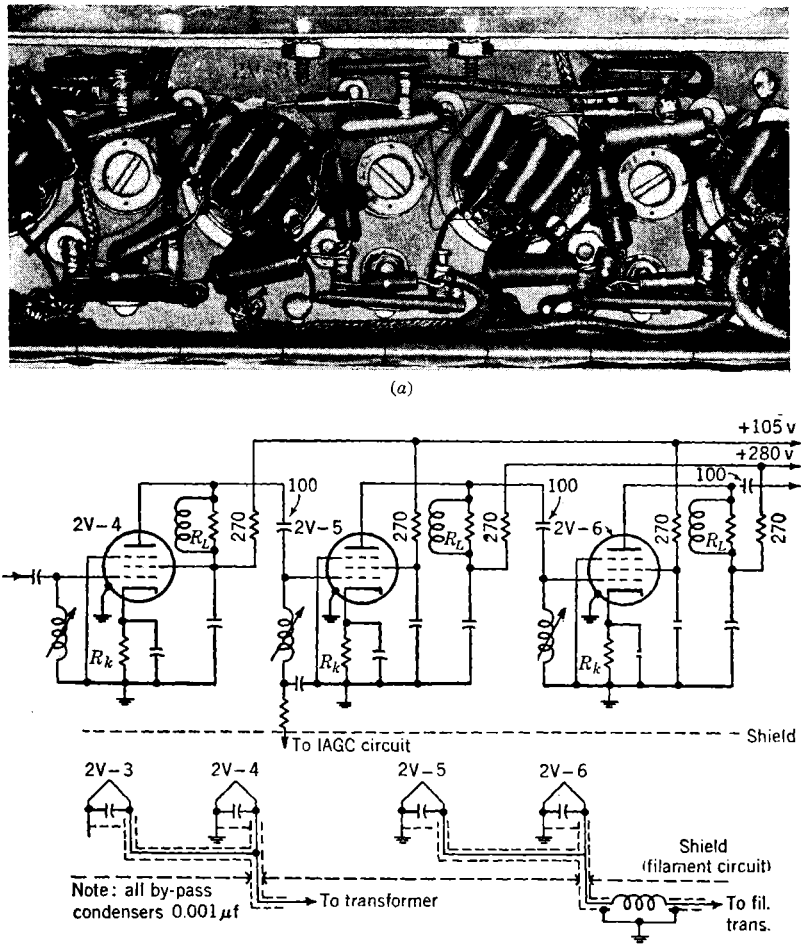


FIG. 10·16.—Component mounting in an i-f amplifier. (*a*) Photograph of a portion of an i-f strip; (*b*) Circuit diagram of portion shown in (*a*).

in the shield leaves the amplifier strip. Outside the amplifier strip it is connected to a series heater choke. It then returns to the shield and goes to the transformer. Each pair of i-f stages is wired in the same way, except that the first two stages are generally the only ones that need to

have a choke in series with them. If more isolation is required, it is usually sufficient to put a choke in the last heater circuit. When the heater circuit is wired in this way, it is easy to determine whether or not feedback is occurring through it and, if it is, what circuits are causing trouble. If all except one of the heater wires from the i-f circuit are fastened to the transformer and that one wire is held on until these tubes have reached operating temperature, then by connecting and disconnecting that wire and observing the effects on the regeneration, the contribution of that circuit can be definitely established.

The last two stages in this amplifier are operated at a higher plate voltage than the rest. The plate condenser is returned to the same ground point as the rest of the components (a return to Pin 3, the suppressor pin, would be preferable; but because there are already too many mechanical connections to that pin, the next best grounding spot is

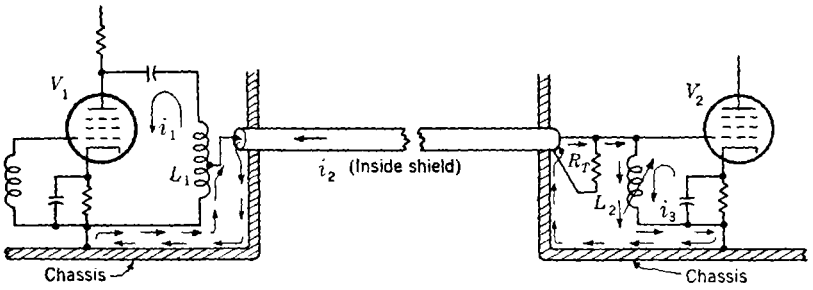


FIG. 10·17.—Connecting i-f cable between two chassis.

chosen). The three condensers mounted on the tube socket are 0.001-μf high-dielectric ceramic condensers. The half-postage-stamp-size silver mica condensers may also be mounted on the tube socket. The resistor across the socket is the cathode resistor. Pins 1, 2, and 3 are toward the bottom of the illustration. The resistor with the coil wound on it is the plate-load resistor. Actually, with the plate operating at a higher voltage than the screen, the coil is unnecessary. It is used here for convenience to permit use of the same component part in each stage, inasmuch as the value of the load resistance is identical.

The slug-tuned inductances are between stages. The components are accessible for servicing, and the wiring is reduced to a minimum.

Special stages, which because of their construction or location do not have short uninterrupted ground paths between the two stages, require special treatment. One such circuit is the coupling network that is used in transmitting i-f signals from a preamplifier chassis to a remote receiver. The two chassis and the connecting cable and circuit are shown in Fig. 10·17. The output stage of the preamplifier drives the cable

that must be terminated (R_T), otherwise reflections and undesirable loading will be experienced in this circuit. The inductance L_2 tunes the input capacitance of the first amplifier in the receiver strip, and L_1 tunes the output capacitance of the last preamplifier circuit. There are three separate circulating currents in this strip, as indicated by the arrows in Fig. 10·17. The first is the circulating current i_1 through the inductance L_1 and through the plate of V_1 to ground. There is also a circulating current i_2 in the cable and through the chassis as indicated by the arrows. The last circulating current i_3 is in the input network of V_2 as shown. To insure stability in this circuit, the three circulating currents must be considered, and the cable connections should be such as to provide a

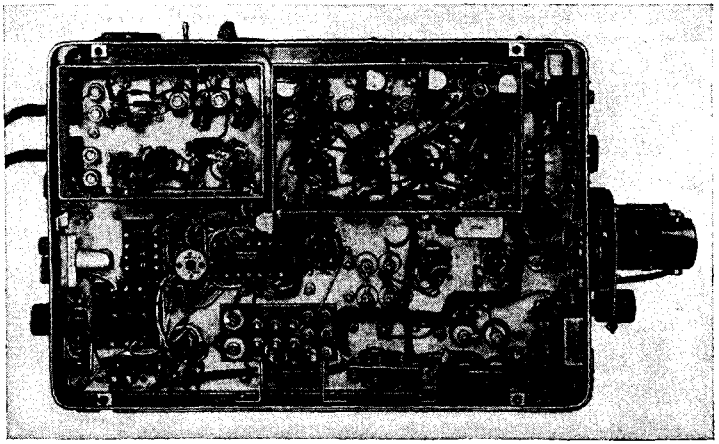


FIG. 10·18.—Bottom view of a receiver with i-f amplifier built in two subchassis.

good circuit for the currents on the inside of the outer conductor as well as to prevent signals on the outside of the outer conductor from entering either chassis. The present trend is to avoid splitting the amplifier in this way.

Another special case arises when there is a connection that bridges two subchassis as that shown in Fig. 10·18. Figure 10·19 shows the i-f current path along the ground for the returning signal. It should be noticed that the ground current must flow along the inside of the subchassis to the hole through which the i-f lead enters, then along the outside of the subchassis to the point (or points) where the chassis is bolted down. From there it goes across the main chassis to the other unit and repeats the same procedure to reach the other i-f grounding point. The completion of the current flow is through the input and output capacitances and along the i-f circuit. This particular type of construction is vulnerable on two counts. First, the very long ground return path makes

this stage much more susceptible to interference and increases the area over which coupling of chassis currents might occur. The second disadvantage of this circuit is that the i-f currents must flow over the studs that fasten the subchassis to the main chassis. This construction introduces into the circuit four junctions that maintain their electrical ground

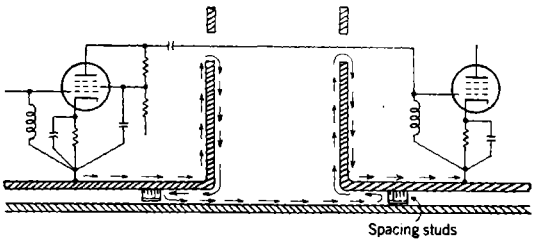


FIG. 10·19.—The ground return path for i-f signal in circuit bridging two subchassis in Fig. 10·18.

by pressure alone. At these four points a small amount of dirt or corrosion can cause an open- or high-resistance circuit for the intermediate frequency. A more reliable circuit might result if a piece of tubing were inserted between the two chassis and firmly connected to each. This construction would make a coaxial junction for the two units and greatly

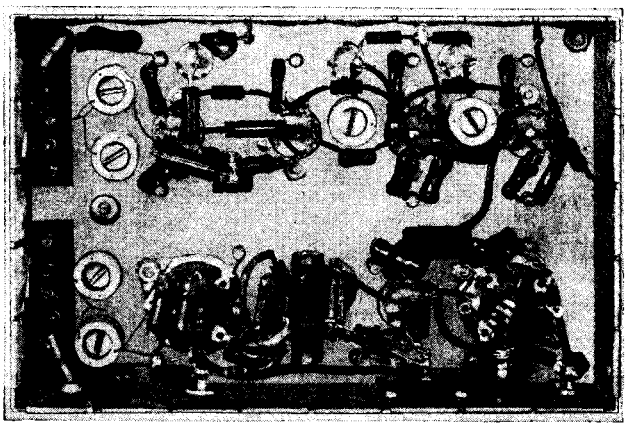


FIG. 10·20.—The input subchassis for receiver in Fig. 10·18.

reduce the length and uncertainty of the ground current path as well as restrict it so that it would not flow on the outside of the subchassis boxes. The addition of this connecting tube would complicate the mechanical construction.

Two-box construction is advantageous in a complicated i-f strip, such as the one in this receiver (see Chap. 14 for the details of the receiver).

The construction used here is adaptable to easy modification and servicing. It should be noted that both subchassis units have covers that are held in place with antirattle clips. (See the detailed photograph of the input box in Fig. 10·20.) All of the power-supply and control leads are brought into the box through button condensers which bypass the circuits at the point of entry. For a detailed photograph of these condensers, see Fig. 10·22*e*. In these units the condenser is located between the center piece and the outside ring. Also illustrated are the standoff condensers (Fig. 10·22*e*) that are used in these units.

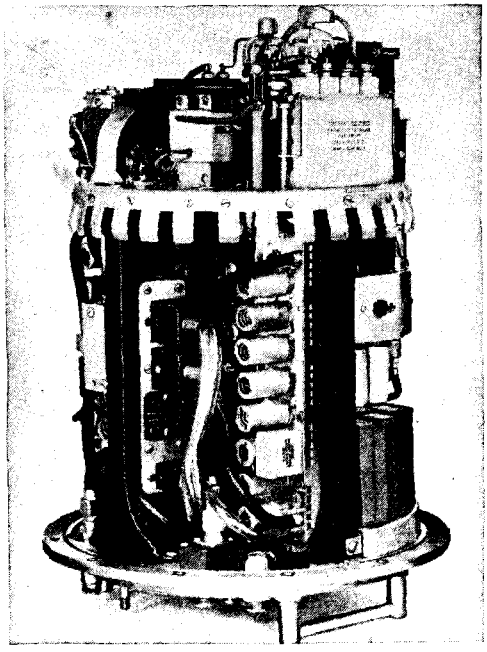


FIG. 10·21.—A small transmitter-receiver unit with the i-f amplifier shown in Fig. 10·9 in place.

10·9. Miniature Tubes and Components.—Unlike the stages using large tubes, stages using miniature tubes have many and varied construction details. Here the circuit components, rather than the sockets, determine the space requirements and construction details in the i-f amplifier. The principles of separating circuit and noncircuit grounds and of arranging the amplifier stages (and grounds) in a straight line still persist.

Most of the miniature-tube i-f amplifiers have been built in strip form and included in a larger shielded unit. The amplifier in Fig. 10·9 is shown mounted in the transmitter package in Fig. 10·21. The heater

circuit often uses series chokes in a progressive filter because of the convenience of mounting the chokes from stage to stage. Wiring of this type is used in the amplifier of Fig. 10·12.

Many of the components used in such strips are pictured in Figs. 10·22 and 10·23. Figure 10·22*e* shows button condensers of various

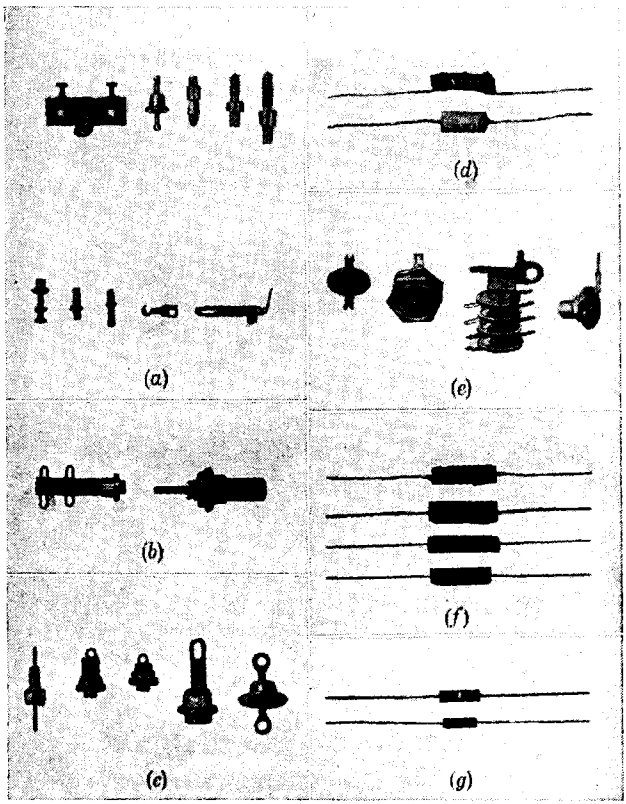


FIG. 10·22.—Small components used in construction of i-f amplifiers: (*a*) lugs; (*b*) coil forms; (*c*) feed-through insulators; (*d*) condensers; (*e*) button condensers; (*f*) fixed-tuned coil forms; (*g*) fixed resistors.

types. The next to the last unit in this row consists of four button condensers mounted in a bracket and may be used to provide all of the bypassing for a single stage. The solder lugs extending out of each side of the condenser form a continuous connection. Figure 10·22*b* shows two types of slug-tuned coil forms used in miniature strips. Figure 10·22*a* and *c* shows tie-point, lead-through, and ground lugs. Figure 10·22*d* shows two condensers, one of which is a high-dielectric (0.001-μf) con-

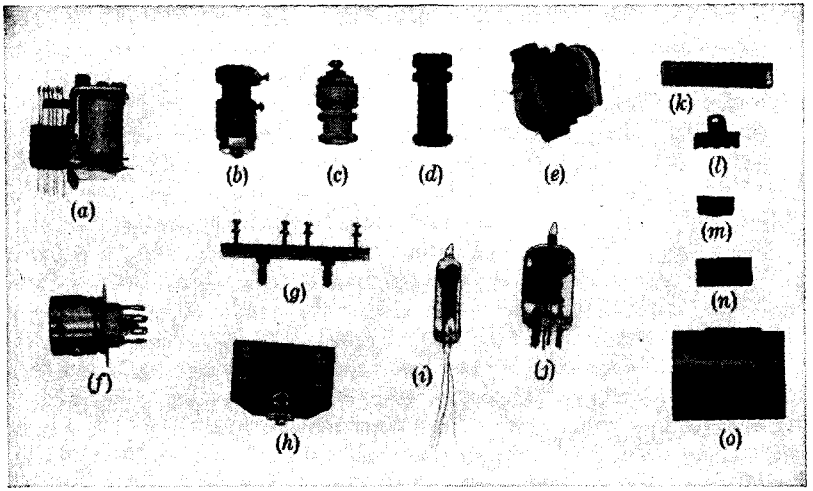


FIG. 10·23.—Receiver components: (*a*) small relay; (*b*, *c*, *d*) slug-tuned inductance forms; (*e*) discriminator transformer; (*f*) miniature-tube socket; (*g*, *h*) assembled tie point and resistor mounting strip; (*i*) subminiature vacuum tube; (*j*) 6AK5 pentode; (*k*, *l*) clips to ground covers in miniature strips; (*m*) miniature antirattle clips; (*n*) regular antirattle clip; (*o*) section of chassis with clip in place.

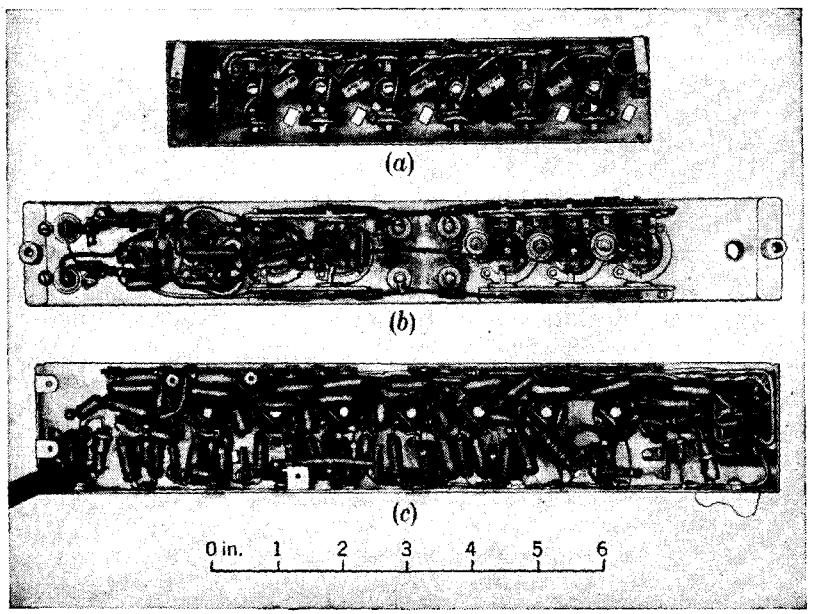


FIG. 10·24.—Three miniature i-f amplifier strips: (*a*) five i-f stages and detector; (*b*) five i-f stages and f-m discriminator; (*c*) six i-f stages, detector, and two video circuits.

denser. The first two components in Fig. 10·22*f* illustrate coil forms on which fixed-tuned inductances are wound. The next two components are heater chokes encased in bakelite. Two sizes of fixed resistors are shown in Fig. 10·22*g*.

In constructing miniature-tube amplifier strips, the choice of condenser types is perhaps the most important item. It is a matter of choosing between the button and the high-dielectric-constant ceramic, or "dogbone" (Fig. 10·22*d*), condensers because these two are much smaller than the regular condensers. The "dogbone" condenser requires a little more mounting space than the button condenser does, but it is easier to replace if defective. Also, since the condenser is mounted by means of pigtails, the mechanical structure required is much simpler. The button condensers, however, make a very neat assembly that seems to be completely devoid of parts. The inductance in the ground lead of the condenser is virtually zero, since the outside ring forms a large grounding surface. The feed-through construction of these condensers is excellent for bypassing and filtering circuits as they enter the box and for reducing some of the confusion around the tube sockets, as illustrated in Fig. 10·24.

The chief difference between the strips shown in Fig. 10·24*a* and *b* and the one of Fig. 10·24*c* lies in the different type of condenser used for bypassing. The first two are undeniably cleaner in appearance, but replacement of a defective condenser is more difficult.

CHAPTER 11

R-F TEST EQUIPMENT

By R. D. O'Neal and J. M. Wolf

11·1. Introduction.—As the operating frequency of a receiver goes up, through vhf into microwaves, the problems and methods of testing the r-f portion change considerably. The change from lumped parameters, and the introduction of coaxial line circuits, of resonators, and finally of waveguide transmission lines all have a profound effect on the way in which test equipment is to be inserted into the circuit as well as the type of tests that it must make. Such special problems and the equipment that has been designed to meet some of them will be considered in this chapter.

In a majority of receiver applications the design is such that sufficient gain is available in the receiver to bring the output due to noise, either external or internal in its origin, up to a point where it masks any weak signal that may be present. Such a receiver may be called *noise limited.* A second class, of somewhat less wide use, has a definitely limited amount of gain available, thus restricting the signals that may be received to those above a critical level. This may be done for reasons of economy or, more commonly, to provide a desired mode of operation in a complete system. Such receivers may be called *gain limited.* A subdivision of noise-limited receivers may be made according to whether the noise originates primarily inside or outside the receiver. At lower frequencies and in ordinary locations, external noise will almost always be the limiting factor.

The effects of slight deterioration in receiver noise figure on over-all system performance will be considerably different in these various cases. Such loss in noise figures may arise from increase in attenuation of the signal between the antenna and the receiver, from attenuation in the r-f parts of the receiver itself, or from increase in the internal noise of the receiver.

Consider first a noise-limited receiver with the noise originating largely outside the receiver. Any increase in attenuation between the antenna and receiver affects both signal and noise equally, as long as it is insufficient to let the internal receiver noise become important, and may be compensated for completely by an increase in receiver gain. Furthermore, any increase in internal noise is unimportant as long as it is a minor

factor in the total noise. Thus in such a situation, receiver noise figure is relatively less important than receiver gain in determining over-all performance. This has often been the case in communication applications at frequencies up to 30 Mc/sec or so. Consider next a gain-limited receiver. Here, by definition, noise plays little or no role in determining performance; accordingly an increase in internal receiver noise will have little or no effect. Increase in attenuation, on the other hand, will make a direct change in performance, since it can no longer be compensated for by an increase in receiver gain. In the last category, a noise-limited receiver where the noise originates internally, deterioration of noise figure produces an equal deterioration in performance no matter how it arises. This last situation therefore is the one demanding the most complete and careful check on receiver properties. It is the situation that prevails in microwave radar systems. The apparatus and techniques developed for testing these have therefore been aimed at the most exacting conditions likely to be encountered. The following discussion will be written largely in terms of radar experience, with the full realization that many receiver applications are less exacting and that some of the specialized techniques are not always applicable.

The term signal generator has been used to denote an instrument that furnishes accurately known amounts of power, usually over a wide range such as 50 to 80 db. The power output may be c-w or modulated. If the instrument performs additional functions, such as measuring the power or frequency of an external source, it is usually called a signal-generator test set.

Signal-generator test sets perform some or all of the following functions:

1. Measure receiver sensitivity.
2. Measure receiver pass band.
3. Aid in aligning receivers.
4. Aid in tuning receiver to transmitter.
5. Check AFC performance.
6. Check beacon discriminators.
7. Measure TR recovery.
8. Measure transmitter power.
9. Measure transmitter pulling.

The first four functions are useful in most applications, whereas the rest are more specialized.

11·2. R-f Components.—Figure 11·1 shows in schematic form the r-f circuit of a typical test set.[1] The *r-f oscillator* is tuned to the required

[1] A more detailed description of some of the components used in test sets is given in Vol. 9 of the Radiation Laboratory Series.

frequency as measured by the *frequency meter*, and the power from the *r-f oscillator* is varied by the *attenuator* or power output adjustment to give reference reading on the *power meter*. Power output of the test set can then be read directly from the *calibrated attenuator*. Normally, in measuring radar receiver sensitivity, this power is decreased until the test signal attains a standard criterion as seen on the radar indicator.

To measure power entering the test set, the *r-f oscillator* is turned off, and the *calibrated attenuator* is adjusted for reference reading on the *power meter*. The input power and frequency is then read directly from the *calibrated attenuator* and *frequency meter*, respectively.

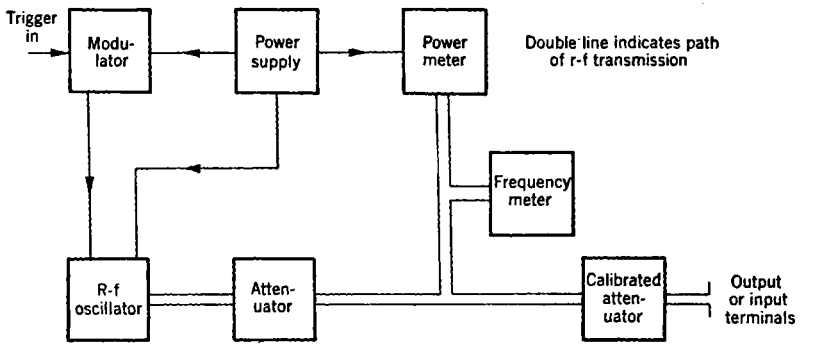


FIG. 11·1.—R-f circuit of typical test set.

The oscillator used in a signal generator is usually the same tube type as is used for the local oscillator of receivers in the band covered by the particular test set. The power output adjustment may be either mechanical or electrical but must in all cases have the following properties:

1. It must be capable of bringing the output of any oscillator tube meeting applicable specifications to the standard reference level.
2. It must produce no appreciable frequency shift, at least within a range of ± 3 db of the reference level.
3. There must be a good impedance match looking back from the T-junction toward the oscillator.
4. There must be sufficient attenuation between the frequency meter and the oscillator to prevent frequency pulling.

With a negative-grid oscillator such as the 446 or 2C40, it is possible to meet the first two requirements by changing power output by means of plate voltage variation. The last two requirements are usually met by the provision of a fixed amount of dissipative attenuation between the oscillator and the T-junction. With velocity modulation oscillators it is necessary to use mechanical means, such as the dissipative attenuators

shown in Fig. 11·2 or 11·3. Attenuation is achieved in one case by moving a strip of resistive material (100 to 300 ohms per square) from the narrow side of the guide, where the field intensity and therefore the attenuation are small, toward the center of the guide, where the field strength and hence the attenuation are a maximum. In the other case, the attenuation is changed by varying the depth to which the resistive material is inserted into the guide. In order that the presence of this material in the r-f line may not cause appreciable mismatch, either the ends of the material are tapered or each end contains an appropriate r-f transformer.

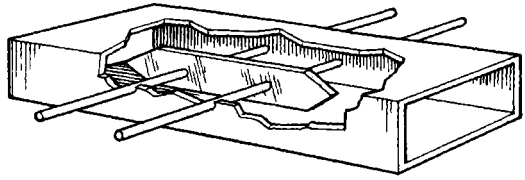


FIG. 11·2.—Resistive attenuator. The attenuation is increased by moving the resistive material toward the center of the waveguide.

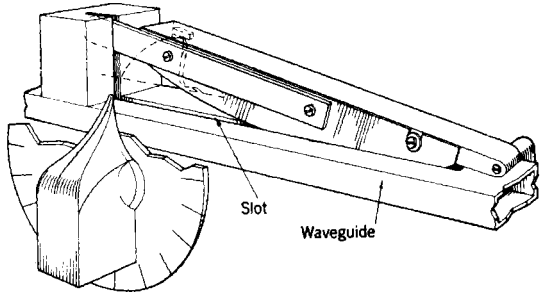


FIG. 11·3.—Resistive attenuator. The attenuation is increased by increasing the depth of penetration of the resistive material.

The power monitor for microwave frequencies must usually be capable of measuring an average power of approximately 2 mw, over the temperature range to which the set may be subjected and over the frequency range required of the set, without appreciable error or drift and without any tuning adjustments.

At microwave frequencies, power is ordinarily measured by its heating effects. This may be accomplished by making use of the thermoelectric effect or the change of electrical resistance with temperature. The latter is more often used. A thermistor (see Fig. 11·4) whose sensitive part is a mixture of metallic oxides having a high negative temperature coefficient of resistance is matched into the r-f line. It forms one arm of a Wheatstone bridge which is calibrated in such a way that the power can be

read directly by the unbalance of the bridge. In other cases the change in d-c power required to effect rebalance of the bridge, after unbalance by r-f power, is measured and used to calculate the r-f power. Measurements may be made with an accuracy of $\pm \frac{1}{2}$ db. To prevent drift, compensating networks may be used. Thermistor mounts can now be made that at 3000 Mc/sec will present a VSWR of less than 1.5 over a 40 per cent band. At 9000 Mc/sec the same can be accomplished over a 12 per cent band.

At lower frequencies (150 to 600 Mc/sec), a diode voltmeter is often used for power measurement, particularly in the case of pulse signals. Indication may be either on a meter or on a calibrated oscilloscope. With

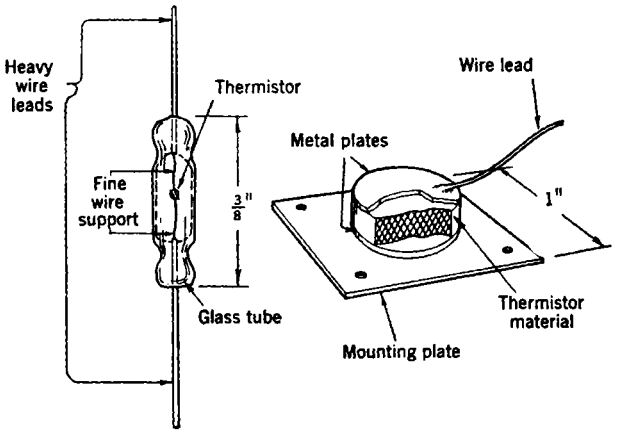


FIG. 11·4.—Typical thermistors for power measurement and temperature compensation.

care, accuracies comparable to the above-quoted $\pm \frac{1}{2}$ db may be attained. Sometimes in the past, careless construction and poor design have resulted in considerably larger errors. Such instruments may be made quite insensitive to changes in temperature or frequency and, if an oscilloscope with a high-gain amplifier is used, may have good power sensitivity.

The frequency meter is usually either a cavity operating in the TE_{011}-mode[1] for frequencies higher than 8000 Mc/sec or a coaxial cavity for frequencies in the range down to 2400 Mc/sec or lower. The two types are shown in Fig. 11·5. In test sets these frequency meters are usually used as reaction indicators; that is, when the cavity is tuned to resonance, a portion of the power is reflected, causing a change in power to the thermistor and, therefore, in the meter reading. At frequencies in the range of 150 to 600 Mc/sec the resonating circuit often consists of parallel

[1] See for example, J. C. Slater, *Microwave Transmission*, McGraw-Hill, New York, 1942.

r-f lines terminated at one or both of their ends in the sides of a rectangular silver-plated brass shield box, tuned by a condenser at one end or at the center. These as well as the two types mentioned above may be used as transmission meters if, in addition to the coupling by which power is fed in, there is a second coupling to which a detector and meter may be attached. At resonance, power is transferred through the cavity to

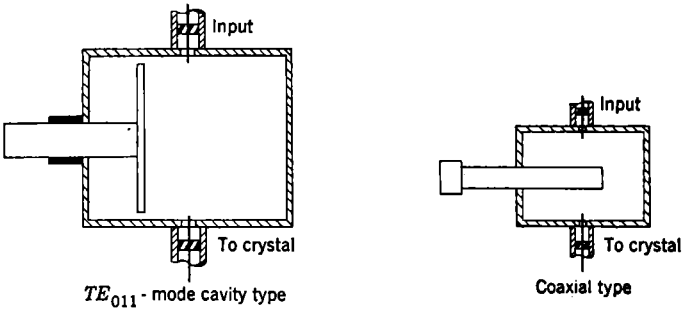


FIG. 11·5.—Frequency meters.

the detector. When used as a transmission cavity, the frequency meter gives an output indication only when it is tuned near resonance, whereas in the reaction method an output reading is obtained at all times.

The variable calibrated attenuator may be either dissipative, as shown in Figs. 11·2 and 11·3, or of the "waveguide below cutoff" type as shown in Fig. 11·6. For a given size of waveguide there is a critical frequency, known as the cutoff frequency, below which signals are attenuated

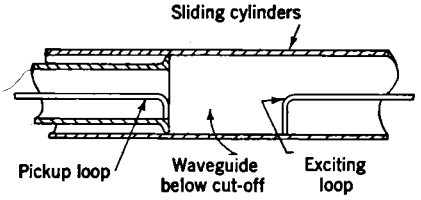


FIG. 11·6.—Waveguide attenuator.

instead of being transmitted freely. For each configuration of electric and magnetic field within the guide, there is a particular cutoff frequency. These configurations are known as "modes." The simplest modes usually have the lowest cutoff frequency. Signals at a frequency below cutoff are attenuated logarithmically at a rate that is different for each mode, the rate being least for the simplest or lowest mode. Considerable control may be exercised over what modes are excited in a waveguide by the proper choice of coupling means, that is, loop or probe, and by the

placement of whatever coupling means is used. For a rectangular waveguide in the lowest mode, that is, with the electric field parallel to the narrow side, at a frequency considerably less than the cutoff frequency, the attenuation is 27.3 db in a distance equal to the width of the guide.

Attenuation in decibels increases linearly down the guide, provided only one mode of excitation is present. Usually, at very short distances, other modes are present, but all except the TE_{11}-mode for loop excitation (Fig. 11·6) and the TM_{01}-mode for disk excitation are very quickly attenuated.

The dissipative attenuator has the advantage of presenting good impedance match at each end and of having low insertion loss. Because of the difficulty of duplicating resistance units in quantity, each attenuator must have an individual calibration. The waveguide type has the advantages that (1) the attenuation in decibels varies linearly with the separation between exciting and pickup loops (strictly true only for separations larger than a certain value), (2) it can be easily reproduced, and (3) it is relatively insensitive to frequency. However, it has the disadvantage that it presents a very bad mismatch (a waveguide below cutoff presents an VSWR of practically infinity) requiring attenuation for matching. Consequently the insertion loss (the ratio of power into the attenuator when it is adjusted for minimum attenuation to the available power out) of such an attenuator is usually high (10 db or greater).

The range of these calibrated attenuators should be such that the signal output may cover the range from about -10 to -85 dbm. This variation of the attenuator may be achieved in two ways, continuously or in steps. A dissipative attenuator continuously calibrated over this wide range is difficult to design. An alternative method is the use of a fixed attenuator, such as a directional coupler inserted in the transmission line in series with a calibrated attenuator. Thus a 40-db coupler and an attenuator calibrated to about 45 db would be suitable.

The previous figures are based on the assumption that the test set will be connected to the receiver through some lossy device such as a directional coupler or pad offering a loss of 20 db or more. If it is necessary to couple directly into the receiver, the attenuator should be such as to cut down the signal output to at least -105 to -110 dbm. This is relatively easy to accomplish with a waveguide-below-cutoff type but often difficult with the dissipative type. These power levels are appropriate for a receiver bandwidth of 1 Mc/sec or so. A narrower receiver would require additional attenuation.

11·3. Pulse-modulated Signal Generators.—In order to test completely a radar receiving system the test set should furnish a pulse equal in length to that of the radar transmitter. In an amplitude-modulated (pulsed) test set the pulse is usually made variable in width to accom-

modate all radar systems in the particular frequency band covered by the set and to permit checking of such special features as beacon discrimination. The pulse is also made movable in time with respect to the trigger so that it can be placed in various portions of the radar scope. For some applications, such as the testing of beacon discriminators, it is preferable to have fixed limit pulses in addition to a continuously variable one. Figure 11·7 shows a block diagram and Fig. 11·8 shows a schematic diagram of a representative circuit for a pulsed-signal generator. Two essentials of this method of modulation are (1) the pulse must have a fast rise time and a fast decay time (about 0.10 μsec for each, or usually

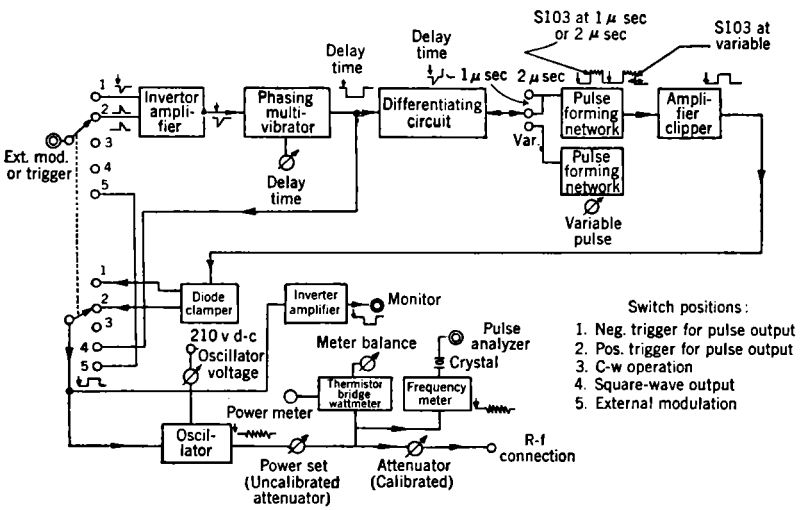


FIG. 11·7.—Block diagram of pulse-modulated signal generator.

about one-fifth the pulse width) in order to simulate a radar pulse, and (2) there must be inappreciable jitter in the pulse (usually less than 0.2 μsec) in order to present a clean picture on an oscilloscope.

The oscillators used in these test sets usually do not generate enough power to permit monitoring of the pulse directly by means of a thermistor bridge. Accordingly an alternative method must be used for monitoring. One possibility is shown in Fig. 11·8. The oscillator is tuned to the correct frequency, and the power level set to the correct value on c-w operation; then, by throwing the switch to + or − *trigger* position, the oscillator tube is thrown out of oscillation except during a pulse, at which time the voltage applied to the tube is the same as the voltage applied during c-w operation. Although the power output is the same on pulse as on c-w, some older sets exhibited a frequency change of as much as 3 Mc/sec due to thermal effects. In later models a factory adjustment of

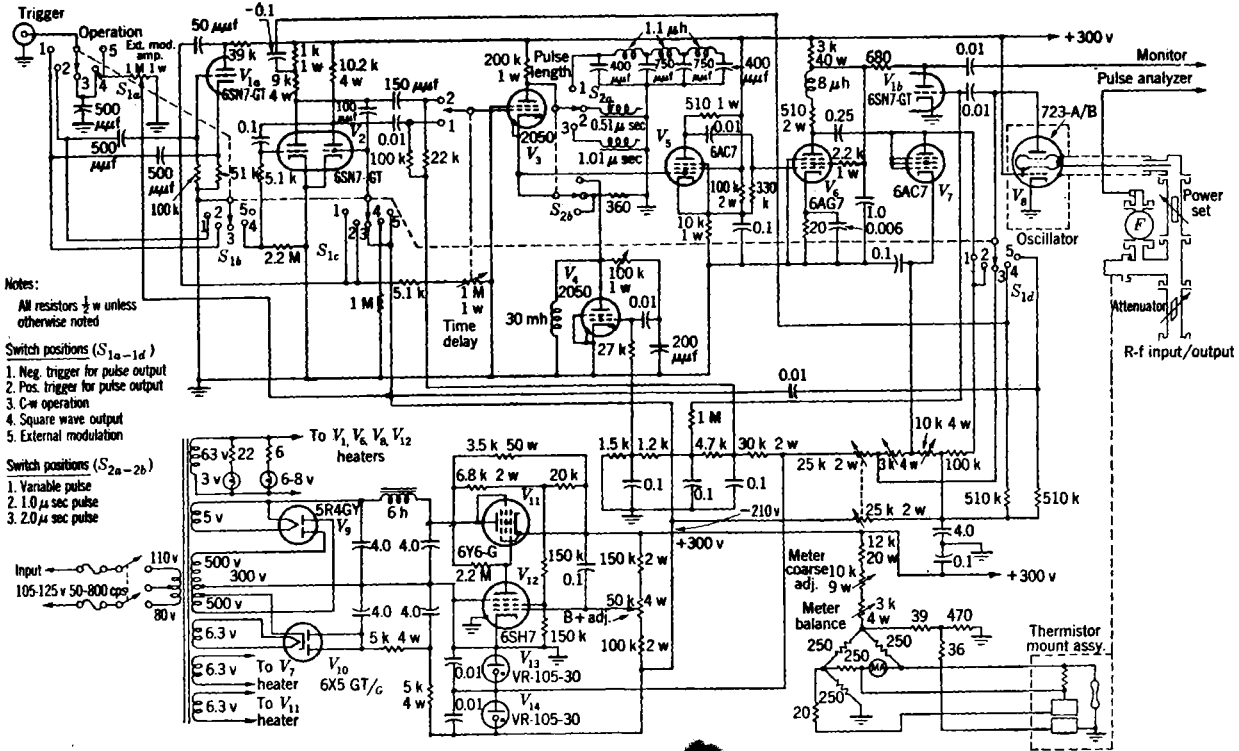


FIG. 11·8.—Circuit diagram of pulse-modulated signal generator.

the divider to which the cathode is attached ensures that there are no frequency shifts. For tubes generating relatively high power, such as lighthouse tubes, monitoring may be accomplished by setting the power level using a long pulse of 15 to 30 μsec at a fixed recurrence frequency. The peak power of this monitoring pulse and the pulse used will be the same, but the frequency in the two cases may be slightly different.

A method of monitoring pulse output directly at microwave frequencies, by measuring the change in the resistance of a bolometer element called a "barretter," has recently been developed. The sensitive element is a very fine platinum wire suitably mounted in the r-f line to offer a matched impedance. The pulses resulting from the resistance changes with each r-f pulse are amplified with a constant-gain amplifier to a level that permits the use of a meter to measure rectified current from a diode detector. The variation of the resistance bears a close analogy to the voltage developed across a parallel *RC*-network receiving current pulses. The time constant of the barretter is about 350 μsec. The amplitude of the pulses is independent of the repetition period, provided the repetition period is long compared with the time constant of the barretter-amplifier combination. The amplifier used must have a fairly wide band (about 10 to 300 kc/sec) in order that the pulse amplitude measurement be independent of the repetition period.

11·4. F-m Signal Generator.—Much of the complication of pulse-modulated test sets arises from the complexity of the circuits necessary for production and control of the modulating pulses. The possibility of producing a high rate of frequency modulation with a reflex klystron has led to the development of another type of test set, known as the f-m test set, which is both simple and versatile. The fundamental arrangement is almost identical with that shown in Fig. 11·1, except that instead of applying a modulating pulse to the oscillator, a sawtooth wave is applied to the reflector as shown in Fig. 11·9*a*. This wave produces frequency modulation together with change in power output as shown in Fig. 11·9*b* and *c*. Figure 11·9*d*, *e*, and *f* shows how the resulting output may simulate a pulse signal in the receiver.

The width of the signal depends on the pass band of the receiver and the slope of the sweep used to produce frequency modulation. It is feasible to use sweeping rates of from 3 to 6 Mc/sec per microsecond. Thus if the receiver has a bandwidth of 3 Mc/sec a signal as short as $\frac{1}{2}$ μsec may be produced, which is comparable in length to the normal pulse signals ordinarily used with such a receiver. If, however, the receiver bandwidth is considerably greater and the normal pulse signals correspondingly shorter, it becomes impractical to sweep the test signal fast enough to simulate the normal signal.

The schematic diagram of an f-m test set for the 10,000-Mc/sec region

is shown in Fig. 11·10, and a photograph of the instrument is shown in Fig. 11·11. Much of the circuit arrangement is similar to that of the pulse-modulated test set. The method of indicating power output requires some comment, however. Since the oscillator is swept completely through one or more of its modes, it is clear that a measurement of average power during frequency modulation would be meaningless. It is necessary, therefore, to remove the frequency modulation by reducing the amplitude of the sawtooth wave to zero, to set the oscillator frequency to the desired value by simultaneous adjustment of the mechanical

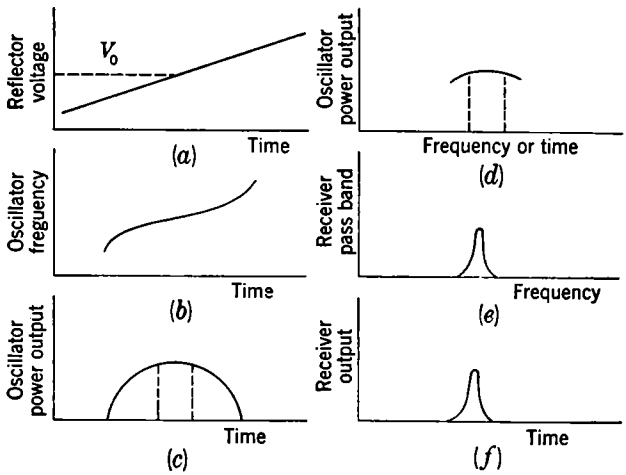


FIG. 11·9.—Scheme of operation of f-m signal generator.

tuning and the steady reflector bias, and to make the adjustment for reference power level under these conditions. As long as the mechanical tuning is not disturbed, any combination of sawtooth wave and steady bias that brings the frequency to the desired value will also bring the power output to the reference level, permitting direct reading of the output power level on the calibrated attenuator.

It should be noted that the process of initial adjustment is much simpler in the case of the f-m signal generator than for the pulsed signal generator. Whereas in the latter case it is necessary that both mechanical tuning and reflector bias be accurately set before any signal is visible in the receiver, in the f-m signal generator it is necessary only that the mechanical tuning be in the general neighborhood of the correct value to produce visible signals. This problem of obtaining an initial indication from the signal generator is, of course, much more serious in the microwave region where frequency bands cover hundreds or even thousands of megacycles per second.

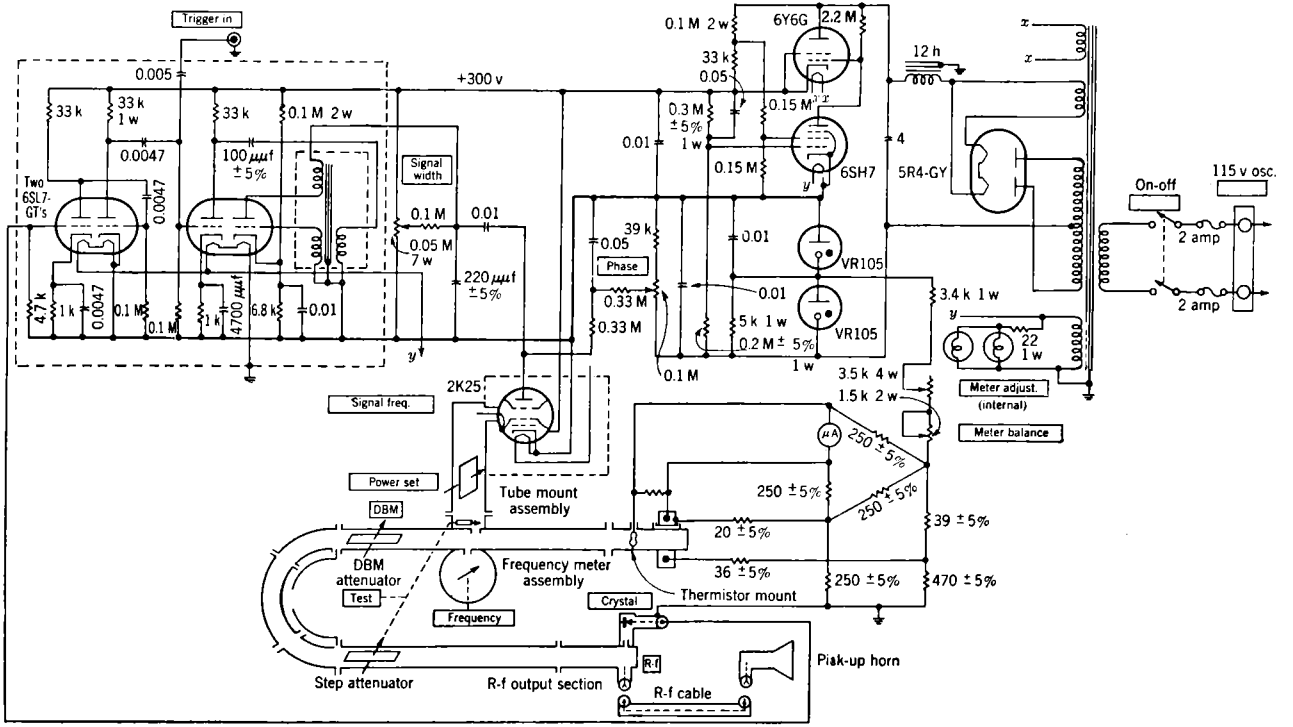


FIG. 11·10.—Circuit diagram of f-m signal generator.

Suppose, for example, that it is desired to measure the sensitivity of a receiver that is known to be tuned to the proper frequency. It is necessary only to connect the signal generator to the receiver, set the amplitude of the sawtooth sweep at maximum, and vary the mechanical tuning until a signal is seen in the receiver output. The output signal is ordinarily displayed on an A-scope. If, after some output signal is found, the mechanical tuning is varied slightly, the signal will appear to move slightly in time or range and will simultaneously vary in amplitude. This behavior is the result of the attainment of the correct frequency through various combinations of mechanical and electrical tuning. The proper setting of the mechanical tuning is the one that gives the largest

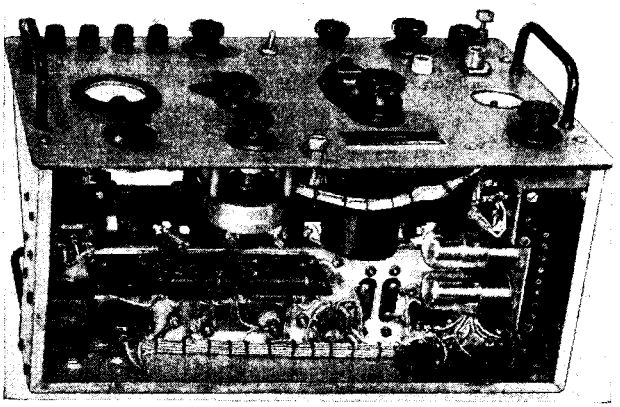


FIG. 11·11.—Photograph of f-m signal generator.

output signal, since this corresponds to operation in the center of the oscillator mode at the desired frequency. With the mechanical tuning so set, the standardization of output power level may be made as described above.

It is apparent that the output signal from the receiver is essentially a curve of receiver response vs. frequency. By reduction of sawtooth sweep amplitude together with any adjustment of reflector bias necessary to keep the signal on the screen of the A-scope, the response curve may be spread out to a width permitting easy examination. If the wavemeter of the test set has a sufficiently high Q, its reaction on the test-set output will be visible in the receiver output in the form of a small pip at the frequency to which the wavemeter is tuned. This pip may be moved about on the response curve by tuning the wavemeter and may be used, for example, to measure the bandwidth of the receiver between half-power points. Another use is found in checking the AFC circuits of a radar receiver. For this purpose, the wavemeter is set and allowed to

remain at the transmitter frequency. When the receiver response curve is viewed, the wavemeter pip should then appear in the center of the pattern. Failure to do so is an indication of misalignment. It should be noted that for this and similar tests on radar systems, the radar transmitter and the test set are operated simultaneously. The interconnections of the units must be such as to permit this simultaneous operation.

11·5. Accuracy of Signal-generator Test Sets.—Inaccuracies may be divided into two classes: (1) inherent errors in the test set and (2) reading errors.

Errors in measurement of power fed into a test set as well as errors in power output may be due to (1) temperature sensitivity of the attenuator and the thermistor mount; (2) frequency sensitivity of the attenuator, thermistor mount, and whatever device is used for taking a fraction of power to the thermistor for monitoring power output, such as a T; and (3) error of calibration. These inaccuracies are such that measurement of absolute power input, as well as power output, cannot usually be guaranteed to better than ±2 db over a −40° to +55°C temperature range and a 10 per cent frequency band at 9000 Mc/sec. At higher frequencies the frequency range will be less; at lower frequencies, greater. These errors for a representative test set are shown in Table 11·1.

TABLE 11·1.—ERRORS IN A REPRESENTATIVE TEST SET

Source of error	Maximum error for power output, db	Maximum error for power input, db
Calibrated attenuation	±1.5	±1.5
Power split variation with frequency	±0.5	0.0
Thermistor bridge effect of temperature	±0.5	±0.5
Thermistor mount—VSWR	±1	±0.2
Calibration of components	±1.0	±1.0
Total	±4.5	±3.2
Over-all error with calibration curves	±2	±2

In addition to the above errors, there may be additional error in measuring receiver sensitivity due to leakage of the test signal from points other than the output of the attenuator to the receiver either through the antenna or through some of the r-f components of the receiver. This leakage is due to a design fault in the signal generator, in that components such as the thermistor mount where relatively high level power (high with respect to minimum discernible signal) exists, are not constructed tightly enough to prevent leakage of appreciable amounts of r-f power. In the case of some signal generators designed in the past, it has been necessary to take great care to locate the signal generator in a place well

shielded from the radar antenna and never try to make minimum discernible signal measurements with the receiver r-f components exposed. The possible errors resulting from this leakage signal are shown in Fig. 11·12 as a function of the ratio of leakage signal to test signal power. The two curves represent the two extremes of relative phase between the two signals. In order to have less than ±0.2-db error from leakage, it is necessary to have the leakage signal at least 30 db below the test signal. Present techniques have not achieved a construction that cuts down leakage to the point where the test set may be placed in the beam of the antenna. If leakage is suspected, the test set should be moved slightly to see if the changing phase causes a changing receiver sensitivity reading. It is to be noted that leakage errors are reduced by using a tangential signal rather than a minimum discernible signal because of the different power levels.

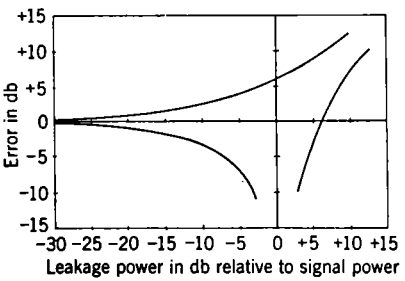


FIG. 11·12.—Possible errors due to leakage in sensitivity measurements.

Inaccuracies in frequency measurements may be due to (1) temperature changes, (2) humidity changes, (3) impedance mismatch of load attached to cavity, (4) extraneous modes, and (5) calibration errors. Errors due to temperature variation are usually made negligible by using invar in the construction of the cavity. Humidity errors due to change in dielectric constant with change in atmospheric water content may amount to as much as 0.04 per cent at a temperature of 55°C and 95 per cent relative humidity. (These are the temperature and humidity limits set up by a joint Army-Navy specification for test equipment.) In most test sets, the frequency meter is not appreciably pulled by impedance mismatches because, for other design considerations, these mismatches are minimized. Extraneous modes are ordinarily eliminated in design. Calibration errors in the microwave region are usually about $\pm\frac{1}{2}$ Mc/sec.

Ordinarily, high absolute accuracy is not required of a frequency meter used in a radar test set. Specifications may call for an accuracy of ±1 Mc/sec at a particular frequency (usually for beacon operation) and an accuracy of ±3 Mc/sec over a frequency band of 10 per cent at about 9000 Mc/sec and 40 per cent at about 3000 Mc/sec, for a specified temperature and humidity. It is usually recommended that for precise fixed frequency measurements at microwave frequencies, as in some beacon applications, fixed-tuned sealed cavities with the critical parts made of invar be used. The loaded Q of the cavity should be high enough

so that accurate readings can be made without causing pulling but low enough so that, as an oscillator is tuned rapidly through the natural frequency of the cavity, a good indication is obtained.

Reading errors may be classed in two categories: (1) those due to differences in ability among observers to determine precisely when a certain condition exists and (2) those due to the use of improper sweep speed, improper pulse width, etc. For the frequency meter, errors are due to reading the indicating meter and the frequency dial. These are obvious sources of error and normally negligible. In measuring receiver sensitivity, however, the inaccuracy of reading is much worse.

Routine field measurement of receiver sensitivity is ordinarily accomplished by applying the signal from a test set to the receiver, displaying the output of the receiver on an oscilloscope, and adjusting the input signal level so that the output indication has a prescribed appearance. Two criteria have been used for setting the signal level: (1) the *tangential signal* and (2) the *minimum discernible signal.* In both cases the signal output of the receiver is compared with its noise output, thereby giving a significant measure of receiver performance. The tangential signal criterion is best described in terms of a sketch such as Fig. 18·1. The noise output of the receiver causes the horizontal trace of the oscilloscope to be widened into a band with diffuse edges. The signal appears on this band and is so adjusted that the bottom of the band at the signal location is level with the top of the band elsewhere. The minimum discernible signal, as its name implies, is the smallest signal that the observer can see in the presence of the receiver noise. The test may be so arranged that the signal can be moved back and forth on the oscilloscope trace in order to facilitate its identification. It should be noted that neither of these criteria establishes the input signal power equal to noise power as used in the definition of noise figure. There is, however, a relatively constant ratio between the three power levels, particularly for fixed conditions of pulse length, sweep speed, etc. Careful measurements have been made, using a number of observers to determine the accuracy of these methods with different types of presentation.[1] Some of the results are shown in Fig. 11·13. Both f-m and pulse-modulated test sets were

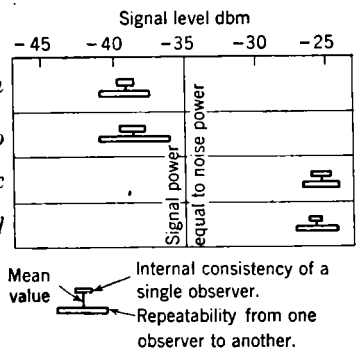


FIG. 11·13.—Reproducibility of various test-signal criteria.

[1] S. Katz, RL Report No. 1023, Jan. 15, 1946.

used, as well as both criteria of signal level. It is seen that for A-scope presentation the tangential signal is slightly more reproducible, particularly from one observer to another. In addition, the higher signal level involved reduces troubles from leakage. On the other hand, the criterion is applicable only when an A-scope display is provided.

11·6. Noise Generators.—It has been possible to build r-f noise sources that may be used to measure the sensitivity of a receiver at 3000 Mc/sec. So far, for field use a single-cavity velocity-modulated tube operated with the reflector connected to the cavity to prevent oscillation has been used. Simple considerations based on shot effect indicate that the noise power P_n is proportional to $I_B B$, where I_B is the beam current and B is the noise bandwidth. The proportionality factor takes into account the efficiency of the grids and the effects of transit time. The noise generator is calibrated against a laboratory-type signal generator and presumably will hold its calibration for long periods.

With tubes manufactured over a year ago and evacuated for 2000 volts, the linear relation between beam current and power output has been found to hold well, and calibrations remain constant over a period of many months.[1]

The noise generator has the following advantages:

1. It operates at a low r-f level and therefore presents little or no leakage problem.
2. It requires no complicated r-f circuits.
3. It is small and simple to operate.
4. It is capable of very precise measurements.

It has the following disadvantages:

1. It has such a low-power output level that power must be fed directly into the receiver line in order to get sufficient indication, which means either using an r-f switch or breaking into the normal line.
2. Correction must be made for receiver pass band if this pass band is very wide (unless new noise generators are designed having a pass band wider than the 2 or 3 Mc/sec that is characteristic of present ones).

In use the receiver gain is adjusted until a convenient reading due to receiver noise is obtained on a meter connected to the second detector output. If sufficient noise from the noise source is then added into the receiver line until the meter reading indicates twice the previous power reading, the beam current (read on another meter) passing through the

[1] However, tubes of later manufacture pumped for only 1000 volts have often not exhibited this linear relationship and have not maintained calibration reliably. This might be accounted for by the insufficient evacuation.

noise source gives the noise power output from which the over-all receiver noise figure may be obtained directly.

Echo Box.—Although primarily a radar development,[1] an echo box, where applicable, is a versatile and useful piece of test equipment for the frequent periodic checks necessary to maintain the performance of a receiver at its peak. To permit its proper use, certain features of the receiver design must be chosen in a definite way. An echo box is a tunable resonator of high Q which is loosely coupled to a radar system in such a way that transmitter pulses set up transients whose maximum amplitude is proportional to the energy in the radar pulse. The transients decay exponentially, giving rise to a diminishing signal called "ringing." The point at which this signal disappears into noise on the radar indicator depends on the initial amplitude of the transients and on the receiver noise level. The duration of this echo-box signal, usually about 40 or 50 μsec, is called "ringtime." The ringtime is proportional to radar performance and is measured in practical units of radar range.

11·7. Miscellaneous Equipment. *Crystal Checker.*—A crystal checker saves time in trouble-shooting and also is very useful in sorting good from bad crystals in supplies of used crystals. By testing a large number of mixer crystals, a statistical correlation has been found between the r-f characteristics of the crystal and the following d-c properties:

1. Front resistance.
2. Back- to front-resistance ratio.
3. Back-current under constant voltage.

These measurements can be made simply and quickly. However, results are not infallible. In the case of X-band crystals, a few per cent of bad crystals may be accepted, although these are not more than 1.5 db worse than those which fulfill the acceptance test. The checker cannot be depended upon to distinguish very good crystals from those which just pass the specification limit.

Spectrum Analyzer.—It is often convenient in trouble-shooting to have a panoramic picture on an oscilloscope presenting the spectrum of the transmitter and LO outputs. This electronic analyzer also may be used to check mode peculiarities such as tuning discontinuties in LO tubes and LO "jitter." It is often used in AFC alignment and trouble-shooting.

The electronic spectrum analyzer normally consists of an electronically-swept oscillator sweeping through a range of 40 to 100 Mc/sec, a narrow-bandwidth i-f amplifier (50 to 125 kc/sec), and an oscilloscope indicator whose horizontal sweep is synchronized with the sweep applied

[1]See *Radar System Engineering*, Vol. I, Radiation Laboratory Series, for a more extended discussion of the echo box.

to the oscillator. A block diagram of such an analyzer is shown in Fig. 11·14. A frequency marker usually is supplied by a high-Q frequency meter. Electronic markers separated by 1 or 2 Mc/sec are sometimes supplied. Usually the receiver has high gain; therefore, an attenuator that may or may not be calibrated may be used to provide a wide working range for the analyzer.

Equipment for Measuring Standing Waves.—The principal purpose for this equipment is to measure the VSWR and sometimes the loss of r-f line components.

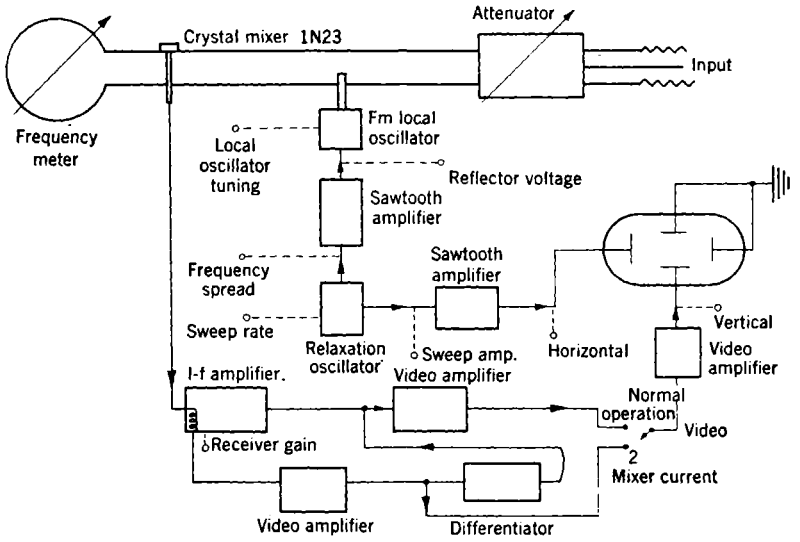


FIG. 11·14.—Block diagram of spectrum analyzer.

Standing wave equipment may consist of any one of the following:

1. A slotted section with a sliding probe moving over a distance of greater than ½ wavelength, together with a suitable detector.
2. A device such as a waveguide squeeze section or movable dielectric causing a phase shift of the standing wave past a suitable fixed detector.
3. A bidirectional coupler with a detector.

A suitable detector for the latter must have an attenuator, since ordinarily there is considerable power-level difference between the incident power and the reflected power. A word of warning should be given against the indiscriminate use of bidirectional couplers. Errors may result from

1. Poor directivity in the coupler coupling out reflected power.
2. Calibration of the two couplers.
3. Attenuator calibration.

The errors contributed by Item 1 may be seen from the following example. If the directivity is 32 db, a VSWR of 1.20 may be measured as 1.26 or 1.16; but if the directivity is 20 db, the same VSWR of 1.20 might be measured as 1.0 or 1.5.

Power Meters.—A sensitive power meter (about 2 mw full scale) is useful in r-f trouble shooting, for it permits measuring of LO power, although, indeed, this may require a special adapter. Also such a power meter may be used as the detector for VSWR measurements with either a slotted line or bidirectional coupler.

Power meters ordinarily consist of (1) a power-measuring head and associated circuit which is the same as that used in signal generators and discussed above and (2) a calibrated attenuator to extend the range of the power head.

11·8. Coupling R-f Test Equipment to the Receiver.—Receiver testing requires that means be provided for coupling the test equipment to the receiver. Since the quantitative measurements made are no more dependable than the coupling device used to link the test equipment to the receiver, particular attention must be paid to the selection of an adequate coupling device for each application. In general any coupling device should have the following properties:

1. It should provide sufficient attenuation to permit measurements of receiver sensitivity. This means that with most signal generators the coupling device must contain at least 20 db of attenuation. This is in case the smallest test signal is -85 dbm, and a connecting cable of approximately 4 db is used to connect the test set to the coupling device. The coupling device then reduces the test signal to -109 dbm. Since the c-w signal equal to noise is seldom better than -100 dbm even for the best receivers, the value of 20 db is adequate. (The receiver bandwidth is assumed to be at least 1 Mc/sec.)
2. The attenuation of the coupling device must be known accurately. It is current practice to specify that the accuracy shall be $\pm\frac{1}{2}$ db. Otherwise, absolute performance measurements cannot be made.
3. The attenuation of the coupling device should not vary appreciably with frequency.
4. Both the input and output of the coupling device must present a good impedance match; otherwise there may be appreciable error due to resonances between either the test equipment and the coupling device or the coupling device and the receiver. In the

latter case the VSWR should practically always be less than 1.2, and usually it is easy to keep it under 1.1. In the former case a higher VSWR can be tolerated, since usually a cable with some loss is used to connect the test equipment to the coupling device. In no case, however, should the VSWR be greater than 1.5.

There are two general classes of coupling devices for radar receivers: (1) the type that is normally used for measurements of both transmitter power and rece ver sensitivity. These are "sampling" devices, taking only a fraction of the power for measurement purposes but not interfering with the main flow of power in the system. (2) The dissipative attenuating pad type which is inserted directly in the transmission line between the antenna and receiver. In this case any flow of power must pass through this dissipative attenuator, and so the normal functioning of the receiving system is impaired. The former type might be used to make tests during normal operation, but the latter type could not. Examples of the first type are (1) line probes, (2) pickup antennas, (3) directional couplers. Examples of the second type are (1) lossy cable, (2) metallized center-conductor coaxial attenuators.

A probe in the r-f line between the antenna and receiver has serious disadvantages which limit its application as a coupling device for receiver performance measurements. These are

1. *Lack of reproducibility.* Insertion of a probe is extremely critical, coupling may change if the probe is removed and replaced or if it is adjustable or mechanically loose. Also, unless individually calibrated, only relative measurements of receiver performance are possible.
2. *Frequency sensitivity.* Even relative measurements with a probe are unreliable, since the coupling depends on the position of the standing-wave pattern in the line.
3. *Main-line reflection.* The VSWR set up by a probe may be difficult to match.
4. *Poor match at test point.* The VSWR looking into the test point usually is high in the case of probe coupling; this contributes to the frequency sensitivity of power measurements.
5. *Power split.* When a test signal is inserted into a line by means of a probe, the power split depends on the relative impedances of the line in each direction from the probe. Only if these impedances are unchanging can the power delivered to the receiver be depended upon.

Various expedients have been contrived to obviate some of these objections. Inserting the probe into the TR cavity tends to reduce fre-

quency sensitivity of coupling. Using double probes appropriately spaced also reduces the influence of standing waves, since these affect the two probes in a compensatory manner.

A pickup antenna taking the form of a horn or dipole has one major advantage as a coupling device. When placed in the antenna pattern, a test point is provided at which the entire receiving system including the antenna can be checked. This is of some importance in the case of a very long transmission line.

Since coupling has to be constant, it is essential to provide a suitable bracket to fix the pickup unit in a reproducible location with respect to

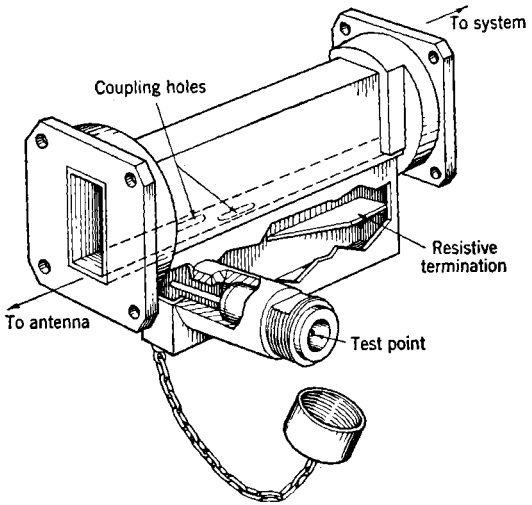


FIG. 11·15.—Directional coupler.

the antenna. A given pickup unit fixed in relation to the antenna can be calibrated against a directional coupler in the line. A serious disadvantage of antenna coupling in the case of systems with long lines is the inconvenience of transporting test equipment to the test point.

A directional coupler is the most convenient and reliable of the coupling devices listed. It consists of a short section of r-f transmission line with a section of secondary line attached. One or more openings between the sections allow power to be coupled from one line to the other. A matched probe in the secondary line feeds the test point—a type N fitting in many designs. Directional couplers are designed to couple to waves traveling in one direction in the main line and to discriminate against those traveling the other way. The device is installed so as to couple preferentially to the wave proceeding from the transmitter and to discriminate against the reflected wave from the antenna. Coupling is

therefore independent of the standing-wave pattern in the line. This unidirectional characteristic is illustrated in Fig. 11·15 showing a "two-hole" design in which two holes (waveguide beyond cutoff) in the narrow side of the guide, spaced an electrical quarter wavelength apart, provide coupling. The components of the forward wave coupled through each hole arrive at the test point in phase. Power from the backward wave coupled toward the test output travels effectively $\frac{1}{2}$ wavelength farther through one hole than through the other; since the waves are 180° out of phase, they cancel and no power from the antenna reaches the test point. Power coupled from the backward wave in the opposite direction is absorbed by the termination placed at the opposite end of the secondary line. To ensure that the coupler discriminates as much as possible against the backward wave, the resistive material used for a termination must be well matched to the secondary line. Coupling depends only on the dimensions of the coupling apertures. The coupling of a given unit does not vary from day to day. Constancy of coupling from one unit to another depends on the manufacturing tolerances. Most designs can be manufactured with coupling tolerances held to ± 0.5 db or better.

Although there are many possible designs for directional couplers, there is more uniformity in the electrical operation of various designs than in their physical appearance.[1] The unidirectional characteristic of each type depends on coupling by at least two paths or two processes in such a manner that there is constructive interference of power flowing in one direction in the secondary line and destructive interference of power flowing in the opposite direction. Coupling to both magnetic and electric vectors, coupling on opposite sides of the waveguide center line (where fields are opposite), coupling at different locations along the primary line—all are used to achieve this unidirectional effect.

Directivity is a measure of the unidirectional property of a coupler and is the ratio (expressed in decibels) of the coupling to the forward wave to the coupling to the backward wave in the r-f line. A minimum directivity of 15 db is adequate for many applications.

Other requirements of a directional coupler are that the VSWR introduced in the main line must be low. With available designs, VSWR set up in the main line by the presence of a directional coupler can easily be made 1.05 or less. The VSWR looking into the test point is the result of the combined effects of reflections from the coupling holes, the termination, and the waveguide-to-coaxial adapter. It can and should be made less than 1.5; for if it is high, measurements tend to be frequency sensitive.

[1] See R. J. Harrison, RL Report No. 724, Dec. 31, 1945, for a detailed discussion of the theory and design of directional couplers.

Directional couplers can be made that will show less than $\pm\frac{1}{2}$-db variation over a 12 per cent frequency band at about 10,000 Mc/sec.

Connecting to the Coupling Device.—Proper connection to the test point is as important as the coupling device itself. The desired accuracy of carefully calibrated test equipment cannot be achieved if a large element of uncertainty is introduced by the connecting cable (patch cord) that links the test set to the directional coupler test point. The nominal attenuation should not be relied upon because of small differences between cable samples. For such lengths, variation due to temperature and frequency sensitivity may also contribute appreciable error. Long cable runs between the test point and test equipment should be avoided, since reliable performance measurements cannot be made in such cases due to temperature and frequency sensitivity of the cable.

Much of the variation in cable attenuation is due to fittings. This is particularly true for extremely short lengths of cable (less than about 3 ft). Even the best fittings (such as the improved type N connectors recently put into manufacture) are responsible for some reflections that may add or cancel depending on relative phase. Losses due to fittings, therefore, vary with frequency, especially when the cable attenuation is low. When an additional reflection is introduced by the mismatch of the test equipment input or by the mismatch of the test point, the frequency sensitivity of the coupling is increased. In general, at least 1.5-db attenuation in the connecting cable is needed to minimize the undesired frequency sensitivity. Damaged connectors can also seriously increase the frequency sensitivity. These considerations militate against the use of very short cable lengths for attaching r-f test equipment to a radar test point.

In some cases it will be advantageous to connect the test equipment to the test point by means of flexible waveguide. The flexible character of this waveguide is meant to accommodate some bending and vibration but not the attachment of a test set at awkward angles.

Precautions regarding too sharp bending apply to flexible cable as well. Three inches is about the minimum radius of bend tolerated by present construction. Sharp bending behind a fitting is especially undesirable. Although mechanically sound in appearance, cables may be damaged electrically.

For special cases directional-coupler-type pads may be connected directly to the test set (preferably guide-to-guide connection) with the secondary output connected directly to the receiver line. The principal disadvantage of such a scheme is the mechanical inflexibility. This may be alleviated by using very short lengths of good flexible waveguide.

11·9. Design Considerations to Facilitate Testing and Alignment.—Wherever possible the receiver designer should facilitate testing and

servicing of the receiver. This involves inclusion of test points, placement of parts to make replacement easy, standardization of components to simplify replacement, and breaking the receiver into accessible and readily removable packages if this promotes ease of servicing.

The test equipment required for maintenance must be considered at an early stage in the design of the receiver. Required test points must be available at locations that are convenient. Necessary test equipment must be provided for and, where feasible, built in, for the advantages of built-in test equipment often are considerable. For radar systems, rather comprehensive recommendations have been formulated, which may often prove useful in other designs.

Test Points.[1]—Most or all of the following circuit and r-f test points should be provided in *convenient* locations with *standard* input-output fittings:

1. Directional coupler (uni- or bidirectional).
2. Video output.
3. Appropriate voltage test jacks or meters for crystal.
4. AFC discriminator output—for systems that provide for discriminator tuning.
5. Additional points—for checking various critical voltages and waveforms (such as triggers and gates) and for use with special equipment peculiar to the particular system.

In addition, some of these test points may be required at more than one place. For example, in a system with a long r-f line, some duplicate points must be provided at the antenna. The importance of locating test points so they will be readily accessible when the equipment is installed cannot be overemphasized.

Built-in Test Equipment.—Most, or all, of the following types of test equipment can be built into a receiving system:

1. Echo box (for radars).
2. Signal-generator test set.
3. R-f noise source.
4. Crystal checker.
5. Frequency meter.
6. Test oscilloscope.

Certain special considerations applying to the design of test equipment that is built into a system are

1. Echo box. The simplicity of operation of an echo box as a radar yardstick for field measurements and its versatility which implements trouble-shooting are advantages that strongly recommend it as a built-in

[1] See Cases JRB-20-25 and JRB-20-26.

component of all radar systems to which it is applicable. The echo box should be located where the cavity tuning mechanism is readily accessible or else be remotely controlled. The weight of present production echo boxes varies from 8 to 18 lb. A feasible future design objective is 4 to 8 lb. About $\frac{1}{4}$ ft^3 of space is required for installation.

2. Signal-generator test set. It is advantageous to have built-in signal generators when receiver performance must be known and maintained very accurately. General-purpose semiportable (25 to 50 lb) signal-generator test sets are now available. However, a considerable saving of weight, often a saving in complexity, and always an increase in accuracy can be effected by building in the signal generator either as a separate box attached directly to the r-f line or as an integral part of the r-f head. In either case the power for the set can be supplied by the

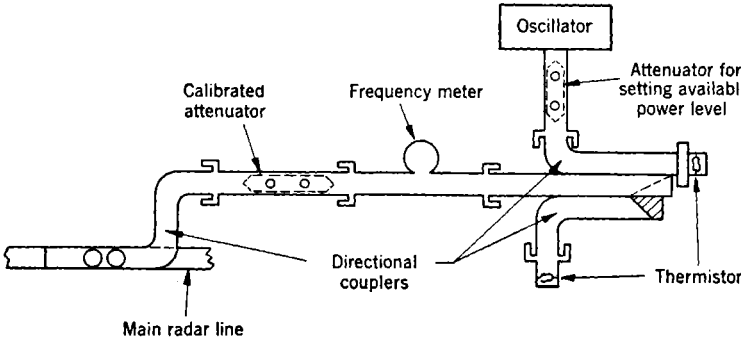


FIG. 11·16.—R-f circuit for built-in signal generator.

power supply for the system. Usually provision will be made both for f-m and c-w operation. Power must therefore be supplied for the test oscillator tube and the power meter (if a thermistor is used), and a sweep must be furnished for frequency modulation. Since the test oscillator tube ordinarily is the same as the radar local oscillator, the same power supply can be used but, of course, must be capable of furnishing the extra power. Regulation must ensure that under all conditions of line voltage variation there will be less than $0.1B$ Mc/sec (B is receiver bandwidth) frequency shift between the local oscillator and test oscillator. The sweep should be supplied from one of the scope sweeps normally used. The rate of change of voltage should be about 1 volt per microsecond. A sketch of the plumbing scheme used in one broad-band built-in signal generator is shown in Fig. 11·16. In general, wherever possible, directional-coupler-type plumbing should be used. By using proper design, test-set plumbing usually can be fitted into waste space in an r-f head. However, this may not always be desirable, since it usually necessitates

remote controls. It is often preferable to have the unit located somewhere between the r-f head and antenna, where all controls can be reached easily. In any case the built-in signal generator should be part of the r-f line, since one of its advantages is that it eliminates cable errors that may occur in the use of most test equipment.

A directional coupler always should be furnished in addition to that contained in the built-in test set, so that the calibration of the built-in set can be checked occasionally and so that other equipment may be used.

Present production built-in test sets weigh between 15 and 20 lb. A feasible future design objective is 10 to 15 lb. About $\frac{1}{3}$ ft^3 space is required.

3. A built-in r-f noise generator requires special coupling such as an r-f switch to permit direct connection into the receiver line. This measures only the receiver noise figure and has the limitations previously enumerated. A built-in noise generator together with the switching arrangement will weigh from 15 to 25 lb and occupy a space of about $\frac{1}{3}$ ft^3.

4. Crystal checker. A simple d-c crystal checker can be built into the receiver using a meter that may be on the panel for another purpose. Leads can be connected permanently to the crystal mount. This facilitates trouble-shooting, since the crystal need not be removed from the system for checking. The weight and space required are negligible.

5. Frequency meter. Fixed cavities are sometimes used as filters or as AFC cavities.

6. Test oscilloscope (particularly for radar application). Some radar indicators are installed a considerable distance from the r-f head. In such cases, r-f tuning and trouble-shooting require two people unless an auxiliary A-scope is available near the r-f head.[1] Other systems give only PPI presentation. To realize the full advantage of an f-m signal generator, an A-scope is required. Also, for best accuracy, echo-box ringtime should be read on an A-scope. A suitable test scope can be provided at a cost of $\frac{1}{10}$ to $\frac{1}{4}$ ft^3 of space and 10 to 20 lb. It should have a sweep of about 5 miles and any other sweeps deemed advisable for the particular system. It should also be provided with some device for accurate ranging (echo-box use).

Weight and Space Requirements.—Most test equipment can be made sufficiently light and compact to meet all reasonable demands. Feasible design objectives were enumerated above in certain specific cases. Some items currently available leave much to be desired in this regard, but test-equipment design has reached the point where accuracy and reliability of operation no longer need be sacrificed in order to comply with space and weight requirements. Also with built-in equipment, weight

[1] To facilitate tuning and testing it is advantageous to locate separate LO reflector voltage and i-f gain controls at the r-f head near the test oscilloscope.

often can be saved by the elimination of special packaging, plumbing, power supply, etc.

Frequency Coverage.—The frequency coverage of an item of built-in test equipment often can be narrower than is usual for that type of instrument when designed for general application. Some receivers are confined to rather narrow frequency bands, such as 100 Mc/sec, but most test sets now available are designed for about 1000-Mc/sec bands. If the test equipment is designed for the narrow band alone, certain design problems may be simplified. Calibration of variable attenuators, pads, and insertion losses is more accurate; test-set plumbing may be simplified; echo-box tuning can be made more uniform; basic echo-box design is simplified in that longer-ringing (higher-Q) cavities are possible; and in a few instances, the operation of narrow-band test equipment is simpler.

CHAPTER 12

I-F TEST EQUIPMENT

By E. Durand and Y. Beers

12·1. General Considerations.—A considerable part of any of the receivers under discussion here is made up of circuits operating at intermediate or video frequencies. Apparatus and methods for testing these circuits are quite different from that used in the r-f portion or for over-all tests. At the same time, the special requirements on transient response and noise figure have resulted in modifications and refinements of the methods and equipment as compared with those previously used in this frequency range. The characteristics of the i-f and video portion of the receiver that are of interest may be listed as follows:

1. *Over-all gain.* This is the ratio between the signal power developed across the output load impedance and the signal power at the amplifier input and is usually expressed in decibels.
2. *Bandwidth and bandpass shape.* The bandwidth B is usually taken to be the interval between the two frequencies at which the amplifier output power has fallen to half of its maximum value, the input power being held constant. The bandpass shape is essentially the curve of amplifier power (or voltage) output as a function of frequency for constant input power. In addition to changes in amplitude response, there are changes in the phase of a signal as it passes through the amplifier, and these also vary with signal frequency. The amplifier phase response is important in determining the response to transients (that is, to pulses); but owing to technical difficulties, it is seldom measured as such.
3. *Pulse response.* Although in principle the exact response of an amplifier to any transient can be predicted from a knowledge of its amplitude and phase responses, it is usually far easier to measure the transient response directly. In addition to effects on the shape of the output pulse, one should note whether there is any *black-out* or other aftereffect on the receiver performance, that is, whether or not the receiver will deliver full normal performance for one signal received immediately following another (strong) signal.
4. *Overload characteristics.* The wide dynamic range of signals to which radar receivers are subjected imposes strict requirements on

the behavior of all circuits at signal levels much in excess of the value required to produce full output from the receiver. Certain of the pertinent features have already been discussed in Secs. 6·11 and 9·3.

5. *Individual stage performance.* Although this information is not needed as an end result, it is often useful in obtaining a desired over-all characteristic.
6. *Dependence on tube selection.* In any lot of tubes of a given type there is necessarily some variation in transconductance and interelectrode capacitance. It is necessary to determine whether or not the shape of the amplifier pass band will be distorted too much and whether or not gain will be adequate under the worst conditions permitted by tube specifications.
7. *Stability.* The most important source of instability is regeneration, or feedback, which causes the passband shape to change with change in gain and in extreme cases may result in oscillation of the amplifier. Other stability factors are the effects of line voltages and of age.
8. *Noise figure.* This is the ultimate measure of the ability of an amplifier to detect weak signals.

To make any one of these measurements it is necessary to have (1) a device for generating signals of the appropriate type, which may or may not have all the refinements ordinarily associated with a conventional signal generator, (2) means for coupling into and out of the receiver under test, and (3) means for measuring or displaying the output of the receiver.

DEVICES FOR PRODUCING SIGNALS

12·2. Swept-frequency Signal Generators.—These are used primarily in amplifier alignment and bandpass determinations. In a typical setup, as shown in the block diagram of Fig. 12·1, the f-m generator frequency is periodically swept across a band of frequencies at some a-f rate. Either the sweep may be a sawtooth, with a rapid flyback to the starting frequency, or it may be smooth, sweeping first up in frequency and then back. It is important that the amplitude of the signal from the generator be constant over the swept band. A synchronized voltage sweep is simultaneously applied to the horizontal plates of a cathode-ray oscilloscope in such a way that a given horizontal position of the CRT spot corresponds uniquely to a single frequency, regardless of whether the sweep is from low to high or from high to low frequency. The relation between spot position and frequency should be as linear as possible. The output of the circuit under test, suitably rectified, is applied to the vertical plates of the oscilloscope. Response curves so traced for typical

amplifiers are shown in Fig. 12·2, and similar curves will be shown in other figures in the chapter.

The frequency modulation may be generated electronically, by means of a reactance tube or by varying the reflector voltage of a reflex velocity-

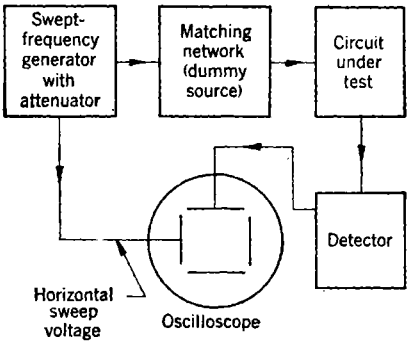


FIG. 12·1.—Block diagram of test setup using a swept-frequency signal generator.

modulation oscillator, or it may be generated mechanically by some device that alters either the inductance or the capacity of the oscillatory circuit. Reactance tube generators are commercially available but operate only in the lower frequencies and generally cover only a narrow band.

Very wide-band coverage together with band-center tunability may be achieved by using a beat-frequency generator. The output of an electronically or mechanically swept microwave oscillator is mixed with the output signal of another microwave oscillator separated from the first by a suitable frequency difference. The band center of the beat output signal may be shifted by retuning either oscillator. If more power is desired, a carefully designed wide-band amplifier operating at the beat frequency may be added.

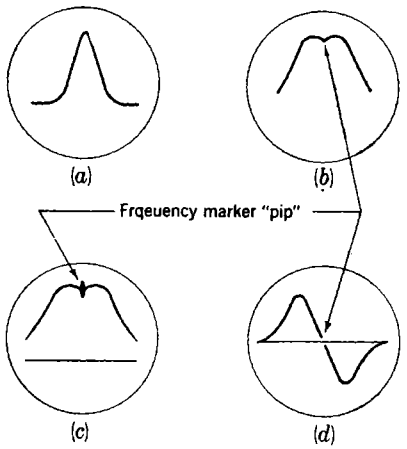


FIG. 12·2.—Typical response curves observed with a swept f-m signal generator.

The generator most widely used at the Radiation Laboratory depends on a motor-driven variable condenser, with a single oscillator operating in the desired frequency range. As originally built, constant amplitude was achieved through the use of a limiter stage following the oscillator.

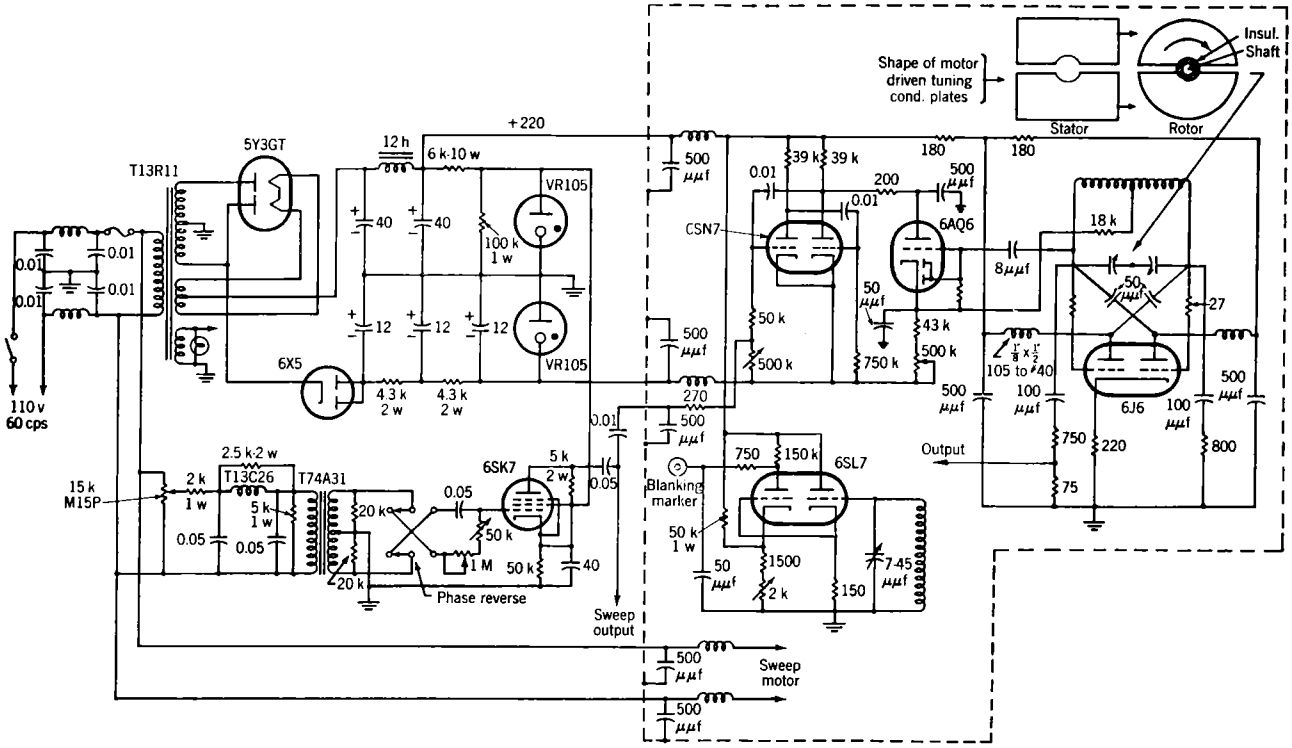


FIG. 12·3.—F-m signal generator.

Recent experiments have shown, however, that although the limiter ensures a constant peak-voltage output, the fact that strong harmonics are present in varying amounts over the band results in variation of the amplitude of the fundamental component of the signal, and the observed pass bands do not have the true shape. Constancy of output is usually checked by applying the generator output directly to a crystal rectifier and noting flatness of response on the trace on the oscilloscope. Because of the fact that with harmonics, the positive and negative peaks of the modulation envelope are not in general equal, when the crystal is reversed in its mount a different output signal is obtained, and the previously observed apparent flatness of output disappears.

In the latest Radiation Laboratory design, therefore, constancy of amplitude is maintained by means of an automatic gain control that adjusts the transconductance of the oscillator tube according to the output amplitude. This type of amplitude limitation results in operation over a fairly linear region of the tube characteristic, so that the output signal is relatively free of harmonics.

A complete schematic diagram is shown in Fig. 12·3. The oscillator is a twin triode type 6J6 in a conventional push-pull Hartley circuit. Frequency modulation is accomplished with a special condenser whose rotor plates are driven at 1800 rpm by a synchronous motor. Outlines of the plates and the electrical connections are shown in Fig. 12·3. Note that the two halves of the rotor are insulated from each other and from ground. The unit consists essentially of two condensers in series, and the capacity used is that existing between the stator sections. Thus, when the stator plates are aligned with the rotor plates, the capacity is a minimum; when they are at 90°, it is a maximum. A complete capacity cycle occurs twice per shaft revolution, so that the frequency is 60 cps. For each revolution of the shaft, the same value of capacity appears four times. It is important that the four angles involved be accurately complementary and supplementary so that the four corresponding spot positions may be fused. A high degree of mechanical symmetry is therefore needed.

For the gain control, a signal from one of the grids, from which the output is also taken, is applied to a diode rectifier. The resultant negative output is fed back through a cathode follower to determine the bias on the grids of the two sections of the oscillator. A single tube, a duplex-diode triode type 6AQ6, serves both functions.

When the circuit under test has appreciable response over the entire swept range, as is often the case when a single stage is being examined, it is hard to tell where the baseline, or position of zero response, lies (see

each rotation of the shaft. For this purpose, a multivibrator is used. The plate of cathode-follower gain-controlling tube is fed from one of the multivibrator plates so that when that section is conducting, the oscillator grids are carried beyond cutoff, effectively reducing the output power to zero. When the same section is nonconducting, however, the gain circuit functions normally. The free-running period of the multivibrator is about 30 cps, and the grid return resistors are unequal, so that one phase is about three times as long as the other. It is synchronized to the frequency sweep by a signal taken from the horizontal sweep.

Several marker devices may be used to tell the operator the relation between spot position and frequency. The simplest is a *passive* or *absorption* marker and consists merely of a tunable high-Q circuit loosely coupled to the oscillator. This absorbs some energy from the circuit at the resonant frequency and produces a small dip at the corresponding point on the trace. It suffers from two drawbacks. (1) It is hard to detect when the output amplitude is low, notably in the determination of the crossover frequency of a discriminator; and (2) if the limiter circuit is really effective, the dip will be wiped out. The effects of such a marker as well as the two markers mentioned below on the oscilloscope are illustrated in Fig. 12·2.

The second device is the *active* marker, which is simply a low-powered tunable oscillator that beats with the swept oscillator to produce an indication at the point where the two frequencies are nearly equal. This likewise is hard to detect when the output amplitude of the circuit under test is low.

The final device, used in the generator of Fig. 12·3, is the *blanking* marker. A small amount of energy is coupled from the oscillator into the high-Q tunable circuit connected to the grid of one half of a 6SL7. This tube, being connected as an infinite-impedance detector, offers little loading, so that a considerable positive voltage is built up at resonance. As the oscillator frequency is swept past this point, therefore, a short positive pulse is generated. This pulse is fed into an amplifier to produce a large negative pulse which may be applied to the intensity-control grid of the CRT to cause a momentary extinguishing of the beam. The sharpness of this blank spot may be increased by having the amplifier biased beyond cutoff so that only the tip of the pulse is effective. This type of marker will produce a clear indication even when the response of the circuit under test is zero and is accordingly definitely superior to the other markers.

The capacity variation with angle is such that the frequency variation is crowded at the high-frequency end of the sweep. This may be offset by making the horizontal deflecting sweep slow at one end and fast

at the other. To obtain such a sweep, a sinusoidal wave is applied to a pentode amplifier which is so biased that it is carried from a region of low g_m to one of high g_m by the signal on the grid, thereby distorting the input wave into the desired shape. A phase-shifting network with coarse and fine controls is provided so that the upsweep and downsweep in frequency can be made to coincide on the oscilloscope.

The power supply uses VR-tube regulation for the sweep-amplifier screen and for the negative supply for the gain-control circuit.

12·3. Video and I-f Pulse Generators.—Although many of the properties of amplifiers can be determined from measurements with c-w generators, their response to transients can best be measured with pulsed signals. In radar work the typical signal pulse is short, flat-topped, and rectangular, as illustrated in Fig. 12·4 Parts (*a*) and (*b*) are idealized video and i-f pulses respectively, and (*c*) and (*d*) show the pulse shapes that are realized in practice.

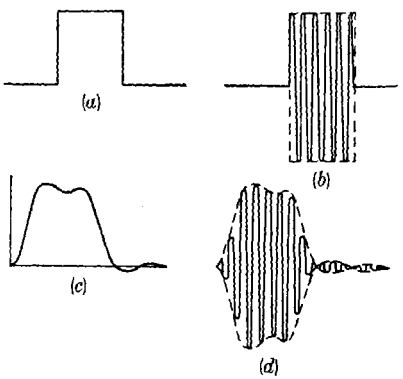


FIG. 12·4.—Idealized and actual video and i-f pulses. (*a*) Ideal video pulse; (*b*) ideal i-f pulse; (*c*) actual video pulse; (*d*) actual i-f pulse.

A good pulse generator should have as short a rise and fall time as possible, and the regions at the top and after the end of the pulse should be flat and free of "wiggles." There should be no frequency modulation during the pulse.

Secondary properties of good pulsers are (1) constancy of the pulse output amplitude and shape with change of line-voltage pulse-repetition frequency (PRF) or output loading, (2) good shielding, and (3) wide range of available pulse lengths.

If the pulses are initiated by a trigger from an external source, they should be free of *jitter;* that is, the delay between the start of the initiating pulse and the start of the output pulse should be constant. In *self-synchronized* pulse generators, there is normally a trigger source for initiating operations in external equipment; these triggers should start quickly and rise rapidly and likewise be free from jitter with respect to the output pulse.

The pulse generator to be described was designed and built at the Radiation Laboratory. It contains not only a pulsed i-f oscillator but also two video pulse generators which actuate the i-f section. Figure 12·5 is a block diagram of the complete generator, and Fig. 12·6, a complete circuit diagram. It is divided into four sections: (1) *power supply,*

(2) *short-video-pulse generator*, (3) *long-video-pulse generator*, and (4) *i-f-pulse generator*.

Power Supply.—The supply provides +400, +195, and −150 volts.

Short-video-pulse Generator.—Operation of this circuit is as follows. A positive trigger applied to the trigger shaper, a gas-tetrode V_1, causes it

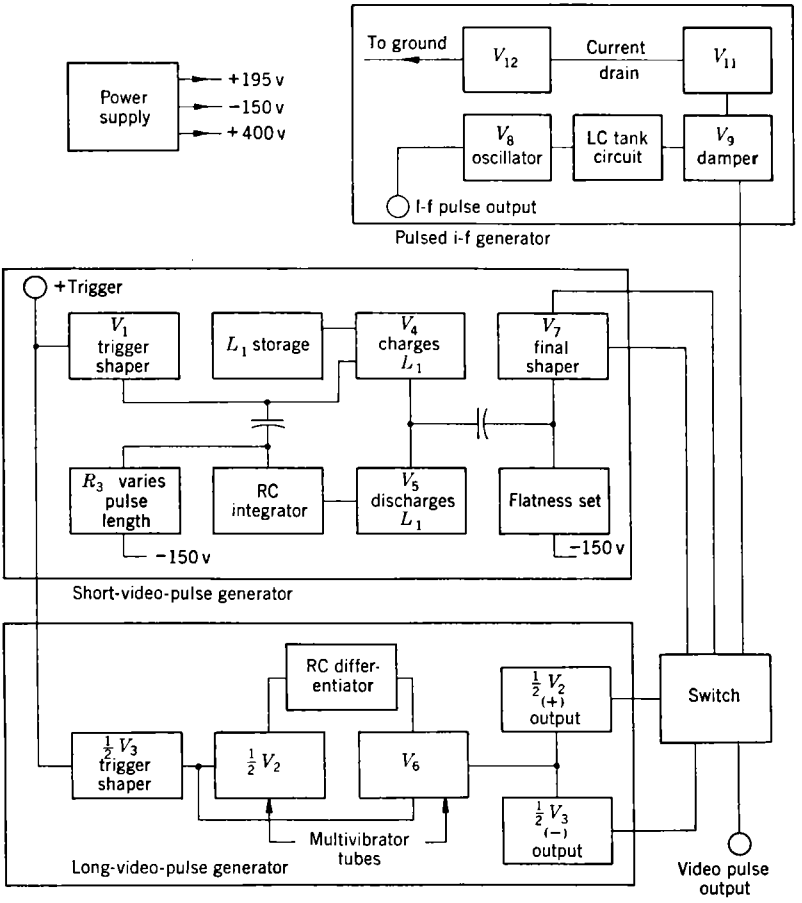


FIG. 12·5.—Block diagram of pulse generator.

to break down or fire. Because C_1 prevents any sudden changes in plate potential, the breakdown results in the appearance of a sharp positive trigger at the cathode. This trigger is coupled in a way that will be considered later to the grids of the two gas-tetrodes V_4 and V_5, which are the actual pulse generators. Before the appearance of the initiating

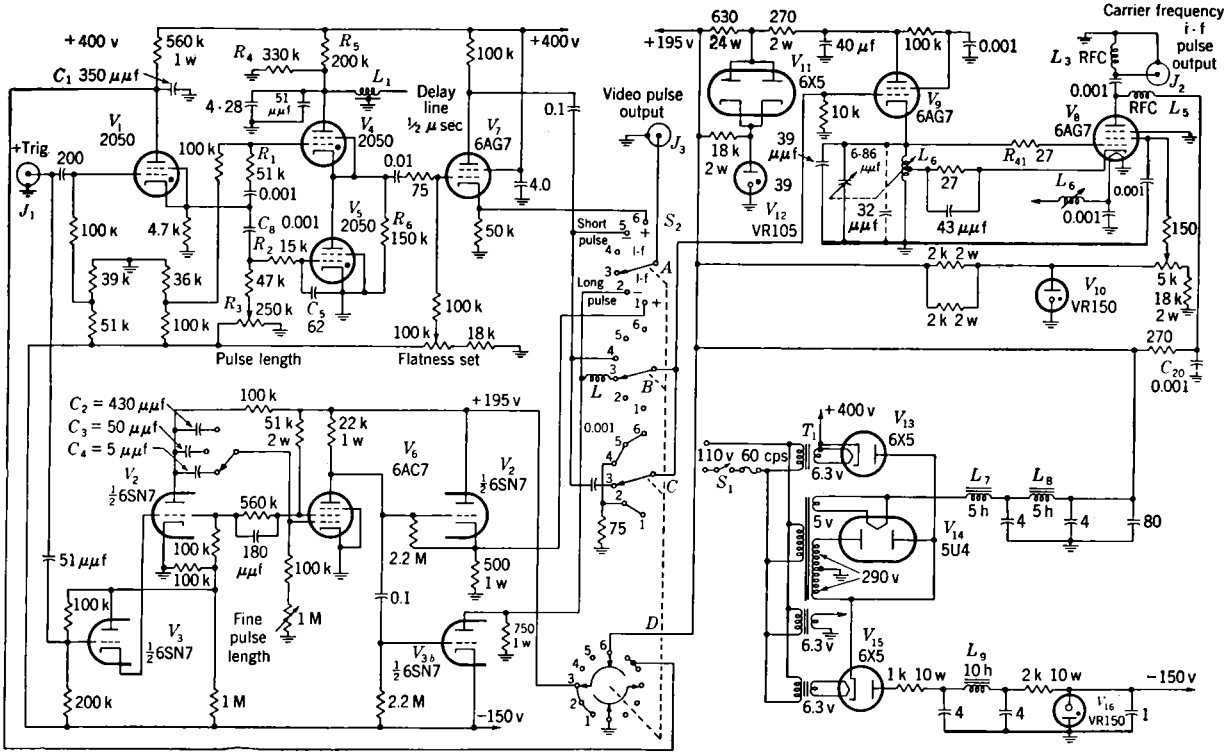


FIG. 12·6.—Circuit of pulse generator.

trigger, these tubes are nonconducting, because of the bias voltages on their grids. The cathode of V_4 and the plate of V_5 are held at ground potential by R_6, and R_4 and R_5 hold the plate of V_4 at about + 250 volts. When V_4 fires, its cathode rises abruptly, causing the start of the actual pulse and simultaneously providing plate voltage for V_5.

Simultaneously, the pulse from the cathode of V_1 is applied through R_2 to the control grid of V_5. The rise of voltage on this grid is exponential, because of R_2 and the shunt capacity C_5 and grid-to-ground capacity of V_5. The amount of bias on this grid is set by the *pulse-length adjustment* R_3, so that the time at which the grid voltage reaches the firing point may be varied. There is thus an adjustable delay between the firing of V_1 and that of V_5.

The firing of V_5 marks the end of the pulse because it results in the potential at the plate-cathode junction of V_5 and V_4 being brought down abruptly nearly to ground potential. An approximately rectangular positive pulse is therefore generated at this junction.

Note that R_1, in conjunction with the total stray capacity between the control grid of V_4 and ground, results in a small but finite delay between the firing of V_1 and that of V_4. This delay is enough so that when the delay in firing of V_5 is set at the minimum value, V_5 will fire at least as soon as V_4. Therefore, the minimum length of pulse that can be generated is determined only by the speed with which the plate voltage on V_5 can rise and the breakdown time of V_5.

The delay line L_1 provides a flat top for the pulse. Earlier pulse generators used a large capacity between this point and ground. The flatness of top was limited by the ability of this condenser to hold its charge. If the capacitance were made large to increase this holding ability, it would become difficult to replace the charge during the interval between pulses. The output was therefore a function of the PRF. The delay line has a small total capacitance to ground and can, therefore, be charged to full potential quickly. On the other hand, it will sustain its voltage accurately for the time required for a signal to travel to the end of the line and back. In this case, the line "length" is either $\frac{1}{2}$ or 1 μsec, and therefore pulses up to 1 or 2 μsec respectively can be generated. The flatness of the pulse top is determined mostly by the quality of the line, although the limiting action of the following amplifier V_7 contributes appreciably.

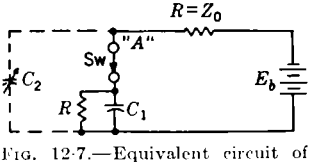


FIG. 12·7.—Equivalent circuit of delay line after firing of V_4.

A delay line may be represented by a battery, whose voltage is the voltage to which the line is charged, in series with a resistance equal to the characteristic impedance of the line, for a period equal to twice the

length of the line after the firing of V_4. Figure 12·7 shows the equivalent circuit for the short interval following the breakdown of V_4. The stray capacity existing between the cathode of V_4 and ground, shown as C_1, may be as much as 50 $\mu\mu$f, since it includes interelectrode capacities of V_4, V_5, and V_7. The voltage at the cathode would therefore rise along the curve shown in Fig. 12·8 with a time constant of about 5×10^{-8} sec. When it reached a value determined by the bias on the grid of V_7, its rise would cease as a result of the diode action of this grid. If, however, a capacity C_2, shown in dotted lines in Fig. 12·7, is added between the plate of V_4 and ground, when the "switch" Sw (V_4) is closed, the junction point A will quickly reach a voltage determined by the ratio of the capacity divider formed by C_1 and C_2. By proper adjustment of C_2, this voltage may be made equal to the desired final pulse-top voltage. In this case, the rise time is limited only by the stray inductance in the leads associated with the condensers and may easily be made less than 2×10^{-9} sec. The rise obtained with such compensation is shown in Fig. 12·8*b*.

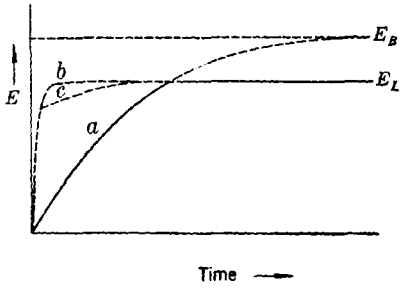


FIG. 12·8.—Shape of voltage rise at the V_4–V_5 junction. (*a*) Without compensating condensers C_2 and C_3; (*b*) compensating condensers set correctly; (*c*) compensating condensers too small. E_L is the voltage at which grid of V_7 starts to draw current.

The fall time of the pulse is quite short, since it is determined by the low effective internal resistance of V_5 at the instant of breakdown or by residual lead inductances.

If the pulse amplitude at the grid of V_7 is so large that a heavy current is drawn by the grid, the drain on the delay line will result in a falling off of its voltage. Therefore, a bias adjustment is installed as a "flatness set."

The output tube V_7 is biased far beyond cutoff during the interpulse interval and therefore provides a sharp termination for the pulse, free from spurious oscillations. Positive and negative output pulses may be taken from its cathode and plate respectively.

Because of the extremely rapid rise and fall times required to give the i-f pulse a good start (see below), the greatest care must be taken in the layout of the pulse tubes and amplifier. The best uhf techniques should be employed, since the spectrum of the pulse contains energy in the regions up to several hundred megacycles per second.

Long-video-pulse Generator.—The heart of the long-pulse generator is a monostable multivibrator, made up of the triode V_2 and the pentode V_6.

As far as the multivibrator action is concerned, V_6 acts as a triode, the screen grid serving as anode. In the normal condition V_2 is fully cut off by the bias, supplied by a resistance network, and V_6 conducts fully. A positive pulse applied to the grid of V_2 through a diode (V_3, cathode and grid) causes it to conduct, and the resulting downward motion of its plate potential, transmitted through C_2, C_3, or C_4, cuts off the control grid of V_6. Therefore the voltage at the plate of V_6 rises; this is the leading edge of the pulse.

Presently, at a time determined by the resistors in the grid circuit of V_6, the negative voltage disappears, and V_6 starts to conduct. The resultant drop in potential of the screen grid of V_6 cuts off V_2 once more, restoring the original condition and terminating the pulse.

The diode V_3 is necessary to prevent the fall of voltage at the end of the initiating trigger from restoring the multivibrator to its original condition. The connections shown at the plate of V_3 have no significance; it was used as a convenient tie point.

Positive and negative output pulses are obtained from the cathode follower V_2 and the amplifier V_{3b} respectively.

I-f Pulse Generator.—The oscillator proper V_8 is an electron-coupled circuit but is normally inhibited from oscillation by the heavy damping of the cathode follower V_9 connected across the tuned circuit. The i-f pulse is initiated by applying a large negative video pulse from the short video-pulse generator to the control grid of the cathode follower. This removes the damping and allows the oscillator to start working. Because of the extremely rapid rise of the video pulse, which occurs in a fraction of an i-f cycle, the standing current of the cathode follower in the inductance of the oscillatory circuit starts the oscillations.

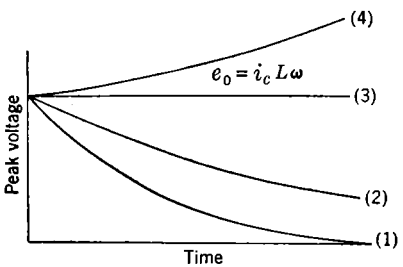


FIG. 12·9.—Dependence of amplitude of oscillatory voltage on time.

Suppose for the moment that the oscillator tube were removed. After the video pulse cuts off the current in the cathode follower, there will exist a damped oscillation in the LC-circuit. The initial current amplitude is given by the standing current in the cathode follower, and the exponential decay is governed by the Q of the circuit. This is shown in Curve 1 of Fig. 12·9. If the tube is replaced, with weak feedback, the falling off of amplitude is delayed, the negative resistance supplied by the tube offsetting part of the circuit losses (Curve 2). Increasing feedback prolongs the delay until a value is reached at which the ampli-

tude remains constant, circuit losses being just offset by the tube (Curve 3). More feedback results in a rising amplitude of oscillation, the rise continuing until limited by nonlinear conditions in the tube, such as grid current or change of g_m (Curve 4).

If the condition producing Curve 3 is met, the i-f pulse will rise to its full value in a fraction of a cycle and will retain constant amplitude for the duration of the video pulse. When the damping is restored, the oscillation will disappear in a very short time because of the heavy loading of the cathode follower.

The energy supplied to the LC-circuit by the tube is adjusted by varying the screen-grid voltage. In order to make the voltage across the tuned circuit independent of frequency, a constant L/C ratio must be maintained. Therefore, both inductance and capacity must be suitably ganged and tuned together. To maintain a flat-topped pulse over the band, compensation must be made for changes in circuit Q. A small bypass condenser across the cathode resistor increases the amount of regenerative feedback at high frequencies. Note that the oscillator always operates over a linear portion of the tube characteristic, so that the output is exceptionally free of harmonics.

To ensure constant output, the supply voltage for the cathode follower is regulated at +105 volts by the VR tube V_{12}. The demand of the cathode follower under low duty-ratio conditions (short pulses, low PRF) is, however, about 100 ma, whereas at the maximum duty ratio, 90 per cent, it is small. Under the heavy-drain conditions, the VR tube might extinguish, as current intended for it was diverted to the cathode follower. A diode V_{11} is therefore provided to allow the plate of the cathode follower to fall below +105 volts without extinguishing the VR tube.

To provide the fast starting discussed above, the short video pulse is always used to initiate the action. The rise time of the long video pulse is probably no better than 10^{-7} sec, a time too long compared with a quarter cycle at the intermediate frequency to give a clean start. By the time the short pulse is over, however, the long pulse will have come in to hold off the cathode follower. It is not desirable that the long pulse should rise too fast, since, owing to the grid-to-cathode capacity of V_9, a "wiggle" would be imparted to the i-f oscillation. In addition, some loading would be placed on the tuned circuit by way of this same grid-to-cathode capacity, causing some frequency modulation and lowering the Q. Both of these effects are overcome by a small r-f choke in the line from the long-pulse generator.

The pulse selector switch S_2 offers a choice of positive and negative video pulse and i-f pulse outputs for both long and short pulses. In addition to switching the output jacks to the appropriate points in the

circuit, the switch interrupts the plate supply for the long-pulse generator in the short-pulse positions and also short-circuits the plate of the trigger shaper V_1 of the short-pulse generator when in the two positions for long video pulses. As was explained above, the short-pulse generator is left in operation to give a quick start when long i-f pulses are required. A second switch S_1 offers three ranges of pulse lengths in the long-pulse position.

The pulse generator gives satisfactory operation for duty ratios up to 90 per cent and PRF up to 2000 cps, the output being independent of the PRF up to this value. Operation at 4000 cps results in a slight sag in the short-pulse output, which may be eliminated by a readjustment of the flatness set control.

12·4. Noise Generators: General Considerations.—There are a number of devices that have been used as noise generators for amplifier measurements. The outstanding feature of all of them is that their available power is proportional to a d-c current. In the case of the temperature-limited diode, the constant of proportionality between the available noise power and the current can be easily calculated, and such an instrument can be used as an absolute instrument up to about 300 Mc/sec. Above this frequency the measurement of impedance (which is required for determining the proportionality constant) becomes difficult, transit-time effects change the noise output, and the diode noise generator can no longer be used as an absolute instrument but is still useful for relative measurements.

Silicon crystals (of the type used for rectifiers and mixers) generate considerable noise when d-c currents are passed through them in the direction of difficult conduction. A practical crystal noise generator will be described in Sec. 12·7. Such a device is very compact and convenient for field measurements, but it is not an absolute device and must be calibrated.

12·5. Theory of Noise Generators Using Temperature-limited Diodes. *High-impedance Noise Generators.*—A temperature-limited diode acts like a constant-current generator of noise because of the shot effect—the fluctuation in the number of electrons emitted by the cathode. The mean square current strength is given by

$$\overline{i_n^2} = 2eIB,$$

where e = charge of the electron
$= 1.60 \times 10^{-19}$ coulomb,
I = current through the diode, amperes,
B = noise bandwidth of device being used to observe the noise.

With oxide-coated or thoriated cathodes, there may be additional noise resulting from the so-called *flicker effect.*

A simple noise generator consists of such a diode connected in parallel (for signal frequencies) with an impedance of the value desired for the signal-source impedance. This device is connected to the input of the amplifier, and the value of I is adjusted by varying filament current of the diode to cause the noise output of the amplifier to double. The available power of the generator W_n is

$$W_n = \tfrac{1}{2}eIR_aB,$$

where R_a is the resistive component of the signal source impedance. The noise figure is

$$F = \frac{e}{2kT} IR_a \qquad (1)$$
$$= 20IR_a$$

if $T = 290°K$.

The absolute value of the noise figure is thus given in terms of two easily measurable quantities, I and R_a.

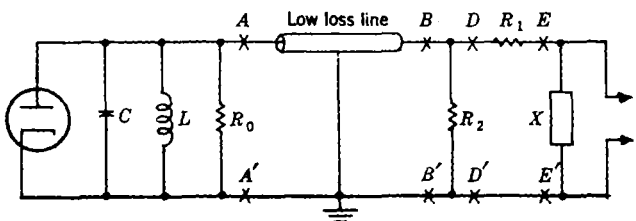


FIG. 12·10.—Matched-line diode noise generator.

This device is very useful provided that (1) the noise generator can be connected in its place with short leads and (2) that the parallel reactance of the noise generator can be made entirely equivalent to that desired for the signal source over the entire range of frequencies of interest. There is always some capacity across the diode. In the cases where the desired shunt reactance corresponds to a capacity greater than the diode capacity, the desired capacity can be attained by adding a padding condenser. In other cases, which occur fairly often in practice, the desired reactance is that of a smaller capacity or an inductance. These cases are somewhat embarrassing, for there is no completely satisfactory way of producing the desired reactance. The best it is possible to do is to resonate out the undesired capacity by means of an inductance. This procedure is rigorously valid at only one frequency.

These difficulties can be largely avoided by the use of a *low-impedance* or *matched-line* noise generator.

The Matched-line Noise Generator.—This apparatus (see Fig. 12·10) consists of a noise diode feeding a lossless line of characteristic impedance

R_0 with a termination at the generator end and a matching network consisting of R_1, R_2, and X so selected that the internal impedance of the entire device equals that of the signal source.

The capacity C is the distributed capacity of the diode, and L resonates with this at the band-center frequency. More exactly, L and R_0 are so chosen that one measures the same impedance at AA' as at BB' (the other components to the right being disconnected in both cases). As R_0 is relatively low, the bandwidth of the circuit C-L-R_0 is large compared with the usual bandwidths of the amplifiers, and therefore the capacity has much smaller effect than in the high-impedance noise generator. (In a typical case where C is 10 $\mu\mu$f and R_0 is 75 ohms, the bandwidth is about 200 Mc/sec.)

It may be shown by a straightforward calculation that the noise figure is given by

$$F = \frac{e}{2kT} \frac{R_0^2 I}{R_a \left(1 + \frac{R_0}{R_2}\right)^2}$$

$$= 20 \frac{R_0^2 I}{R_a \left(1 + \frac{R_0}{R_2}\right)^2}$$

where R_a is the parallel output resistance of the generator, which is made equal to the parallel output resistance of the actual signal source (equal to $1/G_a$ the reciprocal of the conductance of the signal source). The resistances R_1 and R_2 are chosen to give the desired value of R_a by the use of the following relation:

$$R_a = R_1 + \frac{R_0}{1 + \frac{R_0}{R_2}}.$$

Two special values of R_2 are of interest. If it is infinity (that is, not present),

$$F = \frac{e}{2kT} \frac{R_0^2 I}{R_a} \tag{2}$$

$$= \frac{20 R_0^2 I}{R_a}$$

if $T = 290°\text{K}$.

By comparison with Eq. (1), it can be seen that for a given noise figure, the ratio of the d-c current required with the matched-line generator to that required with the high-impedance noise generator is $(R_a/R_0)^2$. To cite a typical example, if R_a is 300 ohms and R_0 is 75 ohms, sixteen times as much current is required with the matched-line generator as with the high-impedance noise generator.

If R_2 is made equal to R_0,

$$F = \frac{e}{2kT}\frac{R_0^2 I}{4R_a} \tag{3}$$

$$= 5R_0^2\frac{I}{R_a}$$

(if $T = 290°K$) and

$$R_a = R_1 + \tfrac{1}{2}R_0.$$

It can be seen by comparing Eqs. (2) and (3) that four times as much current is required to measure a given noise figure if both ends of the line are terminated in the characteristic impedance as if only the generator end is terminated.

12·6. Construction of Diode Noise Generators.—Diodes used as noise generators should have the following characteristics: pure tungsten or thoriated tungsten filament, good saturation (the reciprocal of the slope of the plate-current vs. plate-voltage curve should be at least fifty times the desired generator resistance), short leads, and low capacity. Also, for matched-line use they should have a moderately large emission and plate dissipation. Of all the tube types tested here the British type CV172 seems to be the most generally satisfactory. Although not designed for such use, the Western Electric 708A, the Eimac 15E (the 15R does not saturate very well), and the 801A seem to be more or less satisfactory for matched-line generators. The Sylvania X6030 and the O1A are satisfactory for high-impedance generators. RCA has built some experimental noise diodes in miniature bulbs (R6192), which look promising as far as r-f characteristics are concerned. At 30 Mc/sec, comparisons among many of these types have yielded agreements within 0.05 db. Oxide-coated cathode tubes have proved unsatisfactory where highest accuracy is desired, and indirectly heated cathode tubes are doubly undesirable because of the difficulty with which they are controlled.

A noise generator should be completely shielded and its power leads should have decoupling to prevent picking up of extraneous signals in general and specifically to prevent stray coupling to the output of the receiver, which would cause regeneration. As the diode is operated under temperature-limited conditions, the plate-voltage supply need not have particularly good regulation or filtering. Under these conditions, the plate current varies rapidly as the filament voltage is changed. It is therefore desirable to provide a relatively constant supply voltage such as a storage battery or, if operation from a-c lines is desired, a constant-voltage regulator. Because of the rapid variation of emission with filament voltage, two rheostats for the control of filament voltage should be included in the apparatus, one for rough control and one for fine control.

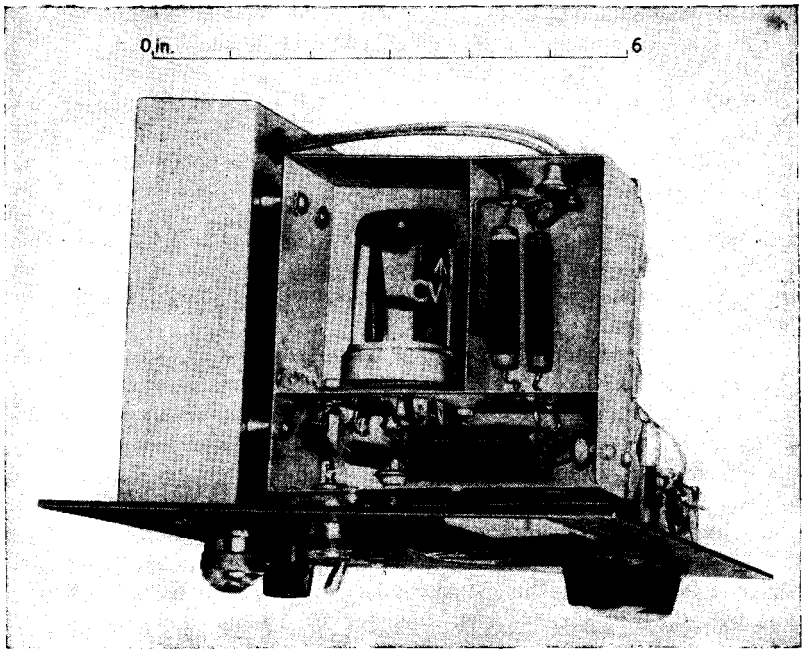


FIG. 12·11.—Interior view of noise generator employing CV172 tube.

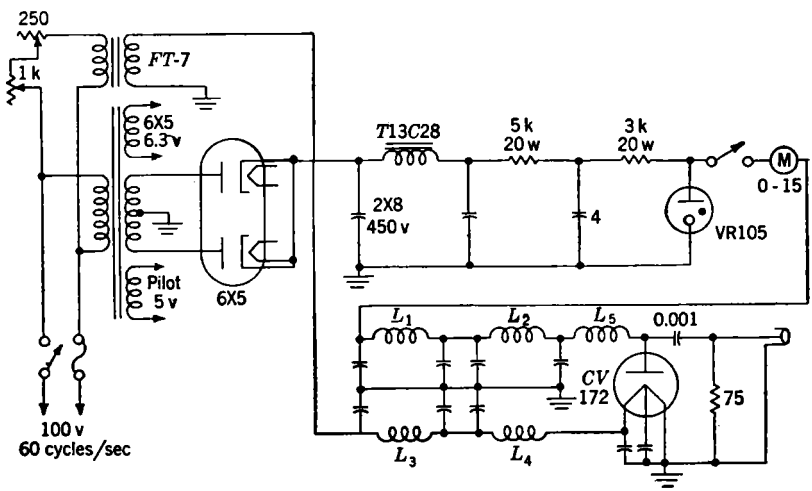


FIG. 12·12.—Noise generator. Coils L_1, L_2, L_3, L_4 are wound with 26 enamel wire on 2W IRC 0.25M; self-resonant at 30 Mc/sec. Coil L_5 is wound with 26 enamel wire on Speer form, tunes 30 Mc/sec with capacitance from plate to ground. Condensers are 500 $\mu\mu$f Erie buttons.

It is important to have firm electrical connections in the filament circuit. The number of connectors and toggle switches should be kept to a minimum; and if alternating current is employed, these and the control rheostats should be placed in the primary of the transformer where slight changes of resistance will have less effect. Sometimes it is necessary to clean the filament pins of the tube and tube socket.

A noise generator[1] employing a CV172 tube is shown in Fig. 12·11, and the complete circuit diagram is shown in Fig. 12·12. It was designed primarily for use as a matched-line generator employing a section of 75-ohm cable (not shown in the diagrams), and therefore a 75-ohm resistor is shown in parallel with the output. This same design could, however, be used as a simple high-impedance generator by raising the value of this resistor and adjusting the reactance in parallel to a suitable value. The inclusion of the large amount of filtering and the VR-105 regulator tube in the power supply is perhaps not entirely necessary.

12·7. Crystal Noise Generators.—Silicon crystals produce considerable noise when a current is passed through them in the direction of lower conductivity.[2] Noise generators of this type have been built that can be used at frequencies up to at least 3000 Mc/sec. The circuit diagram of one that was designed for use at 30 Mc/sec is shown in Fig. 12·13. As it was desired to use a meter with a sensitivity of 1 ma full scale, it was necessary to provide a Π-attenuator consisting of R_2, R_3, and R_4 to reduce the noise output to a convenient value. The values of these resistances are so selected that the parallel resistance at the output terminals is approximately 300 ohms, equal to that of the actual signal source used with the amplifier under test, and C_2 is adjusted to equal the capacity of the signal source. The current is adjusted by R_1 to cause the noise output of the amplifier to double, and the noise figure can be determined from this value of the current and the calibration of the instrument, which must be made experimentally by comparison with a signal generator of another type. Such instruments are very compact and simple in construction but must be calibrated frequently if high accuracy is desired.

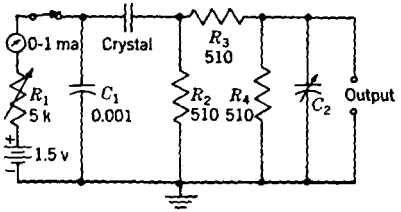


FIG. 12·13.—Circuit of crystal noise generator.

[1] This unit was designed and built by C. Butt and H. W. Brown at the Radiation Laboratory.

[2] J. E. Houldin, "The Crystal Capsule as a Generator of Noise." General Electric, Report No. 8237 (Great Britain), July 9, 1943.

DEVICES FOR COUPLING INTO AND OUT OF THE RECEIVER

Many signal sources are designed to operate into a coaxial line whose characteristic impedance may be 35, 50, 75, or 100 ohms. It is often convenient to have a separate well-shielded attenuator with characteristic impedance equal to that of the line. In addition a *dummy input circuit* is ordinarily required to simulate the normal signal source. In the case of the microwave radar receivers under discussion here, the dummy input circuit simulates the crystal mixer and permits the testing of the i-f amplifier alone. Finally, a so-called *traveling detector* is often used to give easy access to the output of successive stages of the i-f amplifier.

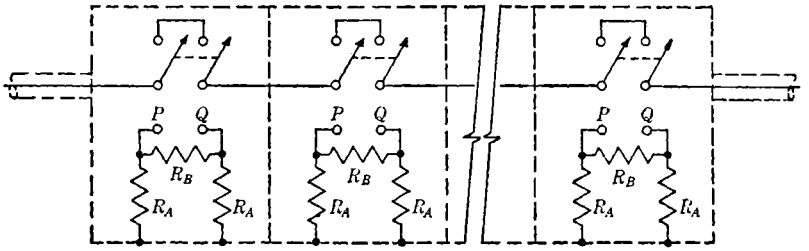


FIG. 12·14.—75-ohm i-f attenuator.

12·8. Attenuators.—A typical circuit is shown in Fig. 12·14. Table 12·1 shows the values that should be used for various sizes of steps for an iterative impedance of 75 ohms. The accuracy of any section may be determined with a bridge by measuring the resistance from P to Q and from P and Q to ground with the switch in the "out" position. The values that should be obtained are tabulated in Table 12·1, together with the values of the individual resistors. For any other iterative impedance, multiply these values by the ratio of the new impedance to 75 ohms.

TABLE 12·1.—RESISTANCE VALUES FOR 75-OHM ATTENUATOR

Attenuation, db	R_A	R_B	Test data	
			Resistance P to Q	Resistance P or Q to ground
1	1305.	8.7	8.7	657.
2	654.	17.5	17.2	330.
3	438.	26.6	25.7	227.
5	267.	45.6	42.1	144.5
10	144.	106.7	78.0	91.9
20	91.7	361.	123.2	76.8

This type of attenuator is reliable for direct current and for any frequency up to some limit determined by stray lead inductance and capacity. By using small composition resistors carefully mounted, satisfactory operation and accuracy of calibration and of match may be obtained up to 60 Mc/sec or higher. One arrangement that works well in this region is shown in Fig. 12·15. The over-all attenuation of this unit is over 100 db. Careful shielding is required if the leakage power due to stray coupling is to be kept well below the desired power. Note that in the arrangement shown, all leads are short, and each section is fully shielded from the adjoining sections.

Wire-wound resistors cannot be used except at very low frequencies because of their inductance. Semiprecision composition resistors are bulky and would have excessive capacity to ground. Small half-watt

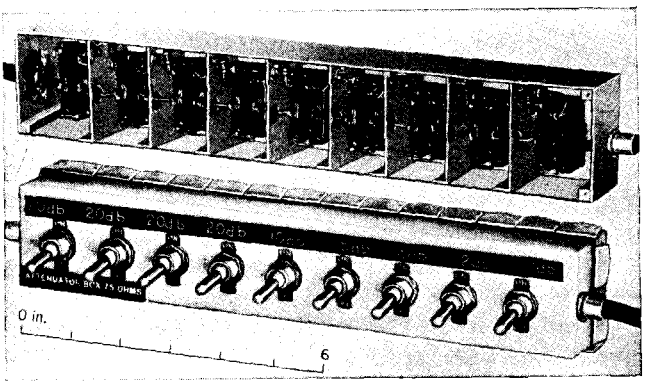


FIG. 12·15.—Photograph of attenuator.

composition resistors, such as the Allen-Bradley type EB, are therefore used. If greater accuracy is desired, resistors may be selected from a larger lot, or resistors measuring less than the desired value may be trimmed to the correct value by filing a notch through the insulating case into the composition until the correct resistance value is obtained, after which the notch is sealed with lacquer. In general, because of aging and temperature effects, it is not worth while trying to achieve an accuracy greater than 2 or 3 per cent.

In making bandwidth or noise measurements, it is often necessary to be able to change the strength of the input signal by exactly 3 db. For this purpose, a single-element attenuator of the type discussed above may be used. It is often mounted in a small metal can provided with input and output cable fittings and is installed in the line from the signal generator to the amplifier under test. Some of the modern signal generators have such attenuators built in.

12·9. Dummy Input Circuits.—The output lead from most signal generators either is a coaxial cable or is intended to work into a coaxial cable of characteristic impedance R_0. It is necessary to provide the proper termination for this cable to ensure that the correct value of attenuation is obtained with the attenuator. The performance of an amplifier is likewise affected by the impedance of the signal source. The impedance of the test source should therefore be the same as the impedance Z_i of the source used in actual operation.

A dummy input circuit, therefore, is used to couple the generator to the amplifier under test. As seen from the generator, the circuit impedance is R_0; as seen from the amplifier, it is $Z_i(= R_i + jX_i)$.

In microwave radar sets the amplifier signal source is a crystal mixer whose i-f impedance is made up of a resistance in parallel with a capacitance. The resistance ranges from 200 to 500 ohms, depending on the

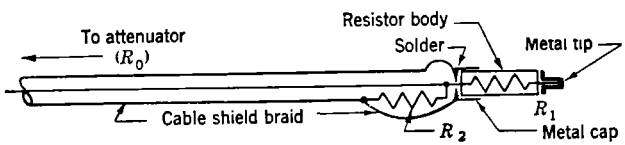


FIG. 12·16.—Diagram of approximate physical layout of dummy crystal.

wavelength and the particular crystal used. The capacitance is from 5 to 15 $\mu\mu$f for the mixer proper; if a cable is used to attach the mixer to the amplifier, as much as 20 $\mu\mu$f has to be added.

Sometimes the mixer is left in place, and the signal is introduced through a *dummy crystal.* Figure 12·16 shows the usual arrangement. The impedance seen looking into the input terminals of the amplifier is usually relatively high. Therefore, the impedance seen from the cable looking into the dummy crystal is only slightly less than the value of the shunt resistance R_2. In practice, R_2 is set equal to R_0, providing the correct termination for the cable and hence for the attenuator. Looking back into the dummy crystal from the amplifier, one sees R_1 in series with $R_0/2$, since the cable itself, being terminated by R_0 at the generator or attenuator end, looks like R_0. R_1 is therefore chosen so that

$$R_1 + R_0/2 = R_i.$$

When a dummy crystal is used, no provision for capacitance need be made, as the mixer itself supplies the right amount. At times, however, it is preferable to remove the mixer and replace it with a complete *dummy mixer.* The circuit used is the same as before, but a variable condenser C_i is added from the output end of R_1 to ground, to simulate the mixer capacity. The dummy mixer is generally mounted in a small metal can. If the amplifier to be tested includes a cable for the mixer, a suitable cable

fitting is mounted directly on the can. If the amplifier is provided with a jack, a short cable may be included as part of the dummy mixer. In either case, the condenser is adjusted to present the same capacity that would be seen at the point of attachment with the regular mixer in place.

Dummy mixers have been built in which various values of R_1 could be selected by means of a switch. In addition to providing values ranging from 200 to 500 ohms, to facilitate the study of the effect of crystal impedance on performance, 0 ohms and 5000 ohms are provided. The former serves to damp out the selective effect of the input circuit in order to study effects in later circuits; the latter provides essentially a constant-current source useful for special applications.

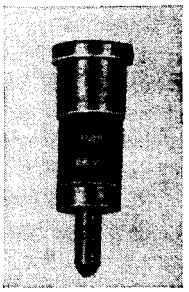


FIG. 12·17.—Dummy crystal.

Some attempts have been made to develop signal generators with push-pull output for testing the balanced input circuits used with balanced mixers. It is not difficult to obtain equal output voltages at the generator, but it is very difficult to provide matched pairs of attenuators that will give equal output voltages at low level. It has been found, fortunately, that all necessary information may be obtained from tests using a single-ended output signal introduced by means of a dummy crystal first into one and then into the other crystal holder, the unused holder being supplied with a second dummy crystal containing a resistor of value R_i, one end of which is grounded. The output that would be obtained with push-pull input is the average of the two outputs obtained in the above manner, except that

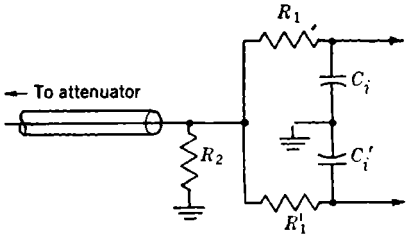


FIG. 12·18.—Double dummy mixer for unbalance tests on balanced-input receivers.

a 6-db allowance must be made, since only one of the input connections is energized at a time. Figure 12·17 shows a dummy crystal suitable for this purpose.

This method is not suitable for the measurement of the degree of unbalance of the circuit, since that involves the determination of the small difference between the two results. However, unbalance may be readily measured by feeding equal signals into both inputs in phase. A

special dummy-crystal pair may be constructed for this purpose, using the connections shown in Fig. 12·18. With such a device, the output of a perfectly symmetrical circuit would be zero, so that the amount of output is a measure of the unbalance.

12·10. Traveling Detectors.—The principal occasion for the use of a traveling detector comes in the alignment of an amplifier where it is desired to measure or observe the pass band of a single interstage circuit. A signal from a swept-frequency oscillator is introduced ahead of the circuit to be tested, in such a way that none of the previous circuits contribute to the selectivity. If the circuit under test is connected to the plate of a given tube, the input signal may be introduced across a 75-ohm resistor at the grid of that tube. The output signal may then be taken from another 75-ohm resistor connected to the plate of the following stage.

Either vacuum tubes or crystal rectifiers are used, but in either case the rectifier input is shunted with a low resistance that provides correct termination for a coaxial cable. The rectifier may then be located at any point far from the amplifier. It is usually located close to the output indicator, which is generally an oscilloscope with a high-gain audio amplifier.

The low impedance of the probe so broadens the circuit to which it is connected that it has a negligible effect on the observed response curve.

This method was widely used for a long time. It suffers from one difficulty, however, that comes from the Miller effect[1] which is present at the input circuit of the tube following the circuit under test. On the average, for 6AK5 tubes, there exists a total capacity C_{pg} from the plate of a tube to its own grid of about 0.07 $\mu\mu$f. Because of the Miller effect, there is an apparent added capacity C_{gk} of $\mathcal{G}C_{pg}$, where $\mathcal{G}$ is the stage amplification. In the functioning amplifier, $\mathcal{G}$ may be from 5 to 10, making C_{gk} from 0.35 to 0.5 $\mu\mu$f. But with the probe in place, the stage gain is less than unity, causing a proportional reduction in C_{gk} and a consequent increase in the resonant frequency of the coupling circuit at the input. The detuning effect may amount to a megacycle or more.

If only the single-stage Miller effect existed, it would be possible to allow for it by aligning each stage above the nominal center frequency by an amount that could be determined experimentally. There is also, however, a "chain Miller effect" in which the change of gain of one stage causes a detuning of the stage two positions down in the line. The result of this is that accurate alignment with the low-impedance probe is difficult or impossible. Such a probe is, however, useful for approximate measurements. For situations requiring accurate alignment the following

[1] F. E. Terman, *Radio Engineers' Handbook*, McGraw-Hill, New York, 1943, p. 468; J. M. Miller, "Dependence of the Input Impedance of a Three-electrode Vacuum Tube upon the Load in the Plate Circuit," *Bur. Standards Sci. Paper* 351.

method has been found useful. Although the ideal amplifier tube operates only over a linear portion of its characteristic, practical tubes are such that with moderately large signal, the operation extends into nonlinear regions; that is, the amplifier acts also as a detector. When a strong signal is applied to the grid of such a tube, the cathode current increases slightly. The value of this current is therefore a measure of the signal amplitude. Furthermore, within limits, the response so obtained will be proportional to the voltage of the input signal. It is therefore suitable for the observation of the pass band.

To make the observation, a lead is connected directly from the cathode of the amplifier to the oscilloscope amplifier input terminal. The output voltage is developed across the impedance consisting of the cathode resistor and bypass condenser in parallel which at these low frequencies is resistive. This method causes no detuning, either from probe capacity or from changing Miller capacity. The output voltage is fairly low, however, so that the gain of the audio amplifier must be high. The only serious drawback is that the connection can be made only with the chassis cover removed. This is not a serious drawback, however, since it is primarily used for tests over a single stage, when regenerative feedback is likely to be small.

DEVICES FOR MEASURING RECEIVER OUTPUT

Except in the case of noise figure measurements, the output indication is ordinarily no problem, a vacuum-tube voltmeter or cathode-ray oscilloscope being used with only ordinary precautions. Accurate measurement of noise figure, however, demands a knowledge of the relation between input power and output indication of the indicating circuit, as well as absence of overload effects ahead of the point of measurement.

12·11. Attenuator and Postamplifier.—This device is the most accurate and generally satisfactory when noise generators are employed. Although it is more complex than the others to be described, it can be operated successfully by relatively untrained personnel. Therefore, it is ideally suited for use on production lines as well as for use in research laboratories.

Method.—The method is illustrated by a block diagram in Fig. 12·19. The reading of the output meter is first observed with the attenuator cut out and the noise generator turned off. Then the attenuator is cut in and the noise generator is adjusted to make the output meter have the same deflection as before. If the attenuator has an attenuation ratio α equal to 2 in power, this available power of the noise generator causes the available output power of the receiver to double. In the event α is not exactly 2, a very simple correction can be made, as will be shown later. Because the postamplifier is operated at constant signal level, no harm is

caused by overloads within it, and this unit can be connected to the amplifier under test at a stage where the signal level is low enough to avoid overload difficulties. This connection can be made by means of an adapter that plugs in in place of one of the tubes in the amplifier.

Construction.—The postamplifier should have the same center frequency as the amplifier under test and a bandwidth somewhat greater. Great care should be exercised in shielding the postamplifier, and it should be operated from its own power supply to avoid any chance of feedback to the amplifier under test. It is convenient to gang with the attenuator switch an SPST switch connected through a cable to the $B+$ supply of the noise generator so that the generator can be turned on and the attenuator can be cut in by the turn of a single knob. However, this switch should be isolated from the attenuator to prevent signals from leaking out. The detector uses any of the common diode receiving tubes.

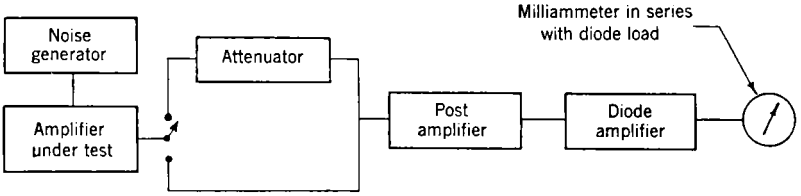


FIG. 12·19. Attenuator and postamplifier method of measuring noise.

A meter indicating 1 ma full scale in series with the diode load resistor is adequate for many measurements, but the over-all sensitivity can be increased by using a meter of 30 or 100 μa full scale, with a dry cell and a variable resistance in series with it connected across the meter to buck out the dark current of the diode. It is useful, also, to have a jack connected to the high-potential end of the load resistor through a decoupling filter for making a connection to an oscilloscope so that the pass band of either the postamplifier or the combination of the amplifier and postamplifier can be observed when a swept-frequency signal generator is connected to their respective inputs.

The attenuator and its immediate circuit are designed with these objectives: (1) to make the calibration independent of the condition of impedance mismatch at the output of the amplifier and (2) to have a value of α as near to 2 as possible. The diagram of the attenuator and its immediate circuit are shown in Fig. 12·20. If the characteristic impedance of the cable is other than 75 ohms, the values of the resistors should be changed in the proper ratio. These resistors are $\frac{1}{2}$ watt carbon, and those in the attenuator are selected to be within 2 per cent of the indicated values. The terminating resistor R and the inductance L (which reso-

nates the input capacity of the tube) are selected so that the same impedance (at the center frequency of the amplifier) is measured across the pins of P_3 with both positions of the attenuator switch. If this condition is met, the calibration of the attenuator is independent of the impedance presented by the cable. However, as a precaution against slight errors in the values of L and R, this cable should present an impedance that approximately matches the iterative impedance of the attenuator to absorb any waves that might be reflected back from R to L. In Fig. 12·20, the amplifier end of the cable is fitted with a plug P_1 which fits into a tube socket of the amplifier (which is assumed to have single-tuned circuits). This plug is designed to connect the cable in parallel with the load of the tube previous to the tube displaced by the plug. The impedance of this load is usually high compared with 75 ohms, and therefore the plug contains a 75-ohm resistor which causes the cable to be ter-

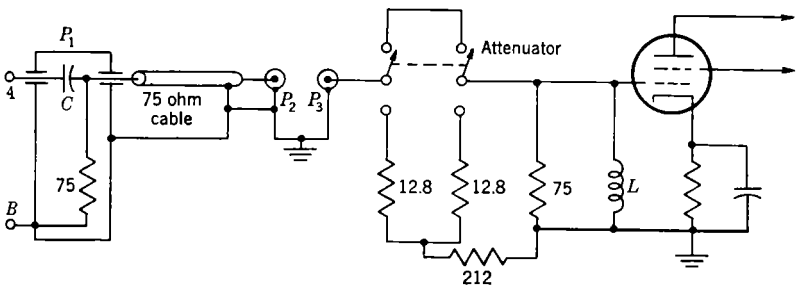


FIG. 12·20.—Details of attenuator.

minated in approximately its characteristic impedance, which in turn, therefore, presents an impedance of approximately 75 ohms to the attenuator. The plug presents a net impedance of 37.5 ohms (75 ohms in the plug in parallel with a 75-ohm terminated line), which is a very heavy load and causes the bandwidth of this circuit to be large. An alternative procedure is to omit the resistor in the plug P_1 but to insert between P_2 and P_3 a T- or Π-attenuator pad having 75 ohms iterative impedance and at least 10-db attenuation. In fact, if amplifiers with different gains or the same amplifier with various settings of the gain control are to be measured, the postamplifier should have sufficient gain to operate with the amplifier having the minimum gain, and it is desirable to employ, between P_2 and P_3, a variable step attenuator to compensate for any excess gain. With such a device, the plug P_1 presents an impedance of 75 ohms, which has the advantage that less distortion of the pass band is produced if the blocking condensers have reactances comparable to 37.5 ohms.

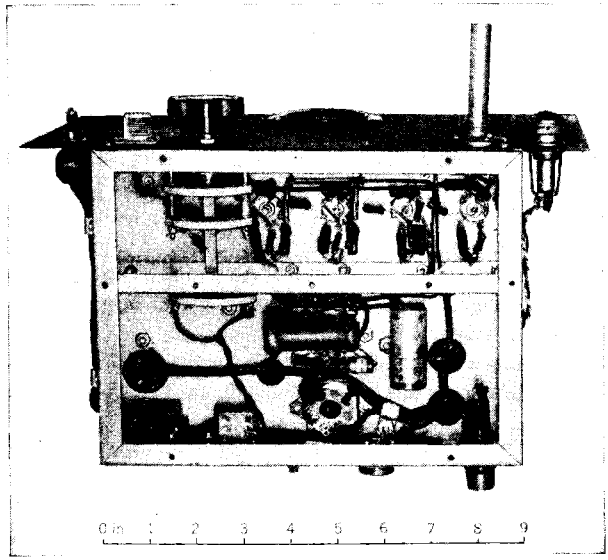


FIG. 12·21.—Bottom view of postamplifier.

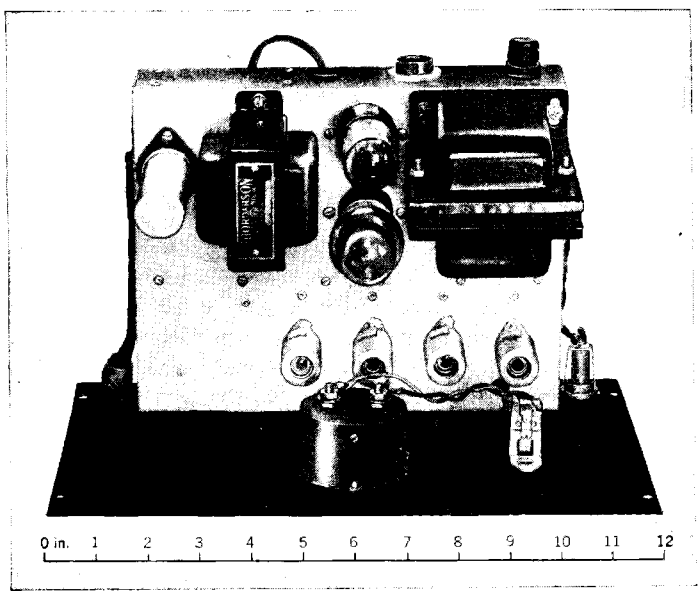


FIG. 12·22.—Top view of postamplifier.

Photographs of a unit of the type described in this section are shown in Figs. 12·21 and 12·22. The circuit diagram is shown in Fig. 12·23. Note how the amplifier is shielded from the power supply and the section of the selector switch that controls the $B+$ of the noise generator. This unit has a bandwidth of 8 Mc/sec centered at 30 Mc/sec, and the voltage gain between the grid of the first tube and the plate of the third tube is approximately 350.

Correction to Be Applied When α Is Not Equal to Two.—The correction to be applied when α is not equal to two is very simple. When a noise figure is measured, it is assumed that at the input of the amplifier there is a single source of noise whose available power is W_n and which simulates all the sources of noise within the amplifier. With the attenuator

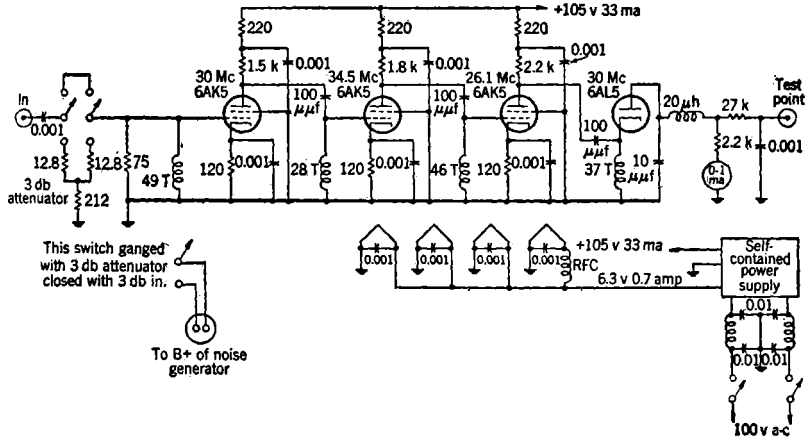


FIG. 12·23.—Auxiliary amplifier for noise figure measurement.

in, the noise source is turned on and adjusted to give an available power W_1 such that the same output is obtained as with the noise generator off and the attenuator out. Therefore

$$W_n + W_1 = \alpha W_n,$$

or

$$W_n = \frac{W_1}{\alpha - 1}.$$

The procedure is to (1) calculate the noise figure as though α were exactly 2 and (2) divide the result by $\alpha - 1$ if the noise figure is expressed as a ratio or subtract $10 \log_{10} (\alpha - 1)$ if it is expressed in decibels. As α is approximately 2, $\alpha - 1$ is approximately unity, half as large. Therefore, the percentage correction in noise figure is approximately twice as great as the percentage deviation between α and 2.

12·12. Crystal and Diode Rectifiers.—Crystal and diode detectors (which include the detectors incorporated in radio receivers with d-c voltmeters connected across their loads or milliameters connected in series with their loads) suffer disadvantages except in regard to availability. Usually they do not respond to a combination of noise and c-w according to the same law as to noise alone, and therefore they are not reliable for highest accuracy when c-w generators are employed. Even with noise generators, one must expend considerable effort, especially with wide-band receivers, to establish the law of the detector. (Such detectors usually have a "dark current" with no signal applied. Depending upon individual conditions it is sometimes necessary to subtract it and other times to neglect it in establishing some simple law.) Generally, one cannot establish any such law very accurately, as it is not possible to hold conditions sufficiently stable. Also the law itself will often hold for only a small range of signal level. Hence this method is usually not very accurate even with noise sources.

Crystal detectors used with indicating devices that have a deflection sensitivity of about 1 μa full scale have been found to be accurately linear in power, but with meters of lower sensitivity they may deviate considerably from such a law.

12·13. Bolometers.—A bolometer is the most satisfactory output-indicating device for use with unmodulated c-w signal generators, as it

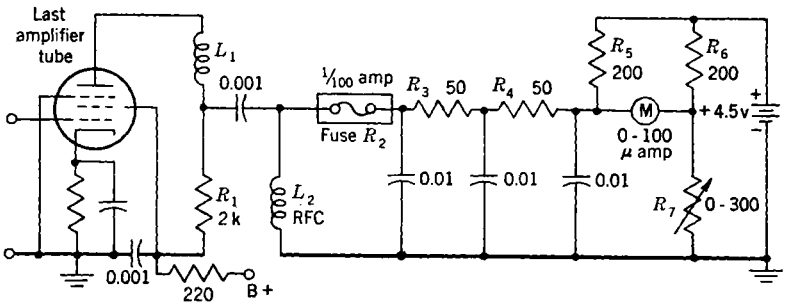


FIG. 12·24.—Unbalanced bridge bolometer.

responds to unmodulated c-w and noise powers according to the same law. It can be used with noise generators as well.

An unbalanced-bridge type of bolometer is sufficient for this purpose, as it is necessary to measure merely relative power and not the absolute power in watts at the output of the amplifier. Such a device is illustrated in Fig. 12·24. R_2 is a $\frac{1}{100}$-amp Littelfuse or a Thermistor which (together with the decoupling resistor R_3 and R_4) forms one arm of a Wheatstone bridge whose other arms are R_5, R_6, and R_7. With the amplifier turned off, the bridge has been balanced by adjustment of R_7. With the ampli-

fier operating, its output, consisting of noise signal or a combination of both, is dissipated in R_2, causing its resistance to change, upsetting the balance of the bridge, and causing the meter to deflect. If the amplifier is linear, the deflection is proportional to the power input.

Generally it is not very convenient to connect such a device to the output of an amplifier that already has been constructed. Therefore it is desirable to mount such a device in an auxiliary postamplifier whose input is obtained from a plug that replaces one of the tubes of the amplifier under test.

Thermocouple Meters.—Experience obtained at 30 Mc/sec has indicated that the deflection of such meters does not depend upon the power according to any simple law nor do they accurately obey the same law as at direct current. Therefore if they are used, they should be calibrated in terms of other devices such as attenuators or bolometers at the frequency at which they are to be used.

CHAPTER 13

A SIMPLE RADAR RECEIVER

By F. M. Ashbrook

13·1. Introduction.—The receiver to be described in this chapter is selected as an example of straightforward design without any of the special features found necessary in radar systems of other types. It is a part of a general-purpose shipborne radar system operating in the region of 10,000 Mc/sec with a peak transmitter power output of approximately 60 kw. As is frequently the case in radar systems, the transmitter, receiver, and duplexing components are mounted in a single waterproof unit known as an "r-f package." Inside this package the receiver unit comprises the mixers for signal and AFC channels, the local oscillator, the AFC circuits, the signal i-f amplifier, the second detector, and the video amplifier. Power supply and operating controls are located remotely.

13·2. Mechanical Construction of Receiver Unit.—To help secure the very high degree of shielding required, it is desirable to mount the mixers as an integral part of the receiver unit. This requirement poses additional mechanical problems, since the whole receiver unit must now be accurately located with respect to the waveguide assembly making up the duplexing system.

Hollow threaded bushings that screw into brackets in the receiver, as shown in Fig. 13·4, are used for mounting the receiver to provide the necessary adjustment for proper alignment. These bushings rest on a mounting frame on which there are four studs that pass through the hollow section of the bushings to hold the receiver in place. The receiver can be moved up and down in space by adjustment of the bushings. The fore-and-aft and sideways movement is made possible by slotted holes in the mounting brackets which allow the brackets to move with respect to the base of the r-f package. When the receiver is mounted in the r-f package, it can be carefully lined up with the waveguide that connects the magnetron with the antenna. The complete r-f package minus cover is shown in Fig. 13·1. The transmitter tube and its transformer have been tipped out of the way for a better view of the receiver.

Since the transmitter and receiver are in the same package, it is necessary that the receiver unit be carefully shielded. The transmitter puts out a signal of about 5×10^4 watts, and the receiver is expected to

detect a signal as weak as 10^{-13} watt, a total power difference of 5×10^{17}. Fortunately the isolation does not have to be complete, for while the transmitter is transmitting, there are no signals to be received. It is important, however, that the receiver be capable of receiving small signals almost immediately after the transmitter ceases. If there were no shielding between the two units, the transmitter power would leak into

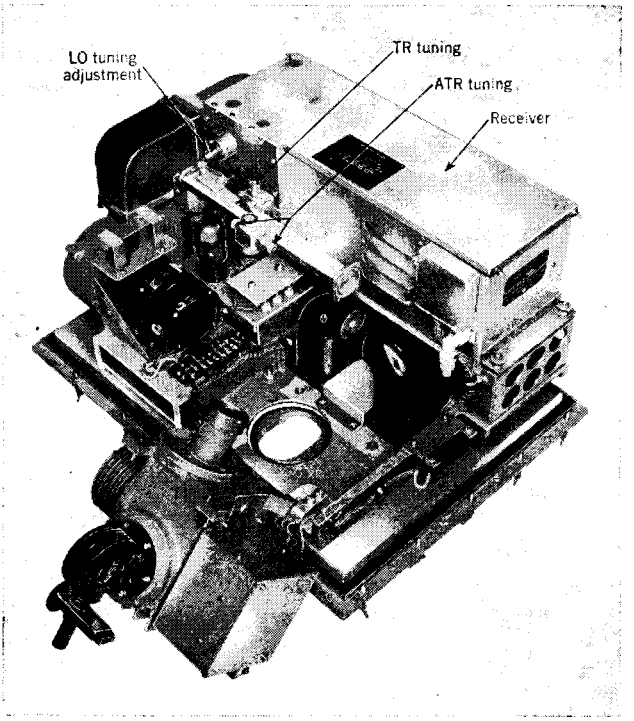


FIG. 13·1.—R-f package with receiver mounted in place. (*Courtesy of Raytheon Manufacturing Company.*)

the receiver circuits and cause them to oscillate and remain ringing for some time after the transmitter ceases. In addition the mixer crystals must be protected from burnout, which may occur at power levels of approximately 1 watt. The TR tube described in Chap. 2 takes care of the isolation of the direct path from waveguide to receiver, but other paths resulting from pickup of transmitter power by i-f components and feedback to the mixers must be stopped by careful shielding.

13·3. Mixer.—A schematic view of the r-f circuit and mixer is shown in Fig. 13·2. The input to the signal channel is coupled into the mixer through the TR tube. The attenuated signal for the AFC circuit is

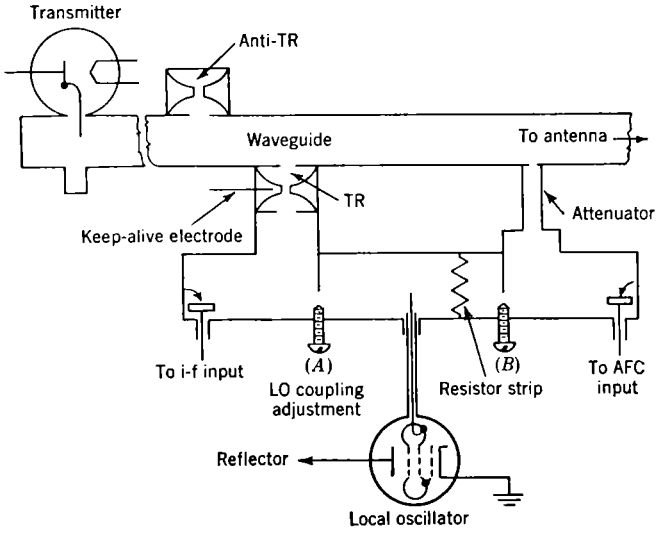


FIG. 13·2.—Diagram of mixer circuit.

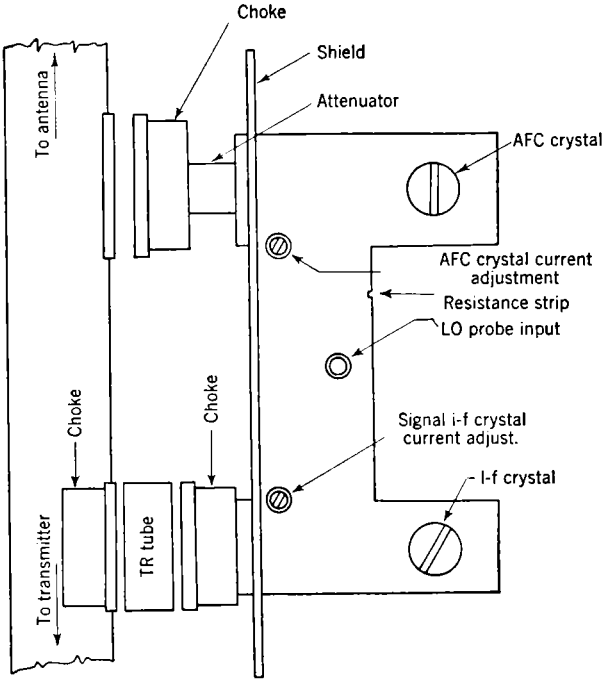


FIG. 13·3.—Partial assembly of duplexer-mixer circuit.

coupled into its mixer through the attenuator (waveguide beyond cutoff). The local oscillator is inserted into the waveguide joining these two mixers. The resistor across this guide is a terminating resistor that reduces the Q of chamber so that it will be less frequency sensitive. The amount of LO power that is injected into each circuit is adjusted by varying the effective width of the slots A and B by means of the screws.

The transmission loss of the attenuator in the AFC mixer is adjusted so that the transmitted pulse will give a signal to the mixer of from 1 to 10 mw. To ensure successful AFC functioning, the leakage into the AFC mixer through the TR tube, the LO chamber, and the two LO-coupling slits must be small compared with the signal that comes through the attenuator. Therefore the leakage into any part of the mixer circuit must be very small. The problem of leakage is serious because, in order that signals shall not leak into the mixer by way of the choke joints on the attenuator and between the TR tubes and choke joint on the mixer,



FIG. 13·4.—Side view of receiver showing mixer assembly.

it is necessary to restrict tolerances and make sure of parallelism between parts of the mixer, duplex assembly, and TR tube (see Fig. 13·3). Compressible metal-mesh gaskets have been used successfully to make up for small misalignments. The metal strip connected to the solenoid in Fig. 13·4 is for the purpose of protecting the crystal when the system is not operating and the keep-alive voltage is not applied to the TR tube. Without the keep-alive voltage, the TR tube does not break down at a power level low enough to protect the crystal. Consequently, near-by radar sets might transmit enough energy into the antenna to burn out the crystal. When the radar set is in operating condition, the solenoid moves the strip across the waveguide until the hole is exactly centered. This hole is slightly larger than the hole in the TR tube and therefore does not interfere with the transmission of energy into the mixer. The shutter was located between the TR tube and the mixer because the first TR tubes manufactured for this frequency had to be broken down by the transmitter once or twice before they functioned properly. The time delay in the opening of the relay circuit is long enough for this breakdown

to occur. From a standpoint of leakage the shutter is in the worst possible location because unless extra precautions are taken, the shutter can act as a very good antenna to couple energy into the mixer chamber.

The socket mounting into which the local oscillator, a 723A/B, is plugged is not shown in Fig. 13·3; all that is shown is the choke into which the probe is inserted. The choke prevents leakage of the r-f signal from the mixer assembly and also serves to keep stray radiations from entering the mixer by way of the local oscillator. In Fig. 13·5 the local oscillator is shown seated in its socket. The stud with the knurled nut carries a bakelite arm that prevents the local oscillator from becoming dislodged from its socket. The r-f output from the tube is taken out through the base on a coaxial conductor, which takes the place of one of the tube pins. The probe extends far enough into the waveguide to give optimum coupling. In Fig. 13·5 the stud to the left of the local

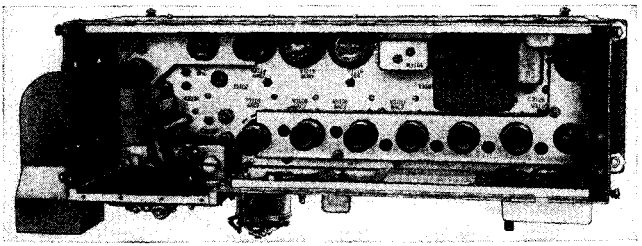


FIG. 13·5.—Top view of receiver.

oscillator is used for adjusting the coupling of the LO signal into the mixer channel as shown diagrammatically by *A* in Fig. 13·2. This adjustment is also provided with an r-f choke to prevent leakage. A similar arrangement is used on the other mixer to adjust the LO injection into the AFC mixer.

The two large caps with the screw slots (Fig. 13·5) permit replacement of crystals. They also hold the crystals firmly in place and grounded. The output of the mixer is taken from the tip of the crystal and passes through an r-f filter circuit that blocks the r-f signal from the i-f amplifier. In this particular mixer the capacity is about 11 $\mu\mu$f and forms part of the input circuit for the first amplifier stage. The LO tube is operated with the cathode at ground; the cavity *and shell* at +300 volts; and reflector electrode at a negative voltage variable from −60 to −180 volts, depending on the frequency of operation and the characteristics of the particular tube. The reflector-voltage circuit will be discussed in more detail in Sec. 13·5.

13·4. I-f Amplifier.—The i-f amplifier operates at an intermediate frequency of 30 Mc/sec and has six stages. It is followed by a diode

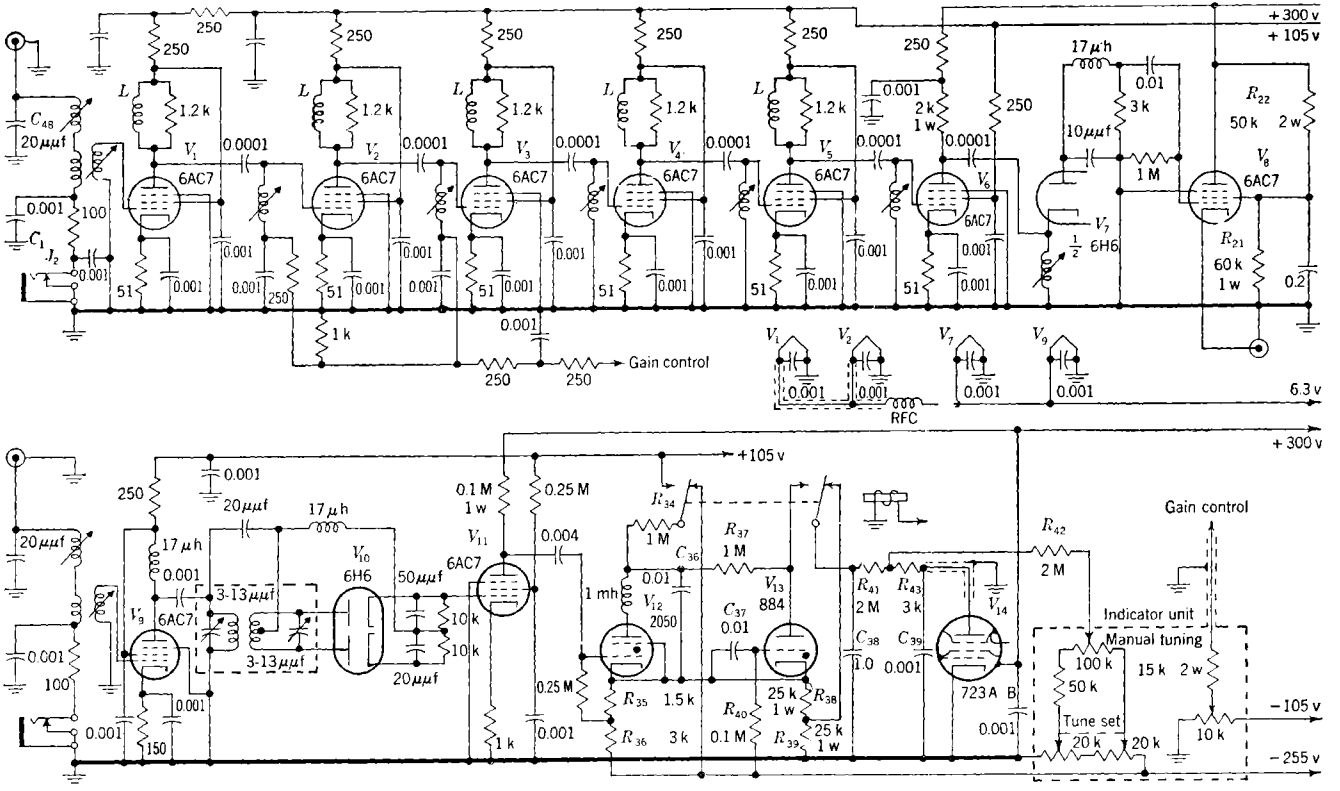


FIG. 13·6.—Circuit diagram of receiver. All resistors are ½ w unless otherwise specified. Heaters of V_3, V_4 and V_5, V_6 are paired and wired the same as V_1, V_2.

detector and a single video-amplifier stage that both limits the signal and drives a terminated 75-ohm line. Except for the input transformer, all of the i-f stages are single-tuned circuits. The circuit diagram of the receiver is shown in Fig. 13·6.

The input circuit is a double-tuned inductively coupled transformer designed for the optimum noise figure in a pentode (6AC7) amplifier. This optimum coupling is, in this case, somewhat looser than would be required for transitional coupling. The coupling is obtained by winding a few turns of the primary transformer on top of the secondary coil and is varied by changing the number of turns of the primary coil that are wound on the secondary form. Greater stability in manufacture is obtained by winding the coupling turns over the grounded end of the secondary coil as shown in the circuit diagram. The low-potential end of the primary coil is filtered and goes to ground through the jack J_2. A 0- to 1-ma meter is plugged into the jack for measuring the crystal

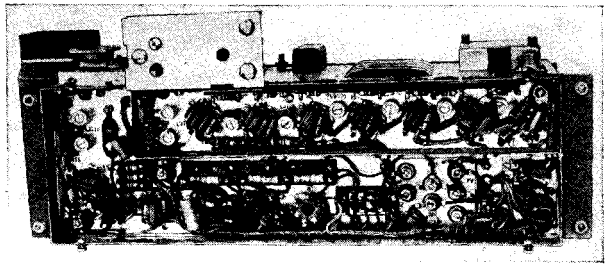


FIG. 13·7.—Bottom view of receiver with covers off.

current. The bandwidth of the input circuit between half-power points is about 10 Mc/sec.

The actual construction of the input circuit can be seen in Fig. 13·7. The output of the mixer is padded with a 20-$\mu\mu$f condenser (C_{48}, Fig. 13·6), which makes the input circuit less sensitive to variation in the mixer capacity. The padding condenser reduces the bandwidth of the input circuit, but in this case the bandwidth is more than enough for the rest of the amplifier. The noise figure is not appreciably affected by the extra primary capacity.

As in all i-f circuits, the points at which the ground connections are made must be carefully chosen. The ground connection on the secondary coil is made at the circuit ground for the tube (Pin 3, V_1). The i-f ground for the primary—the point at which C_1 is grounded—should be as near the mixer as possible, since the path of the i-f circuit is through the crystal to the primary of the coil, then through the bypass condenser to ground. From that ground point the signal must flow along the chassis and up through the mixer to the ground point of the crystal.

The i-f stages are all similar except for the two with gain-control circuits on the grids and the last, which has a higher plate voltage. All of the i-f stages are singled-tuned circuits with a bandwidth of about 5.1 Mc/sec and a voltage gain of 9.6 per stage. The over-all bandwidth of the i-f amplifier is about 1.7 Mc/sec. The calculated values are based on the assumption of an average g_m for 6AC7's of 8000 μmhos and capacity loading in each circuit of 26 $\mu\mu$f. Although the plate-load resistor in the last stage is not the same as that in the other i-f stages, the extra loading of the detector circuit increases the circuit loading so that this stage has about the same bandwidth as the rest.

In the i-f amplifier the circuit ground for each i-f stage has been made on Pin 3 (the suppressor grid), and the noncircuit elements of the tube—the heater and the shell—have separate grounds as described in Sec. 10·7. The heaters for the i-f tubes have been wired in pairs, with a choke in the first pair only. This choke, which can be seen in Fig. 13·7, is located just outside the i-f strip opposite the third i-f tube. The first five i-f tubes are operated with 105 volts on the plate and screen. This operating voltage is such that the maximum operating transconductance of the vacuum tube is realized. The inductor shown in parallel with the load resistor is an i-f choke that eliminates the d-c voltage drop across the load resistance.

The gain control is obtained by changing the bias on the second and third i-f stages. It has sufficient range to reduce the gain of the i-f amplifier so that all the signals can be reduced to zero. Grid gain control, as pointed out in Chap. 9, is the most satisfactory circuit for radar use.

The i-f tubes are held in place by means of a bracket that holds the base projection of the metal tubes (see Fig. 13·5). The holes between the tubes are for access to the tuning adjustment on the interstage coils. Because of the internal structure of the 6AC7 vacuum tube, it is essential that this metal bracket does not make contact with the tube shells. Consequently, this clamp is provided with an insulating liner. Even though Pin 1 on the socket is carefully grounded, the shell will not necessarily be at ground potential. Actually, the internal connection between the tube pin and the shell is comparatively long, because the getter in the tube is flashed through this connection. Ordinarily this length of lead is not enough to be troublesome, but at these frequencies the inductance is sufficient so that an appreciable signal can appear on the shell of the tube. If a metal clamp of large area is allowed to connect shells of successive stages, enough feedback may occur to be troublesome. This same lack of firm grounding of the shell has occasionally required the use of a tube shield in other sets. Such a shield hinders heat dissipation and should therefore be avoided whenever possible.

The detector is a single diode with negative output. At normal operating gain there will be about 1 volt of noise out of the detector. Since the interstage capacity is about 27 $\mu\mu$f, the video bandwidth will be approximately 2.0 Mc/sec—which is somewhat narrow for a 1.7-Mc/sec i-f bandwidth. The condenser coupling to the next stage has a time constant of 0.01 sec (0.01 μf $\times$ 1 megohm), which is sufficient for this application.

The last stage in the signal channel of the receiver proper is a cathode-follower limiter stage. The cathode of this tube drives a 75-ohm terminated line with a negative 0.8-volt signal. The screen of the limiter is operated at about 80 volts and is stabilized by the voltage divider network R_{22} and R_{21}. A signal of about 2.8 volts on the grid of the limiter tube is required to drive the tube to cutoff.

13·5. AFC Circuit.—The local oscillator V_{14} (723A/B) is of the reflex klystron type. The tuning of the oscillator is accomplished by mechanically deforming the resonant cavity for tuning over a wide frequency range and by varying the reflector voltage for small frequency changes. The electronic tuning by reflector-voltage change is of most interest, as both the AFC and manual tuning control act in this way. A more complete discussion of LO tubes appears in Chap. 2.

The AFC circuit in this receiver uses the double gas-tube type of control circuit. The double mixer, having one channel for the signal and an attenuated channel for the AFC, was described in Sec. 13·3. The layout of the AFC circuit may be seen in Fig. 13·7. It starts at the mixer with the two input coils located in the upper left corner (L3110) and continues down the side and across the bottom with the tubes arranged in the same order as that shown in the circuit diagram (Fig. 13·6).

The input circuit for the AFC channel is the same as that for the signal channel. A single i-f stage is followed by the discriminator. Since the discriminator circuit has an input impedance of less than 3000 ohms, it furnishes the plate load for the i-f stage. The discriminator circuit is a conventional Foster-Seeley type. The video amplifier following the discriminator is a narrow-band high-gain amplifier that inverts and amplifies pulses out of the discriminator. The short-time-constant coupling circuit between the amplifier (V_{11}) and the hold gas-tube (V_{12}) prevents power-frequency hum modulation in this circuit from triggering the hold tube. The last tube (V_{13}) is the search tube. The action of the circuit is described in Chap. 3.

It has been pointed out in Chap. 3 that for proper functioning of an AFC system it is vital that the frequency of the local oscillator be the same during the transmitted pulse as in the intervals between pulses. This requirement necessitates considerable care in shielding to be sure that no effect of the high-power transmitter pulse gets to the oscillator.

Both r-f energy and video pulses, arising from rectified r-f power, from transients on power-supply leads, or from leakage of the input pulse to the magnetron, may cause trouble. The reflector circuit is particularly vulnerable to stray pickup because it is a high-impedance circuit. A spurious signal of 1 volt on it will introduce a frequency shift of 2 to 4 Mc/sec. Frequency shifts are also produced by changes in cavity voltage, but the shift per volt is only about one-quarter as great.

Besides shielding the oscillator itself, it is necessary to keep out of the AFC circuit video pulses that are due to stray pickup from the transmitter. If allowed to enter, these pulses will be indistinguishable from the proper pulses coming from the discriminator and may cause improper operation of the circuit.

It will be noted that the plate circuit in the i-f tube (V_9) is shunt-fed. The advantage of this arrangement is that the primary of the discriminator transformer is grounded and any transients that appear in the voltage supply are shunted to ground instead of being transmitted through the discriminator and appearing as a pulse in the output. The sweep tube (V_{13}) sweeps the local oscillator from a high frequency to a lower frequency as the reflector voltage passes through the operating range. In this receiver the discriminator was designed so that the local oscillator would operate at a frequency 30 Mc/sec higher than the transmitter frequency. The bandwidth of the discriminator circuit is described by the separation between the peaks of the discriminator characteristic, which in this case is about 2.6 Mc/sec.

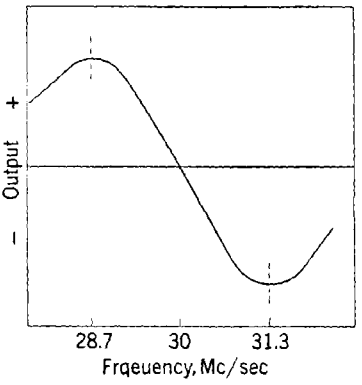


FIG. 13·8.—Discriminator characteristics as observed at plate of V_{11}.

The output of the video amplifier following the discriminator will be either positive or negative pulses—or nothing. The zero output occurs in two cases: when the local oscillator is exactly tuned to the proper frequency and when the local oscillator is detuned sufficiently so that the signal is outside the pass band of the i-f and discriminator circuits. The output of the video tube will be negative pulses if the frequency difference is inside the discriminator pass band and greater than the proper frequency, whereas the output pulses will be positive if the difference frequency is less than the proper frequency. Figure 13·8 shows the discriminator characteristic as amplified and inverted by the video amplifier. The negative pulses out of the video amplifier do not affect

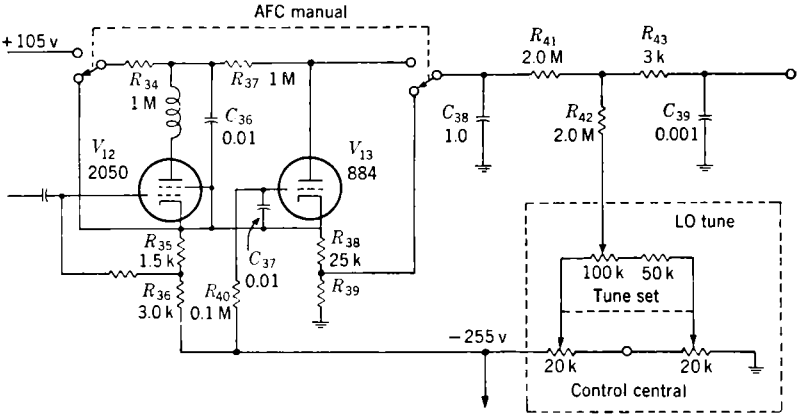


FIG. 13·9. Basic components of gas-tube control circuit. Note that switch is shown in manual position.

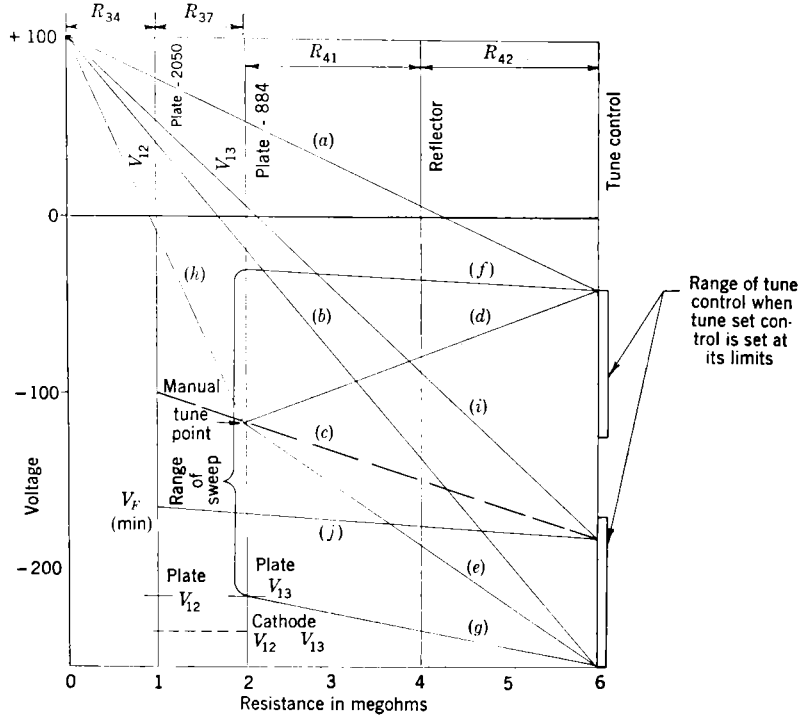


FIG. 13·10.—Resistance-voltage graph.

the gas-tube circuit, since V_{12} (2050) is already biased beyond cutoff. The positive pulses cause the 2050 to fire, which arrests the drift of the oscillator tube as described in Chap. 3.

The circuit diagram of the basic components of the gas-tube control circuit is shown in Fig. 13·9. The bias for the gas tubes is furnished by the resistors R_{35} and R_{36} in the cathode circuit. The bias on V_{12} is about 3.5 volts more than the cutoff bias for the tube. The bias on V_{13} is such that the tube sweeps from about -225 to -35 volts. Variations from tube to tube are such that the upper limit may range from about -70 volts to zero. The voltage at the tap between R_{38} and R_{39} is so chosen that it represents approximately the midpoint of the 884 (V_{13}) sweep. The combination of the tuning control and tune-set control will vary the voltage at the bottom end of R_{42} from -255 to -42 volts. These potentiometers are in another unit which is located at the control position.

Many of the quantities that determine the performance of the circuit may be visualized most easily by the use of the resistance-voltage graph shown in Fig. 13·10. Resistance is plotted as abscissa, and voltage as ordinate. The resistances that are most important in controlling circuit behavior are the series string R_{34}, R_{37}, R_{41}, and R_{42}, which are accordingly plotted in that order. Vertical lines indicate the junctions of the various resistors and are also labeled with the other circuit components that connect at that point. Since the internal impedance of the voltage source, made up of the *tune* and *tune-set* potentiometers, is small compared with the other resistances in the circuit, it has been neglected. If both gas tubes are prevented from firing, the potentials that appear in various parts of the circuit may be found by drawing a line from the point at the upper left, representing the connection of R_{31} to $+105$ volts, to the point on the right-hand line representing the connection of R_{42} to the particular voltage furnished by the *tune* and *tune-set* controls. Lines *a* and *b* represent the limits reached under these conditions. Thus it may be seen from the graph that the voltages at the plate of the 884 tube range from $+55$ to -17 volts at extreme settings of the tuning controls. As explained in Chap. 3, this value is of importance in determining whether or not the sweep tube will fire reliably.

When the AFC-MANUAL switch is in the MANUAL position, the potential distribution throughout the circuit is given by a line such as *c* connecting the point labeled manual-tune point, whose potential is set by the voltage divider R_{38}-R_{39}, with the appropriate point on the tuning-control line. Line *c* is so chosen that the voltage applied to the reflector of the local oscillator is exactly -150 volts and shows that the voltage furnished by the tuning controls must then be -182 volts. Lines *d* and *e* show that the reflector-voltage range in the manual-tuning position is from -80 to -185 volts.

When the 884 search tube is in operation, the voltage at its plate will be within the range indicated on the vertical line. Lines f and g show the limits of the range to which the sweep can carry the circuit with extreme settings of the tuning controls.

The procedure for calculating rates of follow for the circuit is analogous to that described in Chap. 3. The point corresponding to the condition where the 2050 tube (V_{15}) fires on every pulse [designated as $(V_F)_{min}$ in Chap. 3] has been indicated on the vertical line. Line j, connecting this point with the tuning-control point, gives the values needed for

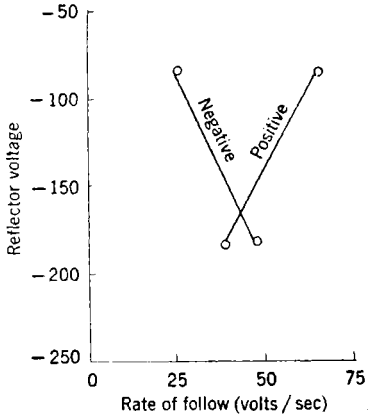


FIG. 13·11.—Follow rates of AFC circuit.

calculating follow rate in the negative direction. Line i gives the necessary values for the positive direction. Both of these lines pertain to a reflector voltage of −150 volts. The results of a series of such calculations are shown in Fig. 13·11. Here the rate of change of reflector voltage in volts per second is plotted against the reflector voltage. The quantity of final interest, the rate of change frequency, is obtained by multiplying these values by the electronic tuning coefficient of the LO tube expressed in megacycles per second per volt. Since this coefficient varies widely from tube to tube, typical values for the 723A/B being in the range of 1 to 4 Mc/sec per volt, the results are left in the form of the rate of change of voltage.

CHAPTER 14

AN AIRBORNE RECEIVER INCORPORATING ANTICLUTTER CIRCUITS

By C. Butt

14·1. General Description of the Receiver.—This receiver was designed to meet the requirements of an airborne S-band radar search system. The transmitter pulse length is 2 μsec; the repetition rate is 300 cps; and the r-f pulse output power is about 1 megawatt. The receiver was planned for ease of production and with a view to later modifications and developments. Two major modifications were incorporated without undue difficulty after production had started, and a receiver for a completely different radar system was readily developed from the same basic design.

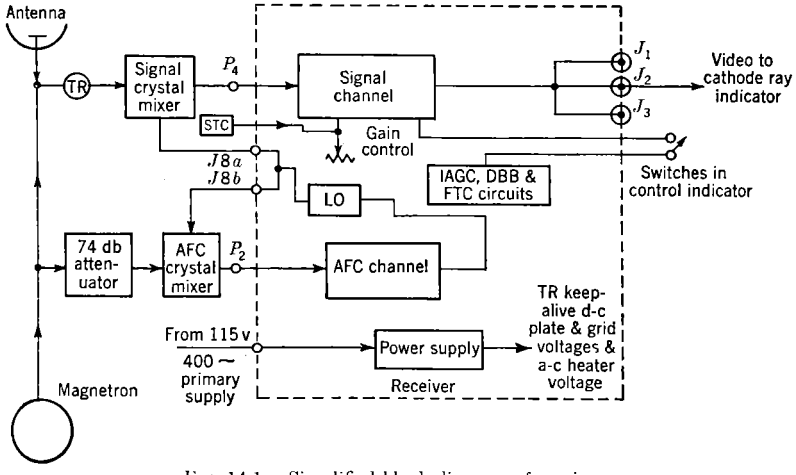
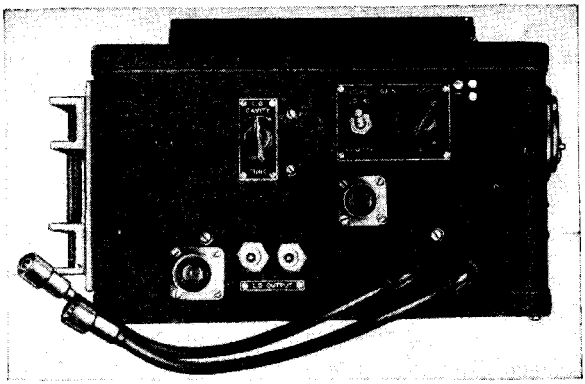
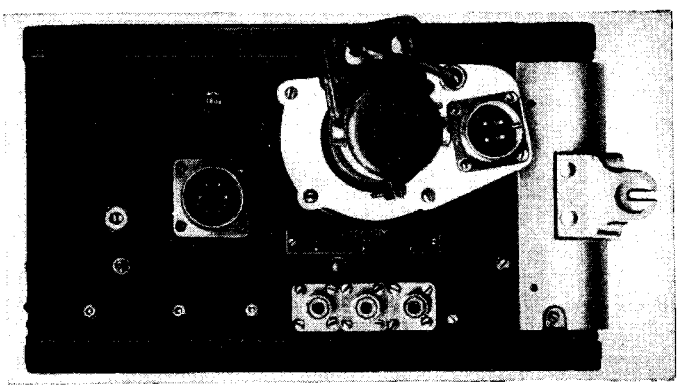


Fig. 14·1.—Simplified block diagram of receiver.

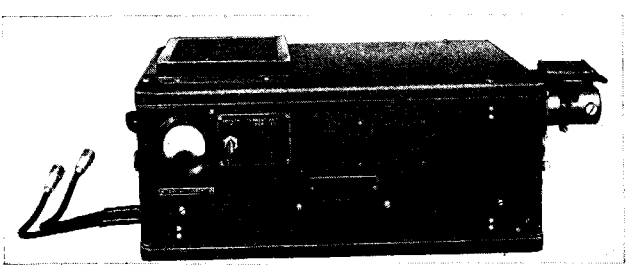
The receiver is a superheterodyne with seven stages of 30-Mc/sec i-f amplification, a second detector, one stage of video amplification, and a cathode-follower output tube. It is equipped with AFC and optional anticlutter circuits known as "instantaneous automatic gain control," IAGC; "detector balance bias," DBB; and "fast time constant," FTC. An additional anticlutter device known as "sensitivity time control,"



(*a*)



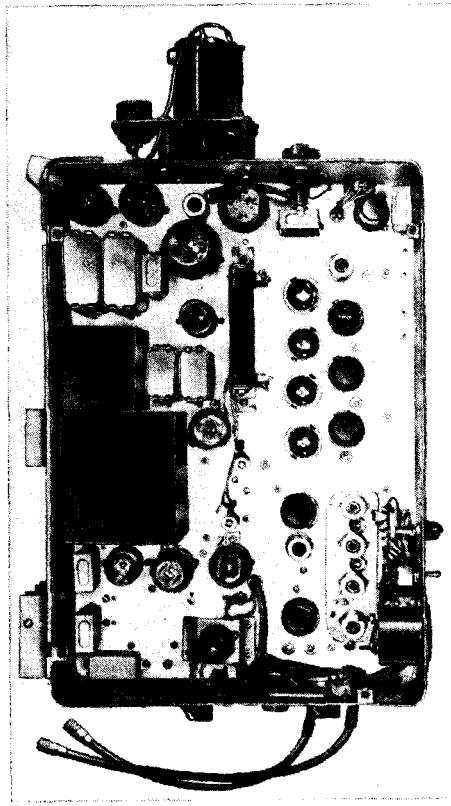
(*b*)



(*c*)

FIG. 14·2.—Receiver. (*a*) Front view; (*b*) rear view; (*c*) side view; (*d*) top view; (*e*) bottom view.

STC, also modifies the receiver operation, but this device is contained in a separate unit. The receiver contains its own power supply operating from a single-phase 120-volt, 400-cps power source. In addition it furnishes keep-alive voltage for the TR tube. The receiver delivers LO power by means of an impedance-matching coupler (splitter) to two



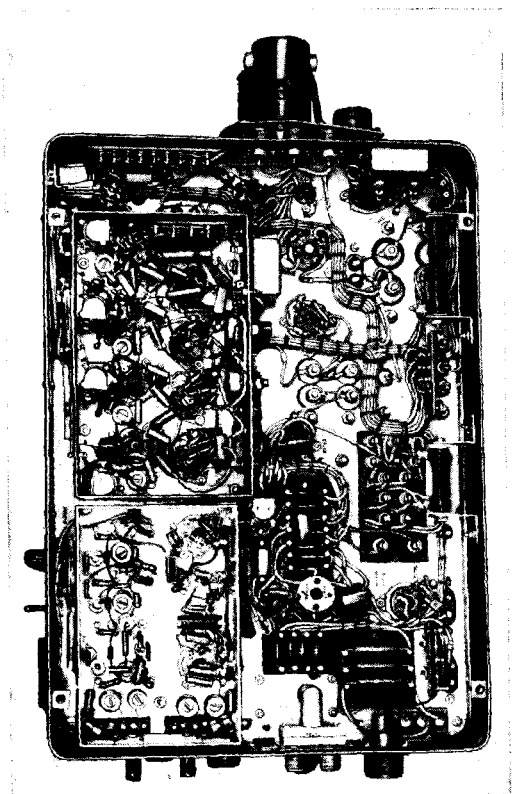
(*d*)

FIG. 14·2.—For descriptive legend see page 350.

external crystal mixers, one for the signal channel and the other for the AFC channel. Figure 14·1 is a block diagram of the receiver showing its relation to the radar system.

A general description of the electrical and mechanical design will be presented first, and the special features and circuitry discussed in more detail subsequently. The receiver design was carefully considered from

the following standpoints: stability, servicing and maintenance, modification facility, and ease of production. In order to satisfy these considerations, part of the AFC channel, the i-f channel, and associated i-f circuits are segregated into two separately shielded subchassis sections. One section contains the first four i-f stages and that portion of the AFC channel composed of the i-f amplifier, discriminator, and video amplifier.



(*e*)

FIG. 14·2.—For descriptive legend see page 350.

The section is referred to as the i-f—AFC chassis. The other section, referred to as the i-f—clutter chassis, contains the last three i-f stages, second detector, and four anticlutter (IAGC-DBB) circuits. The photographs of Fig. 14·2*a* to *e* show in general how this was done.

The segregation of the i-f circuits in this manner is logical, since the circuitry and mechanical layout lends itself nicely to such a design and

splitting the i-f channel into two units of approximately equal gain is advantageous in maintaining stability.[1]

Leads carrying i-f signals are short and component parts are small and isolated to reduce stray capacity and unwanted coupling to a minimum. Ground currents are practically eliminated from the *main* chassis deck, and the separation between the two i-f chassis prevents feedback due to ground current coupling between the output and input circuits. Furthermore, the use of button-type mica feedthrough condensers makes it possible to feed voltages into and out of each subchassis with efficient filtering. Where filtering is unnecessary or undesired, insulated low-capacity Kovar feedthrough terminals are used as, for example, at the second detector output.

The two subchassis are fastened to the main chassis deck by spacer mounting bolts. The main chassis must have clearance holes for the tubes and the tuning screws on the i-f coils. Either subchassis is easily removable as an integral unit by unsoldering leads from the various feedthrough terminals and removing the mounting bolts.

This design has proved to be extremely valuable. Some 72 new subchassis units were installed in field modifications on 52 production receivers without difficulty. These modifications required a complete new i-f—AFC chassis containing low-noise triode input circuits and new i-f—clutter chassis having improved circuits. In addition, complete subchassis were supplied in the kit of spare parts, so that in the event of failure and inability to find or remedy the trouble quickly, the spare chassis could be substituted and the inoperative chassis repaired at leisure or under more favorable conditions. This reduces service and maintenance difficulties considerably. Production was greatly facilitated by the division of labor among each subchassis, the main chassis, and the wrap-around.

The remainder of the receiver circuits and components such as power supply, AFC thyratrons, controls, connectors, blower, etc., are included on the main chassis and wrap-around case. The receiver is cooled by forced-air circulation. A motor-driven blower, mounted externally at the rear of the receiver, sucks out air that is pulled through an air filter mounted on the top cover. This keeps the receiver within the ambient temperature limits and provides dust-free air circulated around the tubes, transformer, power resistors, etc.

The receiver was designed with as close adherence to JAN specifications and approved components as possible. This included complete fungicide treatment and impregnation. In order to determine whether

[1] Further details concerning shielding and mechanical construction are contained in Chap. 10.

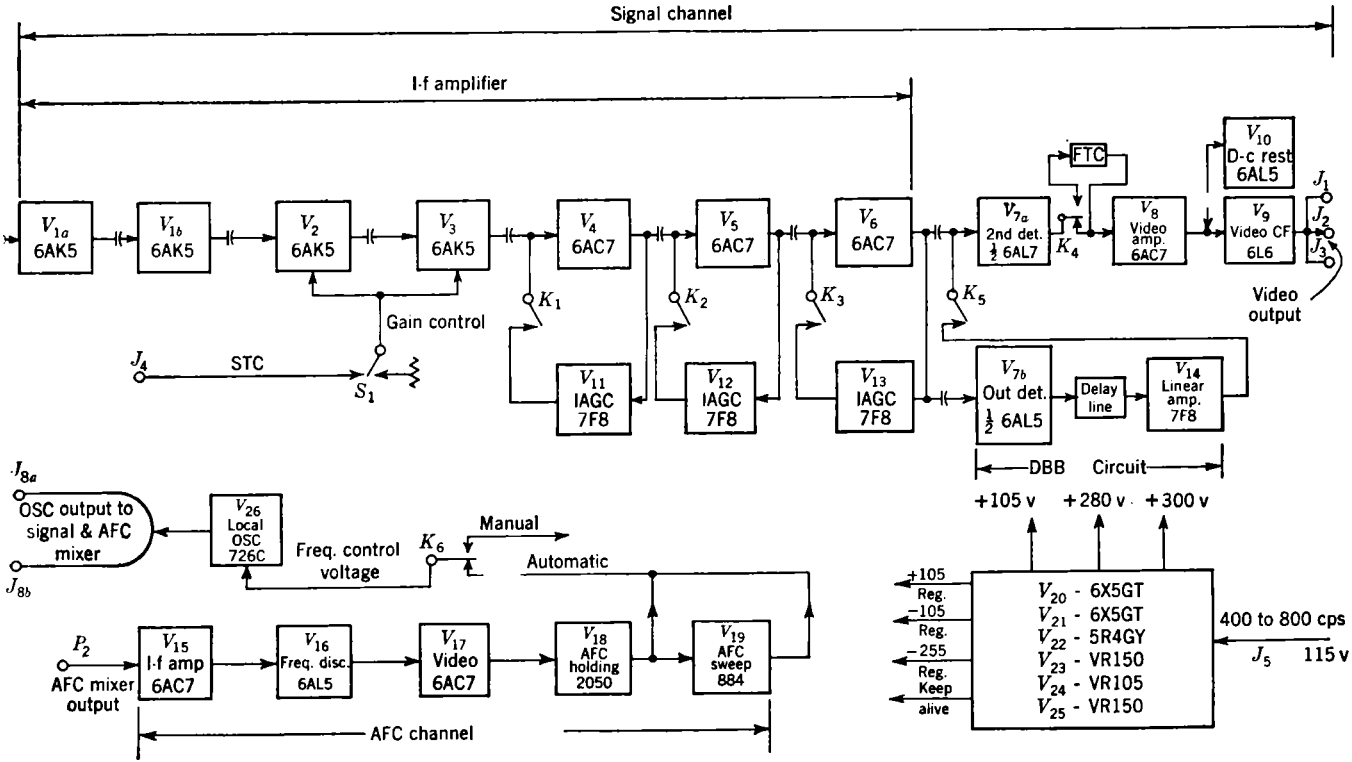


FIG. 14-3.—Functional block diagram of receiver.

or not these specifications were met, type tests were conducted.[1] The tests included operation of the receiver under altitude, temperature, humidity, and vibration conditions. The receiver performance was satisfactory in all cases.

Particular sections, circuits, and special features of the receiver are now discussed in more detail. A functional block diagram is shown in Fig. 14·3.

14·2. Local Oscillator.—The local oscillator used is the type 726C. This is a velocity-modulation tube with self-contained resonant cavity. It may be tuned mechanically by deformation of the cavity or electronic-

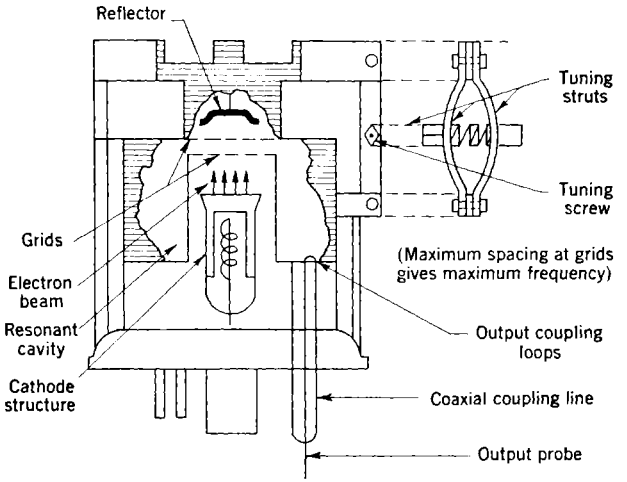


FIG. 14·4.—Simplified sectional view of 726C local oscillator.

ally by variation of the reflector voltage. A schematic diagram of the construction is shown in Fig. 14·4.

The tube will furnish at least 60 mw in the middle of its operating range, which runs from 10.1 to 11.2 cm. The electronic tuning coefficient is approximately 0.5 Mc/sec per volt. Power output is obtained by means of a small coupling loop located in the resonant cavity and connected to a coaxial line terminating in an output probe.

Several factors influenced the selection of the 726C as the local oscillator for this receiver. It is small, easily replaced, and mechanically tunable with a single knob. Outwardly, this tube appears much like an ordinary metal receiving tube, except for the tuning struts. One pin on the octal base is replaced by the coaxial r-f output probe. The corresponding socket hole must be enlarged for clearance of the splitter bush-

[1] RL Report No. 1160, Aug. 20, 1945.

ing into which the coaxial line and probe fits. Once the splitter and socket are assembled, the 726C merely plugs in and out like any ordinary tube. This makes replacement very simple as contrasted to the 2K28 which requires removal of its entire cavity assembly. A simple mechanical arrangement of a knob and insulated shaft, fitted with a retracing spring and socket, engages the tuning nut for expanding and contracting the cavity struts. The mounting of the tube and tuning assembly can be seen in the photograph (Fig. 14·5) of the receiver. The reflector connection is to the top cap. The heater, cathode, and cavity are connected to pins on the tube base.

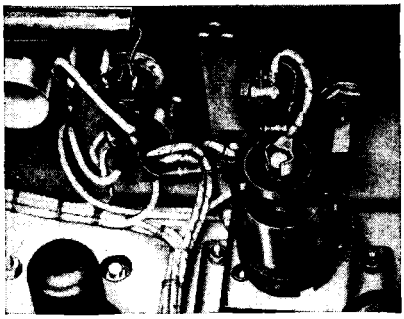
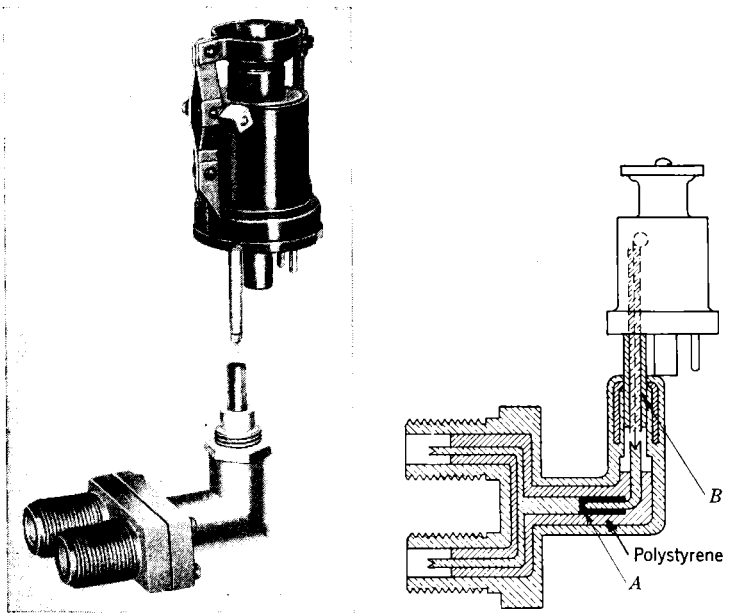


FIG. 14·5.—Detail of receiver showing local oscillator.



(*a*) (*b*)

FIG. 14·6.—Power splitter. (*a*) Photograph of tube and splitter; (*b*) diagram of internal construction.

14·3. 726C Splitter or Coupler.—The power splitter makes it possible to obtain LO power for two separate mixers simultaneously. Power

is fed to the mixers on coaxial cables of 50-ohm characteristic impedance, which are terminated at the mixers by means of resistor disks. At the operating frequency of this receiver, the optimum load for the oscillator is from 22 to 25 ohms. The dimensions of the splitter are so chosen as to provide the small amount of impedance transformation required. Figure 14·6*a* shows a photograph of the tube and splitter, and Fig. 14·6*b* indicates the internal construction. Insulation for direct current must be provided on both the inner and outer conductors of the coaxial line coming from the oscillator, since both are at the potential of the cavity, +300 volts with respect to ground. The quarter-wave choke joints, labeled *A* and *B* in Fig. 14·6*b*, provide this insulation.

For best operation of the AFC system it is desirable that coupling of signal frequency from one mixer to the other by way of the local oscillator be minimized (see Chap. 3). This may be achieved to some extent by proper choice of the length of the section of the splitter between the oscillator and the branching point. Since the oscillator cavity is tuned to a frequency removed from the signal frequency by the intermediate frequency, the impedance seen at the coupling loop for signal frequency will be low. If the electrical length from the coupling loop to the branching point is some multiple of a half wavelength, the impedance at the branching point will also be low and will reduce power transmission from one mixer to the other.

The characteristics of the 726C are such that the optimum load impedance varies quite rapidly with frequency. This makes it difficult, if not impossible, to design a splitter that will permit effective operation over a frequency band much greater than 40 Mc/sec. As far as the splitter itself is concerned, operation should be satisfactory over a much wider band.

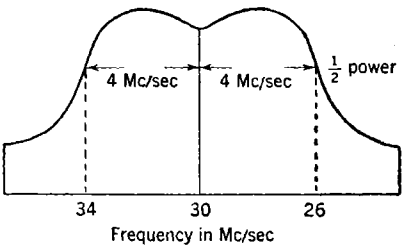


FIG. 14·7.—Bandpass characteristic.

14·4. Input Transformer.—The principles governing the design of the i-f input circuit to the signal channel involving L_1 and T_1 (see Figs. 14·8 and 14·10) have been covered in Chap. 4. The desired response characteristic for optimum noise figure (see Fig. 14·7) has a width of approximately 8 Mc/sec between half-power points. The mutual coupling (in this case inductive) between primary and secondary of T_1 is a little more than transitional, so that there is a small dip of not more than 1 db at the midfrequency.

Both L_1 and T_1 are wound on small ceramic coil forms of the type shown in Fig. 10·23*c*. A movable copper slug permits variation of induct-

ance. The secondary of T_1 consists of a 23-turn single-layer winding. The primary of four turns is wound over the ground end of the secondary. The coil L_1 has six turns.

An extra complication is brought about by the necessity of reading mixer-crystal current and of furnishing a bias voltage to the mixer during operation. If the low-potential end of the input transformer could be grounded directly with a short lead inside the chassis, it would be quite well shielded and isolated. Since it must be brought out on a lead that winds about the chassis considerably, a two-element filter is required between the point where it enters the i-f chassis and the low-potential end of the primary coil on T_1. One element is a self-resonant 30-Mc/sec choke L_2, and the other element is a 100-ohm resistor R_1; each element is bypassed on both sides of 0.001-μf condensers C_{1a}, C_{1b}, and C_{1c}. The inductance in such a filter must be heavily damped to prevent undesirable "ringing" after the transmitted pulse. It is not feasible to use several 100-ohm resistive filter elements because this would insert sufficient resistance in the crystal current lead to bias the crystal excessively and hence reduce the sensitivity. A similar two-stage filter is inserted in the crystal current lead for the AFC input circuit.

It is extremely important that the ground return leads of the input circuits be isolated and have as near zero inductance as possible. In order to accomplish this, the initial ground return for the primary circuit C_{1a} functions as an r-f short circuit and *must* be returned with *very* short leads directly to the point where the i-f input cable enters the subchassis. Grounding this first bypass condenser at some other chassis point introduces a small amount of inductance which is a source of pickup and common coupling between the i-f input circuit and other stray i-f currents. Owing to the high receiver gain, this results in instability and often regeneration to the point of oscillation. It cannot be emphasized too strongly that *provision of proper ground paths for the input circuits are of utmost importance.*

14·5. I-f—AFC Chassis. *I-f Channel.*—This section is composed of two channels, the signal i-f channel and the AFC channel. As these channels function independently, the i-f signal channel will be dealt with first. This consists of four 6AK5 i-f amplifier stages (see Figs. 14·8 and 14·9) of which the first two stages (V_{1a},$V_{1'}$) are operated as triodes (screen tied to plate) and the following two stages (V_2,V_3) as normal pentodes. The input stage operates as a grounded-cathode triode, and the succeeding stage operates with a grounded grid. The reason for using two triodes in this manner is to obtain the low noise figure and consequent higher sensitivity or signal-to-noise ratio that this circuit provides. The theory of this circuit is treated in Chap. 4. This low-noise circuit was not included in the original design, which consisted of three 6AC7 pentode

stages. The noise figure of the complete i-f signal channel was 4 db. After replacing the three 6AC7 stages by the four 6AK5 low-noise input stages, the receiver i-f noise figure is reduced to 2 db. This produces as much improvement in the performance of the radar system as an increase in transmitted power of 50 per cent.

The remaining two i-f stages in this chassis section are standard pentode 6AK5 amplifiers with single-tuned circuits tuned to 30 Mc/sec by adjustment of L_5 and L_7. Any drop in plate voltage due to the plate-load

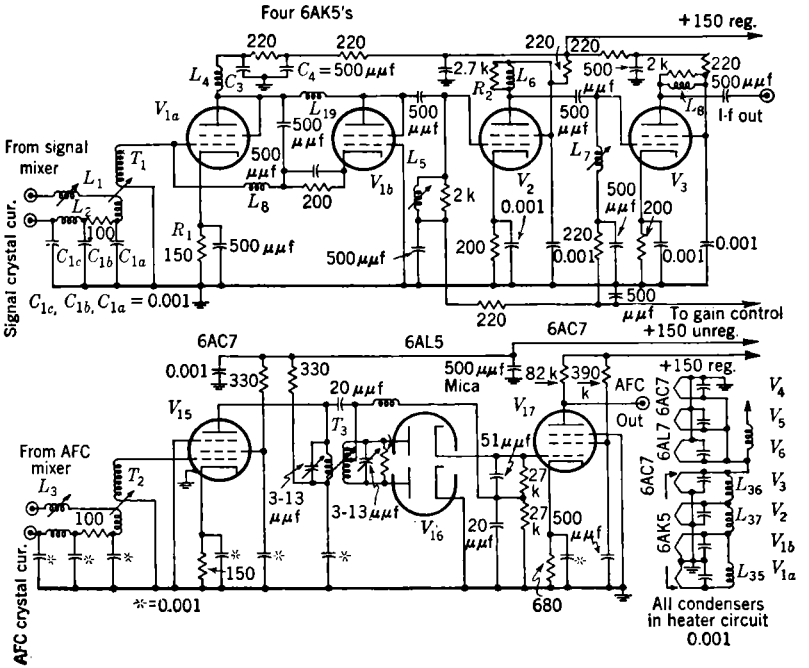


FIG. 14·8.—Circuit of i-f—AFC chassis.

resistor is prevented by a parallel 30-Mc/sec self-resonant choke wound on the resistor as L_6 on R_2. Note the use of button mica feedthrough condensers such as C_3 and C_4 that bypass the supply-voltage leads. High-dielectric-constant ceramic condensers, because of their small size, are used universally throughout the receiver for i-f bypass purposes on screens, heaters, etc., where voltages do not exceed 150 volts and where high Q is not important. Small molded bakelite chokes are used in the heater leads in a progressive filter arrangement. These chokes have low inductance but provide sufficient reactance at 30 Mc/sec to isolate the heaters satisfactorily. Since the 6AK5 is a miniature tube, the parts and

circuit elements must be as small as possible in order to be supported directly on the small socket pins and to fit in a restricted space.

Gain Control.—A switch (S_1, Fig. 14·10) permits control of receiver gain by either a local control in the receiver or a remote control in the control indicator. Ordinarily gain-control voltage would be applied to the second i-f stage (see Chap. 9). However, in this case the gain of the two triode stages combined is only about that of a single stage operated as a pentode, so the following two pentode stages V_2 and V_3 are con-

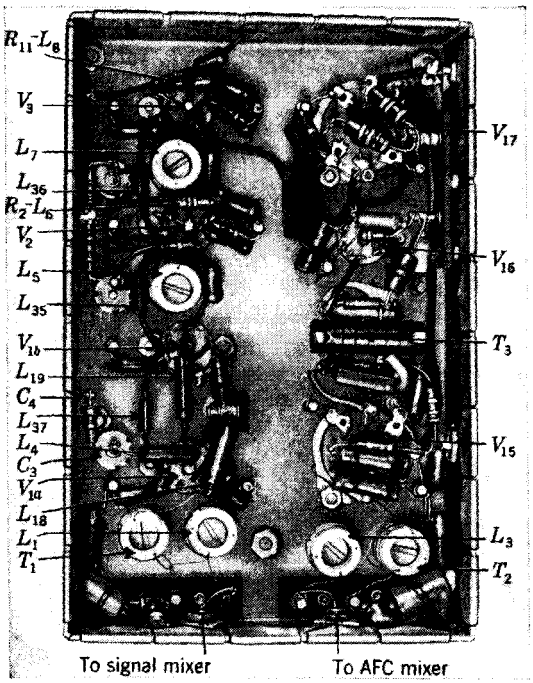


FIG. 14·9.—Photograph of i-f—AFC chassis—bottom view. Part designations are the same as in Figs. 14·8 and 14·10.

trolled. It is important that the grid-to-ground resistance be kept as low as possible. The gain-control lead contains a resistance-capacity filter whose *RC* time constant must be kept as short as possible while providing i-f decoupling. The gain control itself also enters into this time constant and must also be kept low in resistance. If this is not done, large signals of some duration may cause grid current or changes in d-c level which build up on the circuit capacities and block the tube so that it does not recover soon enough after the signals cease. This causes a "dead" time during which the receiver does not respond to weak signals. Refer-

ence to Fig. 14·10 shows that the decoupling resistors R_5, R_9, R_{36} are quite small and that the gain-control lead can never have more than 1800

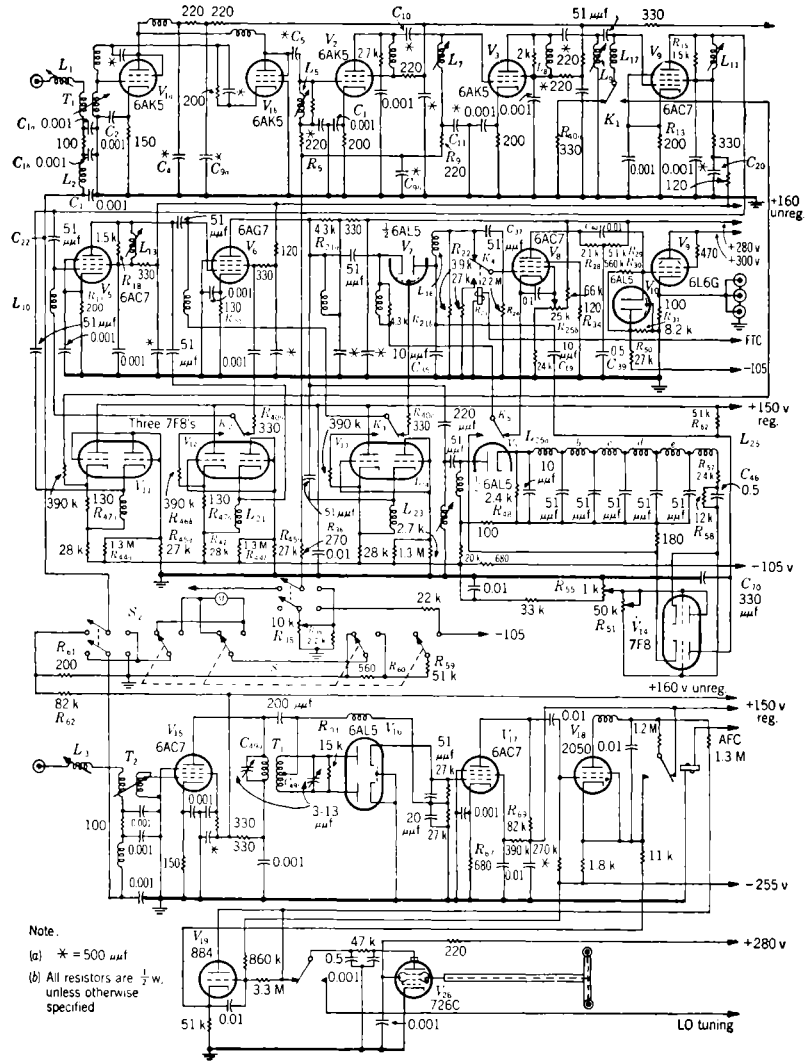


FIG. 14·10.- Circuit of receiver.

ohms to ground, since the 10,000-ohm gain-control potentiometer R_{38} is shunted by the 2200-ohm resistor R_{35}. With the bypass filter con-

densers C_9, C_6, C_{11}, the RC constant is sufficiently short.[1] This assumes that the gain is set at minimum where all the potentiometer resistance is in the circuit. At higher, more normal gain settings, less resistance is used in the gain control and the shunt resistance of R_{35} and R_{38} becomes much lower, thereby producing an even shorter time constant and more rapid recovery.

AFC Channel.—The AFC channel in the i-f—AFC chassis contains an i-f input circuit very similar to the signal i-f input stage except that a 6AC7 is used. Owing to its higher input capacity, the secondary of T_2 has fewer turns in order to resonate at 30 Mc/sec. The pass band should be adjusted for transitional coupling which in this case results in a bandwidth of about 10 Mc/sec. The bandwidth of the input circuit is not governed by noise or sensitivity considerations, since the AFC circuits operate directly from the strong local transmitted pulse converted to 30 Mc/sec. The bandwidth is made quite wide because the shape of the bandpass characteristic will then change very little with variations in circuit constants ordinarily caused by changing tubes, crystals, or mixers. Once properly aligned, the input circuit should require no further adjustments unless L_3 or T_2 requires replacement.

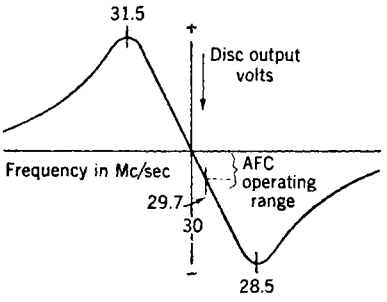


FIG. 14·11.—Discriminator response.

This i-f stage is followed by a standard discriminator circuit commonly used for AFC. The bandwidth of the discriminator characteristic is approximately 3 Mc/sec from peak to peak as illustrated in Fig. 14·11 in which output voltage is plotted against frequency. Loading the secondary of the discriminator transformer T_3 by resistor R_{94} makes the response linear between the peaks. The discriminator output is direct-coupled to the video amplifier. Proper discriminator symmetry and crossover frequency are obtained by adjustment of C_{49a} and C_{49b} which tune the primary and secondary, respectively, of T_3.

Since the local oscillator operates at a higher frequency than the transmitter and normally sweeps from a higher to a lower frequency, the difference frequency resulting will also vary from a frequency higher than 30 Mc/sec to one lower than 30 Mc/sec. Inspection of the discriminator characteristic shows that the resulting output voltage of V_{16} will be first positive as the frequency approaches 30 Mc/sec, zero at 30 Mc/sec, and finally negative as the frequency becomes less than

[1] Approximately 4 μsec. To be strictly correct the capacity contributed by the coupled circuits through C_5 and C_{10} should also be taken into consideration.

30 Mc/sec. It is the negative voltage output that operates the AFC circuits after amplification and inversion by the video stage V_{17} to a positive pulse, which triggers the 2050 control tube and locks the 884 sweep tube.

The video amplifier is operated as an "on" tube, that is, with a large voltage drop across the plate load R_{69} and little applied bias so that a negative pulse on the grid will bias the tube toward cutoff and allow the plate to rise toward the supply voltage. Sufficient bias is maintained on the tube by the cathode resistor R_{67} to prevent grid current. The sensitivity of control of the LO frequency is dependent to a large degree on the gain of the video stage. If the gain is large, only a small negative input voltage is required to trigger the thyratron circuits and lock the oscillator at the correct frequency. The supply voltage to plate and screen of this video stage must be well filtered and regulated in order to prevent any pulses or power-supply ripple from triggering the 2050 tube, which may occur if these effects are allowed to exceed 1.5 or 2 volts.

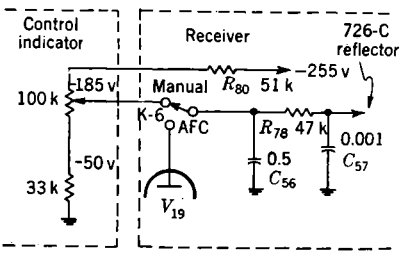


FIG. 14·12.—Circuit for manual tuning of the oscillator.

The thyratron control circuits are external to this chassis section. Since they are very similar to the circuits described in detail in Chap. 3, they will not be described further here. The circuit arrangement for manual tuning of the oscillator is shown in Fig. 14·12.

14·6. I-f—Clutter Chassis.—The second subchassis i-f section contains the three remaining i-f stages V_4, V_5, V_6, which are quite similar, and the second detector V_7. Figure 14·13 shows a circuit diagram of the chassis. The last stage V_6 is a 6AG7 operated at a plate potential of 280 volts. This is done to increase the maximum output of the i-f amplifier under overload conditions. The preceding amplifiers utilize 6AC7 tubes (V_4,V_5), and all three i-f stages are single-tuned circuits adjusted to 30 Mc/sec by inductances L_9, L_{11}, L_{13}, and L_{23}. The bandwidth of the complete i-f channel is between 1 and 1.5 Mc/sec and is determined by the loading resistors, which in this section are R_{15}, R_{18}, and the parallel combination of R_{21a}, R_{21b}.

The three back-bias circuits are closely associated with the last three i-f stages, as inspection of Fig. 14·13 will show. V_{11} operates to bias back V_4; V_{12} and V_{13} operate similarly on V_5 and V_6, respectively. The three IAGC circuits are put into operation simultaneously by means of three interconnected relays remotely operated by a switch at the control indicator. These relays switch the grid returns of the last three i-f

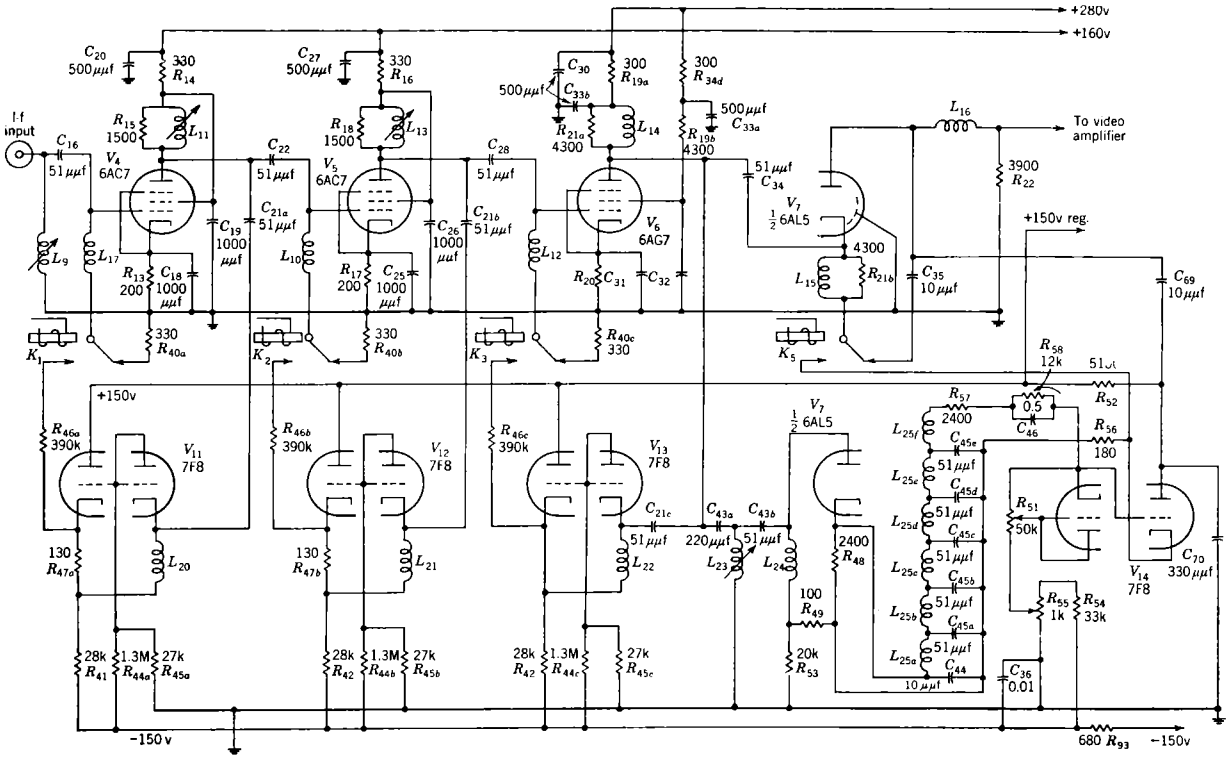


FIG. 14·13.—I-f—clutter chassis circuit.

grid chokes from ground to the cathodes of the IAGC stages. For normal operation, resistors R_{40a}, R_{40b}, and R_{40c} are connected in each grid return to ground to prevent the possibility of oscillation. The three IAGC circuits each utilize a 7F8 tube.

The second detector V_7 is one section of a dual-diode 6AL5. This detector is connected for negative output and converts the amplified i-f

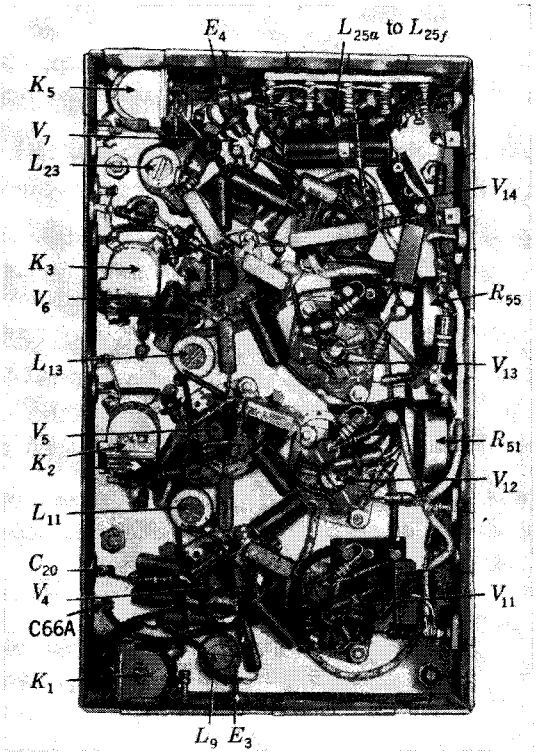


FIG. 14·14.—Photograph of i-f—clutter chassis—bottom view. V_4, V_5, V_6, i-f amplifiers; V_7, dual detector for signal channel and DBB circuit; V_{11}, V_{12}, V_{13}, IAGC or back-bias stages; V_{14}, DBB anticlutter stage; C_{20}, C_{66a}, button feedthrough condensers; E_3, E_4, insulated Kovar feedthrough terminals; K_1, K_2, K_3, IAGC relays; K_5, DBB relay; L_9, L_{11}, L_{13}, L_{23}, 30-Mc/sec tuning coils; L_{25a-f}, delay line; R_{51}, R_{55}, potentiometers for adjusting DBB operation.

signals into video signals for the grid of the video limiter amplifier V_8. The low value of load resistor R_{22} necessary to preserve good pulse response requires that the internal resistance and capacities of the diode be as low as possible in order that most of the signal be developed across the load. The 6AL5 is one of the better diodes suitable for such operation. Filtering for intermediate frequency is provided by L_{16} and C_{35}.

Direct coupling to V_8 is normally used. Direct coupling keeps the limit level of V_8 more constant and eliminates the overshoot or tail that often occurs after a signal pulse of considerable duration when condenser (a-c) coupling is used. For normal operation this condenser would have to be large to prevent differentiation or change in the average voltage across the coupling condenser during a long block of signals. Under certain conditions involving heavy clutter interference, a short-time-constant a-c coupling circuit is advisable and will be described subsequently. For this reason provision has been made for inserting a fast-time-constant coupling circuit of approximately 1.5 μsec by means of an independently operated relay K_4.

The remaining section of the 6AL5 diode V_7 is also used as a detector of the same i-f signal, but its output is applied to the DBB clutter circuit. This DBB detector circuit furnishes signals of positive polarity to a delay line L_{25af} followed by a combined linear amplifier and d-c restorer V_{14}. These amplified signals are then applied to the normal detector to reduce its sensitivity after a delay of $\frac{1}{2}$ μsec or so, as determined by the delay line constants.

The complexity encountered when extensive auxiliary control circuits are combined with the i-f channel entails considerable care in design. Components must be situated close to the circuits with which they are associated, and precautions taken to keep leads short and yet retain as much isolation and reduction of stray coupling as possible. The assembly layout of this chassis in Fig. 14·14 shows in general how this was accomplished. Button mica feedthrough condensers, small ceramic condensers, and self-resonant heater chokes are also used in this chassis. It was found necessary to connect the suppressors of V_4 and V_5 to their cathodes instead of to ground because some 6AC7 tubes were not entirely stable in the more normal arrangement for reasons that are not clear. However, all operated with stability when connected as shown. The cathode resistors R_{13}, R_{17}, R_{20} in the controlled i-f stages should be as high as posssible, without sacrificing gain, in order to provide sufficient d-c degeneration in case the idling back-bias voltage supplied from the cathodes of the IAGC tubes V_{11}, V_{12}, V_{13} are slightly positive. This occurs with some 7F8 tubes as a result of allowable variation in their characteristics. A more detailed description of the IAGC, DBB, and FTC circuits for discrimination against undesired signals is given later.

14·7. Video-amplifier Limiter.—The signal-channel second detector supplies negative video signals to the 6AC7 video amplifier V_8 (see Fig. 14·10). This tube serves as a combined amplifier-limiter as described in Chap. 8. The total voltage drop across R_{28} with no signal is the maximum positive signal available to drive the cathode follower, since any signal applied to the limiter grid greater than the voltage required to cut

the tube off will only maintain the plate current at zero and produce no additional signal output. By controlling the plate current or voltage across R_{28}, the amplitude of the output signal may be adjusted. This is done by adjustment of the screen voltage potentiometer R_{25b}, which varies the plate current over a restricted range. By this means it is possible to set the cathode-follower output signal to its normal level of 6 volts.

Since this receiver is employed with a high-powered radar system, it is capable of receiving echo signals at ranges up to 200 miles when operating with high gain. Under these conditions large blocks of signals may produce limiting output for a period of 1000 μsec, equivalent to 100 miles or more. If the low-frequency response is not sufficiently good, a 100-mile block of signals will not hold up but will gradually fall off in amplitude. When the signal ceases, this may result in a dead time (overshoot) during which a weak signal, if present, would be undetected. Some low-frequency compensation is therefore included after standard practice by the inclusion of C_{39} and R_{29}. Increasing C_{40} would be desirable, but video oscillation or "motor-boating" frequently occurs. The R_{29}-C_{39} compensation functions as if C_{40} were made larger. The suppressor resistor R_{34} is necessary to prevent a parasitic oscillation which occurs with some tubes under pulse conditions.

14·8. Cathode-follower D-c Restorer.—The chief function of the 6L6GA cathode follower V_9 is to act as an impedance-matching device from its high-impedance input to a low-impedance output for driving a low-impedance (100-ohm) cable. Care must be exercised to ensure that the tube ratings are not exceeded. The tube is rated for an absolute maximum of 21 watts dissipation as a triode. The no-signal (idling) current for the tube is 40 ma with -24 volts of fixed bias provided by the bleeder R_{31} and R_{50} (Fig. 14·10). The drop across the cathode 100-ohm load resistor is 4 volts. In order to obtain the 6 volts of limited output signal, the current must be increased 60 ma through the cathode resistor. This means that the peak-current value is 100 ma during the full 6 volts of output signal.

The plate and screen are at $+300$ volts so that the peak power is approximately 30 watts. However, the average power is less than this value, since signals do not stay at saturation or peak current level for the whole elapsed time between each recurrent pulse. With solid land-echo signals, a saturation range of 160 miles may be obtained. The elapsed time between pulses is $\frac{1}{300}$ sec, or 3333 μsec, while a solid 160-mile signal return represents about 1600 μsec. The duty cycle is therefore the ratio of the "on time" to the total time, or 1600/3333, which is about 48 per cent duty cycle.

Thus the average power is 0.48×30, or 14.5 watts. In addition to

the signal power, a certain amount of noise power is also present. For normal gain settings with some of the noise peaks limiting, a good figure for the average noise power is 20 per cent of the peak power, so 0.2×30 watts gives 6 watts of noise power. This, added to the signal power of 14.5 watts, gives a total dissipation of 20.5 watts under fairly severe conditions. Usually the total dissipation will be below this value.

The +300 plate voltage of the cathode follower is secured from the center tap of the power-supply filter L_{34a}, L_{34b} (see Fig. 14·19). Due to the high peak currents of the 6L6GA during signal response, the voltage output from the power supply will be lowered if the plate voltage is obtained from the output of the filter at C_{63a}. This condenser acts mainly as a reservoir condenser, and its voltage will vary with high peak currents of some duration which are also reflected at the same time on V_6, V_8, and the local oscillator V_{26}. This voltage drop produces several undesired effects. The output level of the receiver does not stay constant, and low-frequency motor-boating occurs because of feedback through the common power-supply impedance to other stages. In addition, the accelerating voltage of the 726C local oscillator will become modulated during motor-boating or lowered in voltage during long blocks of signal. This tends to change its operating characteristics and shift its frequency, as well as reduce its power output by changing its operating frequency. These effects are quite serious and are avoided by supplying the plate voltage for the 6L6GA from the midpoint of the filter. Even if the voltage of C_{63b} varies somewhat during signal response, the output voltage at C_{63a} is quite constant due to the isolation afforded by the filter section L_{34a}.

The d-c restorer V_{10} is a 6AL5 diode. Its purpose is to allow only positive polarity signals to operate the grid of V_9 and to maintain a fixed grid operating voltage. It is so connected that for any negative signal or charge on the grid of V_9, the 6AL5 conducts and acts like a short circuit. The low diode resistance of approximately 50 ohms allows the grid to recover quickly, since the time constant of the grid and coupling circuits is very short during conduction. The grid resistor R_{30} is 0.5 megohm and provides the high-resistance input during normal positive signal response.

14·9. Crystal-metering Circuit.—A 0- to 1-ma meter is provided on the receiver for measuring crystal current in either the signal mixer or the AFC mixer in order to adjust the LO power for 0.3- to 0.4-ma crystal current which is the normal operating value. A spring-return double-pole double-throw switch (S_2, Fig. 14·10) is arranged for connecting either crystal to the meter circuit. Normally, the spring return keeps the meter in the AFC crystal circuit, and the switch must be held in the other position to measure the signal crystal current. In normal operation the cir-

cuit is arranged to provide a small positive bias, established by the bleeder R_{61}, R_{62}, to the signal crystal. This bias improves the noise figure of the receiver slightly due to a small reduction in the crystal conversion loss.

The switch S_3 is a four-pole double-throw switch which reverses the meter connections and provides a suitable negative potential for the crystal to measure the "back current." This potential (1 volt) is supplied by the bleeder R_{59}, R_{60}. The "front-to-back" ratio of crystal resistance is a fair indication of good or bad crystals. The forward resistance is usually low and fairly constant, 200 to 400 ohms. The back resistance is quite high and may vary considerably. A high ratio of back-to-front resistance is desirable, and crystals with a low ratio usually do not perform well. If the back resistance of a crystal becomes too low (less than 6000 ohms), it should be replaced. This is indicated on the meter by a reading of 0.15 ma, and crystals in either mixer that show a back current exceeding this value should be discarded. This provides a quick and convenient method of testing the crystals without removing them from the circuit.

14·10. Signal Discrimination Circuits.—The receiver contains some of the most recent circuit features for signal discrimination. Any undesired signals encountered during the reception of echo signals may produce interference. This interference may come from several sources, such as

1. Energy reflected back from the surface of water, commonly known as "sea return."
2. Concentrated land signals as hills, buildings, etc.
3. Dense cloud regions.

Four types of circuits have been found useful. Almost always two and frequently three of these circuits may be used simultaneously to advantage. This is, in part, because the dynamic range for optimum operation of these circuits is limited and additional circuits may be needed to help carry the load. Then, too, a circuit that is used to prevent video saturation and that operates on the last i-f stage or later cannot prevent saturation in the preceding i-f stages; hence, other circuits are needed.

The position of the four circuits IAGC, DBB, FTC, and STC are shown in the block diagram Fig. 14·3. The first three of these circuits are contained in the receiver proper. The STC circuit operates on the receiver, but its circuit elements are located in a separate unit.

14·11. The IAGC Circuits.—A diagram of a single i-f amplifier stage V_5 and its associated IAGC circuits V_{12} is presented in Fig. 14·15 to aid in describing IAGC operation. This is identical with the circuit included in the complete schematic diagram of Fig. 14·10. Symbol designations are also identical.

The IAGC or automatic back-bias circuits are designed to increase the negative grid bias of an i-f stage when the output of that stage gets

above some value determined by the circuit constants. This produces a decrease in gain and prevents saturation of the i-f stages so protected. Back-biasing is preferably applied to individual stages in the last half of the i-f channel. Amplification is needed in the back-bias circuits. This is accomplished together with least phase-shift difficulties (phase shift may cause oscillations) by a diode detector that is coupled to a cathode output amplifier that has zero phase shift.

Figure 14·15 shows that V_{12} is a twin triode 7F8 tube, the first section of which is operated as a diode detector (plate tied to grid) with negative output. The remaining triode section is directly coupled to the output of the diode section in what is known as a "bootstrap" circuit. The output of the i-f stage V_5 divides between the following stage in the i-f channel and its own back-bias circuit. The i-f voltage will, therefore, be developed across the self-resonant coil L_{21} and applied to the cathode of the diode section of V_{12}. The lower end of this coil is connected through the resistor R_{47} to the cathode of the amplifier section. Both cathodes of V_{12} are fed from a supply -105 volts through 28,000 ohms, and $+150$ volts is also applied to the plate of the triode section. The circuit is so designed that the cathode (Pin 5) of the triode section will be at ground potential when no signal is present. R_{44} and R_{45b}, 1.3 megohms and 27,000 ohms, respectively, form a voltage divider between -105 volts and ground that holds the grid (Pin 8) of the amplifier section at about -2 volts. This value of grid bias corresponds to a triode cathode current of approximately 3.5 ma which produces a drop through the 28,000-ohm cathode resistor R_{42} of about 100 volts and makes the cathode stand at ground potential ± 0.7 volt. The resistor R_{45b} also serves as the diode load.

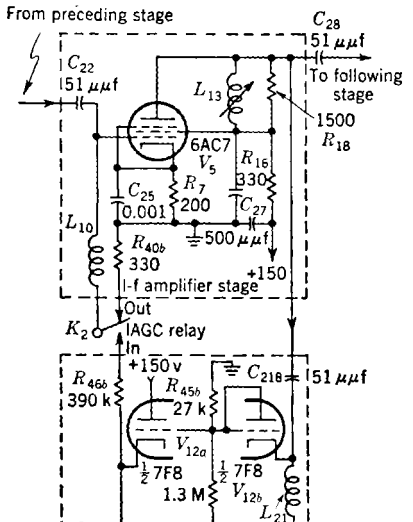


FIG. 14·15.—I-f and IAGC circuits.

When a 30-Mc/sec i-f signal is fed to the cathode (Pin 4) of the diode section, the tube becomes conducting and produces a negative output signal. Since this negative output is coupled directly to the grid (Pin 8)

of the triode section, this grid also becomes negative and is closely followed by its cathode which also feeds back to the diode cathode to make it more negative. The whole circuit is, therefore, a positive-feedback bootstrap arrangement, a given signal tending to reinforce itself. The primary functions of R_{47b} are to provide a small amount of delay bias and to control the amount of feedback. Obviously the back-bias voltage should not operate to bias off the i-f stage until the signal reaches a certain level; otherwise the IAGC circuit would immediately begin to back-bias the i-f stage as soon as signals are present and reduce the gain and output too soon. The 6AG7 in the last i-f stage, because of its larger dynamic range and greater grid cutoff characteristic, requires less delay and more gain in its IAGC circuit. For this reason the resistor equivalent to R_{47b} (Fig. 14·10) is omitted from the cathode of its back-bias amplifier V_{13}.

The output of the IAGC circuit is taken from the cathode (Pin 5) of the 7F8 back-bias amplifier. With increasing signal strength this cathode becomes more and more negative. This negative voltage is fed back to the control grid of the i-f stage V_5, when the relay is in the proper position, by way of R_{46b} and L_{10}. The resistor R_{46b} (360,000 ohms) and the i-f coupling condenser C_{22} (51 $\mu\mu$f) together with the associated wiring and tube capacitance result in a time constant for the circuit of about 30 μsec. The reduction of the i-f gain at the beginning of a very large signal and the recovery to normal gain on cessation are slowed up due to this time constant.

With this back-bias circuit arrangement it is, of course, impossible to cut off completely the i-f stage and reduce its gain to zero, for, in such a case, there would be no signal fed to the IAGC circuit to provide a back-bias voltage.

The voltage gain of a complete IAGC circuit from the diode input to the cathode output is about 20. The maximum back-bias voltage that can be developed, before saturation occurs due to a large c-w signal, is about -40 to -45 volts if this voltage is not allowed to bias the i-f stage and -8 to -10 volts when applied as back bias. It can be seen that the gain of the i-f stage for strong signals is greatly reduced; hence the IAGC circuit does reduce overloading and prevent i-f saturation by strong signals which otherwise would make discrimination of small signals impossible.

14·12. Detector-balance-bias (DBB) Circuit.—The DBB circuit is designed to bias the signal channel detector by an amount sufficient to keep the high-frequency components of clutter from saturating the video amplifier. To accomplish this, the output of the last i-f amplifier is divided between two detectors, the normal signal channel second detector with negative output and the DBB detector with positive output. This

positive output is delayed by a delay line and then fed to a positive-output linear amplifier, the gain of which is different for direct current than for high frequencies. The amplifier output is used to bias the normal signal detector an amount sufficient to remove the d-c term from c-w interference or an amount sufficient to keep any but the highest peaks of clutter from saturating the video. In general, the DBB circuit has a rather small dynamic range and must, therefore, be preceded by an i-f amplifier well protected by IAGC.

The DBB circuit operation is described with the aid of Fig. 14·16. The positive-output DBB detector operates into a $\frac{1}{2}$-μsec delay line

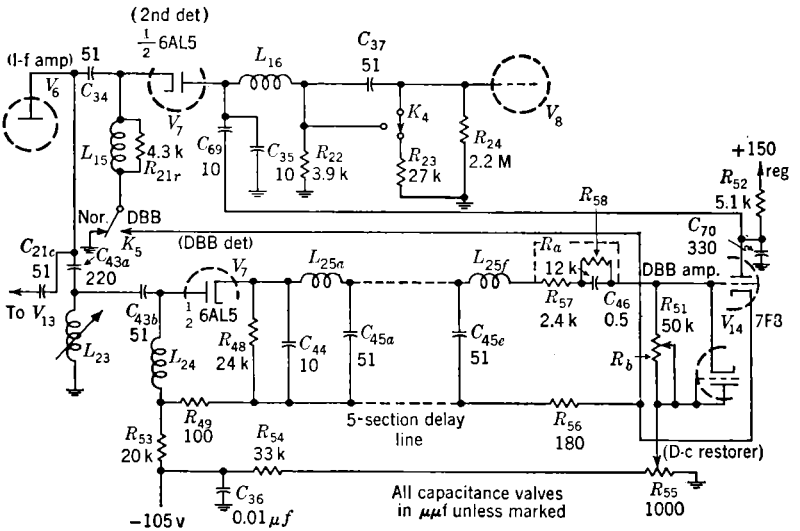


FIG. 14·16.—DBB signal-discriminating circuits.

L_{25af} of about 2400 ohms characteristic impedance. The output of the delay line is applied to the cathode-output amplifier section of V_{14} which feeds a positive voltage back to the cathode of the normal second detector when K_5 is in the DBB position. The remaining section of V_{14} is connected as a diode and employed for d-c restoration at the amplifier grid. R_{48} (2400 ohms) is a termination at the input of the delay line to suppress multiple reflections back and forth in the line, since the output termination is not close enough to absorb all the pulse energy.

The linear amplifier is coupled by a bootstrap circuit arrangement, and the gain of the tube alone is very nearly unity. It can be considered, therefore, that the grid and cathode maintain a fixed voltage difference; that is, they go up and down together. Resistors and voltages employed

in the circuit are such that the cathode of the amplifier is practically at ground potential with no signal present.

I-f signals of 30 Mc/sec applied to the DBB detector result in positive signals which are passed through the delay line with a $\frac{1}{2}$-μsec delay to the network of R_a consisting of R_{57} (2400 ohms), R_{58} and C_{46} in parallel (12,000 ohms and 0.5 μf), and R_b. The signal across R_b (R_{51} set at 40,000 ohms) is applied to the DBB amplifier. The voltage across R_b depends on the character of the 30-Mc/sec i-f signal. The network R_a, R_b has a frequency characteristic such that the output voltage is dependent on whether the i-f signal fed to the DBB detector consists of 30-Mc/sec pulses or c-w. For a given level of i-f signal the ratio of the voltage output of this network to that applied is R_b/R_a. For direct current (resulting from c-w signals) R_a is R_{57} plus R_{58}, and the voltage gain of the network is therefore

$$\mathcal{G}_{dc} = \frac{R_b}{R_a} = \frac{R_{51}}{R_{57} + R_{58}} = \frac{40{,}000}{2400 + 12{,}000} = 2.75 \qquad \text{(approximately)}.$$

The loss of gain through the DBB detector is approximately this same value (2.75); and since the amplifier tube gain is unity, the over-all DBB-circuit gain must also be unity for the case of c-w signals.

For 30-Mc/sec clutter input to the DBB detector the voltage gain of the network is modified because of the high-frequency components involved. R_a becomes only 2400 ohms, since C_{46} (0.5 μf) acts as a short circuit for high frequencies; hence, the pulse gain is

$$\mathcal{G}_{pulse} = \frac{R_b}{R_a} = \frac{40{,}000}{2400} = 16.6.$$

Again the DBB detector loss of 2.75 and unity gain of the DBB amplifier result in an over-all pulse gain in the DBB circuit of 16.6/2.75, or 6.

The delay line is composed of several cascaded sections in order to keep its physical size small and yet maintain a fairly uniform frequency response. Its delay characteristic should ordinarily be equivalent to one-half of the pulse length (τ) used in the radar system. Various other resistors are used in the DBB circuit to provide proper operating voltages. R_{51} is an adjustment for setting the gain in the DBB circuit and is normally adjusted to 40,000 ohms. R_{55} is an adjustment for compensating differences between individual 7F8 tubes after replacement. This potentiometer (1000 ohms) controls the amplifier grid voltage and is adjusted for each tube so that the voltage at the DBB amplifier cathode is +0.5 volt when no signal is present. This value balances out the normal second detector contact potential which averages about −0.3 to −0.7 volt. The d-c restorer is employed to prevent any overshoot, due to the discharge of C_{46} at the end of a long block of signals, from

causing the grid of the amplifier to become considerably negative. If this were allowed to happen, the high negative voltage applied to the second detector cathode would prevent its proper operation.

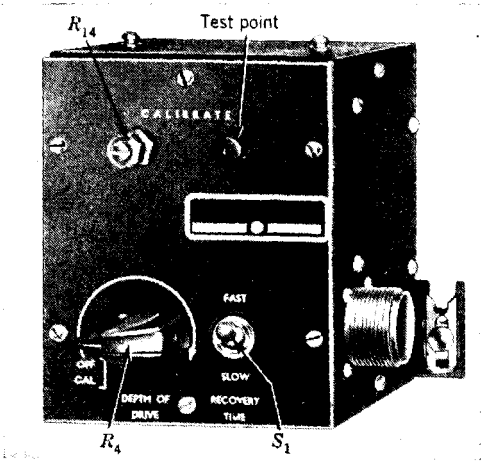
A small amount of pulse energy is supplied to the second detector output out of phase with the normal DBB feedback voltage. This video neutralization is required to balance out a spurious pulse produced when the DBB is operating under pulse signal conditions. When i-f pulse signals are supplied to the DBB detector, the resulting video pulses are fed back to the normal detector cathode circuit. Due to various circuit capacities and resistances, including the interelectrode capacity of the second detector, a differentiated video pulse is present at the output of the signal channel detector. This spurious pulse is canceled out by a pulse of opposite polarity obtained from the plate circuit of the DBB amplifier. The capacitor C_{70} and resistor R_{52} provide the same video response in the plate circuit as in the cathode of the amplifier, and the 10-$\mu\mu$f coupling capacitor C_{69} provides the correct amount of neutralizing voltage.

The DBB operated alone is not satisfactory but when supplemented by IAGC (with long time constant) is quite effective for removing clutter, such as clouds. The IAGC should always be used with the DBB circuits in operation.

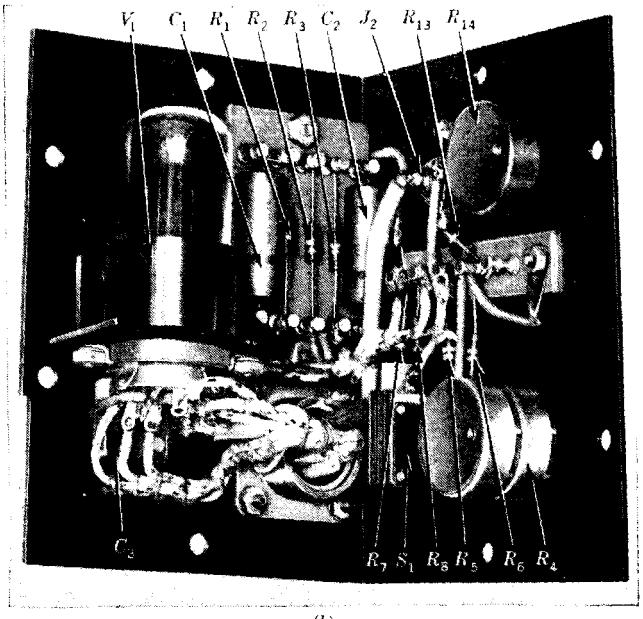
14·13. Fast-time-constant (FTC) Circuit.—The FTC circuit is an *RC*-filter inserted between the second detector and video amplifier. This filter acts as a high-pass coupling network with an *RC* time constant of about 1.5 μsec. Low frequencies or long blocks of signals will not pass through the filter. The time constant of the FTC circuit should be of the same order as the pulse length of the system.

Inspection of Fig. 14·10 shows the simplicity of the FTC coupling circuit. With the relay K_4 switched to FTC operation, the normal d-c coupling from detector to video grid is removed, and coupling is provided through the small 51-$\mu\mu$f capacitor C_{37}. The resistor R_{23} (27,000 ohms) is included in the video input from grid to ground across the 2.2-megohm resistor R_{24}. The low-frequency response is now determined by the 50-$\mu\mu$f capacity and the series resistance of R_{22}, R_{23}. Resistor R_{24} may be neglected, since it is large compared with R_{22} or R_{23}.

14·14. Sensitivity-time-control (STC) Circuit.—The ideal objective of the STC circuit is to vary the receiver gain in such a way that targets of the same effective area will appear on the PPI scopes with the same intensity, regardless of range. This, of course, cannot be fully realized in practice. Assuming a fixed amount of transmitted power and the transmission path to the target free of interfering objects, the received power reflected back from the target is proportional to the reciprocal of the fourth power of the range. This means that there is 625 times as



(*a*)



(*b*)

FIG. 14·17.—Sensitivity-time-control unit (for component designations refer to Fig. 14·18). (*a*) External view; (*b*) internal view.

much power returned to the receiver from a given target at 10-mile range as from the same target at 50 miles and 10,000 times as much as from the same target 100 miles away. If the receiver gain is adjusted sufficiently high to observe distant targets, serious overloading with accompanying loss of discrimination will result for near-by targets in the sea-clutter area. On the other hand if the receiver gain is adjusted for best results in the sea-clutter area, distant targets may not be visible at all.

The STC circuit automatically reduces the gain of the receiver to a suitable value immediately after the transmitted pulse for a time corresponding to the range of the heavy sea-clutter area. The gain is held

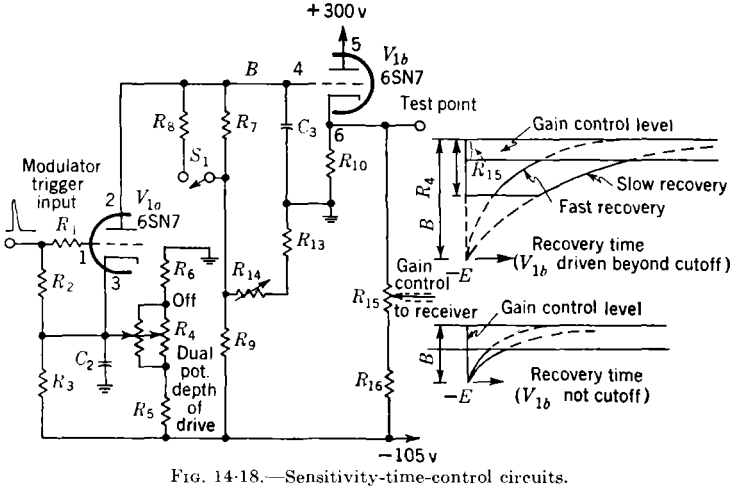


FIG. 14·18.—Sensitivity-time-control circuits.

constant for a period of about 100 to 500 μsec (depending on sea-clutter conditions) and then increases to a maximum value in accordance with a time constant in the STC circuit having one of two possible values. A double triode tube (6SN7) with associated resistors and capacitors constitute the STC circuit. This unit (shown in Fig. 14·17) is separate from the receiver proper and is connected to it by a suitable cable. This permits location of the unit where the controls are readily accessible to the operator. The STC circuit may be omitted, if desired, and manual gain used, in which case the gain is constant.

A schematic diagram of the STC circuit is shown in Fig. 14·18. A positive 150-volt trigger pulse which ordinarily triggers the transmitter is also applied to the STC circuit. This pulse applied to the grid (Pin 1) of the first 6SN7 triode section makes it highly conductive during the pulse and charges the 680-μμf capacitor C_3 to very nearly its cathode

potential. The cathode (Pin 3) potential is determined by the value of R_4 and is adjustable. This potentiometer adjustment controls the depth of drive, that is, the value of negative voltage to which C_3 can be charged. This is also the voltage applied to the grid of the second triode section. The rate at which the capacitor discharges and consequently the change of grid voltage at B will depend largely on the values of R_7 and R_8. The time constant with R_8 alone gives a slow recovery time, while R_7 placed in parallel with R_8 by S_1 produces a more rapid recovery.

With the depth of drive control R_4 in the OFF position, the cathode of V_{1b} should be at zero potential, and this calibration is made by setting the grid voltage by means of R_{14}. Under this condition gain is controlled manually in the normal way by the control R_{15}. When STC is applied by increasing the drive with R_4, the grid of V_{1b} and C_3 fall to a considerable negative value. The tube behaves as a cathode follower, and consequently a negative voltage output from the cathode (Pin 6) is developed in series with the receiver gain control R_{15}. The corresponding reduction in receiver gain conforms to the STC output waveform at point C as illustrated. This waveform indicates that the receiver gain may be reduced to some adjustable value, held constant for an adjustable length of time, and then recover to normal with a selected time constant.

The potentiometer R_4 controls the potential of the grid at point B after the trigger forces conduction in V_{1a}. This potential may be made large enough to cut off V_{1b}.

The resistor R_8 or R_8 and R_7 in parallel together with C_3 determines the recovery time of point B to its normal potential.

R_{10} determines the maximum flat height or the minimum receiver gain if point B is driven sufficiently negative to cut off V_{1b}. The output voltage at point C is then held constant during cutoff until point B is sufficiently recovered to allow the tube to conduct again. The duration of the flat is dependent on the setting of R_4 and also the recovery time constant. Due to the time constant, the potential will continue to rise exponentially to the normal gain value determined by the setting of R_{15}. It is evident that R_4 may be set so that the flat or cutoff point is not reached, in which case the recovery of receiver to normal gain begins immediately after cessation of the trigger pulse.

To summarize, the sensitivity-time-control circuit is triggered at each transmitted pulse, reducing the receiver gain to a predetermined value. This reduced gain may be held constant for a time corresponding to a range of about 10 to 50 miles, thereby preventing overload of the receiver due to the strong near-by or "close-in" sea reflections. The gain then rises to the maximum value in a time corresponding to a range of approximately 50 to 150 miles. Obviously, the sea-return echoes

decrease gradually with increasing range, hence the slow instead of abrupt recovery time. This device is extremely useful for detecting targets in an area of heavy sea clutter.

14·15. Power Supply.—A single transformer T_4 operating from a primary supply of 400 to 800 cycles at 120 volts provides all power used in the receiver. The high-voltage winding supplies a nominal voltage of 415-0-415 to a full-wave rectifier V_{22}, which provides positive plate voltages of +300, +280, +150 unregulated and +150 regulated by a VR150 tube V_{25}. These voltages are suitably filtered by a two-section

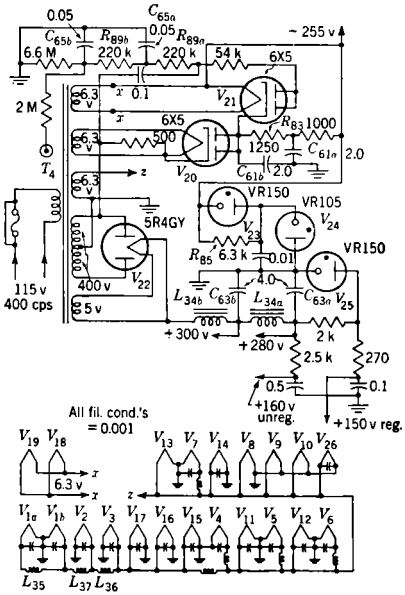


FIG. 14·19.—Circuit of power supply and heaters.

choke-input inductance capacity filter, L_{34a} and $_b$, C_{63a} and $_b$. The dual choke has 3 henrys per section (see Fig. 14·19).

A 5R4GY tube is used for the plate-supply rectifier only because other units of associated equipment employ a tube of the same type in similar power supplies. This reduces the number of different types of tubes required in the complete radar system. A 5U4G will operate quite satisfactorily and might otherwise be used.

A 6X5GT/G half-wave rectifier V_{20} provides negative voltages of −105 and −255 regulated by a VR105, V_{24}, and a VR150, V_{23}, connected in series. A resistance-capacity filter R_{83}, C_{61a}, C_{61b} is sufficient for this supply. The resistor R_{85} across V_{23} is necessary to balance the load on

the negative-voltage supplies so that the current through the VR tubes will remain within the correct operating range.

Another 6X5GT/G half-wave rectifier V_{21} with the resistance-capacity filter R_{89a} and $_b$, C_{65a} and $_b$ supplies a negative voltage of about -800 volts for keep-alive voltage on the external TR tube. This supply is purposely built with poor regulation, since it operates a low-current gas discharge. The voltage at J_6 with the TR tube connected is about -500 volts. The keep-alive rectifier circuit is somewhat unusual, since V_{21} operates as a half-wave voltage doubler but is also connected to the negative d-c output voltage of V_{20}. The d-c output from the bias supply is therefore added to the rectified output of the voltage-doubler circuit. This permits V_{21} to operate within its ratings and still furnish the required high negative voltage. It also reduces the number of tube types, since V_{20} and V_{21} are identical. Furthermore, the need for an additional filament and high-voltage transformer winding is obviated by this circuit.

Any changes of the -105-volt supply to the gain-control circuit or of the $+150$-volt supply on the screens of the first four i-f stages, such as those due to ripple or high peak signal currents, will appear as a modulation of the gain of the receiver and hence of the signal or noise level. Such an effect is undesirable and is eliminated by applying VR-tube regulation to these supply voltages:

When the receiver is operated under conditions of no-signal input, all relays unenergized, and the gain high enough so that receiver noise operates the 6L6GA at a 30 to 35 per cent duty cycle, the power-supply ripple voltages are as follows:

Supply voltage	Ripple voltage
+300 unregulated (6L6GA only)	0.90
+280	0.50
+150	0.30
+150 regulated	0.20
−105	0.25
−255	0.60

The power transformer (T_4) ratings are as follows:

	Volts	Amperes
Pri.	120/115	1.93
Sec. 1	415-0-415	0.220
Sec. 2	6.3	10.
Sec. 3	6.3	1.8
Sec. 4	5.0	3.0
Sec. 5	6.3	0.6

CHAPTER 15

A RECEIVER FOR AN AUTOMATIC-TRACKING RADAR

By C. E. Ingalls

15·1. General Requirements and Description of System.—The radar system of which the receiver to be described in this chapter is a part is a moderately complex shipborne system operating in a frequency range around 10,000 Mc/sec. The system performs automatic tracking of single target objects in azimuth, elevation, and range. The high degree of accuracy required in all three of these quantities places rather extraordinary demands on the design of the component parts, the receiver in particular.

The system operates with a transmitter pulse power of 60 kw and a pulse length of 0.1 μsec. Thus the total length of the pulse corresponds to a range increment of only 49 ft, and resolution of close targets is correspondingly good. It has been found possible to measure the range of individual targets with an average error corresponding to one-fifth to one-tenth the pulse length, or 5 to 10 ft. Such accuracy is maintained nearly to the maximum range, 30,000 yd, at which the system is required to lock on to a target.

Tracking in azimuth and elevation is made possible by the use of a nutating antenna. A fixed paraboloidal reflector is combined with a dipole, which is caused to move in a small circular orbit about the focus of the reflector, to give a radiation pattern in the form of a beam that traces out a small cone centered on the axis of the reflector. The dipole moves in such a manner as to remain parallel to a fixed direction, so that the plane of polarization of the emitted radiation does not change. This general arrangement has come to be known as a *conical scan.* The conical-scanning rate in the system under discussion is about 30 cps. It will be seen that if the reflector is pointed directly at the target of interest, the beam will move around the target, never pointing directly at it. The amplitude of echo signals from the target will not change as the antenna scans because the beam at all times makes a fixed angle with the radius vector to the target. If, however, the reflector is not pointing exactly at the target, the beam will approach nearer to the target in one part of the scanning cycle. Consequently, the amplitude of the echo signals will vary at the scanning rate. This variation may be extracted and compared in phase with a reference voltage generated by the motion

of the antenna to give information concerning the magnitude and sense of the error in the direction of the reflector. This information may then be applied to a servomechanism that positions the reflector so as to cause it to point directly at the target.

If this tracking method is to operate properly, it is vital that any amplitude modulation of the receiver output properly ascribable to the conical scanning be preserved and that any extraneous modulation at the same or near-by frequencies be eliminated. Therefore good AGC is an essential feature of the receiver. Furthermore, if the reflector is to be positioned accurately and quickly, phase delay of the 30-cycle modulation frequency must be maintained small and constant. This requirement placed additional limitations on the AGC circuits.

Extraneous modulation may arise in two principal ways: (1) random amplitude changes resulting from change in the transmission path or change in the aspect of the target and (2) spurious modulation at the scanning rate introduced by the r-f system. Signal fluctuations arising as in (1) may be broadly classed as *fading*. Oscillographic studies have shown that there may be appreciable frequency components as high as 5 cps due to fading. If these components are to be eliminated, a lower limit must be set on the speed of action of the AGC circuit. Fluctuations of Type 2 are most likely to be caused by imperfections in the waveguide or rotating joints, which produce a change in the load applied to the magnetron and thereby result in change of transmitter frequency. This phenomenon is known as "pulling." If the change in transmitter frequency so produced is an appreciable fraction of the pass band of the receiver, a spurious amplitude fluctuation synchronous with the antenna scanning may occur. The servomechanisms positioning the reflector will interpret this as a valid direction-error signal and will point the reflector incorrectly. It is impossible to machine r-f components sufficiently accurately to eliminate pulling altogether. An additional step must therefore be taken by designing the AFC system to have a follow rate sufficiently fast to follow the deviations of the transmitter.

A final requirement on the receiver is a consequence of the small size of target and large range of detection at which the system must be capable of operation. The transmitter power is limited by availability of tubes and to some extent by properties of any r-f system that satisfies other performance requirements of the system as a whole. It is therefore necessary to improve the noise figure of the receiver to the practical limit.

15·2. Radio-frequency Input and Local Oscillator.—Figure 15·1 shows a view of the r-f section of the radar system, together with the receiver circuits. The frequency range covered by this receiver is a 12 per cent band centered at 3.3 cm. A tuned TR tube at 3.3 cm has a bandwidth (between half-power points) of about 30 Mc/sec. This is only slightly

greater than the bandwidth of a single i-f amplifier stage of the construction used in this receiver. If, therefore, such a tube were used, it would limit the bandwidth of the receiving system. Also, any unbalance of crystals will reflect LO power to the TR tube which, because of its narrow bandwidth, reflects the LO power back to both crystals in like phase so that the noise is not balanced out. Furthermore, with high-efficiency crystal converters a substantial amount of i-f energy goes back through

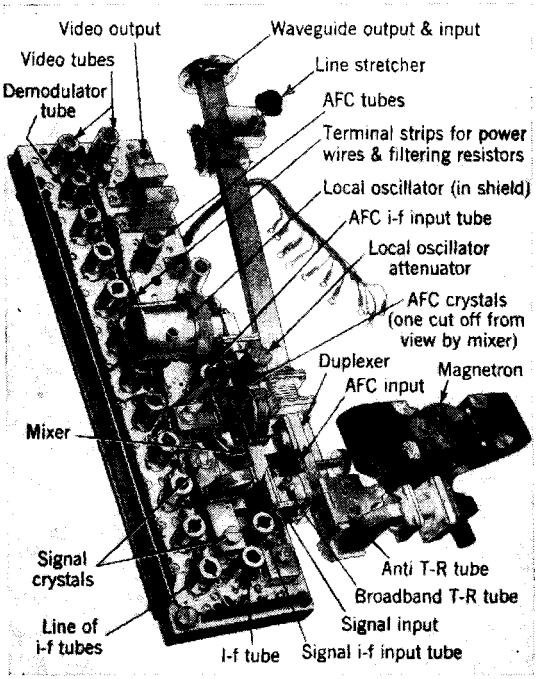


FIG. 15·1.—R-f section and receiver circuits.

the crystal and part of it is converted to image frequency, which is then reflected back by the narrow-band TR tube and reconverted to i-f energy. Since the phases of the voltage and current are usually at variance with those from the normal incoming signal, an apparent mistuning occurs which varies with the tuning of the TR tube. A broad-band TR tube has been designed that has lower loss than the usual tuned design, and it has been used in this receiver. It may be seen in Fig. 15·1.

In order to obtain the lowest possible noise figure, a balanced mixer is used. The unit has been designed so that the signal crystals are well matched into the waveguide. Because of the less critical nature of the

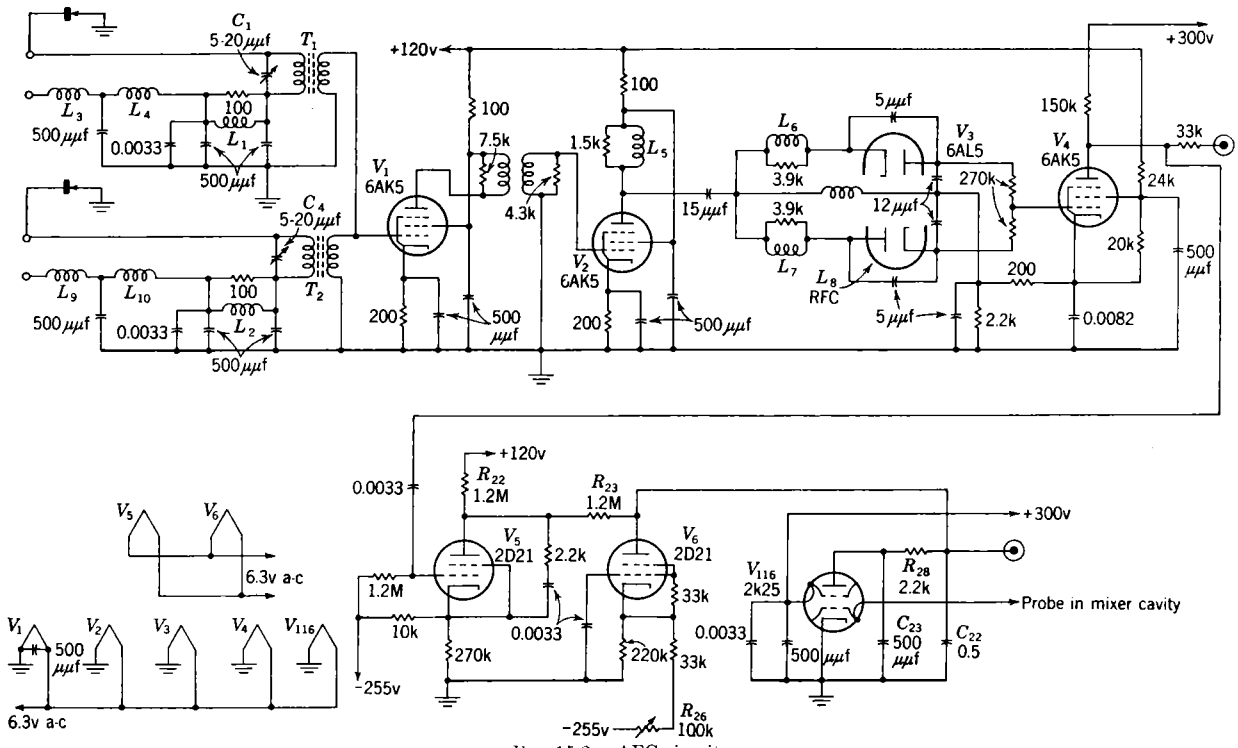


FIG. 15·2.—AFC circuit.

problem and because of space limitations, the crystals in the AFC half of the double mixer have not been so carefully matched into the waveguide. The mixer may also be seen in Fig. 15·1.

Another feature of the mixer is its design for good AFC operation; this design is achieved by attention to details influencing leakage into the mixer, cross attenuation between the signal and AFC channels, and attenuation of transmitter harmonics. The balanced mixer, when manufactured accurately, has good cross attenuation at all frequencies. What little power gets up to the local oscillator and is reflected has to go through the resistive LO attenuator twice. The attenuator from the waveguide to the AFC mixer has been made with 33 db of attenuation caused by a 0.250-in. hole and 39 db caused by a polyiron attenuator. The hole causes sufficient mismatch to prevent appreciable loss to the transmitted pulse, and the attenuation in the polyiron increases with frequency and prevents transmission of harmonics.

The AFC circuit is shown in Fig. 15·2. The input circuit from the balanced mixer consists of the two crystals and two tuned circuits (C_1, T_1 primary and C_4, T_2 primary) coupled to the secondaries of T_1 and T_2 connected in parallel and tuned by the input capacitance of the input 6AK5 amplifier tube. This circuit acts very much like a double-tuned circuit loaded principally on the primary side. It will be described more fully in Sec. 15·3.

The coupling network between V_1 and V_2 is a double-tuned circuit loaded on both primary and secondary. Unlike the other similar transformers in the receiver, no powdered-iron core is used to couple these coils together. The loading is relatively low on primary and secondary, and the two circuits are overcoupled. The two "ears" resulting on the response curve tend to broaden the over-all i-f amplifier and discriminator response.

A discriminator circuit has been developed that is particularly suitable for wide-band work, since a high value of the product of gain and bandwidth is obtained with it (about 6 db more than other types tried). The circuit has been skeletonized and slightly modified for the sake of clarity in Fig. 15·3*a*. Inductance L_6 and capacitors C_{12} and C_{14} form a series-tuned circuit tuned to one side of the intermediate frequency. Inductance L_7 and capacitors C_{16} and C_{15} form a series-tuned circuit tuned to the other side of the intermediate frequency. The frequencies to which these two series-tuned circuits are resonated determine the frequencies of maximum and minimum discriminator output voltage as shown in Fig. 15·3*b*. The proper adjustment of the values of loading resistors R_8 and R_9 straightens the discriminator curve between these maximum and minimum voltages. Inductance L_5 tunes the output circuit capacitance of tube V_2 to the intermediate frequency, 60 Mc/sec.

It may be seen in the discriminator circuit diagram of Fig. 15·3*a* that the two halves of the discriminator are balanced to ground. The transient response of the two halves is therefore the same and will balance out when the input carrier is at the intermediate frequency of 60 Mc/sec. This is an advantage over an unbalanced type, since there is no spike left to operate the AFC when the carrier frequency is at 60 Mc/sec.

In the remainder of the circuit shown in Fig. 15·2, tube V_6 is a gas-filled type 2D21 tube used in a sawtooth generator circuit. The time

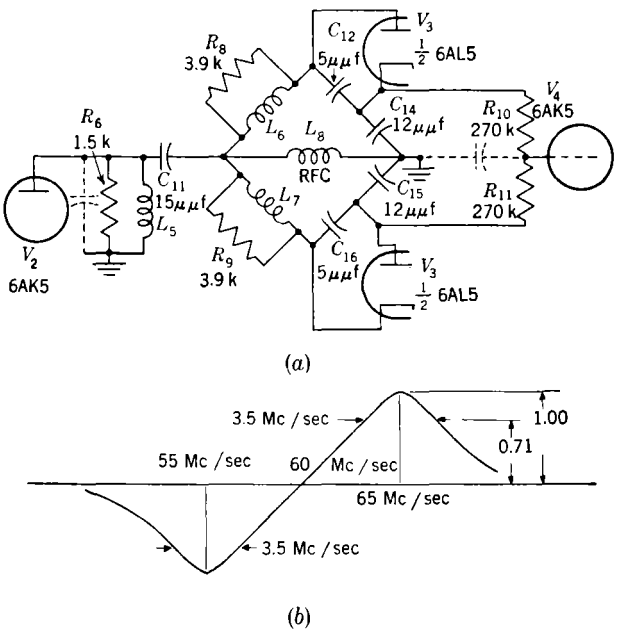


FIG. 15·3.—AFC discriminator. (*a*) circuit; (*b*) output voltage.

constant is 1.2 sec. Tube V_5 is biased so that it does not conduct in the absence of signals on its grid. The type 2K25 oscillator tube V_{16} is frequency-modulated by the application of this sawtooth voltage to its reflector through the filter R_{28} and C_{23} and is thus swept through a frequency range, searching for a signal in this range. The frequency range through which the 2K25 oscillator is swept is adjusted by variation of the 100,000-ohm resistor R_{26} which adjusts the voltage range over which the 2D21 tube V_6 operates. The operation of the rest of the circuit is completely analogous to those described in Sec. 3·4.

It is desired to have the AFC operate only on the transmitted pulse from the magnetron. It is also desired that this pulse should not over-

load the mixer crystals, since the video pulse resulting would contain 60-Mc/sec components which would operate the AFC no matter what the frequency of the 2K25 oscillator. Sufficient attenuation is introduced between the transmitting waveguide and the AFC mixer to allow only sufficient transmitted power to enter the mixer to produce a peak current of 1 ma in the crystals. This produces 0.2-volt rms at the intermediate frequency across the crystal when the proper amount of LO power is applied. The AFC circuit should not operate on less than about 0.02 volt rms, since it might then operate on the side lobes of the transmitted pulse and hold the oscillator off normal frequency. Since the 2D21 tube, V_5 under normal conditions, requires about 13 volts on the grid to operate it, a voltage gain from the crystal to this grid of about 1000 is required. This gain expressed in decibels voltage gain without regard to impedance level is distributed as follows:

	Decibels
Gain from crystals to V_1 grid	$\approx$ 6
Gain from V_1 grid to V_2 grid	$\approx$ 15
Gain from V_2 grid to V_4 grid	$\approx$ 3
Gain from V_4 grid to V_5 grid	$\approx$ 36
Total gain	$\approx$ 60

The 0.1-μsec input pulse is stretched out to about 5 μsec at the plate of tube V_4.

To avoid false directional information resulting from transmitter pulling during the scanning cycle, the AFC circuits should be capable of acting fast enough to follow the transmitter. In Fig. 15·2 tube V_5 may be considered as a very low impedance source when it is operating. The principal time constant of the filter is the one formed by resistor R_{23} and capacitor C_{22}; it must be kept as small as possible. This time constant is 0.6 sec, which is as little as can be tolerated if the sawtooth voltage is not to affect the frequency of the 2K25 oscillator too much. In spite of this time constant long with respect to the error signal period, there is sufficient gain around the AFC feedback loop to eliminate almost completely any 30-cycle frequency modulation of the intermediate frequency. The small amount of frequency modulation remaining does no noticeable harm because of the small curvature in the center of the receiver selectivity curve.

The intermediate frequency of 60 Mc/sec is sufficiently high to prevent almost all difficulty with wrong-sideband operation.

15·3. Receiver Input Circuit.—The receiver input circuit at the intermediate frequency consists of three coupled tuned circuits as shown in Fig. 15·4. Each crystal of the balanced mixer is shunted across a tuned circuit, and appropriate filters are provided so that the d-c crystal cur-

rent may be measured. Each of these tuned circuits is inductively coupled to a secondary coil with the aid of a powdered-iron core. The two secondary coils are connected in parallel across the input of the first amplifier tube. The loading across the tube input circuit is kept at a minimum, and the crystals furnish practically all the loading on their respective tuned circuits. An input circuit of this type is more critical than either the single-tuned circuit or the double-tuned circuit with loading on primary and secondary but has considerably more gain. The r-f input elements will be closely coupled to the i-f input circuit if high-efficiency crystal converters are used. Changes in the r-f elements may

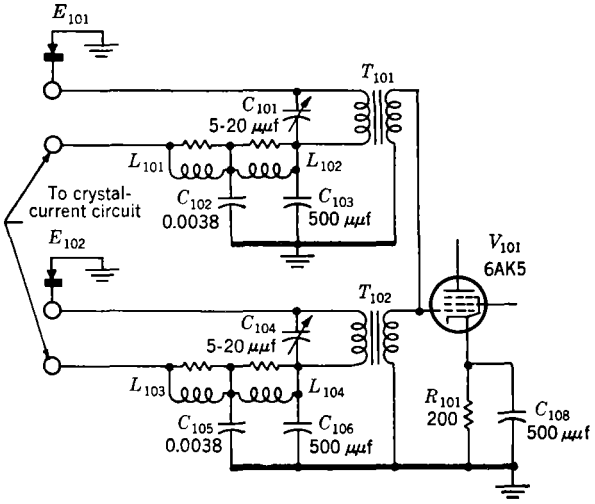


FIG. 15·4.—Receiver input circuit at the intermediate frequency.

change the tuning of the input i-f circuit under these circumstances. The broad-band TR tube and balanced mixer have reduced the possibility of serious changes in the r-f elements; consequently, in practice it is found that the i-f input circuit is quite uniform in response for signal frequencies located anywhere in the 12 per cent r-f band. The i-f input circuit is also quite noncritical in tuning adjustment in manufacture and with tube replacement.

Since a large amount of noise in the receiver is generated by the input tube or tubes, high gain in the input transformer produces a high level of signal at the grid of the first tube to override the noise. As mentioned above, the loading on the secondary circuit is kept at a minimum to give high gain. The primary circuits are made with as low a ratio of inductance to capacitance as is consistent with proper bandwidth, giving higher gain and making the circuit more stable with variation in tube

input impedance and variations in r-f input circuit. The voltage gain of this circuit is approximately 6 db.

15·4. Receiver Input Stages.—The first two stages of the receiver are of the low-noise type using two triodes. The circuit is shown in Fig. 15·5, and an analysis is given in Sec. 4·13. The input triode, or grounded-cathode triode, is a 6AK5 tube with plate and screen connected together. The grid to screen-and-plate capacitance is tuned out by an inductance L_{105} from grid to plate. The second triode, or grounded-grid triode, is one half of a 6J6 double triode which obtains its bias from a 120-ohm resistor R_{103}.

The noise figure of the i-f amplifier is 3.5 db at its frequency of 60 Mc/sec.

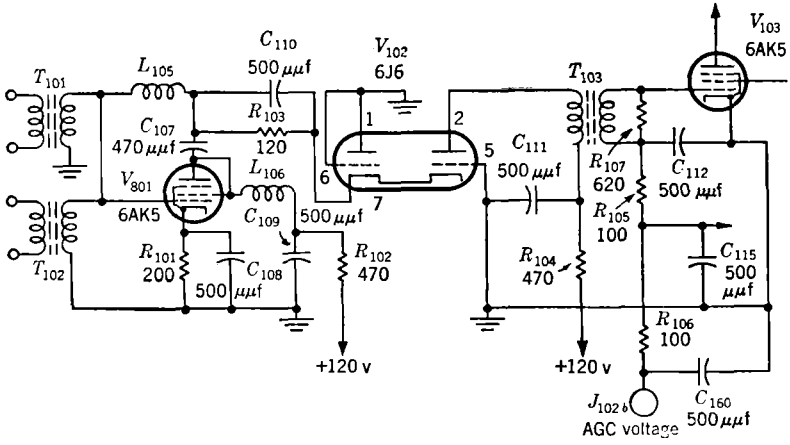


FIG. 15·5.—Low-noise i-f stages of receiver.

15·5. Intermediate-frequency Amplifier.—The general characteristics laid down at the beginning of the design called for an i-f bandwidth of 12 Mc/sec and a total i-f gain of about 100 db. A simple calculation shows that if an amplifier stage has a gain of 6 db, it will take 17 stages to obtain 100 db. If 17 single-tuned stages having an over-all bandwidth of the desired 12 Mc/sec are to be used, the gain per stage or over all is approximately unity, or 0 db. It is clear that single-tuned stages, synchronously tuned, cannot be used.

Staggered triples, staggered pairs, and double-tuned circuits were tried. It was desired to make the tuning of the stages fixed because of the difficulty of alignment in the field. Staggered triples were considered too sensitive to capacitance variation to accept the permissible variation in capacity of either the 6AC7 or 6AK5. Double-tuned circuits with loading entirely on primary or secondary were also found to be too

sensitive to variation. If the loading of the primary equals that of the secondary, that is, if primary and secondary Q's are equal, the amplifier is insensitive to capacitance variation at the bandwidth and frequency used here. However, it was found that with a ratio of primary Q to secondary Q of 2.2, the amplifier is still sufficiently insensitive to capacitance variation and extra gain is obtained.

The double-tuned circuit with a Q ratio of 2.2 was found to have advantages over single-tuned circuits staggered in pairs. The double-tuned circuits have some advantage in over-all gain-bandwidth product. In this receiver the double-tuned circuit was found to be definitely less sensitive to variations in component parameters, and the over-all amplifier was found to be more stable than was the case with the other circuits.

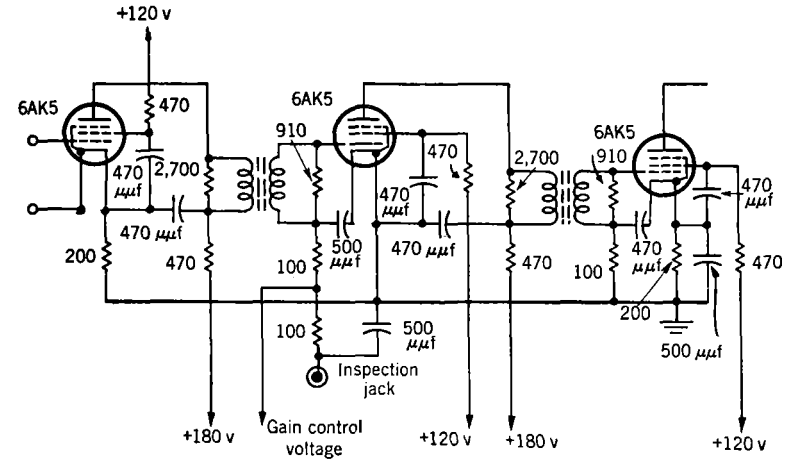


FIG. 15·6.—Circuit of two typical amplifier stages.

Another advantage of the double-tuned circuit is the ability to keep the input and output circuits of an amplifier stage more or less independent of each other, because the tuned circuits do not have to be returned to ground but may be returned directly to an independent cathode lead when the 6AK5 tube is used. Advantage has been taken of this fact in this receiver, as may be seen in Fig. 15·6. It has been found in practice that if the coils are carefully made to dimensions, practically no further adjustment is needed before they are inserted in the receiver, and none is needed after insertion. A further advantage of double-tuned circuits is the elimination of the interstage coupling capacitor—an elimination that decreases the grid-circuit time constant.

The 6AK5 tube was chosen instead of the 6AC7 as an i-f amplifier tube for several reasons. The percentage input and output capacitance tolerances of the 6AK5 which are only half as large as those of the 6AC7,

allow a more critical circuit and, therefore, more gain per stage. The 6AK5 tube has two cathode leads allowing a great reduction in the coupling of the input and output circuits of the tube and a reduction in chassis currents which might couple amplifier stages to produce regeneration. It is simpler to obtain wide bandwidth, and the components, such as bypass capacitors, are smaller at high frequencies than at low. The 6AK5 operates better at high frequencies, takes less space, and uses less power than the 6AC7, all of which considerations are important when the receiver is to be mounted in the r-f head of a radar system.

In a multistage amplifier using double-tuned interstage coupling circuits and having a bandwidth that is as large a fraction of the mid-frequency as the present one, it is advantageous to use unequal Q's in the primary and secondary circuits. If the Q's are equal, it is impossible to obtain sufficient asymmetry to counteract the slope of the top of the response curve caused by the variation with frequency of feedback through the plate-to-grid capacitance of the tube. It is quite simple to obtain the required asymmetry when the Q's are sufficiently different.

Originally the intermediate frequency was chosen as 60 Mc/sec because of its harmonic relation to the 30-Mc/sec i-f amplifiers then in use and the consequent ability to use harmonics of signal generators for testing. A high value was chosen because of the relatively greater ease in obtaining sufficient magnetic coupling between the primary and secondary circuits of the double-tuned circuit and because of the improvement in AFC operation resulting from elimination of wrong-sideband locking.

Figure 15·6 shows a schematic circuit of two typical amplifier stages of this receiver. One has provision for applying gain-control voltage to the grid and the other does not. The data on the interstage coupling network are as follows:

Plate output capacitance	4.8 $\mu\mu$f
Primary loading resistor	2700 ohms
Primary $Q = \omega_0 R_1 C_1$	4.9
Grid input capacitance	7.4 $\mu\mu$f
Secondary loading resistor	910 ohms
Tube input loading resistance	About 8000 ohms
Secondary effective loading resistance	800 ohms
Secondary $Q = \omega_0 R_2 C_2$	2.2
Q-ratio primary to secondary	2.2
Coupling between primary and secondary	38 per cent
Bandwidth	25 Mc/sec
Band center	60 Mc/sec
Gain	11 db

The coupling between primary and secondary is obtained principally by means of an iron core with some capacitance coupling obtained

by having the grid and plate ends of the secondary and primary near each other. The iron core reduces the coil size and the inductive field associated with it.

Intermediate-frequency coupling networks that are different from the typical one are the input transformer network, which has already been discussed in Sec. 15·3; the transformer network between the second and third i-f tubes; and the transformer network between the last i-f stage and the second detector. The transformer network between the second and third i-f tubes has loading only on the secondary, and the transformer coupling coefficient has been changed accordingly. This was done to raise the gain of this stage and thus reduce the effect of noise from the third i-f tube. The gain from the grid of the first i-f tube to the grid of the third is 12.8 db.

The last i-f transformer is the same as the typical i-f transformer described above. The only change in the network is the use of a 3300-ohm loading resistor across the secondary of the transformer. This change compensates for the loading of the diode. There is a loss in the second detector of about 6 db.

The gain of the i-f amplifier from the crystal input to the grid of the first video tube in voltage ratio without regard to impedance is about as follows:

	Decibels
Input transformer gain	6
First two i-f tubes and following transformer gain	12.8
10 i-f stages at 11-db gain each	110
Second detector gain	−6
Total gain (approx.)	122.8

The over-all bandwidth is about 12.5 Mc/sec. The normal output level of the second detector is 0.6 volt, and therefore the gain of the amplifier is sufficient to give full output on noise even when all tubes are down by 20 per cent in transconductance—a condition that may be encountered with tubes that still meet JAN-1A specifications.

The circuit diagram of the signal channel of the receiver chassis is shown in Fig. 15·7. In addition to the features already mentioned, it will be noted that the tubes V_{103}, V_{105}, V_{107}, and V_{110}, which have gain-control connections to their grids, have no cathode resistors. The reason for this, as is explained in Sec. 15·8 of this chapter, is that better gain control is obtained when the bias is small if the cathode resistor is omitted. The initial voltage furnished to these grids by the gain control is −2 volts. When the gain control is disconnected and the receiver power is on, a bias is furnished by the −255-volt supply and the voltage divider resistors R_{31} and R_{32} through resistors R_{29} and R_{30}.

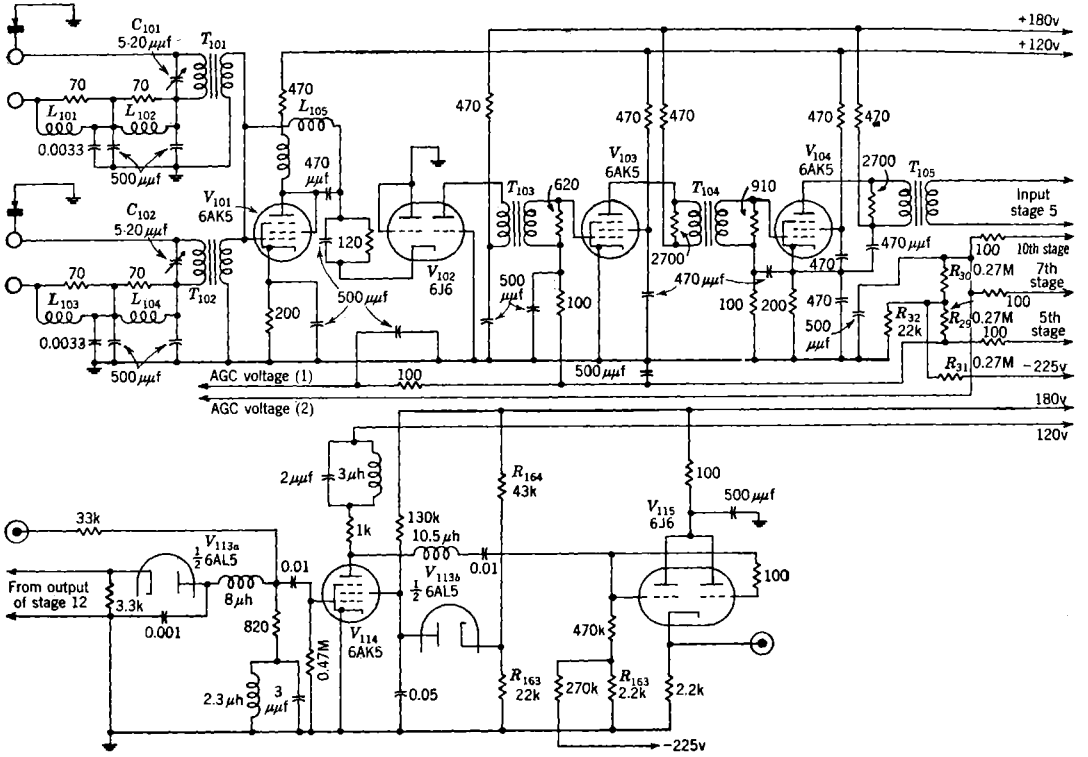


FIG. 15·7.—Circuit of receiver. Eight more i-f amplifier stages (5 to 12) identical with fourth stage except for gain-control stages (5, 7, and 10) which are identical with third stage.

Since the impedance of this network is high, it has no effect when the low-impedance gain control is connected. It will also be noted that tubes V_{103} and V_{105} are furnished gain-control voltage from one source having a long time constant and tubes V_{107} and V_{110} are furnished gain-control voltage from another source with a short time constant.

Figures 15·1, 15·8, 15·9, and 15·10 show the constructional details of the receiver. In Fig. 15·8 which shows a bottom view of the chassis with cover removed, it may be seen that the lower part, devoted to the high over-all gain signal circuits, is disposed in a long narrow tube when the cover is on. To ensure this tube or waveguide effect, spring fingers are attached to the cover to make contact on the partitions,



FIG. 15·8.—Bottom view of receiver chassis.

and spring clips are used along the edges of the chassis. This waveguide is made with as small a cross section as possible to give high attenuation from one end to the other of the chassis (see Sec. 10·2). The electromagnetic fields of the coils are made as small as possible by use of iron cores to make the coils small. In order to meet space limitations, the input stages are doubled back on the rest of the amplifier as may be seen at the left of Fig. 15·8. This portion is shown enlarged in Fig. 15·9. In the boxed-in portion in the upper center part of this figure are the input circuits, the round black objects in the lower corners being the insulators through which the connections to the mixer crystals project. The input transformers T_{101} and T_{102} and ceramic trimmer capacitors C_{101} and C_{104} are connected to the crystal terminals (refer also to Fig. 15·7). The neutralizing coil L_{105} is a single-layer coil on a

bakelite resistor-type form shown at the left of this compartment and projecting through the partition. The crystal-current filter inductances L_{101}, L_{102}, L_{103}, and L_{104} are single-layer coils wound on resistors.

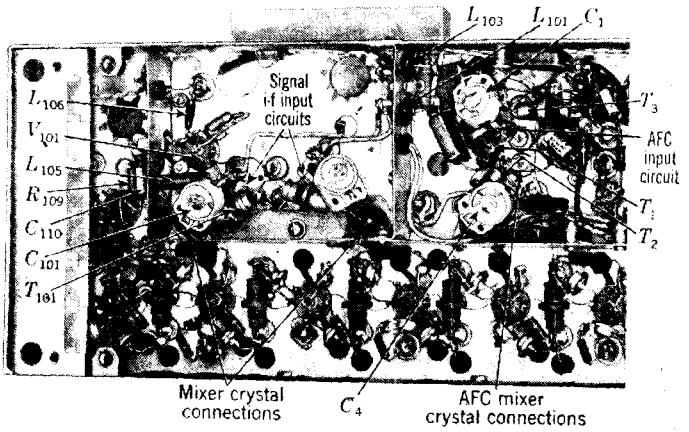


FIG. 15·9.—Bottom view of receiver showing input stages.

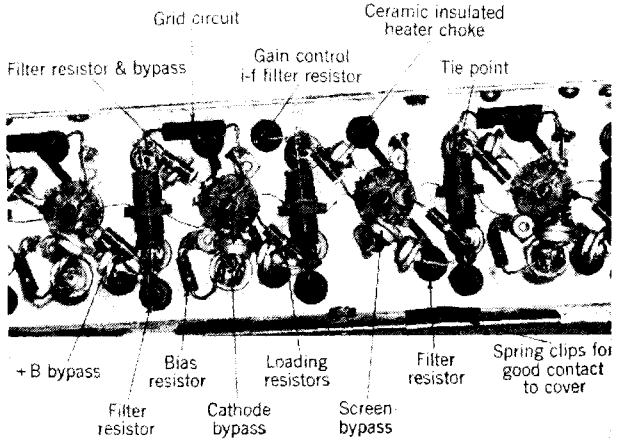


FIG. 15·10.—Bottom view of receiver chassis showing part of signal i-f amplifier.

Two of these may be seen projecting through the partition in the upper right-hand corner of this compartment.

At the right of the input-circuit compartment in Fig. 15·9 is the similar AFC input circuit consisting of the two ceramic trimmer capacitors partially covering the input transformers and mixer crystal connections.

The AFC circuit extends on to the right as shown in Fig. 15·8 in the long compartment in the upper part of the figure. In this compartment are also the filters and connections for the power leads as they leave the chassis. At the right of Fig. 15·8 are the video circuits contained in this chassis.

The construction of each stage of the signal amplifier with its filtering arrangement is best shown in Fig. 15·10. The interstage coupling transformers are mounted with their axes across the waveguide. They are mounted on ceramic-insulated tie points which are made like feed-through insulators so that voltages may be measured with a probe voltmeter from the outside of the chassis. This gives a means of measuring the voltage across the +B-voltage filter resistor and checking the operation of the tube. The grid voltage of the gain-controlled stages and the filtering resistors of both plate and grid circuits may be measured. The low r-f potential ends of the primary and secondary coils are connected to these tie points, and the other ends are separated by a spacer that determines both the inductive and capacitive coupling. In the model shown, the spacer is too narrow because the coils were revised without changing the spacer.

The grid coil is isolated from ground by a 100-ohm filter resistor, and the r-f current is made to return directly to one end of the 6AK5 cathode by means of a bypass capacitor connected directly from the low r-f potential end of the grid coil to that cathode terminal. The low r-f potential end of the plate coil is similarly bypassed to the other end of the cathode to which the screen is also bypassed. This end of the cathode is also bypassed to ground or connected directly to ground when no cathode resistor is used. The filter resistors for plate voltage, screen voltage, and gain-control voltage pass through rubber grommets in holes in the chassis so that one lead is on one side of the metal and the other lead is on the other side. One heater terminal is grounded, and the other heater terminal is connected to the supply through a choke coil which passes through the chassis in much the same manner as the filter resistors described above. In this way the power-supply leads are carried along the top of the chassis away from the coils and therefore are not inductively or capacitively coupled to the coils and cannot couple various amplifier stages. The method of connecting the filter resistors and choke coils to a terminal strip for connecting the power-supply wires above the chassis may be seen in Fig. 15·1. This method also allows the checking of voltage above the chassis so that the cover does not have to be removed and the shielding disturbed for this operation.

15·6. Second Detector and Video Amplifier in Receiver Chassis.—The second detector and video-amplifier circuits contained in the receiver

chassis are shown in the circuit diagram of Fig. 15·7. They consist of tubes V_{113}, V_{114}, V_{115}, and their associated circuits.

The input capacitance of the diode V_{113a}, consisting of one half of a type 6AL5 tube, is 7.4 $\mu\mu$f, which is the same as the input capacitance of a type 6AK5 tube in the i-f amplifier. The loading of this diode and circuit on the i-f transformer is about 1060 ohms. The diode-load resistance is 820 ohms. The gain from the i-f amplifier tube grid driving the diode to the diode load resistor is 5 db (voltage gain), where the input is measured in rms volts (at 60 Mc/sec) during the pulse, and the output is measured in peak-pulse volts.

The normal size of signals while the receiver is operating with automatic gain control is

0.6 peak-pulse volts at the output of V_{113a} (negative).
2.25 peak-pulse volts at the plate of V_{114} (positive).
0.75 peak-pulse volts at the cathode of V_{115} (positive).

The output pulse from the second detector is negative to allow limiting action by the first video-amplifier tube V_{114}. No bias is applied to V_{114}. Any positive voltage on the grid will cause grid current to flow through the 0.47-megohm grid resistor and bias the tube giving d-c restorer action. The time constant of the grid circuit is long enough so that the 30-cycle modulation, caused by the antenna feed nutation when the target is not on the antenna axis, is little affected in phase.

The screen voltage of tube V_{114} is 50 to 55 volts and is obtained through a resistor from the regulated +180-volt supply, since no low-impedance source of 50 volts is available. The 130,000-ohm screen resistor is used to stabilize the tube when the large grid resistor is used as well as to equalize the gain of different tubes. However, when signals occur a large part of the time or when there is a great deal of noise, the average grid voltage becomes considerably negative, and less average current is drawn by the screen. The voltage drop in the 130,000-ohm screen resistor becomes less, and the screen voltage higher. This causes more current to flow in the plate circuit when the grid is instantaneously at zero potential. Under these conditions an input signal that takes the grid from zero voltage to plate-current cutoff will cause a greater output voltage. This is not desirable, since the increased signal may overload later amplifier tubes as well as change intensity of light output of intensity-modulated cathode-ray tubes when the duty ratio changes.

To reduce this effect a lower-impedance (14,500 ohms) voltage divider composed of R_{163} and R_{164} and giving 60 volts is connected by means of diode V_{113b} to the screen of the amplifier tube V_{114}. If the

screen voltage rises above 60 volts, current will flow through the diode and thereby make the screen voltage supply circuit have an impedance about one-tenth its former value. This aids greatly in stabilizing the screen voltage and keeps the maximum output voltage much more constant.

The output cathode-follower type 6J6 tube V_{115} has both halves connected in parallel. The cathode circuit is designed to feed into a 75-ohm line with a d-c path through a 75-ohm resistor connected at the far end of the line. The normal output is 0.75 volt, and the limited output (limited by V_{114}) is 1.5 volts.

The coupling networks between the video stages are of a high-frequency-compensated type that has a product of gain and bandwidth 2.40 times as great as when no peaking elements are used and a rise time response to a step function input 2.48 times as fast. The overshoot is only 0.3 per cent. These networks, however, have little effect on the pulse shape, since their bandwidths are much greater than is necessary. The main consideration is to have the proper signal polarity and voltage and to avoid distortion.

15·7. Video Amplifiers for Oscilloscopes and Tracking Circuits.—This receiver is designed to operate into a 75-ohm transmission line at low level, that is, normally at 0.75 volt with the output limited at 1.5 to 2.0 volts. The receiver is located at the antenna. The low-level video signal is transmitted through a 75-ohm coaxial cable, through slip rings in the turret base, again through a 75-ohm coaxial cable to the video amplifiers 100 to 200 ft away, measured along the cable. The cable passes into each amplifier chassis and out again, connecting by a short low-capacitance lead to the input tube of each video amplifier. In the final chassis the end of the cable is terminated in a 75-ohm resistor.

The bandwidths of the amplifiers for all oscilloscopes and tracking circuits are about the same although the resolution of signal used in each is normally different. This construction provides improved performance in the presence of certain types of interference. If a low-intensity c-w signal of about the same frequency as the incoming normal pulse signal is received, the output will be the normal pulse modulated on the top by the difference in the two carrier frequencies. As the intensity of the c-w signal is increased, the modulation will become deeper and deeper until the output becomes pulses of the beat frequency. The video amplifiers must have bandwidths wide enough to pass any beat frequency that passes through the receiver. Satisfactory operation is obtained when the video amplifiers are not more than 3 db down in response at the frequency at which the i-f amplifier is 40 db down. In the receiver under discussion the video amplifiers should have a bandwidth about twice the semibandwidth of the i-f

amplifier, or a bandwidth at 3 db down of 12 Mc/sec. This bandwidth is closely approached by all of the video amplifiers and exceeded in some, the variation being due to economic factors.

To prevent distortion, all amplifiers are operated on the linear portion of the tube characteristic. Where the d-c level is important, as in intensity-modulated cathode-ray indicators and in the automatic-gain-control circuits, a d-c restorer circuit is used. Since gain controls are unbypassed variable cathode resistors, the negative feedback tends to prevent distortion when the tubes are on the curved portion of the characteristic at low gain.

The low-frequency amplitude and phase characteristics should be such that a 100-μsec pulse will not vary more than about 5 or 10 per cent in amplitude throughout its length. This can be obtained by making all time constants that affect the low-frequency response very large. Equally good low-frequency response can be obtained by means of compensation, with the added advantages of smaller and more economical units and better high-frequency response due to lower capacitance of the units to ground. By analysis of the circuits used and by experiment it was found that if all uncompensated circuits had time constants of 10,000 μsec or more and all compensated circuits had time constants of about 1,000 μsec or more, the desired results could be obtained with components having reasonable tolerances. Shorter time constants would have required smaller tolerances.

The general rules followed are given below. They result in about 5 per cent variation in amplitude of a 100-μsec signal when the deviations of component values are most unfavorable. The screen-voltage supply is made to have sufficiently low impedance by means of a voltage divider (where the screen voltage is not obtained directly from a low-impedance source) such that the total current drawn by tube and divider is twice the screen current. The time constant of the circuit is made about 10,000 μsec with ± 10 per cent tolerances on capacitors and ± 5 per cent tolerances on resistors, the 5 per cent tolerances being dictated by voltage requirements rather than time constant considerations.

The plate-circuit filter resistor is made small enough to prevent the danger of approaching the knee of the plate-current vs. plate-voltage curve. It is made large enough to give a 10,000-μsec time constant with a 1- or 2-μf filter capacitor. The tolerances of both filter resistor and capacitor are ± 10 per cent.

The time constant formed by the product of the plate-load resistor and the plate-circuit filter capacitor is made equal to the time constant formed by the product of the grid resistance of the following tube grid circuit and the plate-to-grid coupling capacitor. The plate-load resistance is determined by high-frequency consideration, and its ± 5 per cent

tolerance is dictated partly by gain requirements. The grid resistor is kept well below the maximum value recommended by the tube manufacturers but high enough to make the plate-to-grid coupling capacitor small in physical size so that it has low capacitance to ground. The grid resistor and the coupling-capacitor tolerances are ±10 per cent unless the time constant is much below 1,000 μsec when the tolerances of one or both are made ±5 per cent. Plate- and grid-circuit compensation of this type does not give the exponential decay of a single time constant, but it gives more constant output where the maximum probable duration of signal is known.

Grid bias is obtained either from a negative voltage supply or from an unbypassed cathode resistor. In general, the latter procedure gives

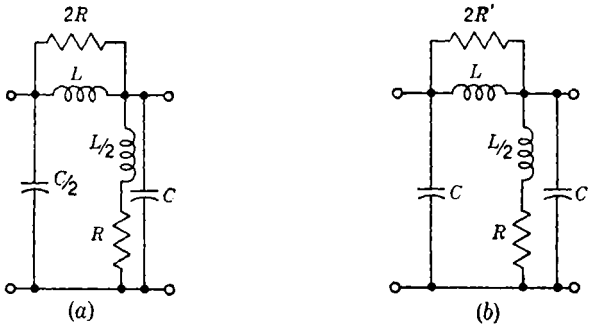


FIG. 15·11.—Video filters. L is wound on resistor $2R$, $L/2$ is wound on a dummy resistor. $L = R^2C$, f_k = frequency at which amplitude is 3 db down.

(*a*) $2\pi f_k RC = 1.47$. Gain-bandwidth product is 2.20 times what it would be without compensation. Overshoot in response to step function = 7 per cent.

(*b*) $2\pi f_k RC = 0.97$. Gain-bandwidth product is 1.94 times what it would be without compensation. Overshoot in response to step function = 3 per cent.

more stable gain characteristics and less distortion. One exception is the use of a 20,000-μsec time constant in the cathode circuit of the type 807 tube where the extra gain is needed.

The high-frequency compensation used in the video amplifiers is of the shunt-series type described in Chap. 8. Figure 15·11*a* is for use where the ratio of input to output filter capacitance or its reciprocal is equal to one-half. In this case, the product of gain and bandwidth is 2.2 times that when no peaking coils are used. Figure 15·11*b* is for use where the ratio of input- to output-filter-capacitance is unity and the product of gain and bandwidth is 1.94 times that when no peaking coils are used. In all of the video stages in this receiver, the capacitance ratio lies at or between these two ratios. Component values for intermediate capacitance ratios were estimated by interpolation. When the capacitance ratio is 0.5, the overshoot of the response to a step-

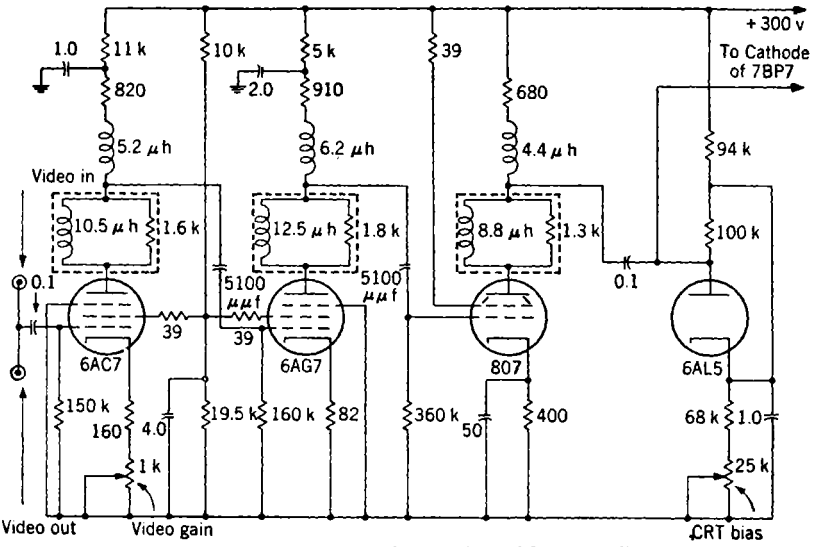


FIG. 15-12.—Circuit of amplifiers for B- and E-type indicators.

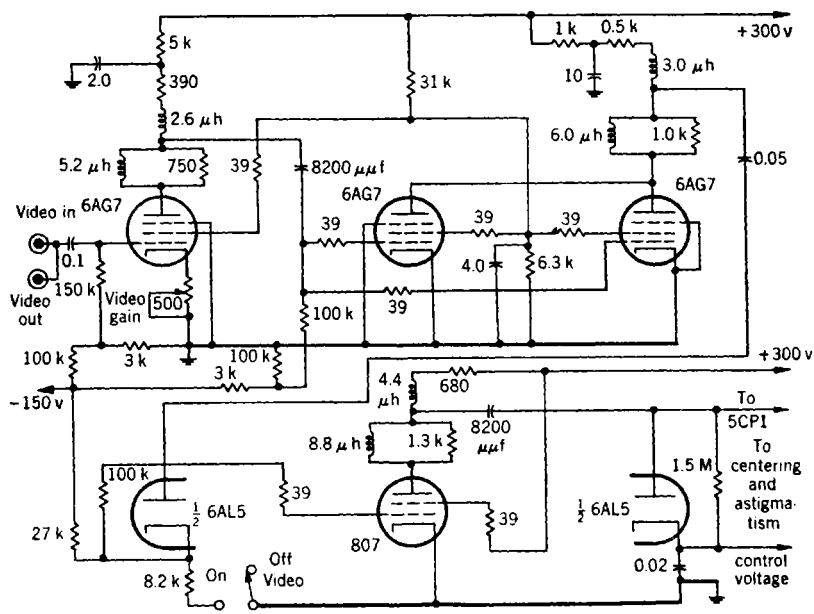


FIG. 15-13.—Circuit of A/R indicator amplifier.

function input is 7 per cent; when the ratio is unity, the overshoot is 3 per cent. Because of the width of the video-amplifier pass band compared with that of the i-f amplifier, only overshoots due to the i-f circuits are noticeable on signals not limited. Overshoots are not objectionable on either limited or undistorted signals. The "shadows" are insignificant.

High-transconductance tubes often tend to oscillate at very high frequencies when leads to the elements happen to be tuned, usually by

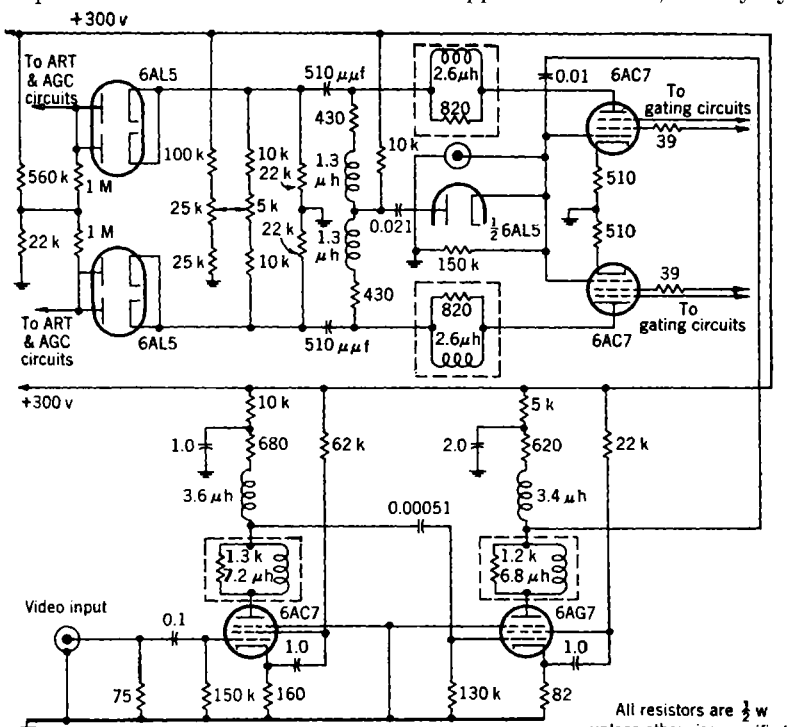


FIG. 15·14.—Video circuits for automatic range-tracking and AGC.

their distributed constants, to the same frequency, or nearly so. Resistors of 39 ohms are placed in some of these leads to damp out such oscillations.

All high-frequency circuits are constructed in such a manner as to keep the capacitances to ground low and the lead lengths short. This gives higher gain-bandwidth products and less feedback. Low-frequency feedback is minimized by use of plate-circuit filters and by a low-impedance power supply that is electronically regulated.

The circuit of Fig. 15·12 is used for the amplifiers for the B-type and E-type intensity-modulated cathode-ray tube indicators. The circuit of Fig. 15·13 is used for the amplifier for the A/R-type cathode-ray tube indicator for deflection of the beam. The circuit of Fig. 15·14 is used for the video amplifier in the automatic-gain-control and automatic range-tracking circuits. It may be noted that the circuit of Fig. 15·14 is that of the final amplifier on the video transmission line and contains the 75-ohm line-terminating resistor.

15·8. Automatic Gain Control; General Considerations.—It has been found that in the presence of large amounts of noise, a desired signal can be more readily distinguished if peak-voltage values are used. A circuit is used in this receiver whereby the peak-pulse voltage is measured each time the desired target signal occurs. This voltage is retained in the circuit until just after the next desired pulse occurs, when the retained voltage is changed to the new pulse voltage. The circuit therefore introduces a delay approximately equal to the repetition period and has no attenuation for any frequencies that may be present in the form of amplitude modulation of the pulses. It is essentially an *all-pass circuit*, which has delay but no attenuation. It is readily shown that if no further attenuation is added, such an inverse feedback loop will oscillate at $f/2$, $3f/2$, $5f/2$, . . . and will have maximum degeneration at $0, f, 2f, 3f, \ldots$ where f is the repetition frequency of the system. It is necessary to provide sufficient attenuation at $f/2$ and higher frequencies to prevent oscillation. Any low-pass filter that will furnish the proper attenuation at $f/2$ will cause still further delay and slow up the AGC action.

The low-pass filter of the AGC circuit must not have too low a cutoff frequency or the delay will be apparent at the direction-error modulation frequency. This delay would not be serious, except for some applications, if it were constant, but, as will be shown later, it is not constant. A maximum of 10° total variation in phase shift in the whole radar system at the direction-error modulation frequency or about one millisecond delay variation at that frequency is allowable.

The low-pass filter could have a cutoff frequency below the direction-error modulation frequency so that the AGC would have no effect at the modulation frequency. The modulation could then be obtained from the video signals. There is, however, too little margin between the fading frequency of 5 cps and the conical-scan frequency of 30 cps to allow sufficiently small variation in phase shift at the modulation frequency of 30 cps in the face of tube variations.

Another disadvantage of a very low cutoff frequency in the AGC filter is that a strong sudden interfering signal will give a high AGC bias that will be delayed so long in reaching the amplifier grids that

the filter storage elements are well filled. The bias then takes the amplifier tubes well beyond plate-current cutoff. Since no signal can then pass through the amplifier, the feedback loop is broken and the feedback system can no longer aid itself in keeping a constant output and therefore cannot aid in restoring the receiver to a condition where it can handle normal signals again. The filter then takes its own time in discharging the capacitors. This takes so long ($1\frac{1}{2}$ to 3 sec) that the radar system may well have lost its target.

Slow time constants in AGC systems are also likely to be so close to the time constant of the radar system that poor target tracking or even oscillation of the radar system as a whole may occur.

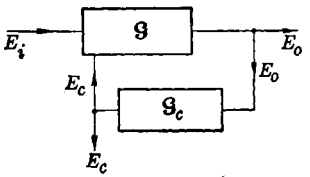


FIG. 15·15.—Feedback loop for AGC.

To obtain good range-error signals it is very desirable to have constant-amplitude output signals in order to eliminate apparent range shift due to nonlinear distortion of the varying amplitude signals that do not have instantaneous rise in amplitude (vertical sides).

The best compromise in AGC-filter design for this system seems to be one that allows nearly complete elimination of the direction-error modulation from the output, together with frequencies as much higher as possible, but one that has a cutoff frequency low enough to prevent oscillation at $f/2$.

It has been mentioned above that tube variations have an important effect on filter design. This effect is due to variation in gain around the feedback loop when the bias on the gain-controlled i-f tubes is changed, as described in Sec. 9·2. In Fig. 15·15, suppose that

E_i = the steady-state input to the receiver,

E_o = the steady-state output of the receiver,

$\mathcal{G}$ = the steady-state gain of the receiver under the conditions of the above E_i and E_0,

$\mathcal{G}_c$ = the gain of the feedback amplifier network and is usually a complex quantity,

$$m = \sum \frac{d}{de_g} (\ln g_m),$$

E_c = the steady-state voltage bias fed back to the gain-controlled i-f amplifier tube grids.

Then

$$E_o = \mathcal{G}E_i.$$

If E_i is increased by a small amount ΔE_i, then E_o will increase by a small amount ΔE_o, and $\mathcal{G}$ will decrease by $\Delta E_o \mathcal{G}_c m \mathcal{G}$ and

$$\left.\begin{aligned} E_o + \Delta E_o &= (E_i + \Delta E_i)(\mathcal{G} - \Delta E_o \mathcal{G}_c m \mathcal{G}), \\ \Delta E_o &= \Delta E_i \mathcal{G} - (E_i + \Delta E_i)\, \Delta E_o \mathcal{G}_c m \mathcal{G}, \\ \Delta E_o &= \frac{\Delta E_i \mathcal{G}}{1 + (E_i + \Delta E_i)\mathcal{G}_c m \mathcal{G}}. \end{aligned}\right\} \tag{1}$$

If $(E_i + \Delta E_i)\mathcal{G}_c m \mathcal{G} \gg 1$,
then

$$\Delta E_o \approx \frac{\Delta E_i}{(E_i + \Delta E_i)\mathcal{G}_c m}. \tag{2}$$

Since

$$\Delta E_c = \Delta E_0 \mathcal{G}_c,$$

then

$$\Delta E_c = \frac{\Delta E_i \mathcal{G} \mathcal{G}_c}{1 + (E_i + \Delta E_i)\mathcal{G}\mathcal{G}_c m} \tag{3}$$

or $\Delta E_c \approx \Delta E_i / E_i m$ if $(E_i + \Delta E_i)\mathcal{G}\mathcal{G}_c m \gg 1$ and $E_i \gg \Delta E_i$ (4)
or

$$\Delta E_c = \frac{M}{m} \tag{5}$$

where M is the modulation factor $\Delta E_i / E_i$.

It is necessary to be able to use signals that are as small as possible compared with the noise level of the receiver. At small signal level, $\mathcal{G}$ must consequently be such that the noise itself will be amplified to full output so that it operates the AGC. The amplification of the AGC loop is made sufficient to accomplish the purpose of the AGC in keeping E_o constant, irrespective of signal level. If E_o is essentially constant, then, since $E_o = \mathcal{G} E_i$ at any given signal level, $\mathcal{G} E_i$ is essentially constant for all values of E_i. If ΔE_i is small compared with E_i, then in Eq. (1) it is seen that $(E_i + \Delta E_i)\mathcal{G}\mathcal{G}_c m$ varies only with $\mathcal{G}_c$ and m. (The pass band of $\mathcal{G}_c$ is assumed to be so small compared with that of $\mathcal{G}$ that variations in delay and gain of $\mathcal{G}$ due to sideband attenuation are considered inconsequential.)

There are two cases that may cause trouble. One is where appreciable phase shift or delay of the direction-error voltage occurs, and the other is where $(E_i + \Delta E_i)\mathcal{G}\mathcal{G}_c m = -1$ so that, according to Eq. (1), ΔE_o becomes infinite and the receiver oscillates. In the first case, if ΔE_i is small and $(E_i + \Delta E_i)\mathcal{G}\mathcal{G}_c m$ is large, and if the modulation is limited to one frequency so that $\mathcal{G}_c$ is constant, then ΔE_c is equal to M/m according to Eq. (5). It is seen that ΔE_c contains the direction-error signal in this receiver. The phase of this error signal will be essentially the same as that of the modulation of the input signal, and the amplitudes will be essentially proportional. However, if m varies greatly with input signal amplitude, then $(E_i + \Delta E_i)\mathcal{G}\mathcal{G}_c m$ may no longer be large compared with unity for all input signals. If $\mathcal{G}_c$ has a large phase angle, and if m is small enough,

according to Eq. (3), ΔE_c will not have the same phase as ΔE_i and a phase shift in the direction-error signal occurs, giving an apparent change in target direction. The magnitude of ΔE_c will also be somewhat less. The plots of m as a function of control grid bias shown in Fig. 9·4*b* indicate that m does vary greatly, so that the gain of the feedback loop must be made great so that even when m is small, the phase shift of the error-signal output will be small.

An increase in gain to prevent error-signal phase shift when m is small may cause the feedback circuit to oscillate at some frequency if $\mathcal{G}_c$ has sufficient phase shift so that $(E_i + \Delta E_i)\mathcal{G}\mathcal{G}_c m$ can become equal to -1. The fact that pulsed data is used has an important effect on this aspect as may be seen from the following explanation.

In this receiver each pulse is measured and the measured voltage is held over until the next pulse from the same target is received. It is then applied as bias to the i-f tubes which are controlled by the AGC

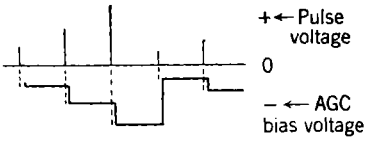


FIG. 15·16.—Relation of AGC voltage and pulse signals.

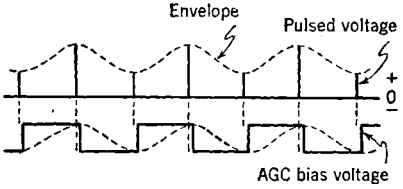


FIG. 15·17.—Pulse signals amplitude modulated at half the repetition rate.

feedback circuit until the next pulse arrives. As soon as this next pulse is measured and the previous pulse has disappeared, the newly measured voltage is used as AGC bias, etc., as shown in Fig. 15·16. When the incoming signals are modulated at a frequency equal to one-half the repetition rate f and the feedback loop is broken where ΔE_c is applied to the i-f tubes, the result is shown in Fig. 15·17. It can be seen that when the incoming signal is large, a small bias is applied to the receiver; and when the incoming signal is small, a large bias is applied to the receiver. In other words, the fed-back voltage is in phase with the incoming signals so that the output will increase indefinitely to the point of oscillation if the loop gain is sufficient. Voltage at one-half the repetition frequency must be attenuated sufficiently to prevent this oscillation. In this receiver a low-pass filter is used to allow degeneration at low frequencies but to provide sufficient attenuation at high frequencies to prevent trouble at half the repetition frequency.

The effect of a single-time-constant low-pass RC-filter is shown in Fig. 15·18. A step function applied to the input of the circuit shown in Fig. 15·18*a* will produce an output as shown in Fig. 15·18*b*. The response to the voltage of Fig. 15·17 is shown in Fig. 15·18*c*. It is seen that the

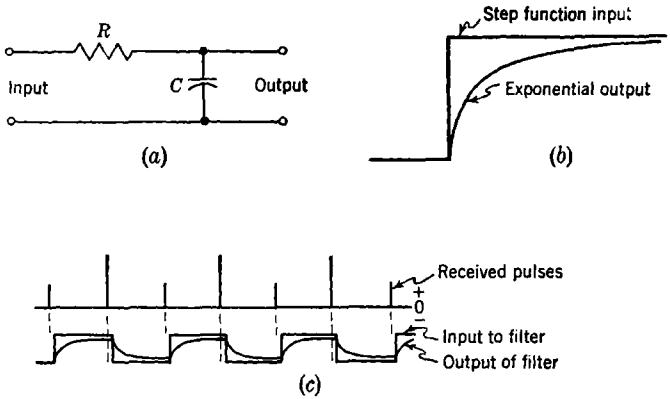


FIG. 15·18.—Effect of single-time-constant low-pass *RC*-filter.

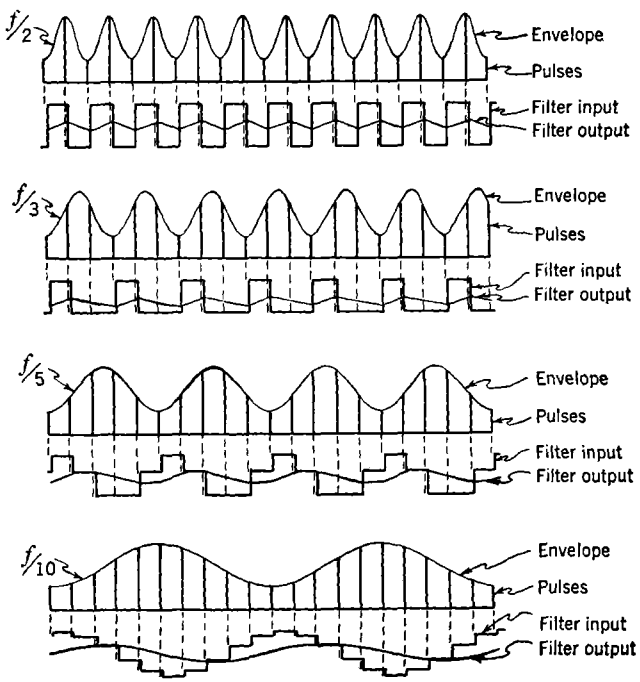


FIG. 15·19.—Illustration of the decrease of the phase shift resulting from pulse-to-pulse delay as the modulation frequency decreases from half the repetition rate.

phase of the feedback voltage is unchanged, that is, the minimum bias occurs at the time of maximum signal; the feedback voltage is in phase with the incoming signal; and the circuit is regenerative at this frequency. The amplitude of the feed-back voltage is reduced, however, and oscillation may be prevented. It can be seen that no matter how long the time constant, the phase is unchanged and the only effect at this frequency is a reduction of amplitude. In other words, the delay caused by the pulsed data and the delay caused by the filter do not add directly.

A series of diagrams in Fig. 15·19 is intended to show that as the modulation frequency is decreased from half the repetition frequency, the phase shift of the error-signal output gradually recedes from 180° when a single *RC*-combination is used in the form of a low-pass filter. This shows that with a network of this type the system cannot oscillate, providing

$$(E_i + \Delta E_i)\mathcal{G}\mathcal{G}_c m$$

is less than unity in magnitude at one-half the repetition frequency and above.

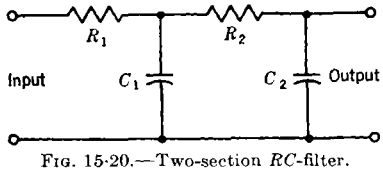


FIG. 15·20.—Two-section *RC*-filter.

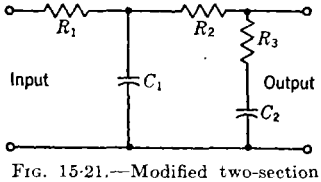


FIG. 15·21.—Modified two-section *RC*-filter.

By a similar method it may be shown that if another section of equal time constant is added to the low-pass *RC*-filter, as shown in Fig. 15·20, so that the filter causes a phase shift of nearly 180° at some frequency below half the repetition frequency, the delay caused by the storage from one pulse to the next will cause sufficient additional delay to make the system oscillate at some frequency in that region if the gain is unity.

If in Fig. 15·20 the ratio of the time constants R_1C_1/R_2C_2 is sufficiently greater or less than unity, the phase shift will not be near enough to 180° for the pulse-to-pulse storage effect to make up the phase difference to 180°. Another method of accomplishing this result is shown in Fig. 15·21. Here the phase shift is prevented by R_3 from approaching 180°, providing R_3 is sufficiently large. The time-constant ratio

$$\frac{R_1C_1}{(R_2 + R_3)C_2}$$

may be made unity, and better cutoff characteristics may be obtained. This circuit was worked out for the receiver but not early enough to incorporate it in the final model in which a single *RC*-network was used. The network shown in Fig. 15·21 has considerable advantage over the

one actually used, since the sharper cutoff allows the gain of the feedback loop to vary more without making it oscillate at half the repetition frequency or without appreciably affecting the phase of the output error signal. Thus the characteristics of the i-f amplifier tubes may vary more without causing trouble.

In order to obtain an output signal of convenient size and still have high gain around the feedback loop, delay bias is provided (see Chap. 9). An advantage is gained by putting this bias after the filter instead of before it. The filter action removes some frequency components with the result that the amount of variation in voltage after the filter is small. When the AGC delay bias is applied after the filter, the bias will merely subtract a d-c voltage from the voltage applied to the i-f amplifier grids. At the input to the filter, however, there are present frequencies that are not yet removed and may be relatively large. If the bias is applied here, the peaks of these higher frequency voltages may go below the bias level and be cut off. During the intervals in which this voltage is cut off, no useful information is being applied to the filter. The loss of this information is a form of distortion that may seriously affect the direction-error signal and hence the automatic-tracking capability.

The purpose of the AGC is to prevent distortion, but it may cause considerable distortion if the feedback voltage is not applied to the i-f amplifier properly. If the feedback is applied to only one i-f tube, strong signals will produce so much bias that the tube will be cut off a large part of the time. This action causes much information to be lost and greatly increases the apparent modulation percentage. It is also inappropriate to apply the AGC bias to the first one or two amplifier stages, since this will impair the signal-to-noise ratio when signals are weak and noise is operating the AGC. On the other hand, control must be applied at a sufficiently early point in the amplifier to preclude the possibility of overload or limiting ahead of the control point. In the receiver under discussion the AGC voltage is applied to the tubes of four amplifier stages, the third, fifth, seventh, and tenth. This arrangement allows the signal to attain a reasonable level above noise, then holds the signal within a region around this level until the final amplification to the desired level for demodulation.

Since the value of m in Eq. (5) is that for the whole amplifier—the sum of the separate values for each of the tubes—its value would be more likely to remain constant if all four tubes were used in like manner. With this arrangement, however, the magnitude of the error-signal output is dangerously near the hum level. A safer margin is obtained if only two tubes are used. It was found that a good compromise consists of two filters, one that passes frequencies from zero to well above the error-signal frequency and one that passes frequencies from zero to a

cutoff well below the error-signal frequency. The AGC voltage for the seventh and tenth i-f tubes traverses only the filter with the high-frequency cutoff, and the AGC voltage for the third and fifth i-f tubes passes through both filters. The gain around the feedback loop at low frequencies is at least twice as great as at high frequencies, and the output is more constant at low frequencies, but the error-signal output has been doubled by this means.

15·9. Description of the Demodulator Circuit of the Receiver.—The purpose of the demodulator circuit of the receiver is to take in all received signals, select the pulses belonging to one signal, and from these selected pulses obtain voltage outputs to operate the direction servos and range servos as well as the voltage to control the gain of the receiver automatically. It also has provision for manually controlling the gain of the receiver.

The circuit is shown in detail in Fig. 15·22. The input signal with AGC operating is 0.75 volt positive for the signal that is controlling gain and about 1.5 volts maximum for any others and comes from a 75-ohm transmission line. The first two tubes are wide-band video-amplifier tubes as described in Sec. 15·7. The third stage of amplification also selects the signal from the desired target. This stage consists of two tubes with control grids operated in parallel and the plates operating two separate but similar circuits. These two amplifier tubes are normally nonconducting except at the time the desired signal arrives at the grids. During the nonconducting time the screen grids are held at a low positive voltage, and the suppressor grids are held at a negative voltage. This arrangement assures that the tubes will not conduct during the off period even if the signals are very strong at the control grids. Just before the desired signal occurs, a comparatively slow positive pulse of voltage, or gate, is applied to the screen grids by means of a 7F8 tube used as a cathode follower. The duration of this gate is slightly longer than the period of the signal. While this gate is on, a second gate supplied by a pulse transformer in a blocking-oscillator circuit is applied to the suppressor grids. This second gate has very steep sides and lasts for about half the period of the signal pulse. The relation of these various gates to the input and output signals is shown in Fig. 15·23. The signal pulse does not have so steep sides as it did on entering the receiver and lasts for about twice as long. Since the signal pulse at the input to the receiver is 0.1 μsec long, the pulse at the control grids of the gated tubes is about 0.2 μsec long from the start of the rise to the end of the fall of the pulse. The gate applied to the suppressor grids is about 0.1 μsec long. The gate is applied directly to one of the suppressors but is passed through a delay line of approximately 0.1 μsec to the other suppressor; therefore the end of the gate to the first suppressor occurs at the time of the start of the

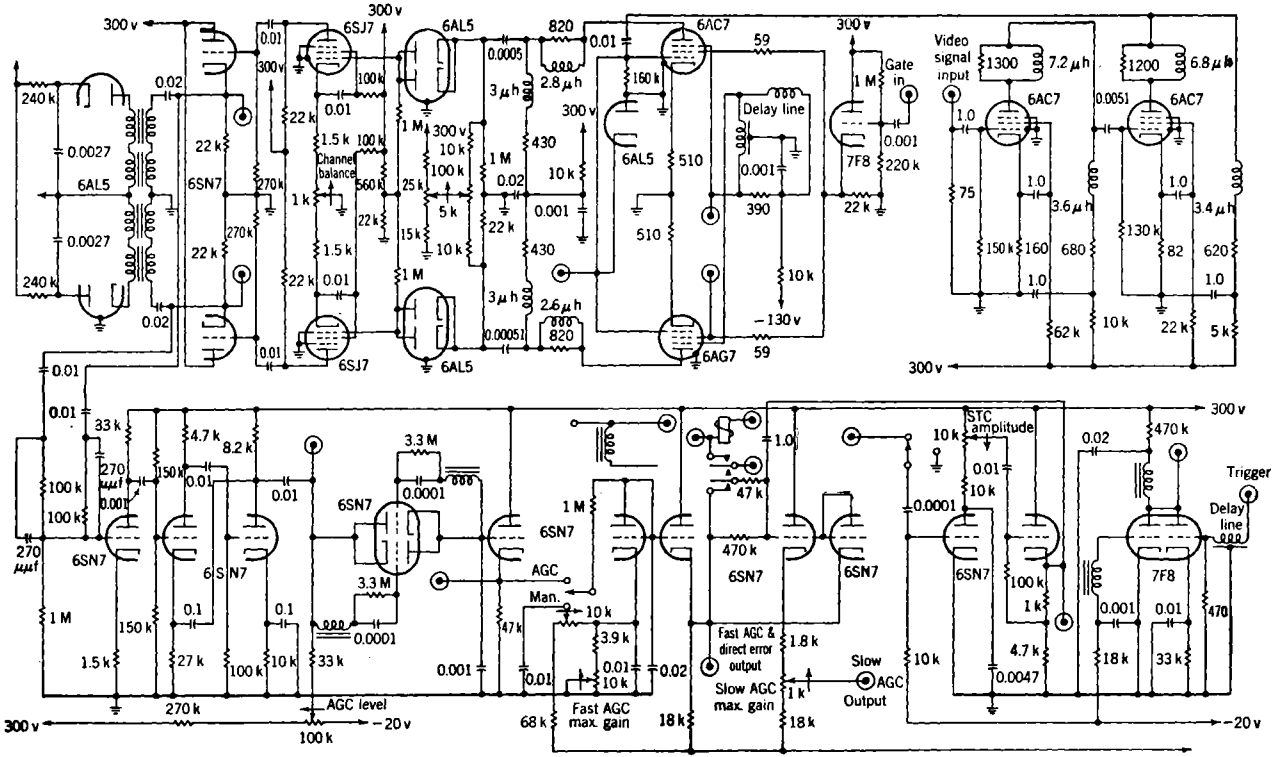


FIG. 15·22.—Demodulator circuit.

gate to the other suppressor. This point at which the end of one gate and the beginning of the other occurs is centered on the peak of the signal pulse applied to the control grids. Thus, the first half of the signal pulse is amplified by one tube, and the last half is amplified by the other. If the point of symmetry of the two gates should be shifted from the center of the signal pulse, for example, as a result of change in range of the target, the height and length of pulse output of one tube will be greater than those of the other. This difference in output is later used to operate the range servos for automatic range-tracking; in addition, the two outputs are also combined to furnish the AGC voltage.

The signal input to the grids of the gated tubes is composed of all received signals and noise. The average voltage input to these tubes therefore varies when any of the signals vary or when the amount of noise relative to the signals varies. The height of the signal from the desired target with respect to the average input voltage therefore varies not only when this signal itself varies but when any other signal or the relative amount of noise varies. To provide a baseline to which the desired signal can be referred and that is not appreciably affected by the other signals or noise, a d-c-restorer diode circuit is included in the grid circuit of the gated tubes.

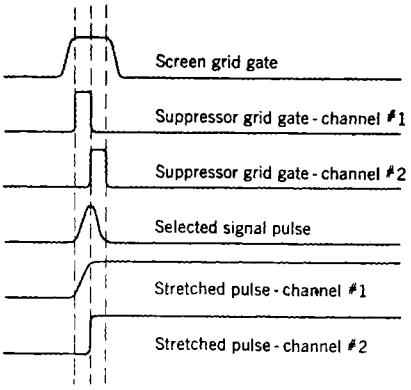


FIG. 15·23.—Gate voltages for signal selection and splitting.

This circuit makes the output of the gated tubes practically independent of all variations except those of the selected signal. For this purpose the time constant of the d-c restorer is moderately long, being about 1,600 μsec.

The output of the two gated tubes goes to two identical pulse-stretching circuits having capacitors that charge very rapidly and discharge slowly. The charging is done through diodes which cease to conduct as soon as the current through them attempts to reverse direction. The charging time of the circuit depends on the plate capacitance of the gated tube, the capacitance of the diode to ground, the grid capacitance of the following tube, the stray capacitance, the plate-load resistance, and the compensation, or peaking, used. The circuit must charge fast enough to have a charging bandwidth as wide as the other video circuits in order to operate in the presence of certain types of interference. This bandwidth is achieved in the same way as in the case of the other video stages.

The discharging time constant of the pulse-stretching circuit is determined by the output capacitance of the diode, the input capacitance of the following tube, the stray capacitance, and the diode load resistance. The diode load resistance must be smaller than the maximum allowable grid resistor of the following tube. In any event, values higher than 1 megohm are likely to be too vulnerable to humidity. The discharging time constant used in this unit is approximately 20 μsec.

The purpose of this pulse-stretching operation is to obtain a voltage proportional to and, as nearly as possible, equal to the peak voltage, the stretched voltage to be of sufficient time duration to allow the charging of a very long time-constant circuit with tubes of reasonable size. The time constant of this very long time-constant circuit must be such as to retain its charge essentially constant from one pulse to the next. The pulse-stretching and very long time-constant circuits combined with their associated circuits may be regarded as a peak voltmeter preceded by a two-stage pulse stretcher. For an input signal as fast as a 0.1-μsec pulse of a 25-Mc/sec wave, the output signal will be a stepped or "staircase" voltage, the height of each step from a baseline being proportional to the corresponding input pulse height or voltage. This circuit should not be confused with the so-called "boxcar" circuit, which has a voltage output momentarily going down to the baseline just before each pulse. Extra filtering is required to remove the high frequencies caused by the boxcar circuit, and thus the over-all response of the AGC circuit is slowed down.

The output of the gated tubes has a large component of the gate voltage. If this voltage were allowed to pass through the amplifier, it would severely tax the dynamic range of the amplifier tubes and require larger tubes. To eliminate this voltage from the rest of the amplifier the diodes of the pulse-stretching circuit are biased to a value that allows only enough to pass through to ensure that the whole signal pulse height is utilized.

The stretched pulses from the diode circuits are impressed on the grids of a pair of type 6SJ7-GT amplifier tubes, one in each of the two channels. The plate resistors are fairly high, since it is no longer necessary to preserve the large bandwidth of the video amplifier. It is necessary only that the stretched pulse reach its peak by the time that the gate for the staircase-voltage generator circuit is applied, which is about a microsecond later than the gates for signal selection. Advantage is taken of this time to obtain extra gain. The cathode resistors furnish sufficient feedback to prevent distortion and grid current. A variable resistor is provided in the cathode circuit to balance the gain of the two channels, and a variable bias on the diodes in the pulse stretcher permits equalization of the gate-pulse heights.

The output of these amplifiers is applied to the grids of two cathode followers which drive two circuits, the range discriminator and the remainder of the AGC circuit.

The range-discriminator circuit is fed through two pulse transformers. As described above, the selected signal is split into two parts by the gates of the selector. When the gates are set at the proper range, the two parts of the signal are of equal voltage; but when the gates are off in range, the two parts of the signal have unequal voltages. These two voltages are compared, and the difference is used to operate the range servos. The output signals of the cathode followers feeding the range discriminator are both positive. The sign of one of these voltages is preserved by the pulse transformer and applied to the plate of a diode; the sign of the other is reversed by its transformer and applied to the cathode of its diode. The terminals of the transformer secondaries that are not connected to the diodes are connected together and to one side of each of the diode output filter capacitors. This connection is used as the low potential side of the output circuit. The load resistors of the diodes are not connected to this low-potential terminal but are connected in series between the high-potential terminals of the diode outputs, and the junction of the two resistors is used as the high-potential output terminal of the circuit. If the pulse voltages applied to the diodes are equal, the diode output voltages will be equal and the junction of the two load resistors will be halfway between these two equal and opposite voltages, or at the same potential as the low-potential terminal of the circuit. If the pulse voltages applied to the two diodes are unequal, the junction of the two load resistors will have a potential with respect to the low-potential terminal of half the difference in the diode output voltages and of the sign of the higher diode output voltage. This is the range-error voltage and is used to operate the range servo circuits. Its sign and magnitude give the direction and amount, respectively, by which the selector gates are removed from the proper range of the signal. The action of this circuit is similar to the discriminator circuit shown in Fig. 15·3.

The cathode-follower tubes that drive the range-discriminator-diode pulse transformers also feed a network in parallel with these pulse transformers to supply input to the AGC circuits. This network consists of resistors and capacitors designed to present a reasonably high impedance to the cathode-follower tubes but to apply to the grid of an amplifier tube a pulse of voltage equal to the average pulse voltage of the two channels and of the same shape. The output of this amplifier is applied as a negative pulse to the grid of one of the tubes of a circuit that forms a power amplifier with very low output impedance. This circuit may be regarded as a two-tube analogue of a cathode follower, in which low out-

put impedance is secured for both positive-going and negative-going signals.

The low-impedance power amplifier is connected to a 0.001-μf capacitor through two tubes which form a switch that is turned on for only a very short time. A blocking oscillator is triggered immediately after the signal-pulse selector gates are turned off. This blocking oscillator turns on these switch tubes slightly after the stretched pulse reaches its peak at the grid of the low-impedance power amplifier. The switch tubes are kept on just long enough for the 0.001-μf capacitor to reach 90 to 95 per cent of the possible final value, which is about 3 μsec. The blocking oscillator then turns off the switch tubes until the next signal pulse is received one repetition period later. Thus the voltage across this 0.001-μf capacitor is the staircase voltage previously mentioned. The action is illustrated in Fig. 15·24.

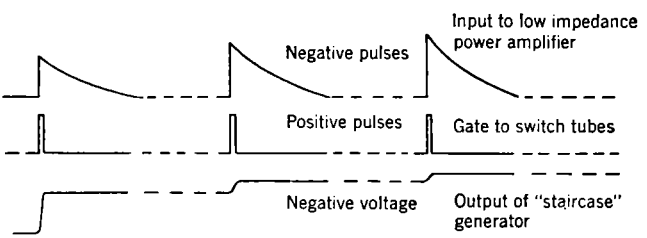


FIG. 15·24.—Action of pulse-stretching circuit.

The switch tubes are connected in reverse directions so that the 0.001-μf capacitor may either charge or discharge readily. The blocking oscillator furnishes sufficient voltage to drive the switch-tube grids highly positive no matter what the potentials of the cathodes may be. The impedance of the grid circuits, however, is made high by means of the 3.3-megohm grid-leak resistors and 100-$\mu\mu$f storage capacitors so that the blocking oscillator can only drive the grids to give a slight grid current. This prevents the blocking oscillator from determining the voltage of the 0.001-μf capacitor. The bias for the switch tubes is furnished by the grid leaks. The time constant of these grid circuits must be long enough to retain sufficient bias to keep the switch tubes cut off when not turned on by the blocking oscillator. This switching circuit is often called a *clamping circuit.*

The d-c level is set at this point by a voltage obtained from an AGC LEVEL control potentiometer on the input side of the switch tubes. The resistance of this voltage source must be low enough that the variable currents through the circuit do not appreciably affect the voltage. Variation of this voltage baseline might be serious if its magnitude and

the associated time constant should cause a phase shift in the direction-error signal voltage.

The staircase voltage is produced across the 0.001-μf capacitor, since the time constant between charging times is essentially infinite. This voltage is applied to the grid of a cathode-follower tube. When the AGC-MANUAL switch is in the AGC position, the output circuit of the cathode follower just mentioned is the fast-time-constant filter for the AGC. This filter consists of a 1-megohm resistor and a 0.02-μf capacitor. The output of the filter is applied to the grid of yet another cathode-follower tube. This cathode follower furnishes voltage to three circuits. It is the low-impedance source of AGC voltage for the i-f amplifier tubes operated through the feedback loop using the fast-time-constant AGC filter only. It furnishes a-c voltage through a capacitor to a tube and filter circuit which filters out the direction-error voltage and passes it on to the direction servos. It also furnishes voltage to the input of the long-time-constant filter for the AGC system.

A diode circuit is associated with the short-time-constant filter for the purpose of furnishing the AGC delay bias and for preventing the grids of the i-f tubes from going more positive than -2 volts. The FAST AGC MAX. GAIN resistor is normally adjusted by disconnecting the receiver video output from the input to this chassis and varying this resistor until the voltage at the fast AGC output terminal is -2 volts with respect to the chassis. A properly synchronized 0.1-μsec pulse of 0.75-volt magnitude is then applied to the input of the chassis where the output of the receiver normally connects and the range gate is properly centered on it. The AGC LEVEL control at the input to the switch tubes is then adjusted to give maximum negative voltage at the fast AGC output terminal. The control is then turned to reduce the fast AGC output voltage until it just reaches -2 volts. The receiver may now be reconnected to the chassis input. The circuits are now so adjusted that the AGC feedback loop will start to operate when the selected signal pulse voltage reaches 0.75 volt at the output of the receiver. The circuit operates as if this delay bias were not present unless the voltage at the plate of the diode at the *output* of the fast AGC filter becomes more positive than that of the cathode. This results in low distortion as explained in Sec. 15·8.

The slow AGC filter consists of a 330,000-ohm resistor and a 1.0-μf capacitor. The capacitor, instead of being connected from the resistor to ground, is connected from the resistor to the output of a cathode-follower tube which has a low output impedance. This cathode-follower tube is fed by a tube that generates a voltage varying in magnitude with time in such a manner as to bias the two slow AGC i-f tubes to give low receiver gain at the time the system transmits a pulse and to increase the

gain gradually with time in such a way that the echo signal from a target has a magnitude at the output of the receiver that is more or less independent of its range. This is called a *sensitivity time control* and reduces the amplitude of strong signals near by so that it is easier to watch the desired signal at close range on intensity-modulated indicators.

The output of the long-time-constant filter is applied to the grid of a cathode-follower tube in order to obtain a low impedance for supplying the AGC voltage to the i-f amplifier grids. Since the grid of the tube is more negative than the cathode, it is necessary to tap down on the cathode resistor somewhat to make the d-c potential applied to the slow AGC i-f tubes the same as that applied to the fast ones. This tapping point is made adjustable to take care of variation between tubes. It is adjusted to give the same d-c output voltage at both slow and fast AGC output terminals.

Manual gain control is secured by switching the input of the fast AGC filter to a potentiometer which furnishes the required variable voltage, manually controlled. Since the manually controlled voltage does not make use of the delay-bias diode, the diode bias circuit must have an impedance low compared with the 1-megohm filter resistor in order to prevent change in maximum gain caused by the change in current in this diode circuit when switching from manual to automatic gain or vice versa. The long-time-constant filter resistor is shunted by a 47,000-ohm resistor when manual gain control is used, to prevent sluggish action of the control.

The voltage gain from the video input circuit of this chassis to the AGC-voltage output terminals is about 20.

15·10. The Receiver Power Supply.—One of the parts of an automatic-tracking radar system that requires careful consideration is the receiver power supply. Voltage of the a-c power-supply frequency or of its harmonics or subharmonics entering the system may modulate the signals or otherwise affect the direction-error signal. The precision of the tracking of the target may be impaired, and the antenna beam may describe a circle about the target or even lose the target entirely.

Since the direction-error modulation of the signals in the receiver is at a frequency of approximately 29 cps, 60-cycle modulation from the power supply will beat with this and cause a resultant at 31 cps. The 29- and 31-cps signals will pass through the circuits and give a resultant that continually shifts in phase, even though the 29-cps component be large compared with the other. As the magnitude of the 31-cps signal approaches that of the proper error signal, the indifference of target-tracking grows. Beating of the harmonics of the 29-cps direction-error signals and the 60-cycle supply frequency will cause similar effects.

Power supplies that are satisfactory in the laboratory may not operate

properly in the field. Subharmonics of the power-supply frequency are usually not present in appreciable quantities on commercial lines. Low frequencies may be found, however, in considerable quantities if the impedance of the power lines is appreciable and if a fair amount of power is used to operate reciprocating machinery such as pumps. Frequencies around 2 to 6 cps are then often found in the outputs of filtered rectified power supplies.

Generators used in small isolated power plants usually have an asymmetrical field structure which causes subharmonics to be generated.

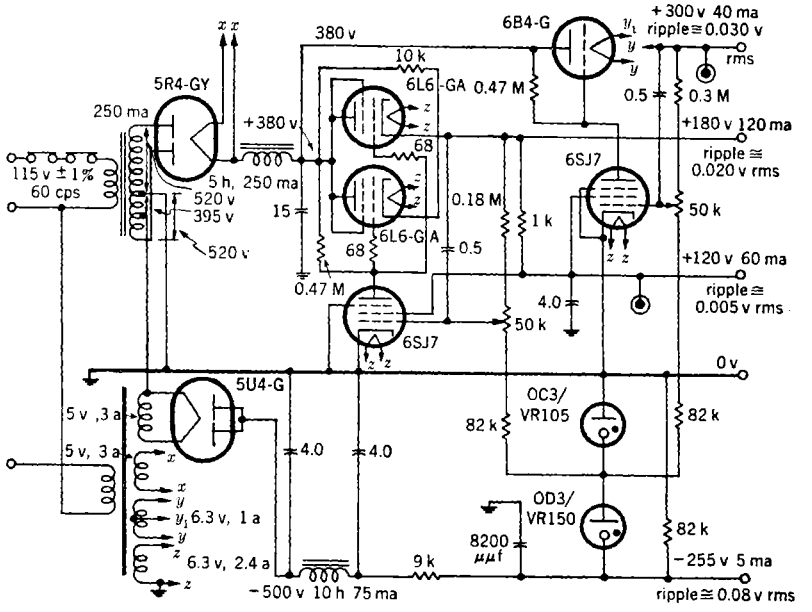


FIG. 15·25.—Power supply.

The frequencies of the subharmonics depend on the number of poles. A four-pole machine generates half the rated frequency; a six-pole machine generates one-half and one-third the rated frequency; and an eight-pole machine generates one-half and one-quarter of the rated frequency. An eight-pole 60-cycle generator may produce a small amount of 15-cps voltage. Since the filter of the rectifier of the receiver power supply is designed for use with a 60-cycle generator, the 15-cps ripple is much less attenuated by the filter and therefore may cause considerable trouble in the receiver circuits.

Filters for eliminating the low subharmonics are heavy and expensive. For these reasons and to prevent low-frequency feedback in the amplifiers

by way of the power supply, electronic voltage regulators are used in the receiver under discussion.

In Fig. 15·25 is shown the power supply for this receiver. It furnishes all a-c and d-c potentials used in the i-f amplifier-chassis part of the receiver. This includes the power for the i-f amplifier tubes, the second detector, the video tubes mounted on this chassis, the AFC circuit tubes, and the local oscillator. The voltages, currents, and ripple voltages are shown when the unit is operated from a good source of 60-cycle input such as a commercial line fed by a central station.

CHAPTER 16

A TWO-BAND 200-MC/SEC SUPERHETERODYNE RECEIVER

By A. B. Macnee

16·1. General Design Considerations.—The design requirements for this receiver were (1) that the sensitivity be the best possible, (2) that the over-all r-f bandwidth be 4.0 Mc/sec, (3) that the receiver be operable at preset spot frequencies in either of two bands, (4) that only one antenna be used for both bands, (5) that adequate gain-control range be available (more than 70 db), (6) that the receiver be capable of receiving pulses of $2\frac{1}{2}$- to 100-μsec duration, and (7) that the receiver provide two video signals, one limited at +40 volts and the other adjustable from 0 to +16 volts.

The two bands of operation selected were from 175 to 185 Mc/sec and from 205 to 215 Mc/sec.[1] The requirement that maximum sensitivity be obtained over these bands ruled out a crystal-video or superregenerative receiver and left the possibilities of a TRF or superheterodyne circuit. A superheterodyne circuit was used because of the smaller amount of r-f gain and the correspondingly fewer tuning controls required and because of the difficulty of obtaining the required pulse response in the trf receiver.

The amount of r-f gain necessary to obtain the over-all sensitivity was then calculated as described in Sec. 1·1. Because of its small physical size, short lead lengths, and excellent transconductance-to-capacity ratio, a 6AK5 was chosen for the mixer stage. With this tube operated as a triode, a noise figure of 13 db or better might be expected at 200 Mc/sec. Grounded-grid r-f stages were selected because they offered the best compromise then available between good noise figure and stability. Since the RCA 6J4 tube is especially designed for grounded-grid operation around 200 Mc/sec, it was chosen for the r-f amplifiers of this receiver. A noise figure of about 6.5 db might be expected for such an amplifier. If a deterioration of $\frac{1}{2}$ db is allowed because of the mixer, the r-f gain required is

$$\mathcal{W}_a = \frac{(20 - 1)}{0.56} = 34.0 \text{ or } 15.3 \qquad \text{db.}$$

[1] These are referred to as "*A*-band" and "*G*-band" respectively.

For a bandwidth of 4.0 Mc/sec, a gain per stage of about 10 db is to be expected from the 6J4's. Two r-f stages are therefore used.

The amount of i-f gain required was set by the consideration that it is desirable to have sufficient gain before the second detector of a superheterodyne receiver to allow it to operate always in a linear portion of its characteristic. Therefore, signals that reach the detector should be approximately 1 volt. The receiver must then have sufficient gain to bring its internal noise up to an rms value of about $\frac{1}{2}$ volt at the second detector. The rms noise voltage at the output terminal of an amplifier may be calculated from the expressions given in Chaps. 1 and 4. Thus, for a receiver having an r-f bandwidth of 4.0 Mc/sec, a noise figure of 7 db, and an antenna impedance of 50 ohms, the total gain required before the detector is

$$\mathcal{W} = \frac{\frac{1}{2}}{3.8 \times 10^{-6}} = 1.35 \times 10^5 = 103 \qquad \text{db.}$$

Since the total gain of the two r-f amplifiers and the mixer will be about 23 db, the i-f amplifier should have a net gain of at least 80 db. In order to provide some safety factor when low-g_m tubes are used, the normal i-f gain should probably be at least 90 db.

Since maximum sensitivity was a primary design consideration in this receiver, the i-f bandwidth was made considerably greater than might normally be expected, a value of 10 Mc/sec being used. By using such a wide i-f amplifier it was possible to design the r-f amplifier bandwidths to the narrowest possible value (4.0 Mc/sec) that gave the best possible noise figure. If a narrower i-f bandwidth had been used, it would have been necessary to widen the r-f bandwidth to maintain the over-all bandwidth at 4.0 Mc/sec, a change that would have caused some deterioration of the r-f noise figure.

The requirement on the video amplifier of this receiver was that it provide two limited signals either individually or simultaneously, one at a fixed level of +40 volts and the other at a variable level of from 0 to +16 volts. The required limiting action was obtained by using a negative-output detector at the end of the i-f strip to drive the first video stage to cutoff for large signals. The video tube chosen was a 6AC7 operating at zero bias and requiring about 2.0 volts to cut it off. The full output voltage of this stage was used to drive a 6AG7 as a cathode-follower output tube. In addition, a voltage divider was used to feed a portion of the output voltage of this stage to a second cathode follower which provided the required variable output voltage. To bring the signal up to the 40-volt level, a video gain of 24 db was required.

Since the requirement of two-band operation without retuning was imposed, two r-f heads were indicated. Simultaneous operation was not

required, however, and the two r-f heads were therefore duplexed into a single i-f amplifier. Switching between r-f heads was accomplished without physical changes merely by switching the voltages on the tubes. Inasmuch as a common antenna was specified, it was necessary that the two r-f heads be so designed as to give satisfactory sensitivity with some sort of antenna-duplexing or switching scheme.

Finally, since this receiver was part of an airborne system, the weight and power consumption were major items to be kept at a minimum. As in most airborne systems, the available power consisted of 28 volts direct current and 115 volts 400-cps alternating current.

R-F HEADS

The heart of this receiver and the point at which it differs most widely from the other receivers described in this book is the r-f head. An antenna was furnished having an internal impedance of essentially 50 ohms over a frequency range from 175 to 215 Mc/sec. The r-f heads,

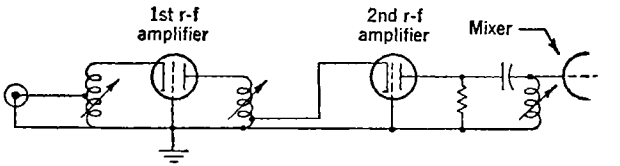


FIG. 16·1.—Equivalent r-f circuit for either of two amplifier sections.

therefore, were designed to operate from a source impedance of 50 ohms. The equivalent r-f circuit for either of the two amplifier sections is shown in Fig. 16·1.

16·2. R-f Amplifier.—The grounded-grid amplifier is essentially a unity current-gain device with a very low input impedance and a very high output impedance. Therefore, to obtain satisfactory gain with two cascaded grounded-grid stages it is necessary to provide some sort of impedance-matching device between them. In this case impedance matching was obtained by utilizing a coupling circuit of the autotransformer type. This circuit has good gain-bandwidth properties as compared with other simple impedance transformers. The amount of step-down in this transformer is best adjusted in practice by a trial-and-error method. With a signal generator on the input terminal of the first stage, the tap on the interstage coil was adjusted for maximum over-all gain. When this adjustment had been made, it was found that the loading of the second stage gave the interstage circuit a bandwidth of approximately 6.0 Mc/sec. This bandwidth is about right for an over-all bandwidth of 4.0 Mc/sec, since the two cascaded identical single-tuned circuits will have an over-all bandwidth of about 0.66 times their individual bandwidths. If this had not been the case, it would have been

necessary to add resistive loading or additional capacity to obtain the required bandwidth. The second grounded-grid stage operates directly into the grid of the mixer stage. A single-tuned circuit was used in this second interstage; and since the input impedance of the mixer stage is very high, it was necessary to add a load resistor to bring the over-all bandwidth up to 4.0 Mc/sec. The value of this load resistor was determined experimentally because of the difficulty of calculating the exact input impedance of the 6AK5 triode mixer with a combination of feedback and transit time loading. The tap of the antenna lead on the input coil was adjusted experimentally for the best signal-to-noise ratio. This adjustment can be made using a signal generator or a diode noise source (see Sec. 12·5). All three tuned circuits in Fig. 16·1 are adjustable through slug-tuned coils, the details of which will be given later.

16·3. Mixer.—The equivalent r-f circuit of the r-f converter is shown in Fig. 16·2. Here, as in Fig. 16·1, all d-c circuits have been omitted.

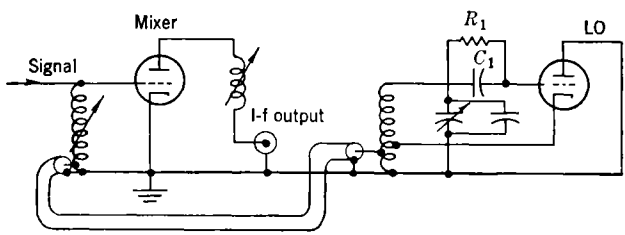


FIG. 16·2.—Equivalent r-f circuit of the mixer.

A 6AK5 connected as a triode is used as a mixer. A 6AK5 connected as a pentode was also tried in this position; but with the two r-f amplifiers used, little change was noted in either the over-all gain or noise figure. The triode connection was chosen because it required fewer components. The LO voltage is injected on the mixer grid. It is brought from the oscillator through a short length of coaxial line which taps into the mixer grid coil. The amount of LO excitation may be adjusted for maximum conversion gain by moving this tap up and down the mixer coil and observing the i-f output with a fixed r-f input signal.

16·4. Local Oscillator.—The local oscillator chosen for this receiver was a grounded-plate Hartley circuit. Satisfactory operation has been attained with this oscillator only with the local oscillator operating on the low-frequency side of the signal. This type of operation has certain drawbacks in that it increases the number and proximity of the image responses over what might be obtained if the local oscillator were on the high-frequency side. With the Hartley circuit, tendencies to instability and parasitic oscillations were observed when higher-frequency operation was attempted. The feedback in this oscillator is obtained through

mutual coupling in the tuning coil which becomes somewhat unreliable and variable when the coil shrinks to the size of two or three widely spaced turns.

A more satisfactory type of oscillator is the Colpitts circuit (Sec. 5·18). By the use of such a circuit with the same tube it is possible to build a satisfactory oscillator for operation up to 250 Mc/sec. A 9002 triode was used in the oscillator of Fig. 16·2. This tube was chosen rather than the 6AK5, used in so many other places in this receiver, because at the time this receiver was being designed there was a considerable shortage of the 6AK5 tubes. It was thought that they should not be used in any circuit where their superior characteristics were not fully utilized. If this receiver were to be redesigned, a 6AK5 connected as a triode would probably be used for the local oscillator.

Bias voltage for the local oscillator is obtained through the grid leak and condenser combination R_1C_1. The minimum value of R_1 that can be tolerated is determined by the maximum average plate current or plate dissipation. The maximum value of the time constant, R_1C_1 of this circuit, is determined by the point at which the tube will go into intermittent oscillation or superregeneration. Therefore, in adjusting the bias, R_1 is varied to obtain the desired tube current whereas C_1 is given the value necessary to prevent superregeneration. Oscillator tuning is accomplished with a small variable condenser. The use of a variable-inductance coil would not be satisfactory here because of the changes in feedback that would result. However, if a Colpitts circuit were used, inductance tuning would be entirely satisfactory. The excitation in the Hartley oscillator is determined by the location of the cathode tap. This is adjusted experimentally to give uniform operation over the entire tuning range.

After an experimental oscillator has been designed and built, the problem of frequency drift must be met. In the design and construction of the oscillator care should be taken to make all assemblies mechanically as strong as possible. First, a calibration run is made on the completed oscillator to determine the rate and direction of its frequency drift with temperature. Then from the data thus obtained a small compensating condenser is chosen to balance the observed drift of the oscillator. This compensating condenser is shown in Fig. 16·2 as the small fixed condenser in shunt with the oscillator-tuning condenser.

16·5. R-f Duplexing.—Besides using a common i-f amplifier, the r-f heads must operate satisfactorily from a common antenna. To accomplish this the duplexing scheme shown in Fig. 16·3 is used.

The length of cable W_1 is so chosen that with tube V_1 biased to cutoff, an open circuit is seen on A-band at the junction J. Similarly the length of W_5 is chosen to present an open circuit on G-band if the input tube to

the A head V_{12} is biased to cutoff. Switching relay K_3 connects a large negative voltage to the grid of the first stage of either r-f head and so shuts off the amplifier that is not being used in such a manner as not to load down the input circuit and cause deterioration of the receiver noise figure. In practice the cable lengths can be chosen to give an open circuit at only one frequency. For this receiver these cable lengths were cut for optimum performance at the center of the tuning ranges, 180 and 210

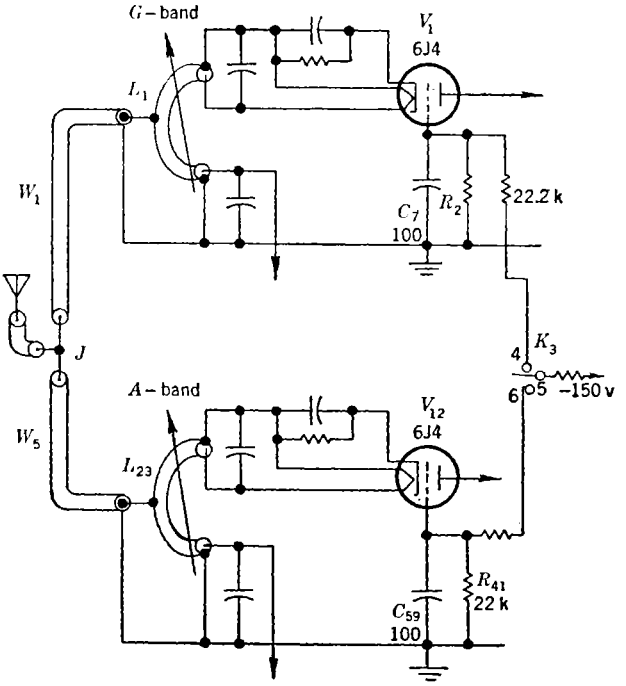


FIG. 16·3.—R-F duplexing circuit.

Mc/sec, and some deterioration in noise figure was allowed at the edges of the bands. The adjustment for optimum performance was made in the following manner. With a signal generator or noise source connected to the input of the receiver and one r-f head turned on, the length of the cable to the opposite r-f head was adjusted by a cut-and-try method until the best signal-to-noise performance was obtained from the operating r-f head. When this optimum was found, the receiver was switched to the opposite band and the second cable was adjusted by a similar cut-and-try technique. In making this adjustment, it is necessary to be certain that the signal generator or noise source is carefully matched into the input cable and that the fittings used throughout the cabling system

are of the constant-impedance type. If fittings that cause appreciable reflections are used or if the generator does not have the proper internal impedance, the two adjustments described above will not be independent of each other and it may be difficult or impossible to obtain the desired duplexing without sacrificing noise figure. In this receiver with type-N cable connectors, it was possible to keep the losses in noise figure due to duplexing to less than 0.5 db at the center of the tuning range and less than 0.8 db over the entire tuning range.

16·6. Construction.—A complete circuit diagram of an r-f head is shown in Fig. 16·4, and photographs of a completed head are shown in Fig.

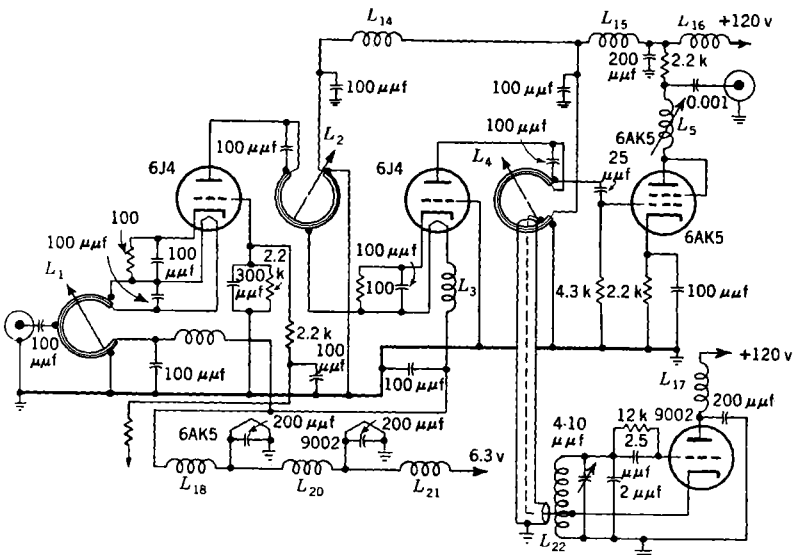


FIG. 16·4.—Circuit diagram of r-f head.

16·5*a* and *b*. The second r-f head is not shown, since both heads are identical except for the sizes of the tuning coils. All the tuning coils are designed especially for grounded-grid construction. In order to keep the cathode-heater capacity out of the tuned circuits, the heaters are run at the same r-f potential as the cathode. This is accomplished in the first stage by using a coaxial coil made of tubing with an insulated wire pulled through. The cathode-bias resistor and bypass condenser are mounted directly on the tube socket. One side of the heater circuit is grounded throughout the receiver. Thus, the outer conductor of the coaxial coil carries one heater return and the cathode current in addition to the signal currents, and the inner conductor carries the other heater

lead. In order to make certain that the inner and outer conductors of the coil are at the same r-f potential, silver-mica button condensers are used at either end of each coil. A photograph of a completed coil together with an enlarged section drawing of one terminal is shown in Fig. 16·6.[1]

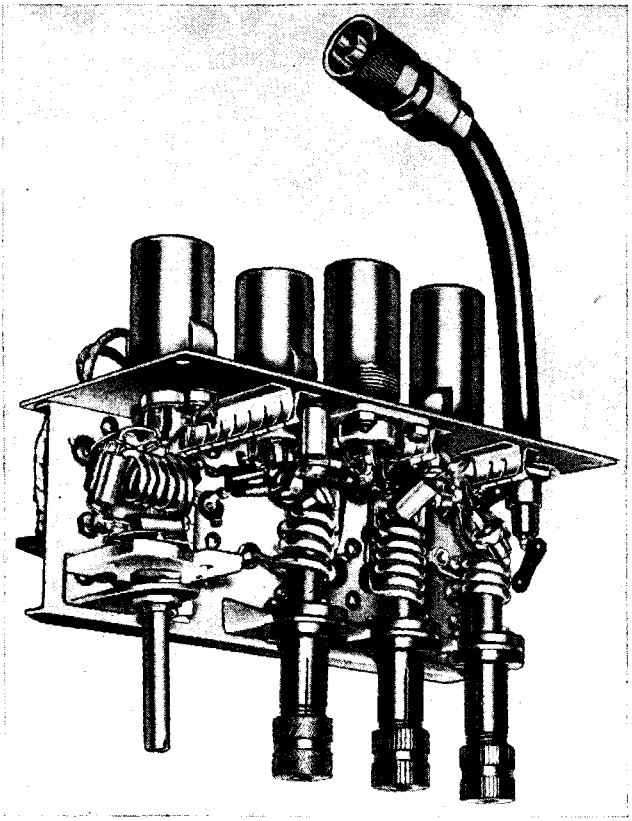


FIG. 16·5a.—Photograph of r-f head, internal view.

The cathode and heaters of the second grounded-grid stage are run off ground with the help of an r-f choke (L_5, Fig. 16·4), because it is not convenient to get a second coaxial coil into this single interstage circuit and because one coaxial coil is required for the plate circuit of the first stage.

The chassis construction of the r-f head with removable cover and shields provides a good compromise between ease of construction and

[1] The idea of using button condensers to bypass the ends of coaxial coils was suggested by E. K. Stodola of Camp Evans.

careful shielding. The shields are riveted onto the cover and then connected to the subchassis through small phosphor-bronze fingers. The chassis and cover are fabricated from brass and then silver-plated to provide the best possible connections. Through the use of the silver-mica feedthrough condensers as bypass units, the decoupling circuits in the heater, plate-voltage, and switching leads are completely isolated below the subchassis, as shown in Fig. 16·5*b*. The r-f chokes used in these decoupling circuits were not found to be critical in size and were wound with the aid of a high-frequency *Q*-meter to be approximately self-resonant at the center of the tuning range. All the adjustable coils are tuned with silver-plated brass slugs. The slugs have threaded shafts and run in split aluminum bushings. They can then be adjusted from the front panel of the receiver and locked in place with the knurled nuts shown in Fig. 16·5*a*. A smooth-bored split bushing is also used to lock the oscillator tuning condenser in position.

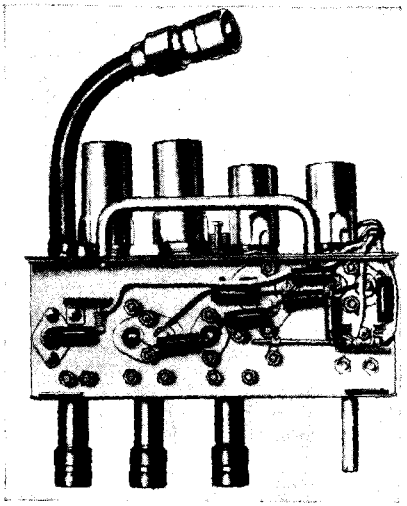


FIG. 16·5*b*.—Photograph of r-f head, back view showing filtering.



FIG. 16·6*a*.—Photograph of coil and one terminal.

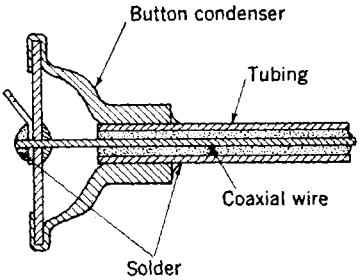


FIG. 16·6*b*.—Enlarged section drawing of one terminal of coil.

When the r-f heads are switched, interference from the local oscillators or the noise developed in the second amplifier and mixer is eliminated by turning off the plate voltage to the unused amplifier. The application of the negative voltage to the grid of the first amplifier is necessary, in addition, to provide the duplexing as previously described,

since otherwise the first amplifier would load the input circuit even though its plate voltage is turned off.

16·7. I-f Duplexing.—Since there are two r-f heads in this receiver and only a single i-f strip, some sort of duplexing scheme, such as that shown in Fig. 16·7, is needed to connect these three units. Each of the r-f heads goes by way of a 5-db T-pad to a tap on the input coil of the first i-f stage. The pad arrangement is located in the i-f amplifier chassis and is designed to have the same characteristic impedance as the coaxial cables connecting the r-f heads to the amplifier. The length of these interconnecting cables is thus rendered uncritical. In order to

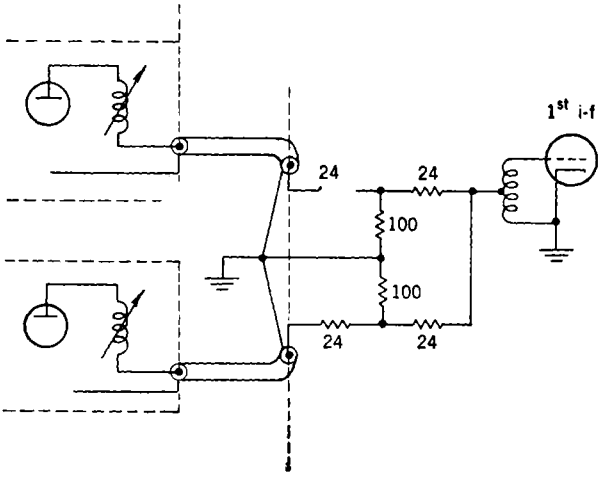


FIG. 16·7.—I-f duplexing circuit.

obtain a satisfactory gain from the output stage of the r-f heads into the low-impedance line (95 ohms), series-resonant matching circuits are used. Since the transfer impedance of this type of circuit is to a first order independent of the impedance of the load into which it works, it will not be affected to any marked degree by changes in the cable length or cable terminating impedance. The impedance, looking back into this type of coupling network from the load resistor (or in this case, the i-f cable), is found to be very nearly a short circuit at the band center. This short circuit would cause a severe distortion of the pass band of the opposite r-f head except for the total of 10-db attenuation provided between the two heads by the pad arrangement.

An alternative coupling circuit, which will work well in situations where the cables between the r-f and i-f units can be kept short, is shown in Fig. 16·8.

In this circuit the output capacity of the two r-f mixers, the interconnecting cables, and whatever stray capacities may be present are all tuned out by the i-f input coil L_1. In order to maintain the required bandwidth with this circuit it is necessary to add resistive loading in the form of R_1. The entire coupling circuit then is equivalent to a simple single-tuned circuit. The maximum transfer impedance that will be obtainable with this circuit will depend upon the bandwidth desired and the amount of capacity present. This circuit will give better results if the interconnecting cables can be kept short as compared with the circuit of Fig. 16·7 and their capacities correspondingly low. If, on the

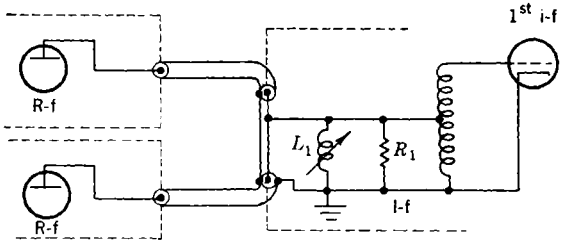


FIG. 16·8.—Alternative i-f duplexing scheme.

other hand, long cables are necessary, the circuit of Fig. 16·7 will probably prove more satisfactory.

I-F AND VIDEO AMPLIFIERS

16·8. I-f Strip.—Because 30 Mc/sec was an intermediate frequency in common use with many radar sets, it was chosen for this receiver. Subsequent developments have indicated that an intermediate frequency of about 60 Mc/sec would have been very helpful in combating the serious image problem that is encountered. Depending upon the r-f bandwidths desired and the image rejection required, either a higher or a lower intermediate frequency might be chosen for a future set. As has been previously indicated, the best noise figure can be obtained from an r-f amplifier by making it as narrow as possible. For this receiver, therefore, the r-f heads were made with as narrow a band as the over-all bandwidth (4.0 Mc/sec) requirement would allow. Then in order to make certain that the i-f bandwidth would not further narrow the over-all bandwidth, it was made 10.0 Mc/sec wide. Although this technique is most desirable in order to obtain the best possible noise figure, it results in a selectivity characteristic that does not drop off as rapidly outside the pass band as might otherwise be possible.

The best pentode tube available from the standpoint of gain-bandwidth product at the time this receiver was designed was the 6AK5 for which $g_m/2\pi C = 66.6$ Mc/sec (Sec. 6·5). This tube also has the advan-

tage of being small and having a fairly low power consumption. For an over-all amplifier gain of 100 db the optimum bandwidth for synchronous single-tuned stages is obtained by using 23 tubes and the over-all bandwidth is only 7 Mc/sec. In addition to requiring a prohibitive number of tubes, the amplifier obtained in this manner is narrower than the required 10.0 Mc/sec. Thus, for an amplifier of the bandwidth and over-all gain required, synchronous single-tuned circuits are out of the question. By using the scheme of staggered tuning of single-tuned circuits, however, it is possible to obtain an over-all gain of 90 db and an over-all bandwidth of 10.0 Mc/sec using only six 6AK5 stages. This is accomplished by using two stagger-tuned triples.

The 6AK5's are assumed to have a g_m of 5000 μmhos and an interstage capacity C of 12 $\mu\mu$f (see Sec. 6·5). The required stage gain is $\frac{90}{6}$ = 15 db, or 5.6x. Under the stated assumptions 6AK5's have a $g_m/2\pi C$ of 66.3 Mc/sec. A stagger-tuned triple will have the same over-all bandwidth as a single-tuned stage having the same stage gain. Thus, the bandwidth of each triple will be 66.3/5.6 = 11.8 Mc/sec. The complete amplifier is to be made of two such triples, which will have an over-all bandwidth of 0.86 times the single-triple bandwidth, or 10.2 Mc/sec. It is now necessary to determine the circuit constants for each 11.8-Mc/sec-wide triple. For each triple, $\Delta f/f_0 = \delta = 11.8/30 = 0.395$; from Fig. 6·4, Chap. 6,[1] $d = 0.200$ and $\alpha = 1.18$. Thus each triple consists of two stages of dissipation factor 0.200 stagger-tuned at

$$30 \times 1.18 = 35.4 \text{ Mc/sec}$$

and $(1/1.18) \times 30 = 25.4$ Mc/sec and a centered stage at 30 Mc/sec, having a dissipation factor of 0.395. From the dissipation factor and center frequency of each stage in the triple, the load resistors may be calculated according to the expression

$$R = \frac{1}{2\pi f_0 C d}, \tag{1}$$

where f_0 = the stage center frequency,
d = the stage dissipation factor,
C = the interstage capacity.

The results for each triple are shown in Table 16·1. The values of load resistance given in this table assume that all the interstage loading is furnished by these resistors. The actual value of the load resistors will be somewhat higher than the table indicates to take into account the interstage loading due to Miller effect, cathode-lead-inductance feedback, and coil losses.

[1] See also H. Wallman, RL Report No. 524, Feb. 23, 1944.

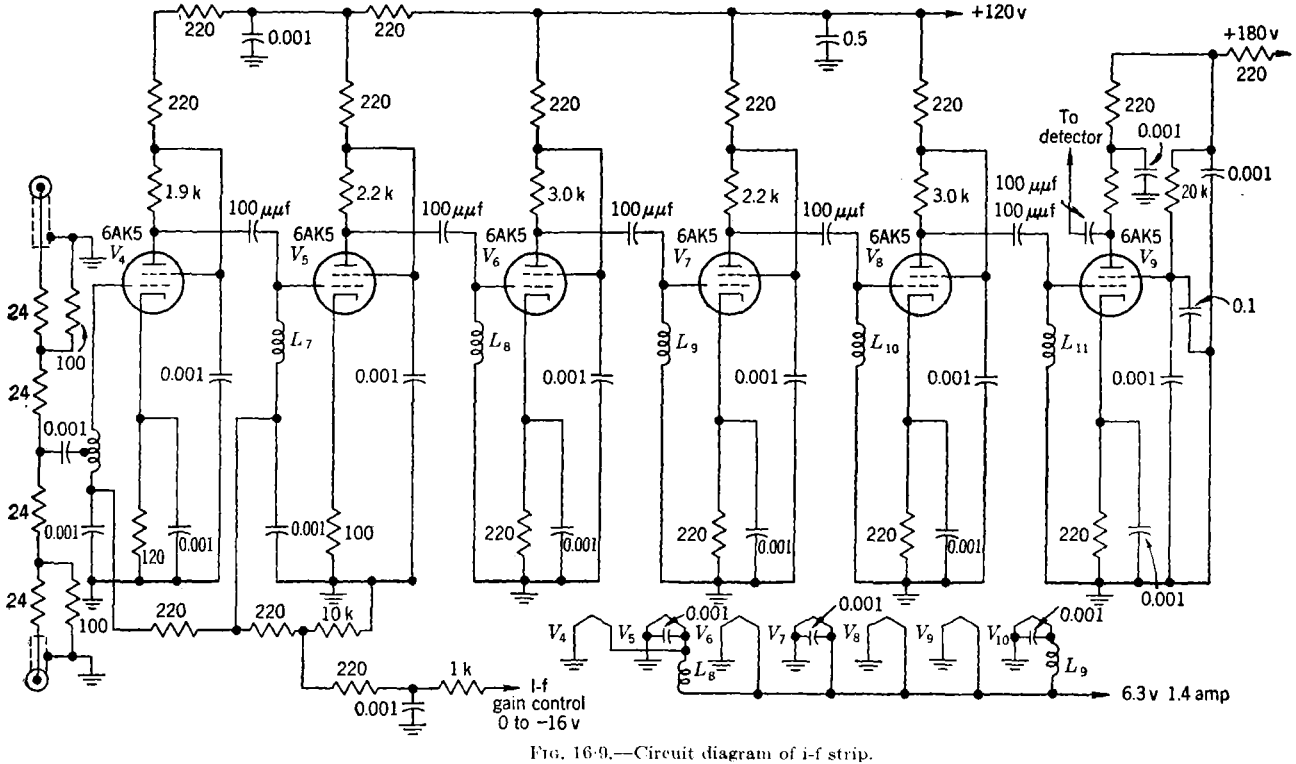


FIG. 16·9.—Circuit diagram of i-f strip.

TABLE 16·1.—DESIGN PARAMETERS FOR STAGGER-TUNED TRIPLE

f_0, Mc/sec	d	Δf, Mc/sec	Load resistance, ohms
25.4	0.200	5.1	2610
30.0	0.395	11.8	1120
35.4	0.200	7.1	1880

The circuit of the actual i-f strip for this receiver is shown in Fig. 16·9. It is apparent that the load resistors actually used are considerably higher than the values of Table 16·1. The first and last interstage circuits of the amplifier are tuned to the center frequency of 30 Mc/sec. The second and fourth stages are tuned to 35.4 Mc/sec, and the third and fifth stages are tuned to 25.4 Mc/sec. No particular significance is to be attached to this order, since the tuned circuits might well be arranged in any other manner. It does usually seem desirable, however, to make the input and output stages the centered stages. Since the exact calculation of the load resistors is not feasible, the line-up procedure is to choose experimentally the load resistor that gives the required bandwidth at the proper center frequency for each single-tuned circuit. It is for this reason that the individual-stage bandwidths were listed in Table 16·1. This experimental adjustment is most easily made with the aid of a swept-frequency signal generator as indicated in Sec. 12·2.

This i-f amplifier was designed to be entirely fixed-tuned. In order that variations in tube capacity may have a minimum effect upon the pass band of the finished amplifier, the adjustment of individual circuits for bandwidth and center frequency should be made with a special set of tubes chosen to have capacity values lying at the center of the possible capacity variations for the tubes. As the loading of the tuned circuit preceding the detector is adjusted, the bandwidth of this circuit will vary with the level of the measuring signal because of the variable loading of the diode detector. The circuit is generally adjusted to have the proper bandwidth with the detector operating at or near its normal operating level; for the receiver under construction, this is an output of about 1.5 to 2.0 volts.

The gain control for this i-f strip is obtained by introducing a negative voltage to the grids of the first two tubes. This voltage can be made sufficiently large to cut off the flow of current in these tubes completely. Since the amount of gain change that can be had in this manner is about 40 db per stage, the control of two tubes provides a gain-control range of about 80 db. If a wider range of gain control were required, a third stage would have to be controlled. Because the shape of the over-all pass band is the product of the shapes of the individual tuned circuits, varying the

level of the gain control will not affect the shape of the over-all pass band as long as it does not affect the characteristics of the individual stages controlled.

Unfortunately, however, this method changes the tube gain by varying the space-charge conditions inside the tube and so influences the characteristic of the tuned circuit connected to the tube grid. The most pronounced effect, if there is no feedback in the stage, will be variation in input capacity. It is found that the input capacity of typical tubes decreases by about 1.5 $\mu\mu$f as they are varied from full gain to minimum gain. This change in capacity will be found very nearly proportional to the g_m of the tube and so may be compensated for by a feedback scheme that introduces a reduction in capacity proportional to g_m. One such feedback scheme is to introduce in the cathode lead of the gain-controlled stage a small unbypassed resistance that introduces a reactive admittance into the grid circuit given by the expression

$$jB = -j\omega g_m C_{gc} R_k, \tag{2}$$

where C_{gc} = the tube grid-cathode capacity,
R_k = the resistance in the cathode lead.

By proper choice of the value of R_k this feedback may be made to compensate for the normal input-capacity change for a considerable range of gain change. This feedback, however, reduces the normal gain of the compensated stage. The second stage of the i-f amplifier in Fig. 16·9 has this sort of compensation. If more d-c cathode bias is required than will be furnished by R_k, a second bypassed cathode resistor may be added in series with R_k to make up the required value.

A bottom view of the i-f amplifier with its cover removed is shown in Fig. 16·10. Of primary interest is the arrangement of the individual stage grounds, the decoupling circuits used, and the insertion of the small shields to reduce the waveguide feedback (see Sec. 10·4). If these shields are to be effective, they must be solidly grounded on at least two sides. In reducing feedback due to stray ground currents between stages, it is extremely important that the individual stages be grounded as indicated in Fig. 16·9. The grid circuit, cathode bypass, screen bypass, and plate circuit should all be grounded to a single point for each stage. The arrangement of parts with respect to the socket used to accomplish this is seen in Fig. 16·10. To prevent feedback between stages due to the fact that a common plate- and screen-voltage supply is used, decoupling circuits are introduced between each tube screen and the common supply line. In addition, for the first two stages, which are at the lowest level and thus most susceptible to feedback, the voltage-supply line has been broken up with two additional decoupling circuits. All plate-decoupling circuits in this amplifier are simple RC-filters using 1000-$\mu\mu$f condensers

and 220-ohm resistors. If greater isolation were required or if high currents were being carried, r-f chokes would probably have been substituted for the 220-ohm resistors. One of the most likely sources of feedback in a high-gain i-f strip is the gain-control circuit for the first two stages. To prevent this feedback the gain-control circuits have been carefully filtered. In addition, as soon as possible the gain-control lead is run away from the input of the intermediate frequency through a shielded lead. Experimentally, the i-f amplifier is tested for regeneration by observing its overall pass band with a swept-frequency signal generator and oscilloscope as the position of the gain control is varied. If the amplifier is free from regeneration, the shape of the pass band will not change appreciably as the gain control is varied from full gain to minimum gain. It is desirable to make the limit level of the last i-f stage as high as possible. This is

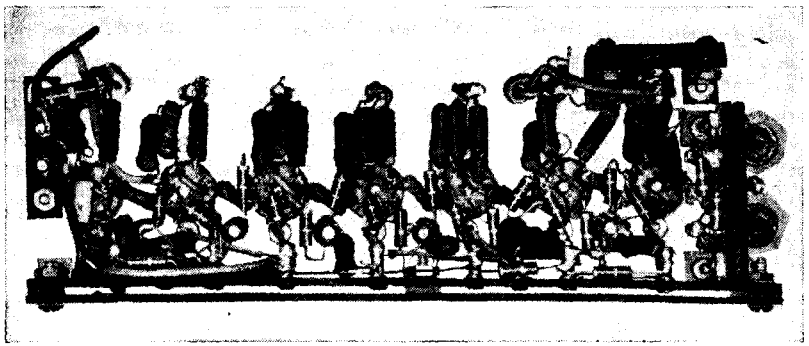


FIG. 16·10.—Bottom view of i-f amplifier.

accomplished by running the last stage at the highest plate and screen voltages allowable. These voltage levels necessitate the use of a higher voltage supply and separate decoupling circuits for the last i-f stage of the receiver.

As has been pointed out, the i-f amplifier was designed to be entirely fixed-tuned. The results obtained in this manner were satisfactory because the i-f bandwidth was so much greater than the r-f bandwidth that some change in the shape of the i-f pass band with various sets of tubes (usually in the form of tilting of the pass band) could be tolerated. If the amplifier were to be used in an application where more critical requirements were put on the shape of the pass band, it would be desirable to introduce a single tunable circuit for trimming purposes. This tuning could be accomplished with a single slug-tuned coil, which would prove most effective if inserted into the input or first interstage circuit.

16·9. Video Amplifiers.—The second detector is one-half of a 6AL5 diode arranged to give a negative output signal. This signal is fed to

the video amplifier where it is amplified, limited, and fed to output cables. A circuit diagram of the video amplifier and the detector is shown in Fig. 16·11. The 6AG7 provides a fixed output level of +40 volts limited, whereas the 6AK5 provides a variable output of from 0 to +16 volts limited. The inductance L_{13} in the plate lead of the detector serves the double purpose of providing some video peaking and acting as a choke

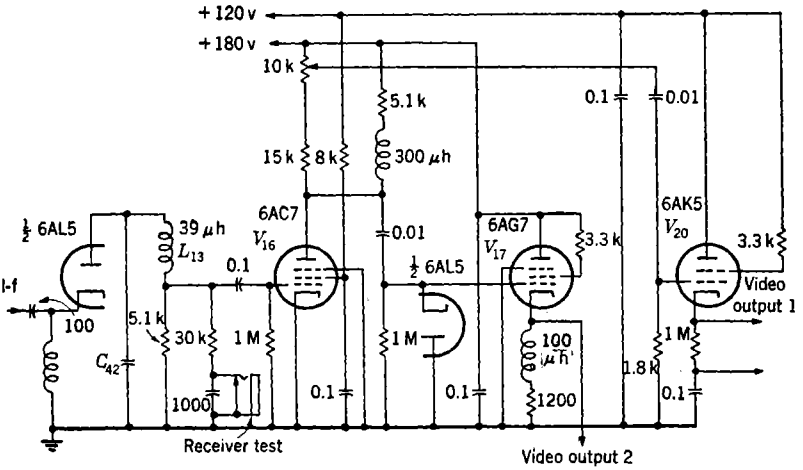


FIG. 16·11.—Circuit diagram of video amplifier and i-f detector.

for intermediate frequency. The condenser C_{42} is introduced to improve the rectification efficiency of the detector. The test jack in series with a 30,000-ohm resistor is inserted to provide a means of observing the d-c output of the detector. For testing purposes a 100-μa meter is inserted in the jack as an output meter. This meter is used for tuning the r-f head and measuring the over-all noise figure and bandwidth.

As was previously indicated, the shortest pulse that this receiver was expected to handle is 2.5 μsec long. The wide bandwidth in the r-f circuits was required because of instability in associated equipment rather than by the need for fast response. As a result the video was designed to have a rise time of 0.5 μsec, which should be sufficient for reproduction of pulses 2.5 μsec long.

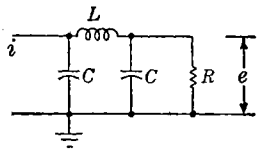


FIG. 16·12.—Coupling circuit.

At high frequencies the coupling circuit between the second detector and the first video stage is equivalent to the circuit shown in Fig. 16·12.

Considerable improvement in the gain-bandwidth product of this circuit can be had by suitably choosing the value of the series inductance

$L = 2CR^2$. In this amplifier, however, the coupling circuit was sufficiently wide without any peaking. Therefore, L was made only large enough, 30 μh, to serve as a good choke for the intermediate frequencies. The total capacity in this circuit is about 30 $\mu\mu$f, which with the 5100-ohm load resistor and without any peaking gives a bandwidth for this circuit of $160/(30 \times 5.1) = 1.05$ Mc/sec.

The interstage circuit between V_{16} and V_{17} is equivalent at high frequencies to the circuit shown in Fig. 16·13.

Since the impedance of the voltage divider feeding the second stage is high, it may be neglected. This circuit is then a common shunt-peaked video circuit (see Chap. 8). An improvement over the simple RC-circuit of about 2 in the figure of merit can be obtained by making $L = CR^2/2$. For a 6AC7 driving a 6AG7 the total interstage capacity is about 25 $\mu\mu$f. The value of the load resistor in this interstage circuit is determined by the gain required. Because it is extremely difficult to maintain the low-frequency response of video stages using a by-passed cathode-bias resistor without resorting to prohibitively large values of condensers, the first video stage is run at zero bias. Under this condition the maximum currents that can be handled by the tube are determined by the screen dissipation. With the screen running at very nearly its rated dissipation and with the maximum plate voltage of about 180 volts available, the no-signal current to the 6AC7 is about 11 ma. After some loss in the output cathode followers and some variability due to tubes were taken into account, the load resistor was chosen to be 5100 ohms. For this value of resistance the optimum peaking coil is calculated to be

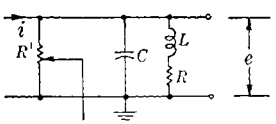


FIG. 16·13.—Interstage circuit.

$$L = (25 \times 10^{-12}) \frac{(5.1 \times 10^3)^2}{2} = 325 \times 10^{-6} \ \mu\text{h}.$$

A coil of 300 μh is sufficiently close to this value. Since the voltage divider used has an impedance of 25,000 ohms, the net interstage impedance at low frequencies is $(25)(5.1)/(25 + 5.1) = 4200$ ohms. Therefore, the nominal output voltage of this stage with a limiting signal (one that will completely bias off the tube) is $11 \times 10^{-3} \times 4.2 \times 10^{+3} = 46.2$ volts. The bandwidth of this interstage circuit is about twice the unpeaked bandwidth, or $2 \times 160/25 \times 5.1 = 2.5$Mc/sec. The peaking network in the output circuit of the high-voltage cathode follower V_{17} is the same as the circuit shown in Fig. 16·13. The voltage gain of the cathode-follower circuit is

$$\mathcal{G} = \frac{g_m R_k}{(1 + g_m R_k)}. \tag{3}$$

For $g_m = 9000$ μmhos and $R_k = 1200$ ohms, this is

$$\mathcal{G} = \frac{10.8}{11.8} = 0.915.$$

Therefore, the normal output voltage for a limiting signal will be $0.915 \times 46.2 = 42.3$ volts. The total capacity of the cable and load to be fed is 140 $\mu\mu$f, so a peaking coil of 100 μh is required. The bandwidth of this circuit is then $2 \times \frac{16}{140} \times 1.2 = 1.9$ Mc/sec.

The variable output voltage was obtained through an adjustable voltage divider consisting of a 15,000-ohm fixed resistor in series with a 10,000-ohm potentiometer. The variable voltage obtained from the arm of this potentiometer is fed to the grid of a second cathode follower V_{20}, which is designed to feed a capacity load of about 80 $\mu\mu$f. Because of the lower capacity load, no peaking circuit is used in the output circuit of this stage.

The low-frequency response of the video amplifier or its response to long pulses is determined by the time constants of the various interstage coupling condensers with their respective grid resistors and by the time constant of the screen bypass condenser and dropping resistor for the first video stage. Since the longest signal that this receiver is expected to reproduce faithfully is a 100-μsec pulse, these time constants were all made considerably longer than 100 μsec. The time constant of the coupling circuit from the detector to the first video grid has to be made somewhat longer than those in the grids of the cathode followers or the screen circuit, because any change in the voltage across this coupling condenser during a long pulse will be amplified by V_{16} and thus have a greater effect on the output signal.

When no signals are present in the receiver, the high-voltage cathode follower V_{17} normally operates at a rather high value of bias. The normal operating point is then on a very nonlinear portion of the tube characteristic. When signals are present in the amplifier, however, the d-c bias on the grid of this tube tends to change, thus causing both a shift in the operating point of the tube and, because of the nonlinearity of the operating region, a change in the gain of the stage and in the output-signal level. To prevent the change in output-signal level, the other half of the 6AL5 diode is connected from the grid of V_{17} to ground in such a manner as to prevent the change in the operating point of the grid. Such a tube is commonly called a "d-c restorer."

16·10. Power Supply.—In addition to the power necessary for the various heaters, this receiver requires three supplies of d-c power. The r-f heads, five of the i-f stages, and the screen of the first video stage require +120 volts. The total current drain is about 120 ma. The last i-f stage, the high-voltage cathode follower, and the plate of the first

video stage require +180 volts at a total drain of about 45 ma. Finally a negative supply is needed to provide a switching voltage for the r-f

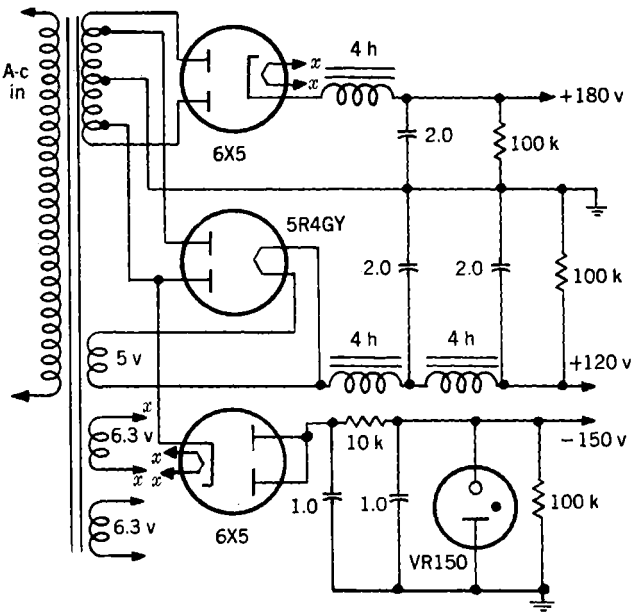


FIG. 16·14.—Power-supply circuit.

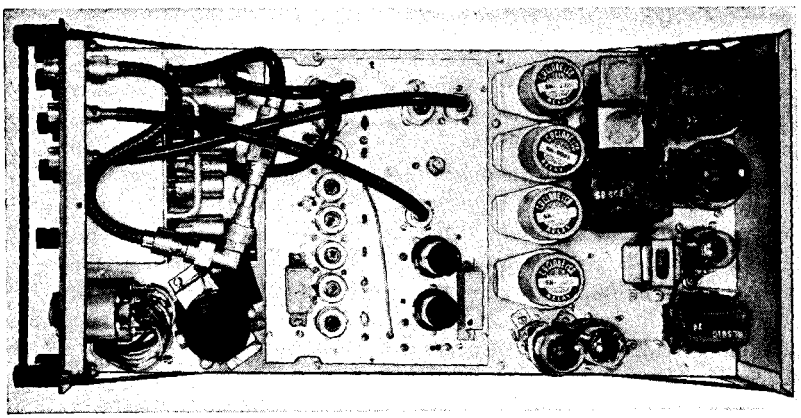


FIG. 16·15.—Complete receiver.

head duplexing circuit and to provide a bias for the gain control on the first two i-f stages. Since this voltage is to be used for the gain control, it should be regulated to prevent changes in line voltage from seriously

affecting the gain of the receiver. Actually, if good regulation with respect to line-voltage changes were required, it would be necessary to regulate the voltages to the screens of the i-f and video amplifiers as well. A single transformer with a tapped high-voltage winding is used to supply three rectifiers which make up the three d-c supplies needed. The circuit of the complete supply is given in Fig. 16·14. Since the current drain on the negative supply is only about 5 ma, a half-wave rectifier is used. Filtering and regulation are obtained through a simple *RC*-network and the VR-150 voltage-regulator tube. The other two supplies are full-wave rectifiers filtered by choke-input *LC*-filters. The 120-volt supply uses a 5R4GY because the 6X5GT does not have a sufficiently high current-carrying capacity. Since this supply feeds all the low-level stages in the receiver, it requires more filtering than the 180-volt supply, which supplies only the output stages. The 100,000-ohm bleeder resistors across all of the supplies serve to discharge the filter condensers. Figure 16·15 shows a photograph of the completed receiver.

CHAPTER 17

WIDE-BAND F-M RECEIVER

By H. Logemann, Jr.

The work to be described in this chapter was occasioned by the need for retransmitting radar video signals over a radio link. It was desired that this link should introduce no appreciable degradation of the signals in either frequency characteristics or noise. The system chosen was frequency modulation with a deviation ratio not much greater than unity. The difficulties encountered in conventional radar receivers as a result of the wide dynamic signal range are, of course, eliminated in this case, since the radar video signals are already limited. Certain special problems were however encountered in the application of pulse signals to an f-m system.

GENERAL REMARKS REGARDING THE DESIGN OF F-M RECEIVERS

17·1. Noise Considerations.—A good insight to f-m receiver performance can be gained by reference to the work of M. G. Crosby.[1] An extension of his work, by H. M. James,[2] includes c-w and pulsed interference problems. In general, they have shown that at low signal levels or strong interference, the a-m system is to be preferred to the f-m system. Frequency modulation, however, does show better performance at high signal levels, and in particular, a large-deviation-ratio f-m system has a decided advantage over amplitude modulation against strong impulse noise.

The relative performance of f-m and a-m receivers on the basis of peak signal and peak fluctuation noise is shown in Fig. 17·1. These curves were derived by adding the mathematical expressions for the f-m carrier and noise, passing this voltage through an ideal limiter by making the amplitude coefficients constant, and then plotting the resultant peak carrier-to-noise ratio as a function of the peak signal-to-noise ratio. The term "carrier-to-noise" ratio is used to denote the signal-to-noise ratio before the second detector in an a-m system or before the limiter and discriminator in an f-m system, whereas "signal to noise" refers to the signal-

[1] M. G. Crosby, "Frequency-modulation Noise Characteristics," *Proc. IRE*, **25**, 472-514, April 1937.

[2] H. M. James, "Notes on Noise, Interference, and Distortion in F-m Transmissions," RL Group Report No. 43-3/23/45.

to-noise ratio after detection. In an a-m receiver, the signal-to-noise ratio is the same as the carrier-to-noise ratio, at least for carrier-to-noise ratio > 0 db, as is indicated in Fig. 17·1 by the straight line through the origin at 45°. However, in frequency modulation the signal-to-noise ratio is greater than the carrier-to-noise ratio at large signal levels. In Fig. 17·1 it is seen that at a carrier-to-noise ratio on the order of 20 db, an f-m receiver with a deviation ratio of 4 has a signal-to-noise ratio approximately 16 db above the signal-to-noise ratio of the a-m receiver and that with a deviation ratio of unity the improvement is about 5 db. This improvement is seen to be constant for high signal levels but then drops suddenly as the signal level is decreased. The point at which the sudden drop crosses the a-m curve is called the "improvement threshold." Note that with a large deviation ratio there is a greater improvement in signal to noise but that the required signal level to reach the improvement threshold is less for the small deviation ratio.

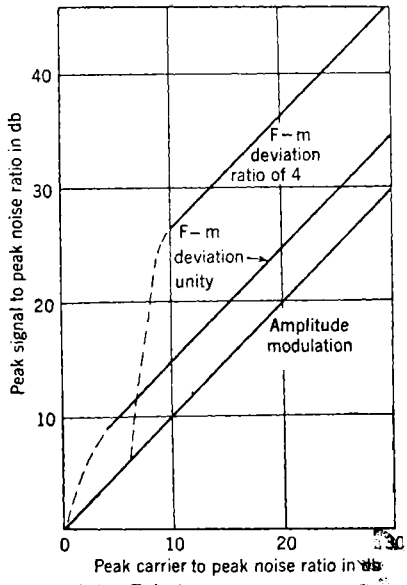


FIG. 17·1.—Relative performance of f-m and a-m receivers.

This is really an oversimplified picture of the relative performance of f-m and a-m receivers, as it considers only peak signal-to-noise conditions. Other factors, such as interference and distortion, come into the picture to alter the appraisal of the over-all system from that given here. Reference should be made to the work of Crosby and James mentioned earlier for a more complete discussion of these relative merits.

17·2. Gain Requirements.—It is assumed that the f-m receiver is to be used in a system where the carrier-to-noise ratio is always well above the improvement threshold, as this is the only case where the better performance of frequency modulation over amplitude modulation is realized. This is equivalent to the statement that the ratio of the input carrier signal level to the equivalent noise level at the input circuit is at least as great as the improvement threshold. Proper operation of an f-m system requires that there be sufficient i-f gain to bring any usable input carrier level up to the point at which limiting occurs. In other words,

the ratio of the voltage at the limiter grid that just produces limiting to the noise voltage at the same point should again be equal to or somewhat less than the improvement threshold. If the ratio is greater, some potentially usable carriers will not be limited and no noise reduction will be obtained for them; if the ratio is much less than the improvement threshold, some excess i-f gain is present. Since the voltage at the limiter to produce limiting and the equivalent noise voltage at the input circuit may both be determined, the above considerations permit calculation of the necessary i-f gain between these two points.

The effective rms noise at the grid of the first tube is given by

$$E = \sqrt{4kTRB}, \tag{1}$$

where k = Boltzmann's constant = 1.380×10^{-23} joule per °K,
T = absolute temperature, °K,
R = effective resistance in grid circuit,
B = noise bandwidth.

If the input circuit is a transitionally coupled double-tuned circuit loaded on the primary only, then

$$R = 2R_1 \frac{C_1}{C_2}, \tag{2}$$

where R_1 = primary loading,
C_1 = primary capacitance,
C_2 = secondary capacitance.

The value of noise given by Eq. (1) is then multiplied by the square root of the noise figure of the i-f amplifier. This noise figure must be estimated in the early stages of design but can be measured[1] after the first model has been built.

17·3. Bandwidth Requirements.—The bandwidth requirements for sinusoidal frequency modulation are well known, and only a brief discussion of bandwidth requirements for a pulsed f-m signal will be given here. It has been shown by James[2] that for $\delta \ll \tau$, where δ is the pulse length and τ the period between pulses, the spectral distribution of a pulsed f-m signal as that shown in Fig. 17·2 can be approximated closely by considering the complete wave train consisting of three parts:

1. A continuous wave of frequency f_c.
2. A sequence of pulses of frequency f_c, duration δ, with intervals of $(\tau - \delta)$, having equal amplitude and opposite phase to the above wave.

[1] See Chap. 12.

[2] H. M. James, "Harmonic Analysis of Rectangular F-m Pulses," RL Group Report No. 43-5/25/45.

3. A sequence of pulses of frequency f_p, duration δ, with interval $(\tau - \delta)$, filling in the gaps in the wave train created by the interference of Items 1 and 2.

The spectrum will correspondingly consist of

1. A few strong lines of frequency f_c, carrying most of the energy.
2. A distribution of lines around f_c, with intensity varying like that for a pulse of duration δ, carrying a fraction δ/τ of the total energy.
3. A distribution of lines near f_p, with an intensity distribution like that for a pulse of duration δ, carrying a fraction δ/τ of the total energy.

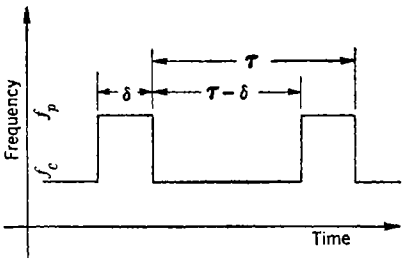


FIG. 17·2.—A pulsed f-m signal.

Thus the bandwidth must be equal to the frequency deviation plus that required to pass a pulse of duration δ.

17·4. Limiter Design.—The function of the limiter in an f-m receiver is to remove amplitude modulation from the carrier. The limiter can be either one or two stages, the two-stage, or cascade limiter, being preferred. The problem to be met in its design is (1) to have the output constant over a large range of input voltage, thus reducing amplitude modulation, and (2) to have the time of operation short enough to reduce impulse noise effectively.

Single-stage Limiters.—The usual design of an f-m limiter is based on *grid limiting* and takes the form shown in Fig. 17·3. If the applied voltage is large enough to draw grid current, the voltage developed across the resistor R will charge condenser C negatively. During the negative half of the cycle, the condenser charge leaks off through R. If $RC \gg 1/f_0$, the negative voltage thus developed across R at the grid of the limiter tube will be proportional to the applied input voltage. Under these conditions the effective range of voltage change in the grid circuit, responsible for plate current change in the plate circuit, will be that lying between the cutoff of the tube and the point at which the tube draws grid current.

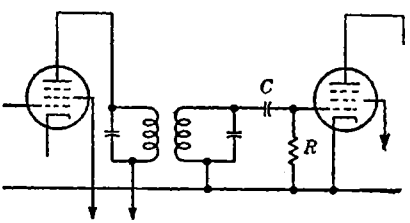


FIG. 17·3.—Usual design of an f-m limiter.

Thus the effective grid swing will be constant and independent of the magnitude of the applied voltage as long as the applied voltage is large enough to draw grid current.

By lowering the plate and screen voltage on the limiter stage, the range of the effective grid-voltage swing can be reduced. This not only reduces the voltage required for cutoff but can also cause plate current saturation, or *plate limiting*, if the plate-load resistor is large enough for its load line to fall well below the knee of the tube characteristics. In wide-band limiter design, the plate-load resistor cannot, in general, be made large enough to take advantage of plate limiting.

Even though the effective grid swing is constant, the output of the limiter is not necessarily constant, because with large input voltages enough bias can be developed to cause Class C operation of the limiter. As the angle of plate current flow is reduced with increase of input voltage, the harmonic content of the plate current increases and the fundamental component, which is the only one passed by the tuned discriminator, is therefore reduced. The limiter therefore has constant output over a certain range of input voltages but has reduced output at higher input voltage levels. The above discussion assumes that the effective grid swing is constant. Actually, this is not the case. The effective grid swing will be constant only if R is much greater than the grid-to-cathode resistance when the grid is conducting. Thus, by reducing the value of R, the grid swing will increase slightly with increased signal level, and a critical value of R can be found where this increased grid swing just balances the loss in output due to the smaller angle of plate current flow. This choice results in a limiter that can be made to have relatively constant output over a wide range of input levels. The limiter is still not completely satisfactory, however, because of impulse noise considerations.

Impulse Noise and Time Constants.—The value of C should be large enough that the time constant RC is much greater than $1/f_0$ (f_0 being the intermediate frequency) in order to assure reasonable efficiency in developing bias voltage but not so large that the circuit cannot recover quickly after a noise impulse. The shortest impulse or the fastest amplitude modulation that can be passed is determined by the bandwidth of the i-f amplifier. When a step function is passed through an amplifier, the rise time, from 10 to 90 per cent of its amplitude, is approximately $0.7/\mathcal{B}$ sec. The duration of the rise and fall of the shortest impulse that can be passed by the amplifier is, therefore, approximately $2/\mathcal{B}$ sec. If an attempt is made to have the time constant be, say, $\frac{1}{6}$ the duration of the shortest impulse, the two requirements stated above imply that

$$\frac{1}{3\mathcal{B}} > RC > \frac{1}{f_0}. \tag{3}$$

Equation (3) can be satisfied only if $3\mathcal{B} < f_0$; often this condition cannot be met in wide-band receivers where it is not uncommon that $3\mathcal{B} > f_0$. If the bandwidth is fixed by the required deviation or transient response, the only thing left to do is to increase the intermediate frequency.

Usually wide-band receivers are used in systems having a deviation ratio greater than unity and, therefore, have video circuits that are too narrow to pass the fastest impulse transmitted by the i-f amplifier. If the limiter time constant is chosen on the basis of video rather than i-f bandwidth, a large noise impulse passed by the i-f amplifier may overload the first video stages. Thus it is desirable to follow the above expression even in the design of the receivers for large deviation ratios.

Another complication in the design of grid limiters is the change of loading on the coupling circuit caused by grid current. A large time constant, obtained by increasing R, reduces the loading but slows down the action of the limiter. The effect of this additional loading can be reduced, however, by using a small L/C ratio in the tuning circuit, resulting in a decreased load resistance for a given bandwidth. This large C will also reduce the effects of changes in the input capacitance of the tube. The above difficulty is not so likely in broad-band receivers where Q of the circuit is less than 10.

Even though single-stage limiters have found considerable use, the compromise required among fast action, constant output, and loading and gain of the previous stage indicates the desirability of two-stage limiters or cascade limiters.

Cascade Limiters.—The cascade limiter consists of two single-stage limiters in series. The first stage is designed for most effective elimination of impulse noise with no regard for constant output, and the second stage is designed for constant output, leaving the burden of removing impulse noise to the first stage. The time constant of the second stage is not important and can be made large. This will result in most constant output, since the first stage does limit to a certain extent and there will be little danger of the output falling off at larger input levels.

It is necessary that the short time constant be in the first limiter, so that fast variations of the amplitude will be removed before being applied to the slow limiter. If this were not done, a noise impulse could develop enough bias to reduce the signal level below the point where the second limiter could follow.

THE RECEIVER AS BUILT

The activities of the Radiation Laboratory were terminated before a finished design for a receiver was available. Two models were, however, assembled from available parts in order to permit some tests of a complete system. It was expected that these would be mounted in a truck for use

in field-strength measurements. The power supply was therefore provided with electronic regulation which would probably be unnecessary in a fixed installation. A block diagram of the complete receiver is shown in Fig. 17·4, and a photograph of the receiver is shown in Fig. 17·5. The r-f assembly, including the magic-T mixer for the signal channel and the frequency-stabilizing circuit for the local oscillator, is mounted on the panel above the i-f amplifier. The LO frequency is controlled by the r-f discriminator described in Sec. 3·17.

The complete circuit diagram of the receiver is shown in Fig. 17·6. The magic-T mixer differs somewhat from those previously described in

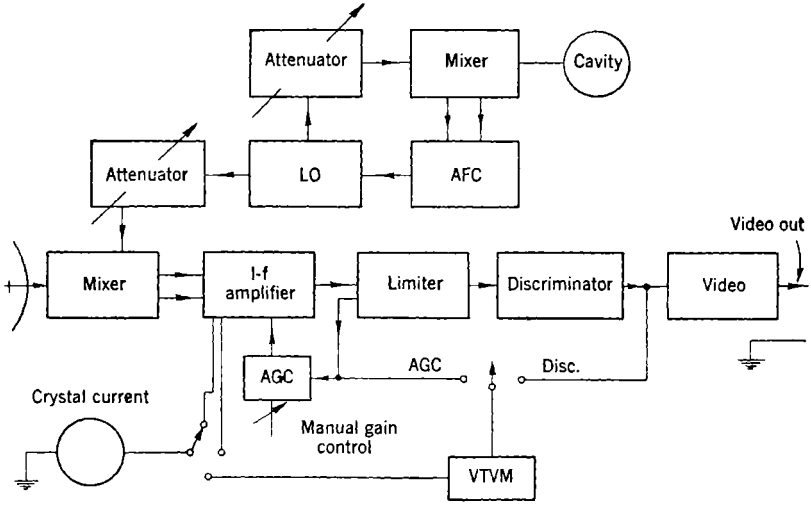


FIG. 17·4.—Block diagram of receiver.

that the two crystals are inserted in the waveguide in opposite orientations. The i-f outputs may therefore be added directly, thereby simplifying the design of the input circuit.

17·5. The Input Circuit.—The input circuit is a π equivalent of a double-tuned circuit and was used because of its ease in alignment and of obtaining separate crystal current outputs for each crystal. It is shown in Fig. 17·7.

The bandwidth of the input circuit was designed to be 20 Mc/sec when transitionally coupled, with R_1 equal to 200 ohms. The i-f impedance of crystals is usually considered to be about 300 ohms, which, with two crystals in parallel, would make R_1 equal to 150 ohms. However, if possible, it is considered good practice to design the input circuit to be transitionally coupled at a higher resistance. This is done to minimize

the change of bandwidth of the input circuit caused by the change of crystal impedance with crystal current. To facilitate adjusting the circuit, a "tuning ring" (shorted turn) has been used on L_a and L_b. In production these could be eliminated. A photograph of the input circuit and part of the i-f amplifier is shown in Fig. 17·8.

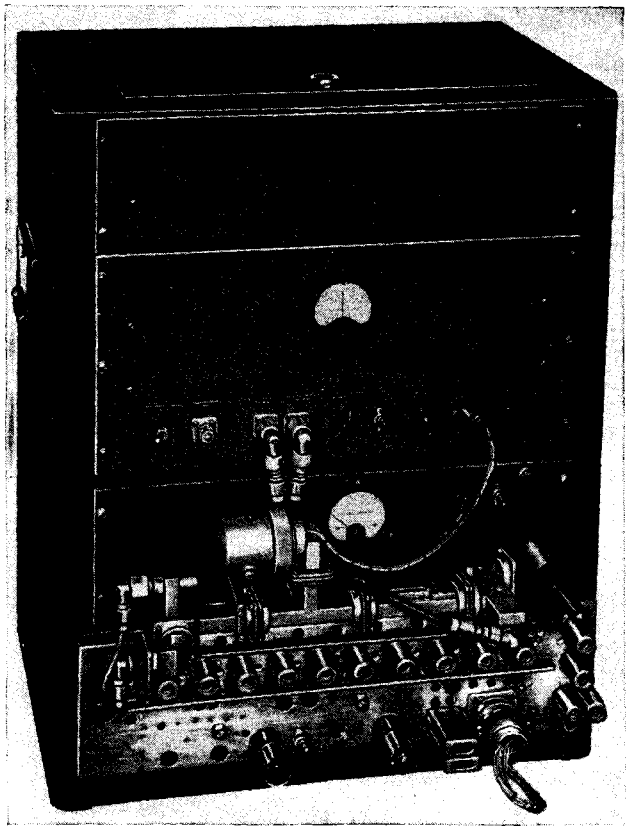


FIG. 17·5.—Photograph of receiver.

The impedance reflected from the primary to the secondary as given by Eq. (2) is about 1000 ohms. This results in about 13 μv Johnson noise and, assuming an optimistic noise figure of 5 db (the noise figures of these receivers have never been measured), will give about 24 μv noise at the grid of the first tube.

17·6. The I-f Amplifier.—The i-f amplifier consists of eight double-tuned stages with an over-all bandwidth of about 11 Mc/sec centered

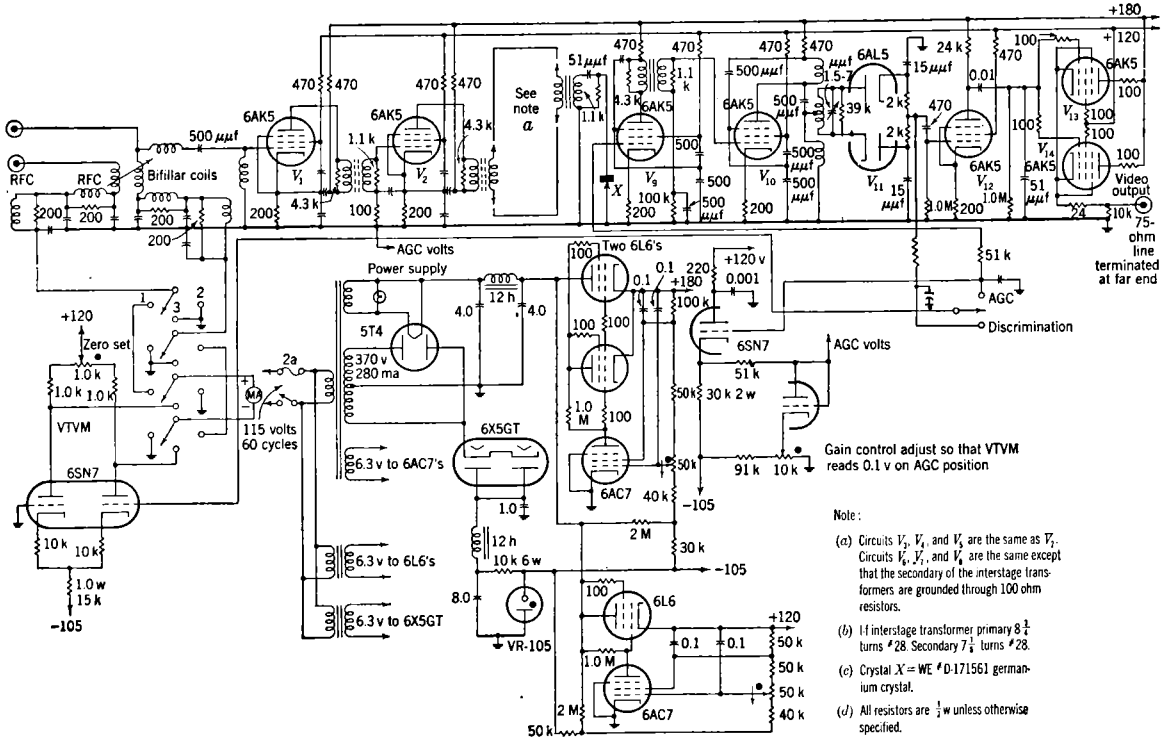


FIG. 17·6.—Circuit diagram of receiver.

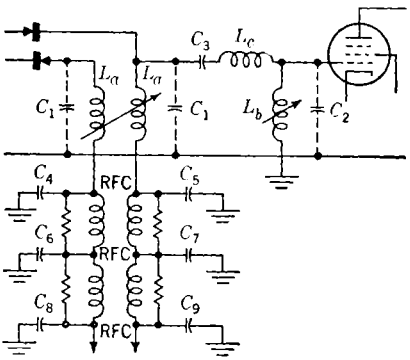


FIG. 17·7.—Input circuit.

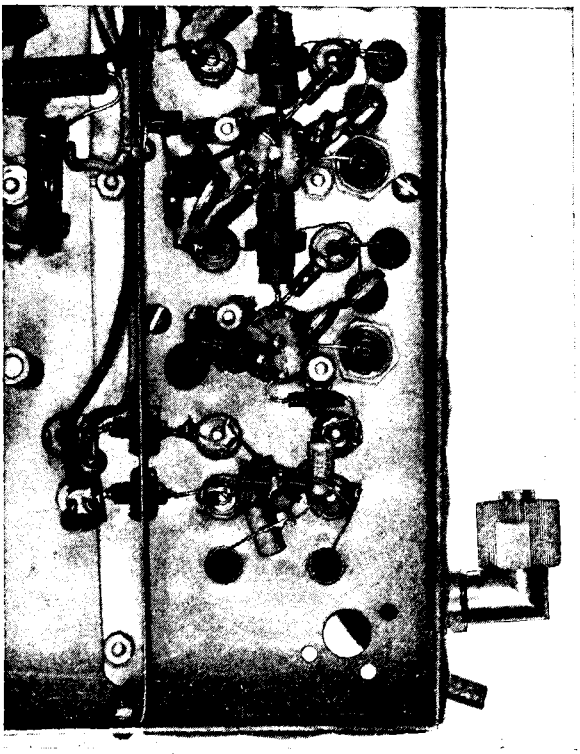


FIG. 17·8.—Input circuit and section of i-f amplifier.

at 60 Mc/sec. The stages are identical, with the exception that the grid-bias gain control is applied only to the second, third, fourth, and fifth stages. Each stage has an average gain of about 14 db and a bandwidth of about 21 Mc/sec. The effective loading on the secondary consists of the input impedance of the 6AK5 tube which is about 7000 ohms at 60 Mc/sec, in parallel with the resistor of 1100 ohms. This results in a secondary loading R_2 of 950 ohms. With a 6AK5 tube having a transconductance of 3600 μmhos, the gain per stage is 3.53 times, or 11 db. Thus the minimum over-all gain that is ever likely to be observed will be approximately 88 db.

By carrying out in detail the procedure suggested in Sec. 17·2 it was found that the rms noise at the grid of the limiter should be at least 0.12 volt. Thus the over-all i-f amplifier gain must be

$$\mathcal{G} = 20 \log \frac{0.12}{24 \times 10^{-6}} = 74 \qquad \text{db.}$$

The gain of 88 db actually provided is thus seen to be somewhat more than is absolutely required. Another design might eliminate one i-f stage.

17·7. The Limiter.—The most successful limiter used in this receiver is shown in Fig. 17·9. Because the design of the interstage transformer

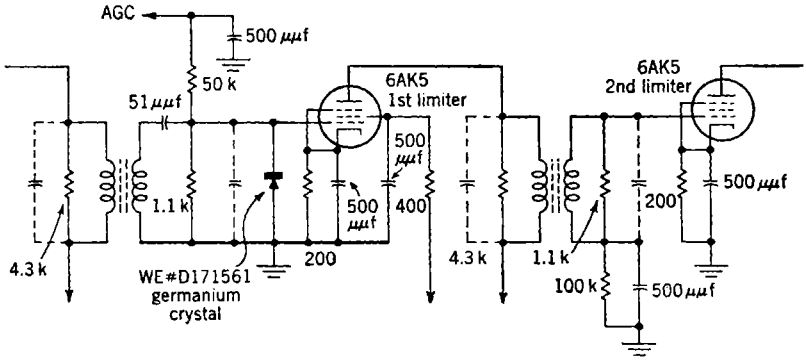


FIG. 17·9.—Circuit diagram of limiter.

in the i-f amplifier called for a secondary loading of 1100 ohms, this resistor was used in the grid circuit of the limiter. The 51-$\mu\mu$f coupling condenser makes the grid circuit time constant 0.056 μsec, or about one-fourth the duration of the shortest impulse passed by the i-f amplifier. Since the 1100-ohm load resistor is the same order of magnitude as the grid-to-cathode resistance when the grid is conducting, very little bias will be developed. The use of a germanium crystal as a diode overcomes this difficulty, as the resistance of the crystal when conducting is appreciably

less than that of the load resistor. The same interstage transformer was used in the plate circuit as in the other i-f stages. Inasmuch as the load is too low for plate limiting, the plate and screen voltages were made the same as the other stages. The gain of each limiter stage is about 3, so that even with full plate and screen voltage, the grid of the last limiter stage is never driven to the point where the output of the second limiter decreases. Placement of the time constant of the second limiter in the ground side of the secondary coil was merely for convenience; its value of 50 μsec was chosen arbitrarily. When the gain of the i-f amplifier is so adjusted that noise develops 0.1 volt bias at the grid of the first limiter, the ratio of peak limit level to noise is about 6 or 7 db.

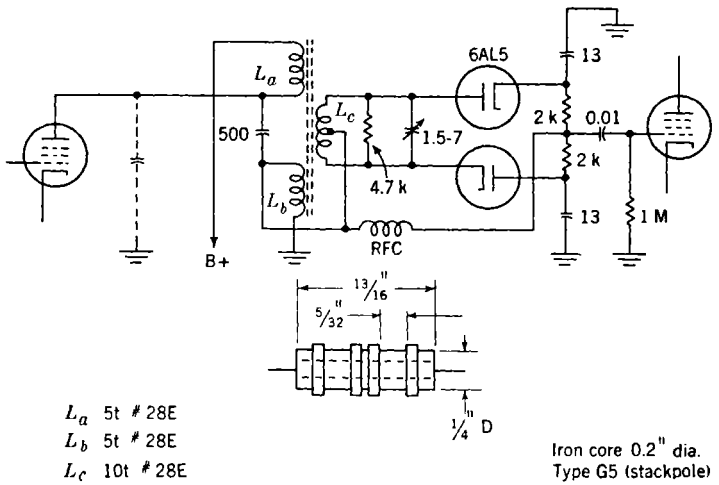


FIG. 17·10.—Discriminator circuit and transformer assembly.

17·8. The Discriminator.—Two types of discriminators have been used successfully in these receivers. The first is the conventional Foster-Seeley discriminator with the primary split so as to achieve coupling on both sides of the balanced secondary. The coils are wound on a type G5 (Stackpole) iron core. Details of the coil form and assembly are given in Fig. 17·10. Discriminators of this type, centered at 60 Mc/sec, have been made fairly linear over a range of about 15 Mc/sec, the bandwidth between peaks being about 20 Mc/sec.

The second type of discriminator used is shown in Fig. 17·11. This discriminator is much easier to adjust in that there is no mutual inductance and all three coils can be adjusted independently by the use of "tuning rings." Although this particular discriminator is not quite so wide as the discriminator of the Foster-Seeley type mentioned above,

there is no reason to believe that this represents an upper limit to its bandwidth. It is about 17 Mc/sec wide between peaks. Because of the relative ease in its adjustment, the second discriminator is probably to be preferred. However, it should be possible to make either discriminator fixed-tuned in production.

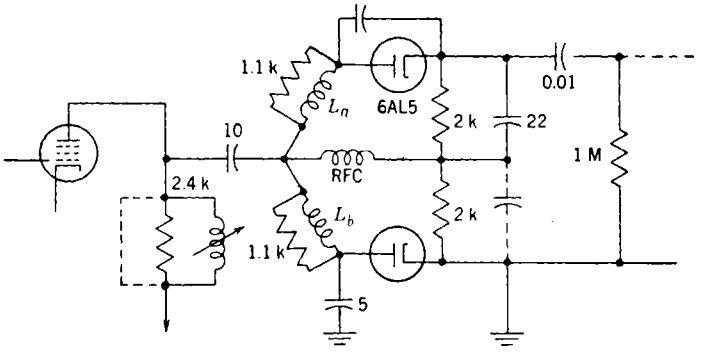


FIG. 17·11.—Alternative discriminator circuit.

In Fig. 17·12 the over-all receiver pass band is shown at two different signal levels. With a signal that just produces limiting, the discriminator is seen to be linear to ±2 per cent over a range of ±6 Mc/sec. Note also that as the signal level is increased, the bandwidth between the peaks of the discriminator increases. This behavior is a consequence of the presence of the limiter and the fact that the bandwidth of the discriminator is greater than the bandwidth of the i-f amplifier. The useful discriminator bandwidth is then determined not by the actual discriminator bandwidth but by the bandwidth of the i-f amplifier. The action of the limiter, however, must also be considered. For a constant input voltage of variable frequency, the magnitude of the output voltage of the i-f amplifier will be determined by the i-f amplifier pass band. However, the input to the discriminator will be constant and independent of the i-f amplifier pass band as long as the signal is large enough to limit. For a given input voltage, then, the input voltage to the discriminator will fall as soon as the signal deviates far enough from resonance to be reduced by the pass band of the i-f amplifier to the point where it no longer limits.

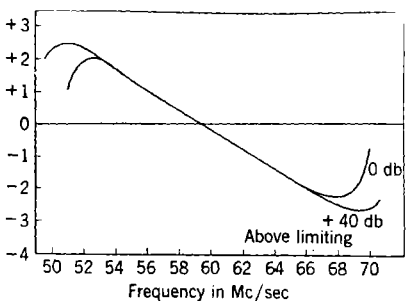


FIG. 17·12.—Receiver output characteristic at two signal levels.

Thus the over-all receiver pass band is seen to be a function of the input signal level, although the slope of the discriminator characteristic will remain constant.

17·9. The Video Amplifier.—The video amplifier is of conventional design. Two 6AK5's in parallel are necessary to obtain the required 2 volts of video output across the 75-ohm line. So little gain is required of the video amplifier that it is necessary to add shunt capacitance to the coupling circuit to make the bandwidth sufficiently narrow.

17·10. Automatic Gain Control.—The negative voltage developed at the grid of the first limiter can be used to advantage for AGC purposes. Even though the output of the limiter will be constant over a large range of input voltages, it may be advisable to use AGC to protect the later stages of the i-f amplifier from overloading. The AGC voltage is taken directly from the grid of the limiter through a high resistor, as shown in Fig. 17·9. Since it is desirable to limit on low signal levels, it is necessary to incorporate an amplitude delay of about 20 or 30 db. This delay was obtained by the use of a 6SN7, half of which was used in a diode clamp

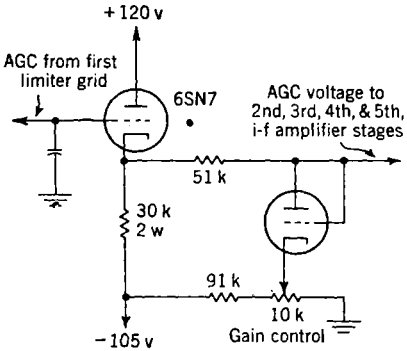


FIG. 17·13.—Schematic diagram of circuit for applying AGC voltage.

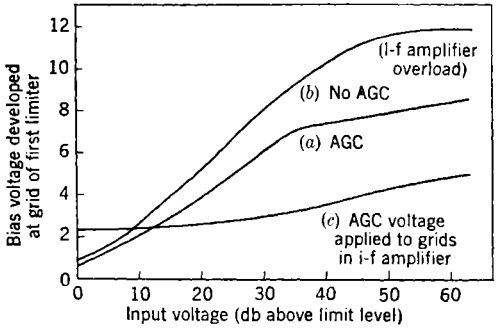


FIG. 17·14.—Performance of AGC.

circuit and the other half as a cathode follower to bring the AGC voltage to the grids of the four controlled stages. The circuit is shown schematically in Fig. 17·13. In Fig. 17·14 the performance of the AGC is shown. The voltage developed at the grid of the limiter is used as the ordinate

rather than the actual receiver output because the output of the limiter would be constant in any event. From Fig. 17·14 it may be seen that the delay is about 30 db and that enough AGC voltage is available to prevent i-f overload.

17·11. Vacuum-tube Voltmeter.—A conventional zero-centered VTVM is built into the receiver to read either the voltage developed at the grid of the first limiter or the output of the discriminator. The manual gain control is set so that thermal noise develops about 0.1 volt at the limiter grid. At this value the limiter is known to operate at the improvement threshold. The voltmeter is used in the discriminator position to facilitate tuning the receiver. The 0 to 1-ma meter is also used to measure crystal current.

CHAPTER 18

BEACON SUPERHETERODYNE RECEIVERS

By J. H. Tinlot

INTRODUCTION

The problems arising in the design of superheterodyne receivers for covering a very wide band of frequencies (so-called "wide-coverage" receivers) with relatively high sensitivity will be treated in this chapter. The term "wide coverage" effectively describes the function of such a receiver because it is intended to receive simultaneously signals having carriers located anywhere in a given band of frequencies, whereas the customary broad-band receiver is made to reproduce with good fidelity a signal of constant carrier frequency.

There are cases where a superheterodyne receiver of conventional design can be made broad enough to cover the required frequency range. Where this range is approximately 50 Mc/sec or greater, however, new methods of attack are needed.

The general requirements of this type of superheterodyne introduce special problems, one of which is the analysis of the theoretical maximum sensitivity. Although the i-f bandwidth of such a receiver must be very large, the video bandwidth after the second detector need be only large enough to give the fidelity required by the particular application. Since the i-f bandwidth may be, in some cases, forty or fifty times the video bandwidth, the calculation of the noise figure of the whole receiver becomes complex.

In a conventional receiver the midfrequency for the i-f amplifier may be chosen more or less arbitrarily and in particular may be made large enough to minimize noise originating in the local oscillator. Here, however, it is intimately concerned with the frequency coverage desired. Similarly, the input circuit to the first i-f stage is determined by bandwidth considerations instead of optimum noise figure.

The receivers that have been built have been intended to receive a signal in the form of a pulse 2 μsec long. Since the i-f bandwidth is much larger than the video bandwidth, modification of a receiver for a different type of signal is simply a matter of adjusting the characteristics of the circuits following the second detector. Therefore, the treatment can be made general. Because of the specific applications of the examples treated, the measure of sensitivity is in all cases taken to be the power of

a 2-μsec pulse, which produces at the output of the receiver what is called a "tangential signal." Such a signal is illustrated in Fig. 18·1. Although this definition is easy to interpret visually from an oscilloscope presentation, a tangential signal may be defined more rigorously as a signal 8 db above the rms noise power; from this standard any other definition of sensitivity can be translated without much trouble.

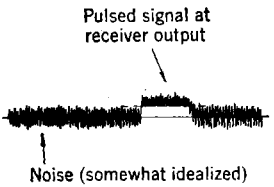


FIG. 18·1.—Tangential signal.

18·1. Methods of Approach.—Three methods of approach that make possible a coverage several times greater than is possible by conventional means have been developed. These have arbitrarily been named (1) the single-sideband method, (2) the switched-LO method, and (3) the frequency-modulated-LO method.

Examples of each of these are shown in Figs. 18·2 to 18·4, in which typical figures have been used for ease of explanation.

The single-sideband method would at first seem to be merely an extension of a conventional superheterodyne for larger bandwidths. A receiver is built using an i-f amplifier whose bandwidth is equal to the required r-f coverage. Although heterodyning produces r-f acceptance bands on both sides of the local oscillator, only one of these is used, the other merely adding a problem in reducing erroneous response and interference. The difference from the conventional superheterodyne arises from the fact, previously mentioned, that although the i-f bandwidth must be made large for wide coverage, the video bandwidth need be only large enough for correct fidelity. In this case, the video bandwidth was made about 1.5 Mc/sec, whereas the i-f bandwidth was made 70 Mc/sec. Furthermore, the shape of the i-f pass band is unimportant as long as it results in fairly uniform sensitivity over the band to be covered (70 Mc/sec in this case), and the location of the pass band in the frequency spectrum is completely arbitrary as long as it does not overlap into the video-amplifier pass band. In the example shown in Fig. 18·2, the i-f amplifier has a pass band extending from 6 to 76 Mc/sec. It therefore allows the use of an i-f amplifier that makes possible much larger bandwidths, with large gain, than may be achieved by any conventional means (Sec. 18·2)

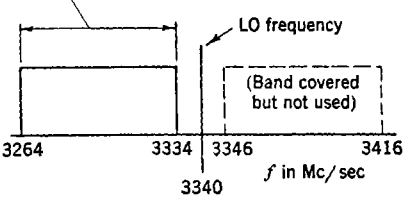


FIG. 18·2.—Single-sideband receiver. I-f band limits are 6 and 76 Mc/sec.

The switched-LO method (a typical example of which is illustrated

in Fig. 18·3) covers an r-f band much wider than the i-f bandwidth at the expense of some sacrifice in performance. For instance, Fig. 18·3 shows a scheme whereby a bandwidth of about 126 Mc/sec is covered, which is much greater than can be covered by the first method. Using the numbers of Fig. 18·3, consider the operation of the receiver with the local oscillator tuned to frequency A (9390 Mc/sec). Since the i-f amplifier passes frequencies from 8 to 48 Mc/sec, r-f signals of frequencies 9398 to 9438 Mc/sec and 9342 to 9382 Mc/sec will be amplified. If, now, the local oscillator is tuned to frequency B (9360 Mc/sec), the frequencies in the ranges 9312 to 9352 Mc/sec and 9368 to 9408 Mc/sec will be covered. If the local oscillator is switched rapidly between frequencies A and B, a band between 9312 and 9438 Mc/sec will be covered although considerable regions will be covered for only half the time. This half-

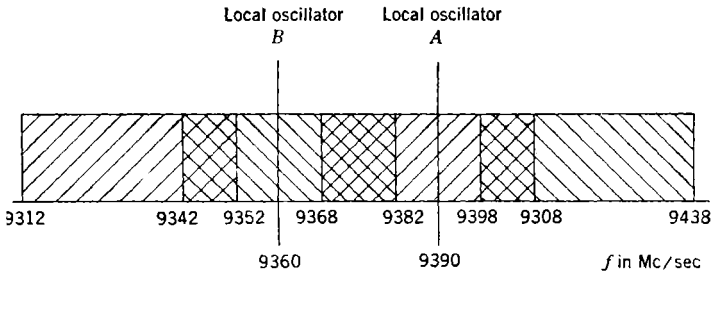


FIG. 18·3.—Switched-LO receiver. I-f band limits are 8 and 48 Mc/sec.

time response is the sacrifice in performance mentioned earlier. Often, by adjusting the LO switching rate to a particular value, the loss of response because of the time-sharing property of the receiver can be made unimportant, although it is always a fundamental limitation.

In a sense, the most versatile broad-coverage scheme is the f-m LO method. In the single-sideband method, the total r-f coverage equals the i-f amplifier bandwidth, which is sharply limited by existing components and techniques. In the switched-LO receiver, the i-f bandwidth is about one-third of the resulting coverage; therefore the practical maximum coverage is essentially multiplied by three. In the f-m LO method, the i-f bandwidth need be only one-sixth, one-tenth, one-fourteenth, etc., the r-f coverage, depending upon how far the scheme can be carried. If more than one LO frequency is simultaneously injected into the mixer, the r-f coverage will be multiplied by the number of local oscillators if conditions are right. A possible way of doing this is illustrated in the

example of Fig. 18·4. Here a receiver is used having an i-f amplifier with band limits of 1.0 and 21.0 Mc/sec. If a local oscillator A is tuned to 9375 Mc/sec, then the r-f coverage is obviously 9376 to 9396 Mc/sec and 9354 to 9374 Mc/sec. If, in addition, local oscillators B and C are also present and their frequencies also injected into the mixer, then bands of 9314 to 9334, 9336 to 9356, 9394 to 9414, and 9416 to 9436 Mc/sec will be covered simultaneously. Thus, except for three 2.0-Mc/sec gaps, a total band of 122 Mc/sec (9314 to 9436 Mc/sec) will be covered by an i-f amplifier having only a 20-Mc/sec bandwidth.

Although it is possible to use multiple local oscillators, a much simpler means of accomplishing the same result is to apply frequency modulation to a single oscillator. In the velocity-modulated oscillators used in

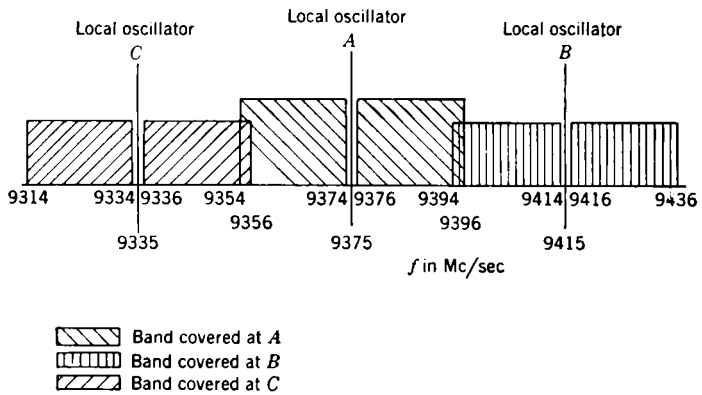


FIG. 18·4.—Frequency-modulated-LO receiver. I-f band limits are 1 and 21 Mc/sec.

microwaves this is fairly easy. It will be remembered that an oscillator of frequency f_o, frequency modulated at a rate f_m, will produce discrete side frequencies at f_o, $(f_o \pm f_m)$, $(f_o \pm 2f_m)$, $(f_o \pm 3f_m)$, etc. Thus, by choosing f_m as the required separation of the local oscillators, the end effect is the same as if several oscillators were simultaneously injected into the mixer.

It is desirable that the multiple local oscillators used in the manner explained above have approximately the same amplitude. It is a relatively simple matter to adjust the amplitude of frequency modulation (that is, the deviation) to result in having f_o, $f_o + f_m$ and $f_o - f_m$ about equal; the difficulty of such an adjustment increases as the required number of side frequencies increases.

In the example of Fig. 18·4, f_m is chosen as 40 Mc/sec. From the mathematics of f-m theory, it is found that a deviation of ± 57 Mc/sec will produce about equal fundamental (f_o) and first side frequencies

($f_o \pm f_m$). The problem then reduces to tuning the local oscillator, without modulation, to frequency f_o, then applying the required amount of 40-Mc/sec frequency modulation.

Although it is more difficult, it is possible to make use of further removed side frequencies and thereby increase the receiver coverage (with the same i-f amplifier design) almost at will. The main obstacle to great extension of this procedure is that to obtain fair conversion efficiency in the bands due to side-frequency local oscillators, the LO power must be above an amount determined by the type of mixer. Furthermore, since each frequency acts like a separate oscillator, each will cause the crystal to put out excess noise due to it and will contribute LO noise, just as if the frequency were due to an entirely separate oscillator. Thus a severe limitation of sensitivity is encountered.

The problem of the gaps that appear in Fig. 18·4 can be solved in different ways, depending upon the type of signal to be received. The

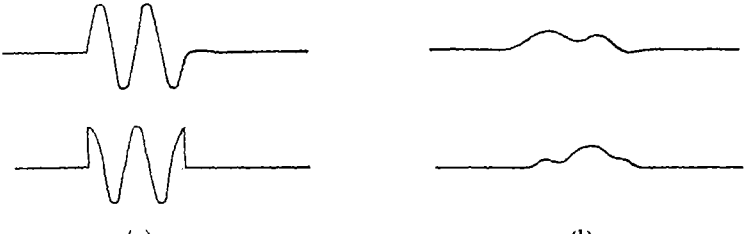


FIG. 18·5.—A 2-μsec pulse width 1-Mc/sec i-f carrier. (*a*) I-f pulse idealized; (*b*) possible detected pulse.

reasoning so far has inferred the reception of c-w signals, whereas it has been stated that 2-μsec pulsed signals were considered in all of these examples. If a 2-μsec pulse has its carrier at, for example, exactly 9375 Mc/sec, a small amount of energy will be received, since the frequency spectrum of an ideal pulse extends to either side of the carrier with decreasing amplitude. The loss (compared with the sensitivity in the middle of the bands) in the case of such a 2-Mc/sec gap will be considerable; therefore some time-sharing scheme may have to be adopted. One possibility is a low-frequency wobbling or switching of all the local oscillators so that frequencies in the gaps are covered a substantial fraction of the time.

The consideration of what happens to 2-μsec pulses in the gaps and near the gaps brings up one of the great limitations of this type of wide-coverage receiver—that of limited fidelity. A 2-μsec pulse whose carrier has a frequency of 9376 Mc/sec, when heterodyned with the 9375-Mc/sec local oscillator, will produce an i-f pulse with i-f carrier of 1.0 Mc/sec. This pulse will, therefore, contain no more than two cycles, such as the

waves shown in Fig. 18·5*a*. Since no possible means of detection can give a good 2-μsec pulse from such an i-f pulse, the receiver can at best give waveforms at the second detector such as those shown in Fig. 18·5*b*. Therefore, most conceptions of good fidelity must be abandoned. For instance, if the signal at the receiver output is to be used merely for triggering succeeding circuits with fairly lenient requirements on delay, then this arrangement can be made practical. One of the first two schemes outlined previously must be considered for any application requiring good reproduction of a pulse or other video signal.

In the third example, the f-m LO type of receiver, the i-f amplifier needed must be conceived on entirely different principles from those of conventional amplifiers. Although a 20-Mc/sec-wide amplifier can be built by the usual means, the location of this band (1.0 to 21.0 Mc/sec) is so unusual as to present a considerable problem. On the other hand, this requirement lends itself very well to the scheme explained in detail in Sec. 18·2.

18·2. I-f Amplifier Requirements.—The i-f amplifier for wide-coverage applications must have

1. Sufficient gain. This, in general, means that the amplifier must have gain such that the noise (and consequently the smallest signal) drives the second detector into the linear region. For a diode second detector, this signal must be approximately 0.5 to 1.0 volt.
2. Very large bandwidth. The examples enumerated in Sec. 18·1 are typical of conditions that may be met; the i-f bandwidths stipulated, even in the f-m LO receiver, are very great compared with those achieved with ease in customary radar i-f amplifiers.
3. Specified minimum variation in gain over the acceptance band. This springs directly from the fact that the receiver, for optimum performance, should have as nearly constant sensitivity as possible over the band to be covered. Very great variations in gain will result in reduced sensitivity at the points of weakest gain. This effect cannot be compensated for by increasing the total amplifier gain, because the signal-to-noise ratio at the weakest gain points will be much worse than in the case of constant gain.

Unlike the radar-receiver case, no restrictions whatever are made on the shape of the i-f amplifier pass band for reasons of fidelity. It is well known that in the case of a narrow i-f amplifier (that is, one in which the bandwidth is comparable to the width of the signal-frequency spectrum), a pulsed signal will have overshoots or "ringing" unless the shape of the pass band is carefully designed. In the wide-coverage case, however, the i-f pass band is much wider than the pulse-frequency spectrum. The

ringing introduced is therefore of extremely high frequency and is completely eliminated in the video amplifier, which is of normal width for usable reproduction of the pulse. Therefore, the necessary bandwidth can be achieved by the use of circuits that, if used in a conventional radar i-f amplifier, would cause impossible distortion. Thus, the construction of an i-f amplifier using a coupling circuit embodying shunt-series peaking or compensation is absolutely practical, although it is automatically ruled out in the radar case for the reasons given above. The remainder of this section will give briefly the theory of i-f amplifier design using shunt-series peaking and necessary design data for typical problems.

The general procedure is to "synthesize" a bandpass amplifier by making use of a low-pass filter as a coupling circuit and rejecting a band of frequencies from zero frequency up to the frequency chosen as the desired low-frequency cutoff point of the amplifier. That this scheme possesses considerable advantage may be shown qualitatively, as follows:

Any attempt to improve the gain-bandwidth product of an amplifier stage almost invariably involves increasing the complexity of the circuit, that is, increasing the number of components necessary to construct the desired filter. Thus, although the gain-bandwidth product may be increased, there is greater difficulty of adjustment. Furthermore, added stray capacitance and inductance will at some point largely nullify the resulting gain in efficiency of the filter. A practical limit in gain-bandwidth product, which will in general be relatively far from the theoretical upper limit, will therefore be reached. For a four-terminal network, this limit is equal to $\pi^2/2$ in units of $g_m/2\pi C$. However, a low-pass filter having the same gain-bandwidth product as the corresponding bandpass filter will be considerably simpler, particularly in the number of components. Thus, it may be possible to go much farther toward the theoretical limit before the practical limit in complexity is reached.

DESIGN CONSIDERATIONS

The gain per stage of an i-f amplifier that is sufficiently wide for the applications contemplated here will necessarily be small; consequently a large number of stages will be needed. The choice of a coupling circuit must be made with this in mind. For example, suppose that two coupling circuits were available having response curves such as those shown in Fig. 18·6. If only one stage were to be used or if transient response were an important factor, the circuit giving curve (*b*) would be far superior. However, if 10 or so such circuits were put in cascade, the over-all bandwidth would be less than would be obtained with circuits giving curve (*a*). It is easy to compute the rate at which the over-all bandwidth of cascaded uncompensated *RC*-coupled stages falls off. If f_c is the 3-db cutoff fre-

quency for a single stage, n stages in cascade will have a cutoff frequency f_n given by

$$f_n = \sqrt{2^{1/n} - 1}\, f_c.$$

For n greater than 5 this may be written approximately

$$f_n = 0.834 \frac{f_c}{\sqrt{n}}.$$

Thus for $n = 10$, $f_n = 0.27 f_c$. On the other hand the series-shunt-peaked coupling circuit gives a response curve much more like curve (*a*) in Fig. 18·6, and the bandwidth of cascaded stages is only a few per cent less than that of a single stage. Although the loss of bandwidth with cascaded stages can be reduced by such schemes as using staggered tuning or feedback chains, not only are these solutions inferior to shunt-series peaking in over-all gain-bandwidth product, but the greater difficulty of adjustment and construction makes such schemes of questionable practical value.

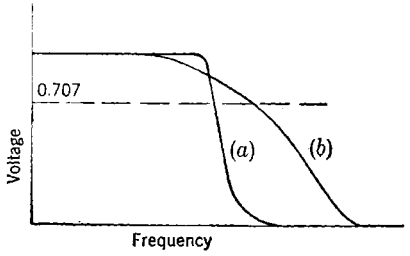


FIG. 18·6.—Frequency response of hypothetical coupling circuits.

Computations concerning the general case of low-pass filters, including the case of the shunt-series-peaking filter, were made by O. H. Schade.[1] The usable system for designing such filters, as outlined by Schade, was actually a combination of conventional filter theory and the compilation of experimental data for representative practical cases. In brief, Schade's method is as follows. He first pointed out the advantage of a four-terminal network over a two-terminal network, considering only the bandwidth and neglecting the flatness of response (that is, the necessity for constant transfer impedance). Since for use in vacuum-tube coupling circuits the shunt arms of the filters must include the tube and circuit capacitances, Schade reduced the number of types of filters to be considered to a workable number and then treated the case of the circuit given in Fig. 18·7. This is the circuit for the shunt-series-peaking filter, but without any damping of the series arm. Experimental deter-

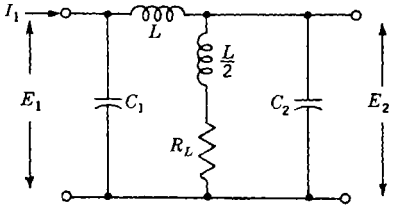
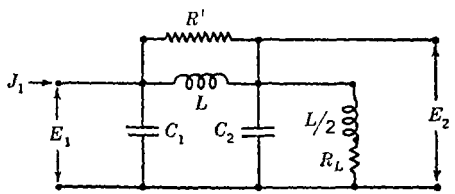


FIG. 18·7.—Schade's original four-terminal network.

[1] RCA License Laboratory Report, 1938.

mination of the resulting transfer-impedance characteristics (supplemented by calculation of a few points as a theoretical check) revealed that the filter was remarkably insensitive to changes in the ratio of capacitances in the two shunt arms although all the response curves showed a sizable peak in transfer impedance near the cutoff frequency. By further experimental manipulation, Schade established the amount of damping of the series arm required to form a very nearly flat transfer-impedance characteristic (Fig. 18·8) for several ratios of input to output capacitances.



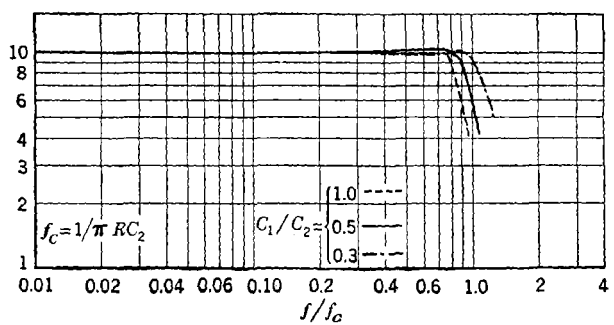


FIG. 18·8.—Frequency-response curves of shunt-series-peaking filters.

The insensitiveness of the filter to these ratios is evident from the curves of Fig. 18·8. This does not, however, mean that for given values of resistance and inductance the capacitance ratio can be varied without altering the characteristic but that a filter can be designed for a widely varying ratio of capacitances with very nearly the same advantage in gain-bandwidth product.

18·3. High-frequency Cutoff.—Table 18·1, which is compiled from Schade's report, gives the information necessary to design a shunt-series-peaking filter for ratios of output to input capacitance (C_1/C_2) of 0.3, 0.5, and 1.0. These ratios cover the range ordinarily to be expected in vacuum-tube applications. It is, however, possible to interpolate for ratios other than those given above with reasonable assurance of success because of the insensitiveness of the filter to these ratios. Of the

TABLE 18·1.—DESIGN DATA FOR SHUNT-SERIES-PEAKING FILTER

A $\frac{C_1}{C_2}$	B $\frac{R'}{R}$	C $\frac{R_L}{R}$	D $\frac{\Delta f}{f_c}$	E GB factor $\left(\text{units of } \frac{g_m}{2\pi C}\right)$
0.3	5.66	1.088	0.96	2.0
0.5	5.66	1.075	0.85	2.5
1.0	18.90	1.00	0.76	3.0

six components involved in the filter, only one can be independently determined in a coupling-circuit application. This one variable is determined either by the bandwidth or by the gain (whichever is to be the limiting factor). All the other parameters either are fixed or are functions of the chosen parameter. This fact follows because C_1 and C_2, the output and input capacitances, are ordinarily invariable because they have presumably been reduced to a practical minimum.

As an example, for a wide-coverage-receiver application, it is generally desirable to design the filter for a stipulated bandwidth and to cascade the necessary number of stages to obtain the required gain. Thus, if Δf = desired bandwidth, f_c (the cutoff frequency of the filter) can be found immediately from Column D, Table 18·1 for the correct ratio C_1/C_2. Then the quantity R can be calculated immediately from the equation

$$R = \frac{1}{\pi f_c C_2}.$$

Knowing R, the actual load resistor R_L is obtained from Column C, Table 18·1, and similarly the series-damping resistor R' is determined by use of Column B. The series-inductance arm L is then given by

$$L = R^2 C_2.$$

The filter is then completely designed.[1]

For the common high-frequency pentodes (6AC7,6AK5, 6AG5) the ratio C_1/C_2, computed for the capacities within the tube itself without any wiring, is about 0.5. The gain-bandwidth factor is then approximately 2.5 (Column E, Table 18·1). It is seen from Column E that for $C_1/C_2 = 1.0$, the greatest gain-bandwidth factor (3.0) is obtained. If it is possible, therefore, the circuit should be so laid out as to make the total capacities approach the ratio 1.0, by adjusting the placement of

[1] To design an amplifier using a large number of cascaded stages, the required bandwidth of one stage can be obtained by assuming a 1 per cent reduction in bandwidth for each stage.

parts to provide the minimum stray capacitance on the input side. Needless to say, no extra capacity should be added in the process.

The incorporation of the shunt-series-peaking filter in a high-gain amplifier is most simply accomplished as shown in Fig. 18·9. Since the load resistor R_L will ordinarily be small for the bandwidths contemplated, the plate and screen voltage can come from a common supply. The drop in voltage across R_L due to plate-current flow will usually be small and will not impair the pentode characteristics. A single decoupling network (R_d and C_d) will therefore suffice. The choice of a grid-return circuit will usually be influenced by several considerations, some of which are treated in detail in the succeeding sections.

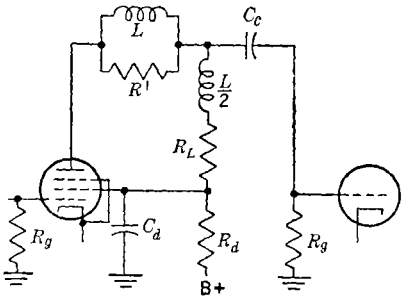


FIG. 18·9.—Practical pentode coupling circuit using shunt-series peaking.

18·4. Low-frequency Cutoff.—The use of the shunt-series-peaking type of coupling filter for a bandpass application requires additional circuitry to provide the necessary rejection of a specified range of the low frequencies.

Although numerous methods of accomplishing this rejection are available, the choice of a particular scheme will be dictated by two considerations. The first is simplicity, and the second is the amount and the sharpness of attenuation of the low frequencies required by the particular application contemplated. In the interest of simplicity, a means of low-frequency rejection involving circuit elements independent of the coupling-circuit filter is greatly desirable. In cases where this is not suitable, additional coupling filters providing selective attenuation are indicated.

A simple method of providing limited attenuation over a range of frequencies is the introduction of an appropriate network in the cathode circuit of the amplifier tube. If Z_k is the magnitude of the impedance in the cathode circuit of the (pentode) amplifier stage, then the gain is reduced in the ratio

$$\frac{\mathcal{G}'}{\mathcal{G}} = \frac{1}{1 + g_m Z_k},$$

where $\mathcal{G}$ = gain without network,
$\mathcal{G}'$ = gain with network,
and g_m = transconductance of tube.
Thus, if $g_m Z_k \gg 1$, $(\mathcal{G}'/\mathcal{G}) \propto (1/Z_k)$.

One suitable network is shown in Fig. 18·10*a*. The resistance R_k may be determined by the desired d-c operating conditions for the tube. The inductance and capacitance may be determined from the following considerations. Denote by f_0 the frequency at which the reactance of the condenser C_k is equal to the resistance R_k; that is,

$$f_0 = \frac{1}{2\pi R_k C_k}.$$

Furthermore, suppose that $2\pi f_0 L = pR_k$, $L = pR_k^2 C_k$.

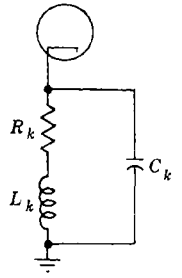


FIG. 18·10*a*

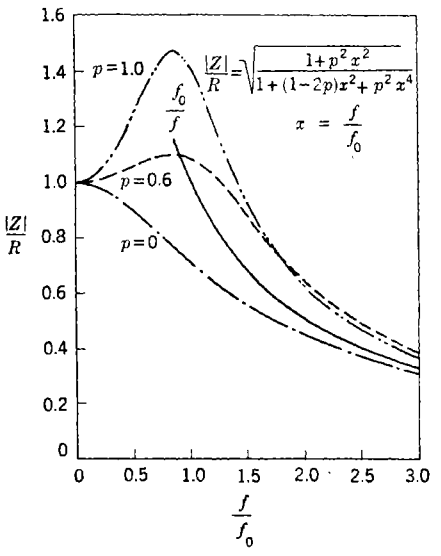


FIG. 18·10*b*

FIG. 18·10*a*.—Cathode degenerative filter for low-frequency rejection.
FIG. 18·10*b*.—Impedance curves for cathode filter.

Figure 18·10*b* shows curves of the impedance of the network as a function of frequency for different values of p. For all these curves, $Z_k/R_k \to 1$ at $f = 0$, and $Z_k/R_k \to f_0/f$ for large values of frequency.

Suppose now that f_c' is to be the low-frequency cutoff for the whole amplifier, of which n stages are to have cathode degeneration. Then $(\mathcal{G}'/\mathcal{G})^n = 1/\sqrt{2}$.

Now if f_c' is larger than about $3f_0$, the asymptotic expression

$$Z_k = R_k \frac{f_0}{f}$$

may be used.

From these expressions one obtains

$$\frac{\mathcal{G}'}{\mathcal{G}} = \frac{1}{1 + g_m R_k \dfrac{f_0}{f_c'}} = \frac{1}{2^{1/2n}}.$$

For $n \geqq 4$, this may be written with sufficient accuracy as

$$f_0 = \left(\frac{0.35}{n}\right)\left(\frac{1}{g_m R_k}\right) f_c'.$$

Thus, if $ng_mR_k > 1$, the expression used above for Z_k is valid. Therefore

$$C_k = \frac{g_m n}{2.2 f_c'},$$

and

$$L_k = \frac{p R_k^2 g_m n}{2.2 f_c'}.$$

The value of p may be chosen more or less arbitrarily, approximately 0.6 being most common. An inspection of the curves for Z_k/R_k shows that for this value of p the degeneration at f_0 will be nearly the same as at zero frequency where it is

$$\frac{\mathcal{G}'}{\mathcal{G}} = \frac{1}{1 + g_m R_k}.$$

Thus, if $ng_mR_k > 1$ just holds, the total reduction in gain will be only about 8 db, which in most cases is insufficient. If, however, $g_mR_k > 1$, more than 6 db per stage will be obtained, which is usually adequate. This then may be taken as the limiting condition for the type of cathode degeneration.

Where larger low-frequency attenuation is required than is practical with cathode degeneration, a high-pass coupling filter may be used. A filter with relatively sharp cutoff is desired, although simplicity is very important. The addition of any components in the coupling circuit adds both to the difficulty of minimizing stray capacitances and to problems of stability. A simple solution is shown in Fig. 18·11. Since a single capacitance-inductance filter is efficient in itself, it will seldom be necessary to use more than one section of filter per stage. The appropriate cutoff frequency can be calculated by conventional filter theory.

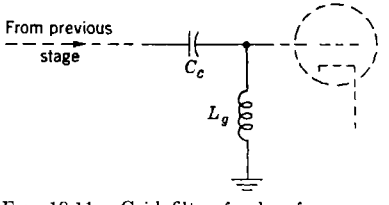


FIG. 18·11.—Grid filter for low-frequency rejection.

18·5. Limits of Performance.—As for all types of amplifiers, the shunt-series amplifier has definite practical limits on bandwidth, gain, and flatness of response. Obviously, the extreme limit on bandwidth is reached when the stage gain reaches 1.0. For shunt-series peaking using 6AK5's operated at a g_m of 5000, the gain-bandwidth product is approximately 180 Mc/sec. For bandwidths greater than 180 Mc/sec, therefore, the amplifier would result in a loss rather than a gain.

The practical limit on gain cannot be established so easily. With present techniques, a gain of above about 110 db in a single amplifier becomes exceedingly difficult to achieve because of stability problems. A more definite limitation on gain follows from the limit on flatness of response.

Reference to the curves of Fig. 18·8 for the shunt-series-peaking filter shows that although the frequency characteristic is very nearly flat, there is, for each ratio of C_1/C_2, a slight rise in gain near the high-frequency cutoff point. There does not seem to be any simple way of adjusting the values of the filter to give an exactly flat response without losing considerably in gain-bandwidth product. This slight peak is insignificant when only three or four stages are cascaded but becomes much more important when double this number of stages is contemplated. The values of the components appearing in the circuit also have an important effect on flatness of response. Tube capacitance variations and manufacturing tolerances of resistors and chokes all add a probable error which results in sizable variations in the shape of the frequency-response curve.[1] A practical limit to the actual number of stages is, therefore, reached when the combination of the peaked response of the filter and unavoidable variations of components combine to produce the maximum allowable variation in gain over the pass band. Since the gain per stage is fixed, the limitation of gain follows immediately if the amplifier is designed for a particular bandwidth.

18·6. Input Circuits.—The design of a receiver using a shunt-series-peaking type of i-f amplifier leads to a special problem in selecting a satisfactory form of coupling between the mixer and the grid of the first amplifier stage. Because of the very large bandwidth usually encountered and the particular frequency range, it usually is impossible to obtain any voltage stepup in the input circuit and practically impossible to use any of the conventional bandpass input circuits.

Following the same line of reasoning as that given in Sec. 18·2 for a low-pass instead of a bandpass filter for i-f coupling, it is evident that an input circuit of the low-pass type is advantageous. The simplest solu-

[1] A system of tuning to overcome such variations is theoretically possible, although the problem of making such adjustments in a multistage high-frequency amplifier using a complicated filter is usually of prohibitive difficulty.

tion would be to connect the output of the mixer directly to the grid of the first amplifier stage, possibly through a blocking circuit. The 3-db bandwidth of the input is then given by

$$\mathcal{B} = \frac{1}{2nR_m(C_m + C_2)},$$

where R_m and C_m are the mixer resistance and capacitance respectively and C_2 is the tube input capacitance. Where it is impossible to use such a scheme, more complicated filters can be considered. The number of applicable filter types is limited, however, because of the fact that any additional resistive loading of the mixer results in a loss of signal-to-noise ratio at the input and corresponding loss in actual receiver sensitivity.

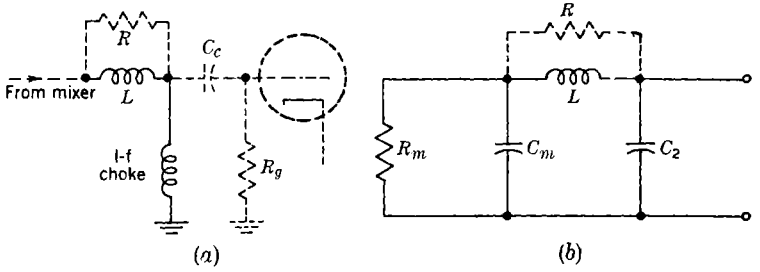


FIG. 18·12.—(*a*) Series-peaking input circuit; (*b*) equivalent circuit.

The circuit most practical for this application is a series-peaking filter (Fig. 18·12*a*), which has the advantage of great simplicity (one or, at the most, two added components) and considerable advantage over direct coupling. From the equivalent input circuit (Fig. 18·12*b*), the value of the series-inductance arm L can be calculated by the formula

$$L = R_m^2 C_m \frac{(1 + k)}{k},$$

where $k = C_2/C_m$, and the bandwidth by the formula

$$\mathcal{B} = 1/2\pi R_m C_m.$$

The frequency characteristic of such a filter will show a sizable peak (sometimes as much as 6 db) near the high-frequency cutoff. When this peak is objectionable, damping of the series inductance L can be introduced without affecting the bandwidth or the sensitivity too much. The amount of damping for the desired characteristic can, in general, be determined most easily by experiment.

18·7. "Echo Suppression" Circuit.—All amplifiers are, in the strictest sense, nonlinear devices. Furthermore, their gain at any instant depends to some extent on the previous history of signals to which they have been

subjected. Although such nonlinearity is ordinarily thought of as undesirable, it is often used to accomplish special tasks. There sometimes arises a need, for example, to construct an amplifier that upon the recep-

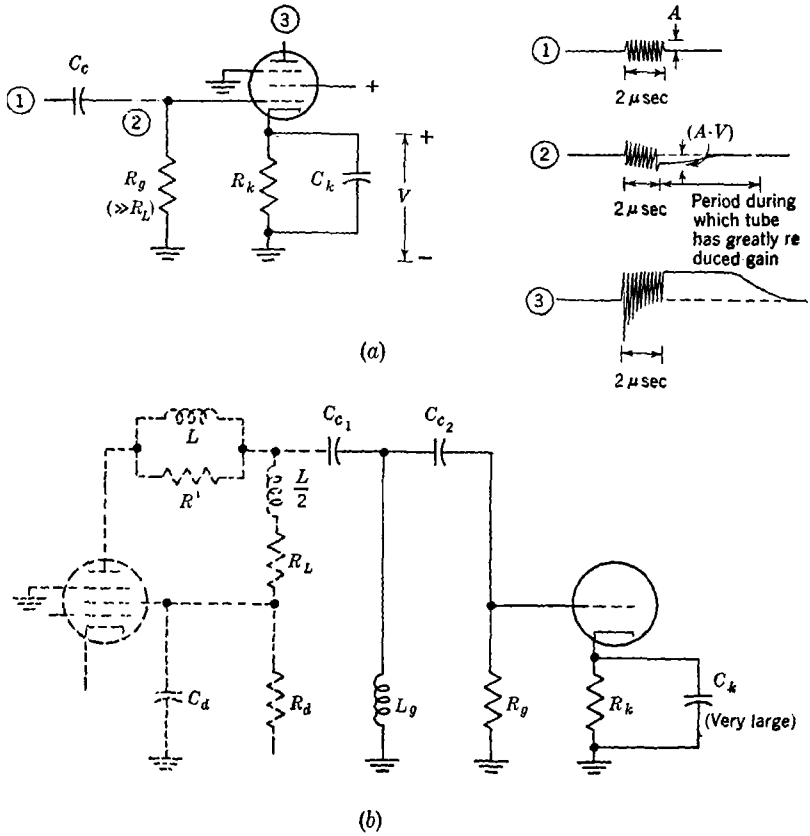


FIG. 18·13.—Echo suppression. (*a*) Simplified circuit illustrating waveforms; (*b*) practical coupling circuit.

tion of a signal of specified strength becomes comparatively insensitive for a definite period of time. The "echo-suppression"[1] circuit is one possible scheme to accomplish this purpose.

[1] The term "echo suppression" originated from a particular application in which the r-f echoes of a pulsed signal were to be suppressed in the i-f amplifier and their effect minimized. Since these echoes necessarily occurred after the main signal and were known to be of lesser amplitude, the problem resolved itself as one of desensitizing the amplifier for the period of time in which undesired echoes were to be expected. This problem was considered by Dayle Collup, of the Naval Research Laboratory, who is largely responsible for the solution described here.

Although the necessity for this type of response is not peculiar to applications involving the shunt-series-peaking type of amplifier, modification of this amplifier for such performance is particularly simple. However, the general method of attack can be applied to most of the other types of amplifiers.

The echo-suppression problem immediately suggests the use of some means of automatic gain control. Although so-called back-bias schemes could be adapted to accomplish the desired variation in amplifier gain, a much simpler solution involves only the circuit elements shown in the circuit of Fig. 18·9. By correct selection of the value of the grid return resistor R_g and the coupling condenser C_c, an actual IAGC effect can be obtained. The grid circuit has been redrawn in Fig. 18·13*a* along with the waveforms showing the response to an i-f pulse of 2-μsec duration. If the pulse has an amplitude A greater than the normal grid-to-cathode negative bias V, the grid will be driven positive, and grid current will flow, charging the condenser C_c. Immediately after the pulse the grid will be biased negatively by the charge on C_c, and the tube will have reduced gain. The tube gain will return to normal as the grid potential returns exponentially toward ground. This effect is exactly what is sought for echo suppression. For best operation, the tube must be operated in such a manner that the tube input resistance (with the grid positive with respect to the cathode) is minimized, in order that the greatest possible grid current may flow during the pulse. Therefore either the tube should be operated at zero bias or, where this is not practical, the grid should be biased in such a way that no degeneration results.

The time constants involved in the "charge" (change of condenser charge during the signal) and "discharge" (return of condenser to original charge) must be correctly adjusted for the desired characteristics. For instance, for maximum efficiency the charge cycle should be, for all practical purposes, complete before the end of the pulse, whereas the discharge time is dictated by the length of time during which the tube is to have reduced gain. The charge time constant is seen to be the product of C_c and $(R_L + R'_g)$, where R'_g is the effective input resistance of the pentode when operated with its grid positive. For the case of 6AK5's, the input resistance can be made as low as 200 ohms. If R_L is given the value 500 ohms, then a value of C_c may be calculated so that the resulting time constant is less than the duration of the pulse (2 μsec). For instance, if 0.5 μsec is selected for the charge time constant, then

$$C = \frac{t}{R} = 0.5 \times \frac{10^{-6}}{700} = 710 \qquad \mu\mu\text{f}.$$

The discharge time constant is equal to $C_c(R_L + R_g)$, where R_g is the grid-return resistor. Since C_c has been set by the above calculation, R_g

may then be determined so that the discharge time constant causes reduction of the gain of the tube for the required length of time. If this period is, for instance, 20 μsec, then

$$(R_L + R_g)C_c = 20 \times 10^{-6} = (R_g + 500) \times 710 \times 10^{-12}$$

Solving,

$$R_g = 27{,}700 \quad \text{ohms.}$$

In practice, the circuit of Fig. 18·13*b* has been used. Here a condenser-choke filter (C_{c_1} and L_c) has been added. A filter of this type is essential for the rejection of low frequencies, since the use of cathode degenerative circuits is impossible because of the requirement of high grid current for correct echo suppression. For the above calculation, C_c then becomes the series combination of C_{c_1} and C_{c_2}. The inductance L_c ordinarily has no effect on the operation of the suppression circuit, since it has a relatively high impedance for the frequencies concerned.

For true suppression over large ranges of signal and echo amplitude, it is not sufficient to use the circuit of Fig. 18·13*b* in one stage only. The stage for which the suppression circuit should first be considered is the last amplifier stage at which point the signal is largest (that is, suppression will start for the smallest signal input to the amplifier). If, however, signal and echo input to the amplifier are of sufficient amplitude to limit before the last stage of the amplifier, then the signal appearing at the grid of this stage will be a single elongated pulse. Reduction of tube gain will then be effective only at the end of this elongated pulse instead of immediately after the main signal, and the purpose will be defeated. Theoretically, therefore, the suppression circuit must be included in all stages. It can usually be determined, however, at which point in the amplifier limited signals will not be obtained at the maximum expected input to the amplifier. Suppression circuits in stages preceding this point will then be unnecessary.

PRACTICAL EXAMPLES OF WIDE-COVERAGE RECEIVERS

The three examples of practical receivers that are described in detail in this section illustrate the application of the principles described earlier and in particular the schemes shown in Figs. 18·2 to 18·4. The characteristics of these receivers (based primarily upon production specifications) are listed in Table 18·2. Comparison of these figures will show some difference in performance, particularly in power requirements, but, in general, the sensitivity and bandwidth attained are of the same order.

SINGLE-SIDEBAND RECEIVER

A block diagram of the single-sideband receiver is shown in Fig. 18·14. This receiver was designed for an application requiring high

TABLE 18·2.—CHARACTERISTICS* OF PRODUCTION RECEIVERS

Receiver	Coverage desired, Mc/sec	Actual coverage, Mc/sec	Sensitivity (at weakest point in band), w	Variation in sensitivity over the band, db	I-f band limits, Mc/sec	Video bandwidth, Mc/sec	Pulse rise time (at output), μsec	Power		
								I-f, ma (unreg.)	Video, LO misc., ma (reg.)	Heaters, amp. @ 6.3 v a-c
Single sideband	3267–3333	3264–3334	1×10^{-10}	3– 8	6–76	1.7	0.5	162 @ +105 v (unreg.)	56 @ +250 v (reg.)	6.6
Switched LO	9320–9430	9312–9438	2×10^{-10}	3– 5	8–48	1.7	0.5	96 @ +150 v (unreg.)	64 @ +250 v (reg.)	6.5
F-m LO	9320–9430	9314–9436	1×10^{-9}	6–10	1–21	0.3	2.0	60 @ +65 v (unreg.)	35 @ +330 v (reg.)	2.25

* All characteristics based on the reception of a 2-μsec pulsed signal.

sensitivity and good fidelity, with physical size and power requirements secondary in importance. Because of the bandwidth requirements, it was found practical to use a single-sideband design making possible the solution of the tuning problem by the use of an AFC circuit. The fidelity requirement necessitated fairly careful construction of the video amplifier, particularly in view of the circuits that were to follow the receiver.

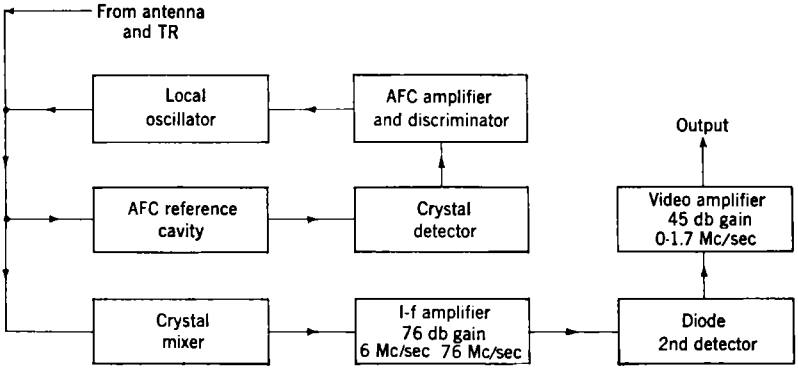


FIG. 18·14.—Block diagram of single-sideband receiver.

18·8. Local Oscillator and Mixer.—As shown in Fig. 18·2, the local oscillator is to be tuned to 3340 Mc/sec in order to cover the required band of 3267 to 3333 Mc/sec. The oscillator used is the 2K28, a reflex klystron with external cavity resonator. It is tuned either by varying the position of tuning slugs in the cavity (so-called "mechanical tuning") or by varying the reflector voltage ("electrical tuning"). As will be seen in Sec. 18·9 on AFC, automatic tuning is accomplished by controlling the electrical tuning, provided that the mechanical tuning has previously been adjusted to approximately the right point.

A 1N21A crystal is used for frequency conversion. The exact r-f system used is of some interest and is shown in semipictorial form in Fig. 18·15. Besides feeding the crystal converter with signal from the antenna and TR tube and LO power, the r-f system must couple a suitable fraction of the power to a reference cavity for AFC purposes. As is seen in Fig. 18·15, the received signal passes, essentially without interference, to the 1N21A converter through a $\frac{3}{8}$-in. coaxial line. A directional coupler of the coaxial type is used to couple LO power into this line. The coupler is formed by constructing a $\frac{7}{8}$-in. line using the outer conductor of the $\frac{3}{8}$-in. line as its inner conductor. The two coupling holes in the $\frac{3}{8}$-in. line outer conductor result in about 17-db attenuation in transferring LO power from the $\frac{7}{8}$- to the $\frac{3}{8}$-in. line. The 10-db attenu-

ator (see Fig. 18·15) approximately terminates the $\frac{7}{8}$-in. line and prevents unwanted reflections of the LO power. Beyond the attenuator an adaptor couples into a reference cavity that is used for AFC. The power reaching this point, although attenuated by 10 db, is sufficient to operate the AFC circuit.

The main advantages of this scheme are (1) that the local oscillator will not be affected by anything that occurs in the AFC circuits (because of the 10-db attenuator) and (2) that approximately constant input to the AFC reference cavity will be obtained. The second point is obvious when it is remembered that for optimum operation the mixer should be operated with a specified LO power that is adjusted to fairly narrow limits.

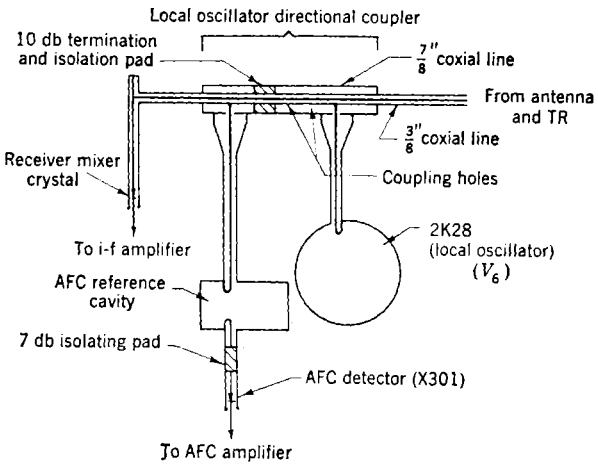


FIG. 18·15.—R-f connections of single-sideband receiver.

This adjustment can be accomplished by varying the coupling at the point where power is taken from the local oscillator itself (Fig. 18·15). If, therefore, the power reaching the mixer (and hence the power reaching the directional coupler) is held almost constant, the signal reaching the AFC cavity will also be almost constant. This property is important in that it facilitates the design of the AFC circuits.

The AFC reference cavity is of the transmission type and results in attenuation of about 6 db at its resonant frequency. Its output is connected through a 7-db isolating attenuator to the AFC detector (a 1N21A crystal).

18·9. AFC Circuit.—The scheme used for AFC is similar to that described in Chap. 3 under absolute-frequency AFC systems, the actual circuit being that shown in Fig. 18·16. Its operation is similar to the example treated in Chap. 3 except that the frequency modulation of the

local oscillator (necessary to provide error signal) is accomplished at a 60-cps rate instead of at 400 cps. The resulting frequency modulation is, in this case, less than ± 0.2 Mc/sec, and the total error in LO tuning including the effect of hunting that results from the use of the gas-tube control circuit is not more than ± 0.5 Mc/sec. This tuning error is not objectionable in this application, since a margin of safety is included in the actual frequency range covered by the receiver. In ordinary operation no manual adjustment of tuning is necessary. A visual check on the operation of the AFC circuit is given by the mixer crystal current

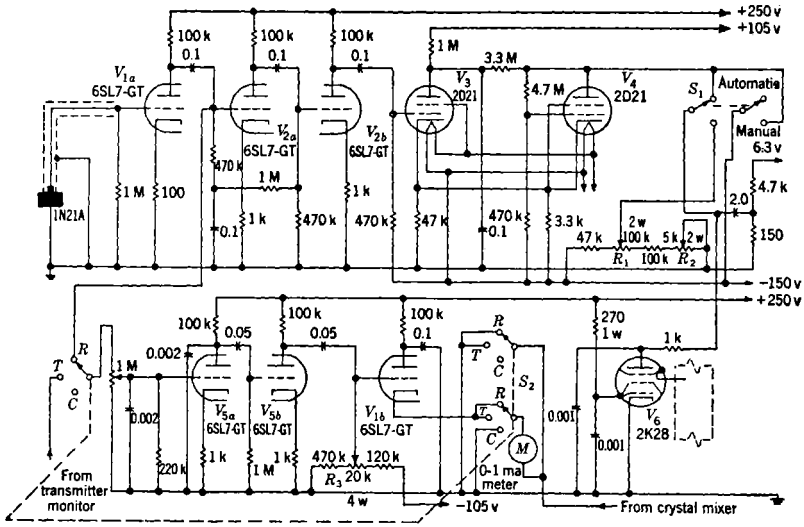


FIG. 18·16.—AFC and manual tuning circuits—single-sideband receiver.

(which is read on meter M, Fig. 18·16, when switch S_2 is in position C). When the gas-tube circuit is "searching" (that is, sweeping the reflector voltage of the 2K28) a periodic variation in crystal current occurs about once every 2 sec. When the AFC has locked (that is, is holding the local oscillator to the frequency of the reference cavity), the crystal current will exhibit no such variation. If adjustment of the mechanical tuning is required as a result of insertion of a new oscillator tube or other change, the same meter is used. The tuning slugs in the external cavity of the 2K28 are all turned so that they do not protrude into the cavity. The AFC will then be in the searching condition. Each tuning slug is then turned in very slowly. At some point searching will be observed to stop and the crystal current will remain stationary, indicating that the mechanical tuning is such that the oscillator has been tuned to the right frequency by automatic adjustment of the reflector voltage. The tuning

slugs are then set to maximize this steady crystal current, indicating operation at the center of the oscillator mode where it has greatest power output and maximum tuning range in both directions.

On occasion, it may be desirable to tune the local oscillator by manual means. This is accomplished by the use of the potentiometer R_1 and R_2 (Fig. 18·16) with switch S_1 in the manual position. If the mechanical tuning is known to be approximately correctly set, switch S_2 is turned to position R, which enables reading the amplified output of the AFC detector X_1 on meter M. The potentiometer R_3 is first set so that a reading of 0.02 or 0.03 ma is obtained on the meter. The coarse adjustment R_1 is turned until an indication is obtained. After setting the amplifier gain control R_4 so that the meter reading does not go off scale, a final adjustment is made with R_2, the fine control.

18·10. I-f Amplifier.—The i-f amplifier used is of the shunt-series-peaking type and uses selective cathode degeneration for low-frequency rejection. Its circuit is shown in Fig. 18·17*a*. This amplifier is a good illustration of the advantage of the shunt-series-peaking circuit and also of some of its limitations. A gain of about 8.5 db per stage or an over-all gain of 76 db is obtained in a nine-stage amplifier using 6AK5 pentodes

TABLE 18·3.—PARTS LIST FOR CIRCUIT OF FIG. 18·17*a*

Part No.	Value, μh	Turns	No. of wire (enameled)	Form	Method of winding	Remarks
L_1	3.80	43	30	$\frac{3}{16}$″ diam. × $\frac{5}{8}$″ long	Single layer, close wound	Values not critical
L_2	0.72	25	30	$\frac{1}{8}$″ diam. × $\frac{3}{8}$″ long	Single layer, close wound	Resonates with 37 μμf at 30 Mc/sec
L_3, L_{31}	20.00	76	36	$\frac{9}{16}$″ diam. × $\frac{3}{8}$″ long	Single pie, universal wound	Resonates with 30 μμf at 6.5 Mc/sec
L_4, L_7, L_{10}, L_{13}, L_{16}, L_{19}, L_{22}, L_{25}, L_{28}	8.90	72	35	$\frac{3}{16}$″ diam. × $\frac{5}{8}$″ long	Single layer, close wound	Resonates with 30 μμf at 9.7 Mc/sec
L_6, L_9, L_{12}, L_{15}, L_{18}, L_{21}, L_{24}, L_{27}, L_{30}	0.72	7	30	Hollow, $\frac{3}{8}$″ diam. × $\frac{3}{4}$″ long	Single layer, close wound	Resonates with 37 μμf at 30 Mc/sec
L_5, L_8, L_{11}, L_{14}, L_{17}, L_{20}, L_{23}, L_{26}, L_{29}	1.50	33	32	2.7k $\frac{1}{2}$-w Allen-Bradley resistor	Single layer, close wound	Resonates with 18 μμf at 30 Mc/sec

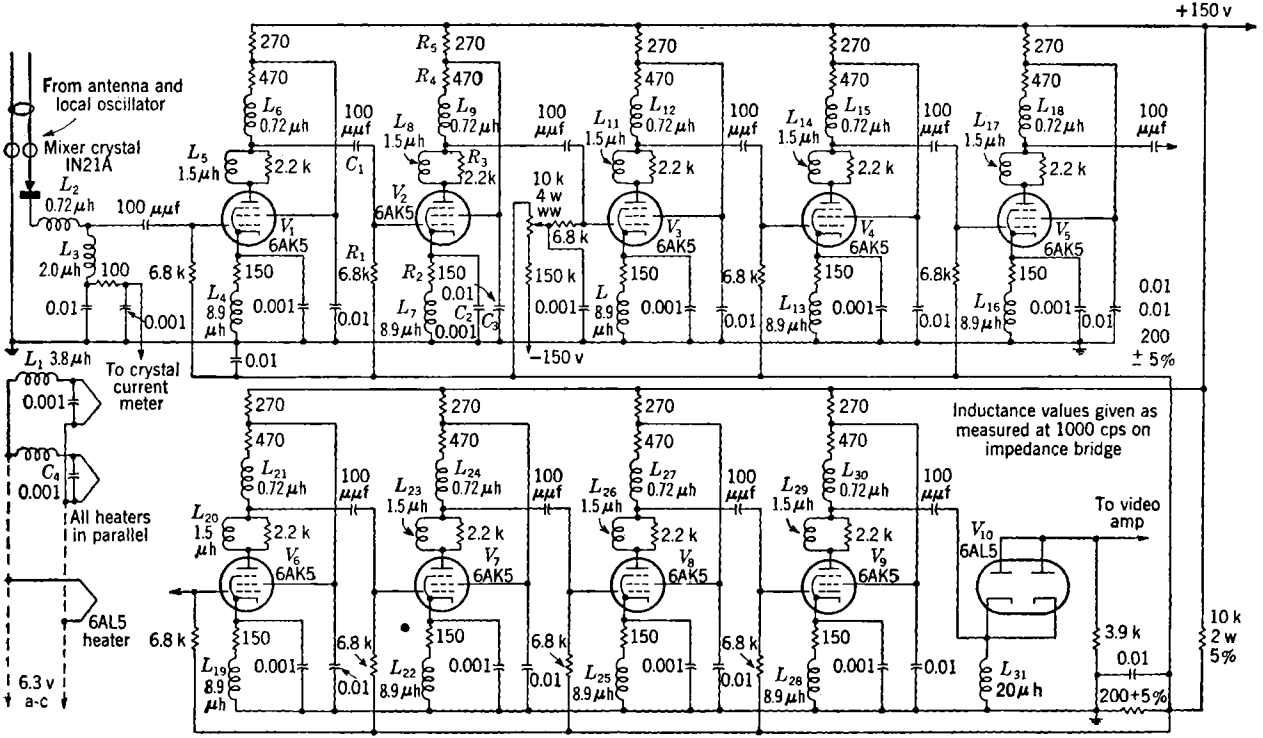


FIG. 18·17*a*.—I-f amplifier—single-sideband receiver (see Table 18·3 for parts list). Inductance values given as measured at 1000 cps on impedance bridge. All resistors ½ w and components ±10 per cent unless otherwise noted.

and having the pass band of 6 to 76 Mc/sec. The choice of such band limits means that in transforming a video amplifier whose pass band is 0 to 76 Mc/sec to a bandpass amplifier, only $\frac{6}{76}$, or about 8 per cent, of the bandwidth must be sacrificed. The resulting circuit still gives a

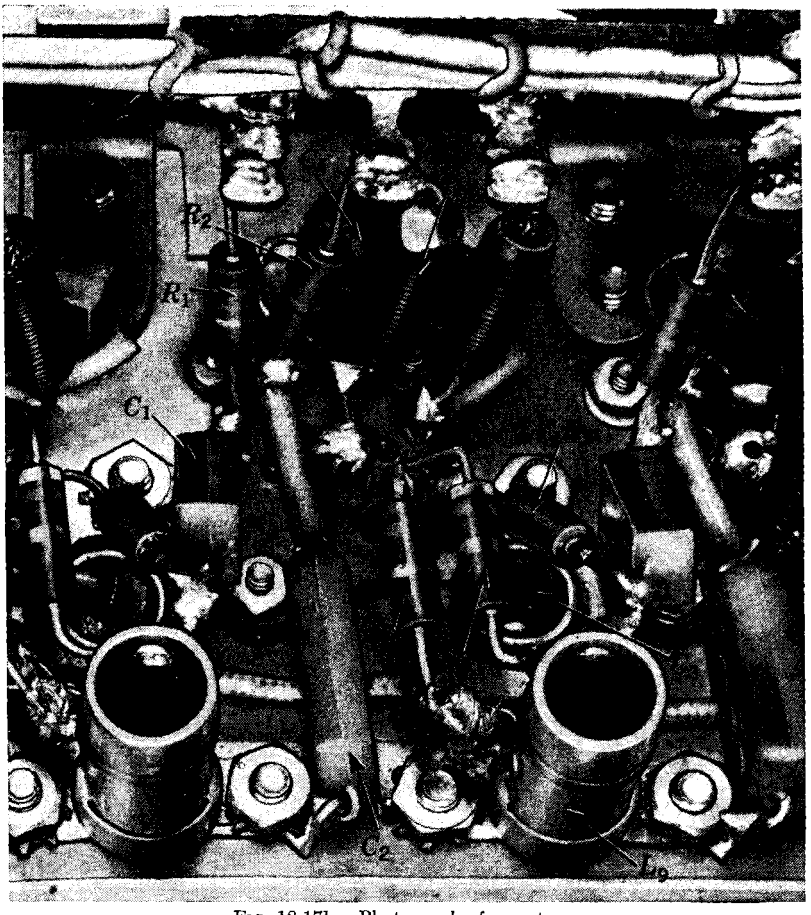


FIG. 18·17*b*.—Photograph of one stage.

tremendous advantage in over-all bandwidth over any of the usual true bandpass schemes.

It was stated in Sec. 18·2 that the practical limit on gain is closely connected with the limit on flatness of response. This amplifier illustrates this limitation in that the combination of bandwidth and high gain results in considerable uncertainty in the shape of response curve.

Tolerances in tube capacitances (about ±15 per cent for 6AK5's), in resistors and condensers (±10 per cent), and in coils (±5 per cent) are such that a variation in gain over the band of as much as 8 db must be expected. Inasmuch as tube capacitance variations have the greatest effect, little improvement can be obtained by narrowing tolerances of other components.

A somewhat unconventional scheme for providing grid bias is used consisting of a combination of cathode self-bias and fixed positive grid bias. This scheme has the advantage of allowing the use of a fairly large cathode-bias resistor (necessary for proper low-frequency rejection and for minimizing the effect of tube variations) while still running the tube at a low effective grid bias. The actual grid-to-cathode bias is adjusted to 0.6 volt which, with a 105-volt screen supply, results in a mutual conductance of 6500 μmhos (design center value). A low-limit g_m of 5700 μmhos is thus obtained for design calculations (from the tolerance on g_m), compared with 3900 μmhos usually assumed for 6AK5's under normal operating conditions (120 volts on the screen, −1.5-volts grid bias). It will be noted that a bus supplies all the grids from a bleeder with no decoupling between grids. Satisfactory stability under these conditions can be obtained only with extremely careful construction. The exact layout of one stage is shown in Fig. 18·17*b*.

18·11. Second Detector and Video Amplifier.—The fidelity of the wide-coverage type of receiver is determined almost entirely by the characteristics of the video amplifier, since the i-f amplifier has such an extremely large bandwidth. The statement of the fidelity requirement for this receiver is given in Table 18·2 as a pulse rise time of 0.5 μsec. The video bandwidth must therefore be about 1 Mc/sec.[1] The video amplifier (see Fig. 18·18), although it has no power-output stage as such, is arranged so that the last stage V_2 can drive a monitor amplifier V_{3a}, a series diode V_{3a} and a succeeding 6AC7 stage V_4. The bandwidth of this portion of the circuit is actually over 2 Mc/sec, and the first video stage V_1 has a bandwidth of about 1.7 Mc/sec.[2]

The circuits that the video amplifier drives are such that the minimum signal for correct operation must have about 10-volt amplitude. Since, for maximum sensitivity, signals approximately equal to noise must be received and amplified to useful level, the noise output of the video amplifier must necessarily be about 10 volts also. The gain of the video amplifier (about 45 db) is such that for the conditions of lowest i-f amplifier

[1] This value is obtained by making use of the well-known approximation giving the relation between video bandwidth and rise time by the following equation: Rise time (in μsec) = 0.35/bandwidth (in Mc/sec).

[2] The diode V_{3a} and the driven stage V_4 are really part of the circuits that this particular receiver was to operate and are not strictly receiver stages.

gain (that is, with tubes having low g_m) at least 10 volts of noise will be produced at the receiver output.

Although fidelity (and corresponding bandwidth requirements) and gain were the important factors influencing the design of the video amplifier, it was also necessary to assure correct operation over the whole range of signal strength to which the receiver would be subjected. This range is essentially from minimum signal strength (a signal of the order of noise amplitude) up to a signal strength sufficient to limit in the last i-f amplifier stage. The signal appearing at the output of the second detector in the limiting case is estimated to be about 5 volts in amplitude. Special

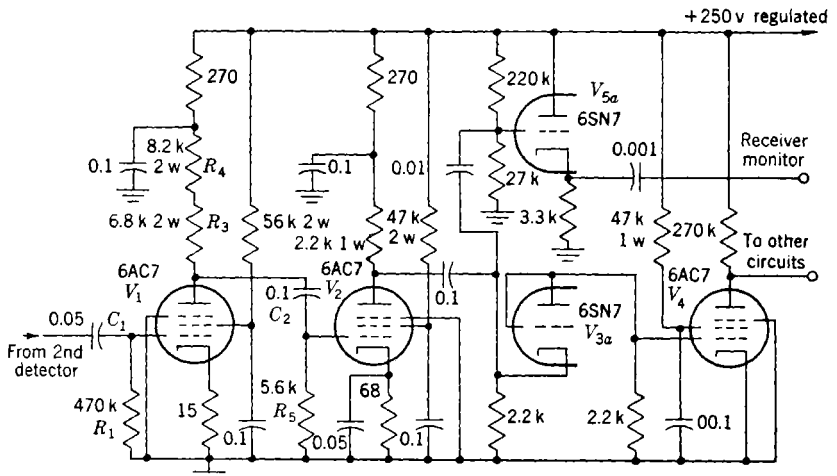


FIG. 18·18.—Video amplifier and monitoring circuit—single-sideband receiver. Note: One side of heater is grounded.

precautions are necessary to assure that the receiver will recover in the shortest possible time after the reception of a signal as large as this. Since the pulse at the output of the second detector is negative, the problem of coupling into the first video stage V_1 is to make the coupling time constant ($C_1 \times R_1$) large enough to prevent the formation of a harmful overshoot. The problem of coupling V_1 to the second video stage V_2 is complicated by the fact that the pulse at this point is positive. It is therefore practically impossible for large signals to avoid driving the grid of V_2 positive and causing long paralysis of the amplifier. This problem (which is a general one of high-gain multistage video amplifiers) is analyzed in detail in Chap. 19. As there shown, the plate resistor R_3 and R_4 should be made large, and the grid resistor R_5 small. The parallel combination of the plate and grid resistors is to give the correct load

resistance, calculated from bandwidth considerations. This combination results in negligible paralysis effect when the correct value of the coupling condenser C_2 is chosen.

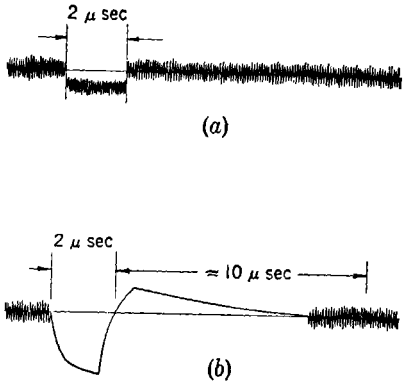


FIG. 18·19.—Waveforms obtained at monitor point. (*a*) Signal 6 db above tangential signal; (*b*) signal 60 db above tangential signal.

Approximate waveforms as observed at the monitor point under minimum and maximum signal conditions are shown in Fig. 18·19. Since the monitor amplifier V_{5a} is really a negatively driven cathode follower, the waveforms are not an accurate representation of the actual waveforms to be expected at the output of V_2. It is seen from Fig. 18·19, however, that complete recovery of the receiver after a strong signal occupies a period of about 10 μsec. This period was not considered objectionable in the application at hand.

18·12. General Information.—Inasmuch as there is appreciable variation in i-f amplifier gain over the band, different absolute sensitivities will result for signals of different frequencies. A logical convention therefore is to specify the sensitivity of the whole receiver as that sensitivity which is measured for a signal whose frequency corresponds to the point of minimum i-f amplifier gain within the pass band.[1] In addition, the band coverage of the receiver is arbitrarily defined as that range of frequencies over which the receiver has at least this sensitivity. A quick check of receiver performance can be made with a pulsed-signal generator of variable frequency and absolute pulse-power calibration. A measure of signal output can be accomplished visually by connecting a synchroscope to the video monitor point. The first step is to locate the frequency corresponding to minimum i-f gain. Figure 18·20 shows a representative response curve for the single-sideband receiver. If the

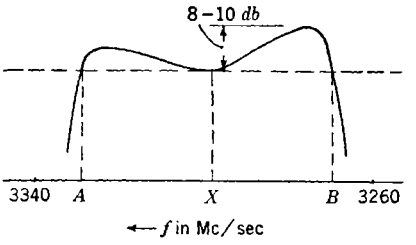


FIG. 18·20.—Representative response curve of single-sideband receiver.

[1] This convention applies only to receivers of the wide-coverage type; more accurate definitions of sensitivity are generally employed for ordinary broad-band receivers.

carrier frequency of the signal generator pulse is varied from approximately 3260 to 3340 Mc/sec and the pulse power is adjusted to the order of magnitude of the minimum signal (that is, about 10^{-9} watt), the signal appearing at the video monitor point will vary in magnitude, the minimum amplitude occurring at point X (Fig. 18·20). A careful measurement of signal-generator power is then made, with the generator tuned to frequency X, and its power adjusted to produce a tangential signal as observed on the synchroscope. This measurement gives the sensitivity of the receiver as defined above. With the signal generator adjusted to produce a pulse of exactly the same power, its carrier frequency is varied until at some frequency A (Fig. 18·20), the pulse on the synchroscope is again seen to be of tangential amplitude. This frequency A is defined as one limit of the frequency coverage. The procedure is repeated until another point B at the other end of the band is located. The difference of frequencies A and B is then the total coverage. Because of the means of measuring bandwidth, it is assured that over this coverage the receiver will have at least the sensitivity measured previously.

SWITCHED-LO RECEIVER

The basic requirements for this receiver were essentially the same as for the single-sideband receiver described previously (that is, high

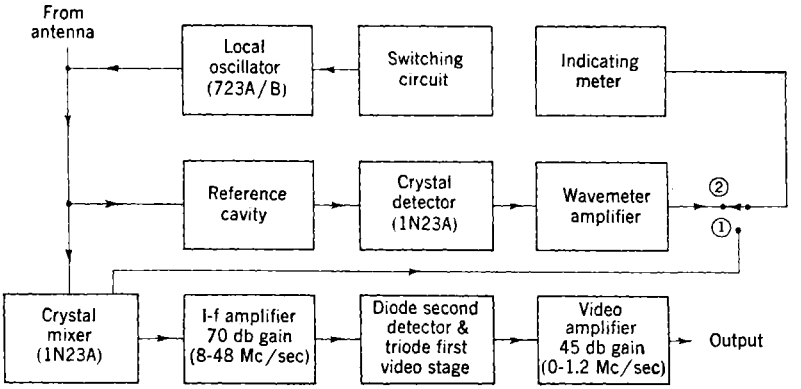


FIG. 18·21.—Block diagram of switched-LO receiver.

sensitivity and good fidelity), but the range of frequencies to be covered necessitated the use of the more complicated switched-LO technique. In addition, it was desired to have minimum variation in gain over the coverage band and to provide for good echo-suppression characteristics (see Sec. 18·7 for echo-suppression theory). Although, theoretically, the solution pictured in Fig. 18·3 is simple (provided the time-sharing aspect is not objectionable), a great practical problem arises in establish-

ing a workable and reliable procedure for tuning the receiver. AFC is difficult if not impossible, since the local oscillator must operate alternately at two accurately fixed frequencies and should give equal power at these frequencies. For reliable operation, temperature changes, mechanical shock, and power-supply variations must not detune the receiver appreciably over a specified period of time. In spite of all efforts at

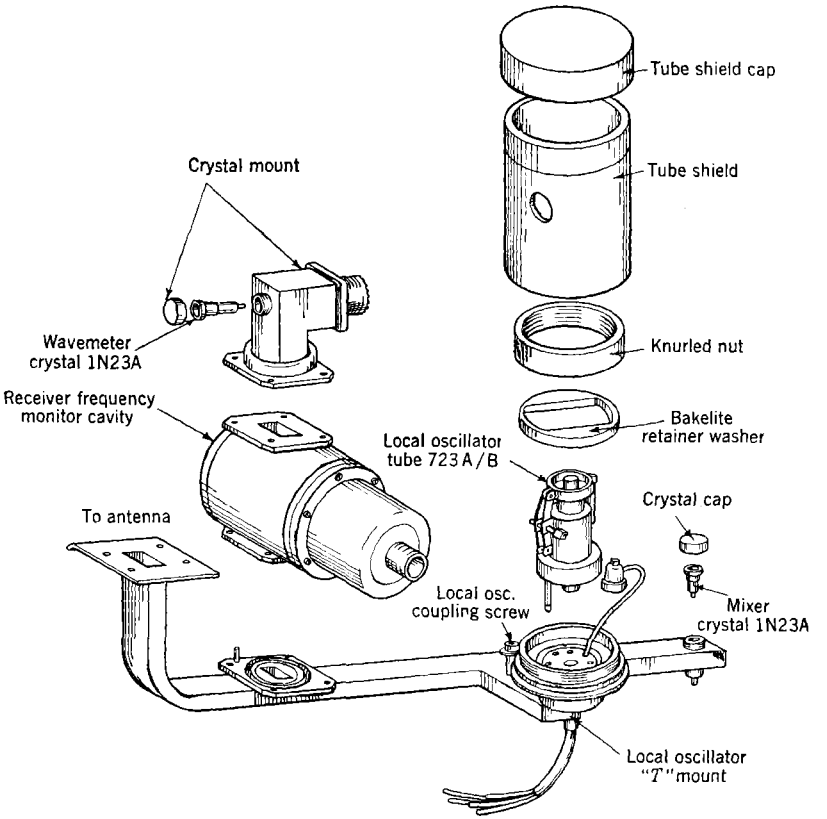


FIG. 18·22*a*.—Receiver r-f components.

minimizing frequency drift, however, it is found that periodic adjustments or at least checks of the receiver tuning are necessary, and a simple tuning method is therefore imperative. A block diagram of the complete receiver is given in Fig. 18·21.

18·13. Local Oscillator and Mixer.—The r-f assembly is shown in pictorial form in Fig. 18·22*a*. Since no TR tube is used, the r-f problem is simple, the signal being transmitted through waveguide directly from

the antenna to the crystal converter. The local oscillator is mounted on a shorted stub, which is coupled to the main guide as shown in Fig. 18·22*b*. A screw-driver adjustment is provided which varies the coupling

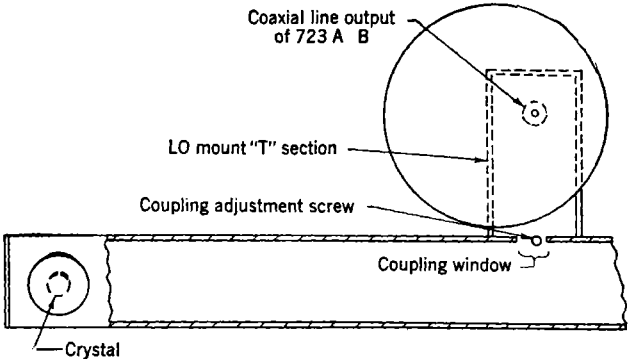


FIG. 18·22*b*.—LO mount T-section and crystal holder.

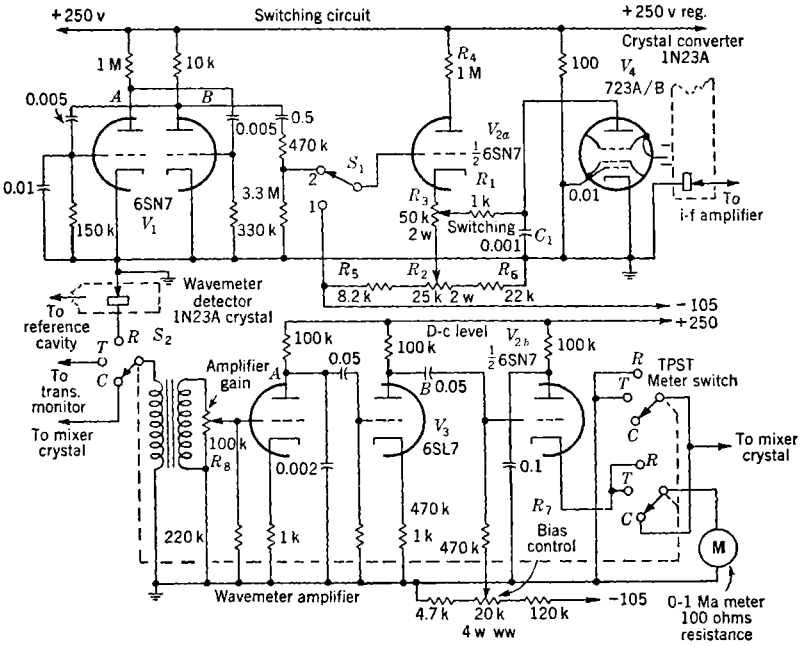


FIG. 18·23.—Tuning circuit. Switched-LO receiver.

of the local oscillator by inserting a pin into the coupling window. For tuning purposes a frequency-monitor cavity is coupled into the main guide by a waveguide T in such a manner that at resonance about 10 per

cent of the LO power is transmitted to the wavemeter crystal. The mixer crystal is a 1N23A crystal mounted in the waveguide at such a distance from the shorting endplate that for most crystals a good impedance match is obtained. The placing of the monitor cavity and LO mount along the guide is therefore not critical. The LO power upon entering the main waveguide splits approximately equally between the mixer and antenna. This method has the advantage of allowing placement of the reference-cavity coupling anywhere along the main waveguide but has the disadvantage of considerable LO radiation from the antenna with subsequent problems of interference with other systems.

As shown in Fig. 18·3, the local oscillator must be operating alternately at two frequencies, 9360 and 9390 Mc/sec, each accurately determined. The reference cavity is constructed so that its resonance frequency can be varied between these two limits, the limits having an accuracy of about ± 0.5 Mc/sec. An approximate reference point at 9375 Mc/sec is also provided, the purpose of which is explained subsequently. There is a 1N23A crystal in the output of the cavity whose output goes by way of a coaxial cable to the tuning circuits (Fig. 18·23).

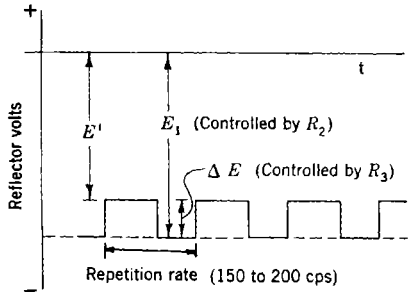


FIG. 18·24.—Operation of switching circuit controls.

18·14. Tuning Circuits.—The requirement of having the local oscillator operate at two alternate frequencies is best fulfilled by switching the frequency electronically. The difference in the two operating frequencies (30 Mc/sec) is such that this switching may be done simply by applying the proper square-wave voltage to the reflector of the 723A/B[1] or its equivalent, the 2K25.

Tubes $V_{1a\text{-}b}$ and V_{2a} (Fig. 18·23) comprise the switching circuit. The output of the circuit is obtained at the junction of R_1 and C_1 and is connected directly to the reflector of V_4 (the 723A/B local oscillator). The waveform at this point, as well as the basic function of the switching circuit, is shown in Fig. 18·24. Although it would be desirable to be able to control the absolute levels E_1 and E' completely independently (and thus adjust the two LO frequencies independently), this is not easily accomplished. Instead, a compromise has been adopted. The level E_1

[1] The specifications on the 723A/B require a width of at least 30 Mc/sec between half-power points in mode 3; it is thus assured that when the tube is adjusted to give equal power at the two frequencies, this power will be at least half the maximum power obtainable.

is controlled by the setting of potentiometer R_2, and the quantity ΔE, the difference in voltage of the levels E_1 and E', is controlled by potentiometer R_3. The setting of R_3 will not influence level E_1, and potentiometer R_2 may be varied somewhat without affecting ΔE. Sections V_{1a} and V_{1b} make up a free-running multivibrator, which generates a square wave of about 150-cps repetition rate. When switch S_1 is in position 2, this signal is coupled to the grid of V_{2a} at such an amplitude that the positive half-cycle tends to drive the grid positive, while the negative half-cycle renders the tube nonconducting. Since R_4 is so large, the current through V_{2a} during the positive half-cycle is determined chiefly by this resistance and the applied plate voltage and will not be affected much by small changes in R_2. During the negative half-cycle, with V_{2a} cut off, the potential applied to the reflector of V_4 is determined entirely by the voltage divider R_5, R_2, and R_6 and corresponds to the level E. During the positive half-cycle, there is added to E_1 the drop across that part of R_3 which is in the circuit, which corresponds to ΔE. Accordingly, there is the condition that E_1 is entirely independent of ΔE and ΔE is nearly independent of E_1.

The portion of the circuit of Fig. 18·23 that is labeled "wavemeter amplifier" consists of a two-stage amplifier V_{3a} and V_{3b} and a detector V_{2b}. The meter M serves three functions, depending upon the setting of switch S_2. With S_2 in position C, the meter indicates the mixer crystal current and in positions T and R acts as a tuning indicator for the transmitter and receiver respectively. For greatest sensitivity the detector V_{2b} is adjusted by means of the bias control R_7 so that its cathode current with no signal input to the amplifier varies from about 20 to 40 μa. An a-c signal of 10 to 20 mv at the detector crystal will result in half- to full-scale reading on the meter. In the event that the meter is driven off scale, the amplifier gain can be reduced by the use of potentiometer R_8. This means of indication depends upon the presence of an a-c signal at the crystal detector, and the r-f signal must therefore be modulated in some way. This requirement plays an important part in determining the best tuning procedure.

The controls that are used in the tuning procedure are the following:

1. Wavemeter control. This is a shaft control which varies the reference-cavity frequency between the limits of 9360 and 9390 Mc/sec with an approximate set point at 9375 Mc/sec.
2. Meter switch. Switch S_2 (Fig. 18·23).
3. Switching control. Potentiometer R_3 (Fig. 18·23). This potentiometer is ganged with switch S_1, so that in the OFF (completely counterclockwise) position, S_1 is in Position 1, and is in Position 2 for any other setting of R_3.

4. D-c level control. Potentiometer R_2 (Fig. 18·23).
5. Amplifier gain. Potentiometer R_8 (Fig. 18·23).
6. Bias control. Potentiometer R_7 (Fig. 18·23).
7. Mechanical tuning. This is a shaft control which deforms the LO resonator cavity and in this way changes the LO frequency mechanically.

18·15. Complete Tuning Procedure.—The steps that will enable an inexperienced operator to tune the receiver correctly and accurately are as follows:

1. Turn the meter switch to position R and the switching control to the OFF position. Set the amplifier gain for minimum gain. Make sure that the meter M indicates a current of 20 to 40 μa; if this is not so, adjust the setting by means of the bias control.
2. Set the meter switch to position C (reading mixer crystal current); adjust the d-c level control for maximum reading.
3. Turn the switching control slowly clockwise; the mixer crystal current will decrease to about half of its former value and stay there over a moderate range. Set the control for the middle of this range.
4. Set the wavemeter control in the middle of its range (that is, a frequency of about 9375 Mc/sec). Set the meter switch to R. Set the amplifier gain at maximum. Turn the mechanical-tuning control very slowly, watching for an indication on meter M. Do not turn it more than one-half turn in either direction from the starting position. When a deflection occurs, adjust the mechanical-tuning control to give maximum meter reading. If the meter goes off scale, turn down the amplifier gain. If no meter deflection occurs within one-half turn of the mechanical-tuning control, as may happen when a new 723A/B tube has been installed, turn the mechanical-tuning control counterclockwise as far as it will go. Then turn it clockwise, one turn. Set meter switch on C; adjust d-c level for maximum crystal current; and return meter switch to R. Turn the mechanical-tuning control slowly clockwise. If no meter deflection occurs within one-half turn, set the meter switch to C, readjust the d-c level for maximum crystal current, return the meter switch to R, and continue turning the mechanical-tuning control clockwise up to another half turn. Repeat this procedure until a deflection has been found and is maximized for both d-c level and mechanical-tuning adjustments simultaneously.
5. Turn the wavemeter control to 9390 Mc/sec. Turn the d-c level slowly clockwise until a meter deflection occurs. Adjust for maximum meter reading.

6. Turn the wavemeter control to 9360 Mc/sec. Turn the switching control slowly counterclockwise. When a deflection occurs, maximize the reading.
7. Check the frequencies once more; turn the wavemeter control to 9390 Mc/sec, and adjust the d-c level for maximum meter reading; then turn the wavemeter control to 9360 Mc/sec, and adjust switching for maximum. The two readings should be approximately equal, indicating equal power from the local oscillator at both frequencies.

The sequence of events accomplished by these steps may best be explained by use of Fig. 18·25, which shows approximate characteristics of a 723A/B or 2K25 for two settings of the mechanical-tuning control. Frequency and power output are plotted against reflector voltage. The dashed curves represent a possible initial condition with the mechanical tuning set to too low a frequency, and the solid curves represent the final desired state, with the voltage level E_1 corresponding to oscillation at 9390 Mc/sec, ΔE shifting the operating point to 9360 Mc/sec, and the mechanical tuning being correct to produce approximately equal power output at the two frequencies.

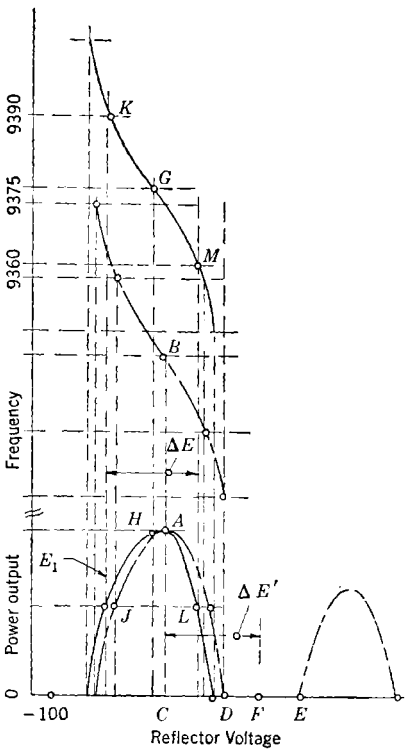


FIG. 18·25.—Typical characteristic curves for 723A/B tube.

Step 1 is concerned only with setting the operating point of the detector in the wavemeter amplifier to the correct value. Step 2 brings the power output of the oscillator to a maximum at some arbitrary frequency—in other words, sets conditions to correspond to points A and B of Fig. 18·25. Step 3 now introduces a value of switching voltage $\Delta E'$ sufficient to take the operating point from A to some point F in the range between D and E where the tube does not oscillate. Since half the time is now spent at A and half at F, the power output, or mixer crystal current,

will be about half what it was before. Furthermore, the r-f output of the oscillator will be modulated so that the wavemeter amplifier may be used. The oscillator will remain in this oscillating condition only half the time until Step 5, so that its output is at one frequency instead of two. Step 4 now brings the operating conditions to the points G and H, corresponding to maximum power output at the intermediate point of 9375 Mc/sec. Step 5 then sets E_1 to the point K, operation at 9390 Mc/sec; and step 6 sets ΔE to bring the conditions to point M, operation at 9360 Mc/sec. Step 7 merely rechecks to see that none of the previous adjustments have been disturbed.

After the completion of the tuning procedure, an adjustment of the mixer crystal current may be necessary in order to assure maximum conversion efficiency. The desired current has been found to lie between 0.3 and 0.7 ma. This condition can be obtained by adjusting the amount of LO power coupled to the mixer, which is determined by the setting of the coupling screw (Fig. 18·22*b*).

18·16. I-f Amplifier.—Although the necessary bandwidth (a theoretical minimum of 30 Mc/sec is needed to cover 120 Mc/sec) presents a simpler problem than in the case of the single-sideband receiver described above, an added difficulty is introduced by the requirement of small variation in gain over the band. As shown in Sec. 18·2, the total number of stages rather than the total gain usually determines the expected uncertainty in the i-f pass band. The use of a minimum number of i-f amplifier stages is thus indicated. It is still desirable, however, to have high gain in the i-f amplifier in order to have good second-detector efficiency and to reduce the problems connected with the video-amplifier design resulting from microphonics and recovery time.

The amplifier shown in Fig. 18·26*a* is a result of these considerations. It consists of nine 6AC7 pentode stages giving a gain of about 75 db, bandpass limits of 8 and 48 Mc/sec, and a maximum variation in gain over the band of 3 db. The particular band limits result in considerable overlapping between the operating pass bands (see Fig. 18·3) which allows a safety factor to compensate for any inaccuracies in oscillator tuning.

The first five stages of the amplifier use the coupling circuit (shunt-series peaking) shown in Fig. 18·9 and a cathode-degenerative circuit for low-frequency rejection. The circuits of the last four stages include specific provisions for echo suppression. The cathode circuit of V_6, for instance, provides for effective bypassing of low frequencies, which is necessary to minimize the tube input resistance (as explained previously). C_4, C_5, and L_4 form an effective high-pass filter to reject the low-frequency signals produced at V_6 in the presence of large signals. The circuit of the eighth stage (V_8) is arranged in a similar manner. The ninth stage (V_9) necessarily has the greatest echo-suppression effect, since input

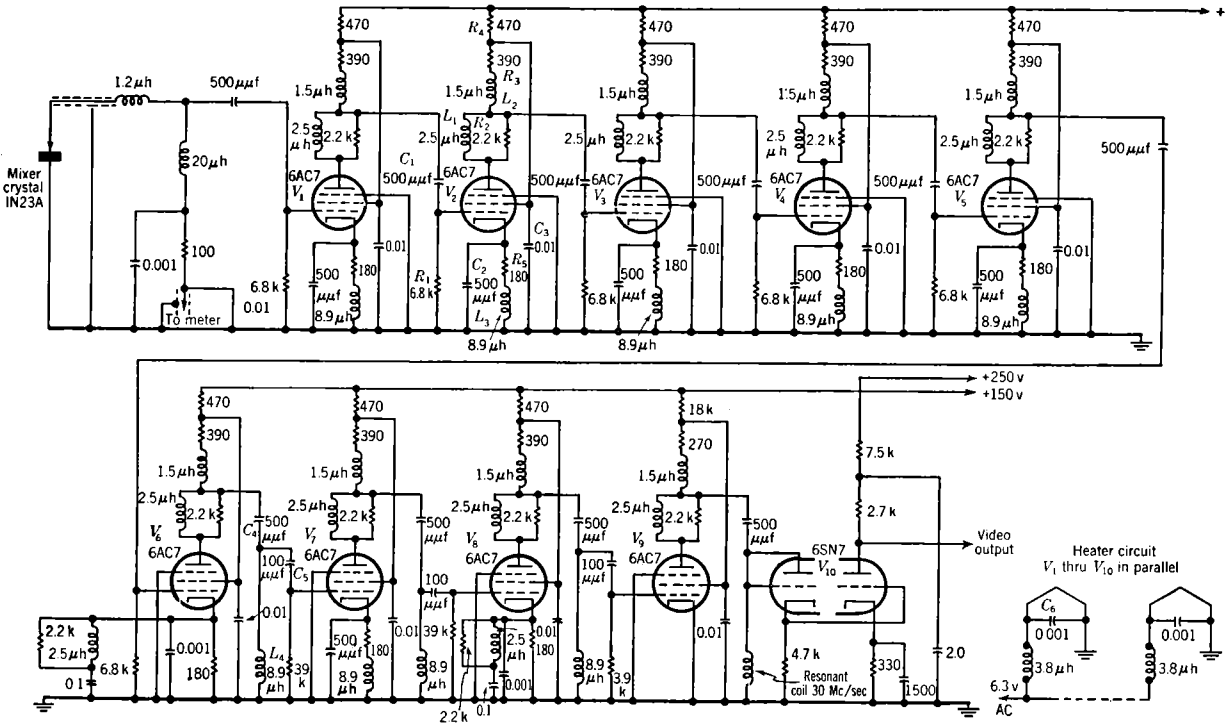


FIG. 18·26*a*.—I-f amplifier—switched-LO receiver.

signals of only a few decibels above noise amplitude have been amplified by the eight previous stages to an amplitude sufficient to draw considerable grid current. Consequently, V_9 is operated at zero grid bias and at a very low screen voltage (about 35 volts) to reduce its grid input resistance to a minimum.

A rough calculation similar to that given for the single-sideband i-f amplifier is used to determine the expected minimum gain of each stage.

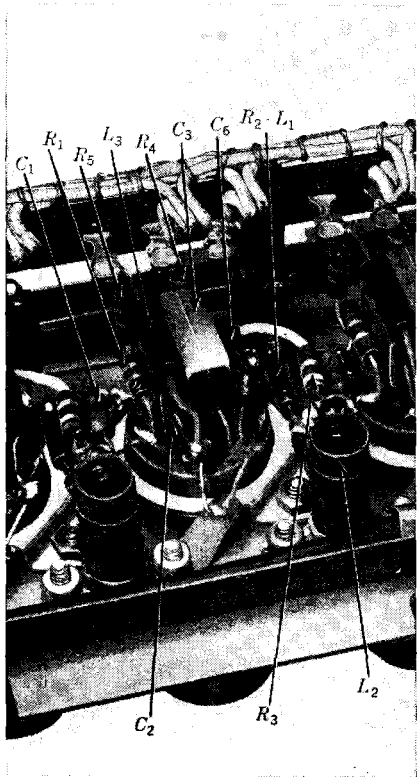


FIG. 18·26*b*.—Photograph of one stage.

V_1 through V_8 are operated with 146 volts on the screen (about 5 volts drop through the decoupling resistors) and 180 ohms in the cathodes. In this condition, the mutual conductance (design center) of 6AC7's is estimated to be 8300 μmhos. Considering the tolerance on g_m (about ± 20 per cent) and the degenerative effect of the cathode resistors, a low-limit g_m of about 7700 μmhos is to be expected. This value gives a minimum gain of 9.4 db per stage, or 75 db for the first eight stages. The

gain of the last stage is very small (about 1 db) because of the peculiar operating conditions required for echo suppression resulting in a g_m of about 4000 μmhos.

Considerable care must be taken in the mechanical construction of the amplifier. The placing of parts in one stage is shown in Fig. 18·26*b*. In addition to assuring minimum stray capacitance and inductance, this method of construction also assures very small variation in characteristics from one stage to the next, thus contributing to the flatness of i-f amplifier response. In order to reduce the number of coupling-circuit components, the series inductance and resistance arm (for instance, R_2, L_1) are constructed as one unit, with the inductance produced by a single-layer coil wound on the resistor itself.

No gain-control provision has been included in the i-f amplifier, since for best operation of the echo-suppression circuit it is desirable to have as large a signal as possible at the grid of the last i-f stage. Gain control is therefore accomplished after the last i-f stage. The arrangement used here involves controlling the gain of the video amplifier.

18·17. Second Detector and Video Amplifier.—The second detector consists of one-half of the double triode V_{10} (Fig. 18·26*a*) connected as a diode. The other half of the triode is operated as a conventional amplifier, the output of the diode being directly connected to the grid. Although the gain of the triode-video stage is low (about 14 db), it was included to allow the use of the same video amplifier as was used for the single-sideband receiver. The output of the triode is connected directly to the amplifier of Fig. 18·18. The bandwidth of the triode stage is sufficient (over 2 Mc/sec) to result in essentially the same video bandwidth achieved in the single-sideband receiver.

The only modification of the circuit of Fig. 18·18 is the inclusion of the gain control in the form of a potentiometer in the cathode of V_3. This potentiometer provides about 15 db of control.

18·18. General Information.—Since from the second detector onward the switched receiver is almost identical with the single-sideband receiver, the same remarks as to sensitivity and bandwidth measurements apply. Some mention should be made of additional problems introduced by the switching scheme itself.

The requirement that the local oscillator be operated so that equal power is produced at the two frequencies resulting from switching might be reached intuitively, but it may also be justified by a simple analysis. It is known from a study of crystal-converter characteristics that the sensitivity (that is, signal-to-noise ratio) will be almost constant over a fairly wide range of LO power. On the other hand, the rms noise produced by the mixer is very dependent upon LO power. It can thus be expected that if the power output of the switched oscillator were unequal

at its switched frequencies the noise pattern at the input to the second detector would appear as in Fig. 18·27*a*. The noise produced at the output of the video amplifier would then have the waveform shown in Fig. 18·27*b*. The envelope of the pattern of Fig. 18·27*a* has been partly differentiated by the coupling time constants of the video amplifier to produce "spikes" at a rate of twice the switching repetition rate. The result is that either the gain of the i-f amplifier must be reduced so that the spikes are no longer large enough to influence the circuits following the receiver or the spikes must be allowed to have the same effect as normal signals. This situation is obviously undesirable.

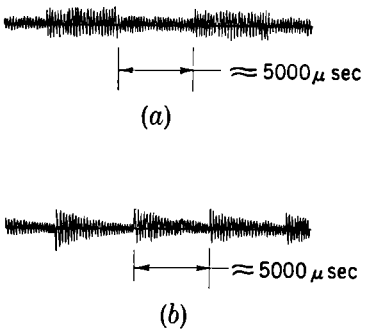


FIG. 18·27.—Appearance of receiver noise with unequal LO power at switched frequencies. Switching repetition rate ≈ 200 cps. (*a*) I-f noise; (*b*) noise at video monitor point.

FREQUENCY-MODULATED LO RECEIVER

The application for which this receiver was designed had somewhat different requirements from those treated in the previous receivers. In this case, high sensitivity was needed, but specific limitations were placed on the power-supply drain and on physical size. Fidelity was unimportant as long as a signal of sufficient amplitude was available to trigger succeeding circuits. The f-m LO scheme is well suited for an application of this kind.

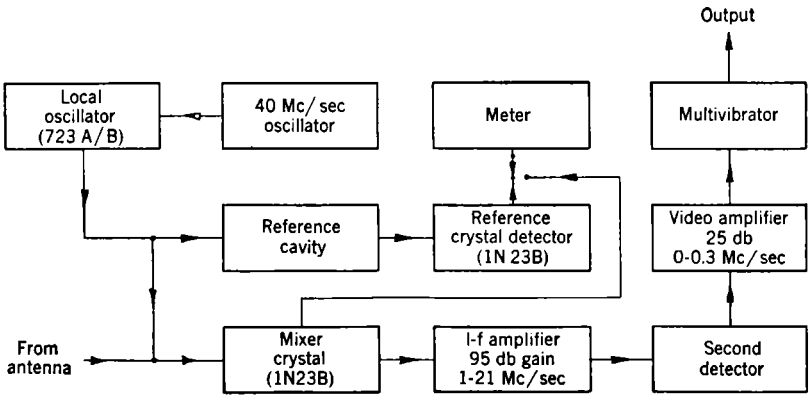


FIG. 18·28.—Block diagram of f-m LO receiver.

The receiver to be described has the characteristics pictured in Fig. 18·4 and is shown in block-diagram form in Fig. 18·28.

18·19. Local Oscillator and Mixer.—The mechanical assembly of the r-f components is shown in Fig. 18·29. Note the compact construction that can be achieved. The 723A/B LO tube is mounted on a piece of waveguide (*a*, Fig. 18·29) which is short-circuited at one end and is connected at the other end to a transmission-type reference cavity (*e*) of relatively low Q (about 2000 at 9375 Mc/sec). In the output of the cavity is a 1N23B crystal (*f*), whose rectified current is indicated directly on a 500-μa meter (not shown). A portion (about 20 per cent) of the LO power enters the coupling waveguide (*l*) and splits about equally

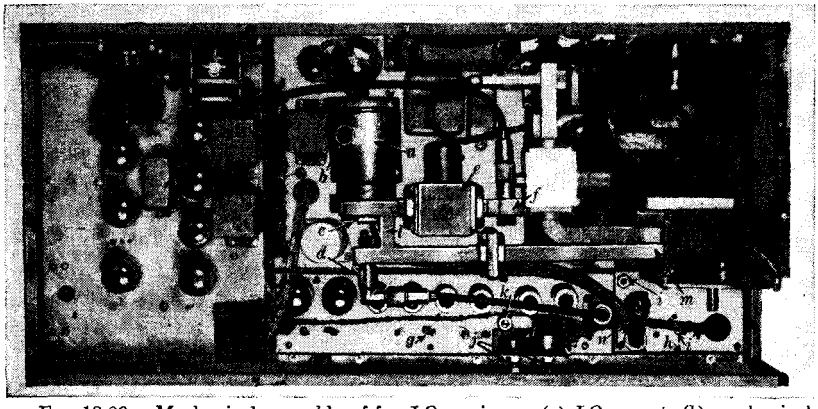


FIG. 18·29.—Mechanical assembly of f-m LO receiver. (*a*) LO mount; (*b*) mechanical tuning shaft; (*c*) LO coupling; (*d*) mixer-crystal mount; (*e*) reference cavity; (*f*) reference detector crystal mount; (*g*) i-f amplifier; (*h*) 40-Mc/sec oscillator; (*i*) 40-Mc/sec oscillator controls; (*j*) tuning controls; (*k*) i-f amplifier gain control; (*l*) coupling waveguide; (*m*) antenna input; (*n*) cable.

between the antenna and the mixer crystal (*d*). The amount of power reaching the mixer is adjustable by means of the coupling screw (*c*).

The necessary frequency modulation of the local oscillator is accomplished by applying the correct sinusoidal modulating voltage to the reflector. Since side frequencies spaced 40 Mc/sec from the fundamental frequency are required (see Fig. 18·4), modulation at a rate of 40 Mc/sec is used. This modulation produces such a deviation as to approach the optimum condition of equal power at the fundamental and side frequencies. The operation of the tube under these conditions is therefore closely tied up with its electronic-tuning characteristics. These characteristics have already been considered in some detail during the discussion of the switched-LO problem, and it may be remembered that the specification on the 723A/B requires at least 30 Mc/sec between half-power points of electronic-tuning range under specified operating condi-

tions. On the other hand it is stated in Sec. 18·1 that a deviation of ±57 Mc/sec (at a modulation rate of 40 Mc/sec) is theoretically required to produce equal fundamental and side-frequency power. This means a total frequency excursion of 114 Mc/sec, which is obviously impossible in the case of a tube with the minimum 30-Mc/sec tuning range.

The tuning characteristics of the 723A/B are much better for this application when it is operated under somewhat unusual conditions. The tuning range (between half-power points) is at least doubled in going to the mode that exists in the region of −60 volts reflector voltage, and another slight increase is obtained in raising the accelerator voltage from the customary 300 to 330 volts. This method of operation is used in the receiver herein described and makes the f-m scheme realizable.

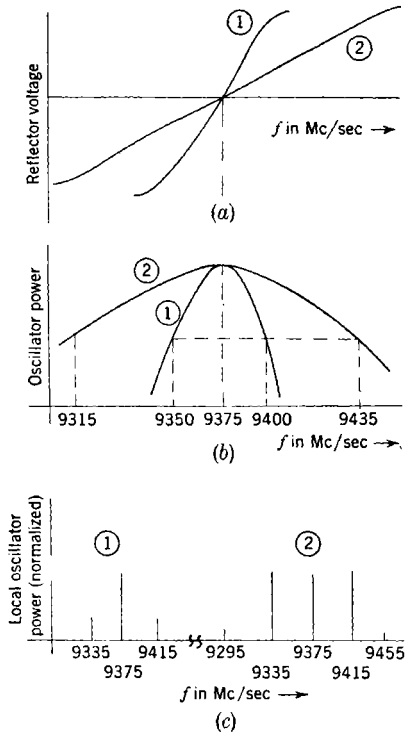


FIG. 18·30.—Characteristic curves of 723A/B tubes for use as f-m local oscillator. Tubes labeled 1 and 2 represent the extremes of allowable variation. (*a*) Frequency vs. reflector voltage; (*b*) power output vs. reflector voltage; (*c*) distribution of sideband power.

The next requirement is a method of operationally adjusting the amount of frequency modulation. Two of the easiest quantities to measure during adjustment of the receiver are (1) the magnitude of the reference-cavity crystal current, which is a measure of the power transmitted by the reference cavity, and (2) the rectified mixer-crystal current, which is a measure of the LO power reaching the mixer crystal. Consider what information can be obtained from these indications. Suppose that a 723A/B oscillator tube (Tube 1), which has the minimum expected electronic-tuning range (about 50 Mc/sec being a conservative estimate under the particular operating conditions) is chosen. If Tube 1 is tuned (without any modulation) so that the maximum power point of its mode corresponds to a frequency of 9375 Mc/sec, its characteristic curves may be represented by Curves 1, Fig. 18·30*a* and *b*. A reading of mixer-crystal current will then be a measure of the oscillator power at

the peak of the mode. Suppose now that an a-c voltage is superimposed on the existing reflector d-c voltage, creating frequency modulation. The operating point of the oscillator will travel symmetrically along the mode of Fig. 18·30*a*. If the limits of the frequency excursion (that is, travel of the operating point) coincide with the half-power points (the frequencies 9350 and 9400 Mc/sec), the reading of mixer-crystal current will now be lower by some factor n, since the average LO power reaching the crystal is decreased.

Tube 2 should be chosen for the maximum expected electronic-tuning range (about 120 Mc/sec), so that its characteristic curves are represented by Curves 2, Fig. 18·30*a* and *b*. If now the amount of modulation is adjusted for this tube so that the mixer-crystal current is observed to change by the same factor n, it may be assumed that the limits of its frequency excursion correspond approximately to the half-power points of its mode (provided that the power output curve is symmetrical and is similar in shape for the two tubes[1]).

This kind of adjustment may then result in producing an excursion of 50 Mc/sec for Tube 1 and 120 Mc/sec for Tube 2. It is now known, however, that a similar adjustment made on any other tube will result in some excursion between these two limits, since Tubes 1 and 2 were chosen for extreme characteristics. Since it can be shown that both Tubes 1 and 2 will operate satisfactorily when the factor n is chosen to be about 0.7, a usable and simple method of adjusting the amount of frequency modulation is available.

The tuning procedure[2] is somewhat analogous to that described for the switched-LO receiver, although simpler, since only the center frequency needs to be fixed. Furthermore, the method of power indication (crystal-current meter) responds to a c-w signal, so that some adjustments may be made without modulation. The first step is to place the operating point at the peak of the mode (corresponding to point A in Fig. 18·25). The frequency modulation is removed; the meter is connected to read mixer current; and the reflector voltage is varied to maximize the crystal current. The second step sets the mechanical tuning at the proper point. The meter is connected to the reference crystal, and the mechanical tuning is varied until some current is indicated. If it is necessary to turn this control more than about half a turn, voltages should be reset to the peak of the mode by repeating Step 1. When an

[1] This assumption is not exact, since the symmetry depends to a considerable extent upon the construction of the oscillator mount as well as on details of the tube structure, which may change from tube to tube.

[2] If either the oscillator tube or the r-f assembly has an accidental sharp resonance in the vicinity, this adjustment procedure may not work out easily. Such resonances should not be present.

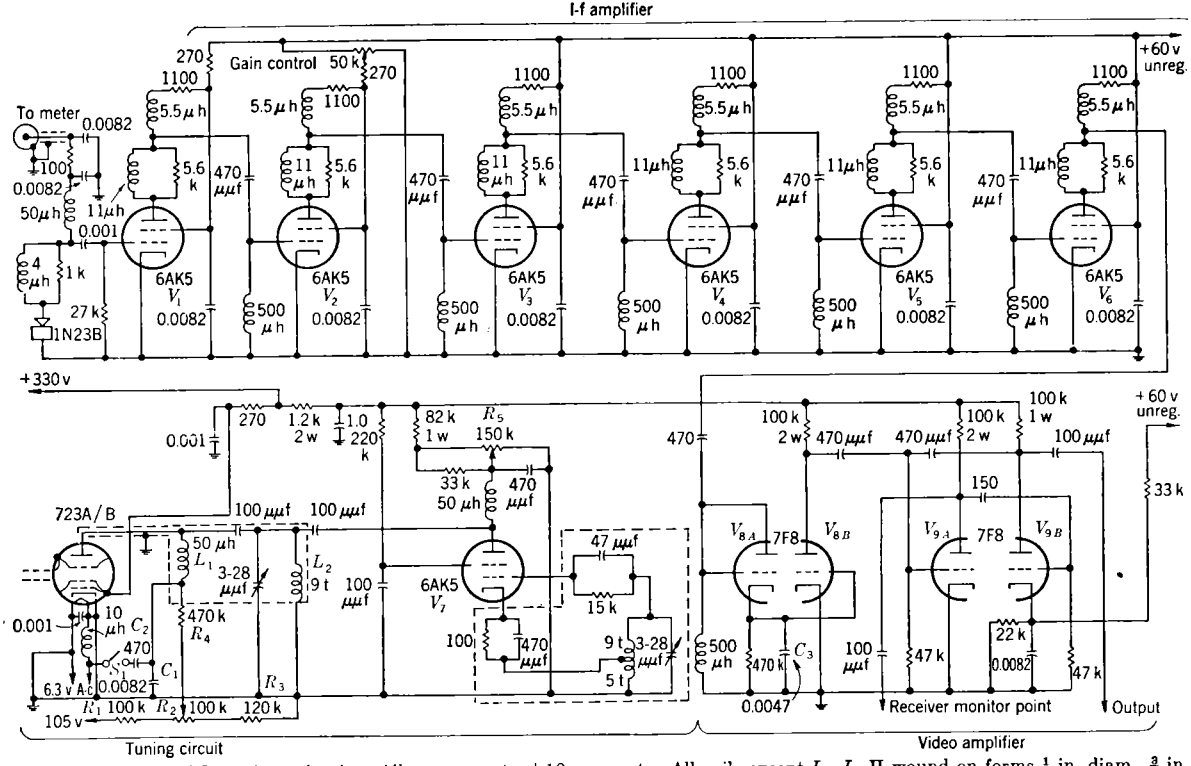


FIG. 18·31.—F-m LO receiver circuit. All components ±10 per cent. All coils except L_1, L_2 Π-wound on forms $\frac{1}{8}$ in. diam., $\frac{3}{8}$ in. long. L_1 and L_2 No. 28E, close wound on $\frac{5}{16}$ in. polystyrene form.

indication is obtained, it should be maximized for both of these adjustments; that is, mixer-crystal current should be maximized by means of the reflector voltage, and reference-cavity-crystal current by means of mechanical tuning. The final step is to increase the value of the modulating voltage to the point where the mixer-crystal current drops to 0.7 times its value in the absence of modulation.

A tuning check may be made, after frequency modulation has been applied, by connecting the meter to read reference-crystal current; if the mechanical tuning is now varied, three maxima will be produced, corresponding to the fundamental and first side frequencies. The middle reading is maximized, thus setting the fundamental to exactly 9375 Mc/sec.

18·20. Tuning Circuit.—The tuning circuit must have three functions. It must (1) provide an adjustable d-c voltage of the right magnitude, (2) provide for superposition of a 40-Mc/sec sinusoidal voltage on the d-c voltage, and (3) include a means of accounting for the gaps in the receiver-coverage band (see Sec. 18·1).

The portion of Fig. 18·31 labeled "tuning circuit" fulfills these requirements. From available data on the operation of the 723A/B, it is known that a range of -40 to -70 volts is needed in order that all tubes may be accepted. This voltage is supplied from a bleeder R_1, R_2, and R_3, with potentiometer R_2 providing the necessary range of variation. This voltage is connected to the reflector of the local oscillator through the isolating networks R_4, L_1, and C_1. The 40-Mc/sec modulating voltage is obtained from an electron-coupled 6AK5 oscillator. Its amplitude is controlled by the potentiometer R_5, which varies the plate voltage of the pentode. This arrangement (with the particular coils used) results in a possible variation of 0 to 10 volts in amplitude. The 40-Mc/sec signal is connected directly to the reflector and is isolated from the d-c voltage source by the 50-μh choke L_1. In order to fill in the gaps in the coverage band, which are inherent in this type of receiver, a small power-frequency voltage is applied to the reflector (in addition to the other voltages). This voltage is obtained from a capacitance divider (C_1 and C_2) across the heater supply. The result is to shift the whole coverage band at the power-frequency rate by an amount of 1.5 to 3 Mc/sec, so that signals received at frequencies in the vicinity of the gaps will be amplified at least a portion of the time at full sensitivity.

Switch S_1 is provided to eliminate the power-frequency signal during the tuning procedure and is ganged to R_5. With R_5 in the counterclockwise position, therefore, both the 40-Mc/sec and the power-frequency modulation are eliminated.

18·21. I-f Amplifier.—The determination of the optimum i-f amplifier band limits poses a different problem from that met in the two receivers

described previously. In this case the upper limit is set by the total band coverage and by the f-m rate. In other words, for a coverage of 120 Mc/sec and a modulation rate of 40 Mc/sec, the i-f amplifier upper limit must be at least 20 Mc/sec. The choice of the lower limit must be determined by the characteristics of the i-f amplifier and the video amplifier. Evidently, since the gap in response (see Fig. 18·4) is equal to twice the lower-frequency limit, it is desirable to set this limit as low as possible in order to minimize the problem of providing response in the gaps. Since, however, the amplifier cutoff cannot be absolutely sharp, there is danger that the i-f pass band will overlap into the video-amplifier pass band. The video amplifier will be shown to require a bandwidth of about 300 kc/sec, so that it is not found practical to make the lower limit[1] of the i-f amplifier less than 1.0 Mc/sec.

The i-f amplifier bandpass limits have actually been set at 1.0 and 21.0 Mc/sec, the extra 1.0 Mc/sec in the upper limit acting as a safety factor to account for slight errors in LO tuning.

Because of the comparatively small i-f bandwidth required, the problem of minimizing the variation in gain over the band is much less severe than in the switched-LO receiver. The limitation on i-f gain arises from stability considerations. The i-f amplifier shown in Fig. 19·31 provides a gain of about 90 db, and variation in gain over the band may be kept to less than 5 db with no other precautions taken than to use components of 10 per cent tolerance. The stages are all similar, using the shunt-series-peaking type of coupling circuit with grid filters for low-frequency rejection.

The operating conditions of the 6AK5 i-f amplifier tubes warrant some discussion. It is important in this case not only to obtain the correct gain and bandwidth but also to accomplish this with the minimum of power drain. If the power drain of the plates and screens were the only thing that needed minimizing and the number of stages could be as large as desired, a figure of merit defined as the ratio of the logarithm of the transconductance to the cathode current would be a true measure of such operation. It is found for 6AK5's, as well as for many other pentodes, that this ratio increases as the cathode current and screen voltage are reduced, the maximum occurring for a cathode current much smaller than that obtained under normal conditions. Since, however, the g_m also decreases, a compromise must be made between the saving in power and the increase in the number of stages that would be required for a given over-all gain. The operating conditions here adopted are as follows: screen voltage, +60 volts; grid voltage, 0 volts; g_m (design center),

[1] It must be remembered that this limit is defined by convention as the frequency at which the i-f amplifier gain (and receiver sensitivity) is equal to that measured at the point of lowest gain within the pass band.

5000 μmhos; total cathode current (plate plus screen), about 8 ma. Such operation is considerably removed from the conditions specified for test of the tubes, and greater variability of gain from tube to tube must be expected. This variability has not been found to be serious with six i-f amplifier stages. The saving in power is obvious, since under normal conditions the 6AK5 is rated at 5000-μmho mutual conductance with a screen and plate voltage of 120 volts and a cathode current of 10 ma.

Gain control[1] of the i-f amplifier is accomplished by varying the screen voltage of the second stage (V_2) to provide a range of control of about 25 db. The input circuit is of the series-peaking type and results in a bandwidth of about 25 Mc/sec if the mixer is considered to have a resistance of 300 ohms and a capacitance of 15 $\mu\mu$f.

18·22. Second Detector, Video Amplifier, Output Stage.—The video circuits, including the diode second detector (see Fig. 18·31), have a bandwidth of about 300 kc/sec. This value was selected in view of the lax requirement on fidelity and, after careful consideration of the optimum bandwidth, for good signal-to-noise ratio. The actual waveform of a 2-μsec pulse signal at the output of the video stage is very poor by conventional standards. However, except for the problem of delay, this waveform has no deleterious effect on the operation of the circuits following the receiver, since the output stage is such as to produce a signal independent of the waveform of the video signal.



FIG. 18·32.—Waveform at second detector showing effect of C_3 on strong pulses.

The second detector uses the first half of a 7F8 double triode V_{8a} connected as a diode, and its output is directly coupled to the grid of V_{8b}, which is a triode video amplifier. The bandwidth at the second detector is purposely limited by the effect of the condenser C_3, this being necessary for reasons of stability and to prevent overlapping of the video and i-f amplifier pass band. Another important function of C_3 is to resolve waveforms, such as are shown in Fig. 18·5*a* and *b*, into usable pulses. This effect is shown in Fig. 18·32. The only disadvantage of this is that the pulse will be elongated, depending upon the signal strength, to as much as 20 μsec, resulting in a problem in recovery of the receiver.

The output of the video amplifier is used to trigger a biased multivibrator $V_{9a\text{-}b}$ which produces a gate of about 30-μsec duration. The waveforms obtained here are shown in Fig. 18·33. A convenient way of regarding this portion of the circuit is to consider V_{9a} as an additional video stage that produces a positive pulse at the grid of V_{8b}. The cathode of V_{9b} is connected to a bleeder across the +60-volt i-f supply

[1] The considerations that normally make this method of gain control inadvisable do not operate here.

and ground in such a way as to provide a bias between the grid and cathode of about 20 volts, which is sufficient to make the tube normally nonconductive. A positive pulse of about 15 volts amplitude on the grid will initiate conduction, and $V_{9a\text{-}b}$ will then act as a conventional biased multivibrator. The actual signal at the grid of V_{9a} can be observed on a synchroscope at the receiver monitor point. This waveform is shown in Fig. 18·33*a*, under the condition of an incoming signal

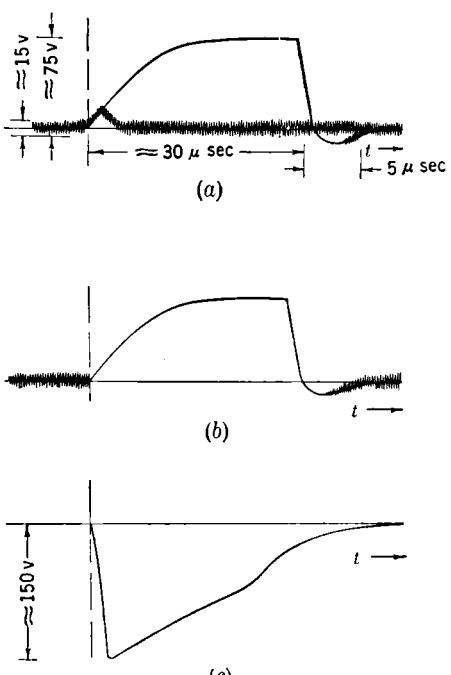


FIG. 18·33.—Output waveforms. (*a*) Pulse initiating gate intermittently, waveform at monitor point; (*b*) gate resulting from strong pulse, waveform at monitor point; (*c*) waveform at receiver output for reception of strong signal.

just large enough to cause the multivibrator to operate intermittently. Figure 18·33*b* shows the gate resulting from the reception of any signal larger than that of Fig. 18·33*a*, and the corresponding waveform at the plate of V_{9b} is given in Fig. 18·33*c*. This last signal is used to trigger succeeding circuits, since it has a good rise time and is produced at comparatively low impedance (the plate resistance of the triode).

18·23. General Information.—Two points that have been mentioned in connection with the video amplifier are of considerable importance and should be considered in some detail. These are (1) the delay prob-

lem and (2) the recovery time. The question of absolute delay (that is, the time elapsed between the start of the r-f pulse and the start of the output gate) may or may not be significant, depending upon the ultimate application of the receiver. In this as in the other receivers, the signal waveform is not influenced significantly by the shape of the i-f pass band but is controlled by the video-amplifier characteristics. Because of the small video bandwidth, the signal will have the crude triangular shape shown in Fig. 18·33*a*, so that it is evident that the gate will be initiated a maximum of 2 μsec (the pulse width) later than the start of the r-f pulse. For very strong signals, however, this delay may be as low as 0.1 or 0.2 μsec. An inherent variation in delay of about 2 μsec is therefore present. However, this variation is not linearly related to the signal strength, the greater part of the change taking place over a range of 10 db of signal strength (from minimum signal to 10 db above minimum.) If, therefore, the receiver is to be operated primarily on signals 10 db or more above the minimum acceptable signal, the delay problem becomes very much less important.

The length of the multivibrator gate plus the length of time for the multivibrator itself to recover determines the minimum time that the receiver will be made inoperative after the reception of a signal. This differs from the condition existing in the first two receivers described, in which the recovery of the video amplifier was the limiting factor. Although the multivibrator or output scheme would therefore seem undesirable, this is not strictly true. The recovery of the receiver has already been shown to be limited by the elongated waveform produced at the second detector (Fig. 18·32). The effect of the multivibrator is to blank out completely anything that occurs after the initiation of the gate, thus eliminating any possibility of mistaking the tail of the elongated pulse for another signal with corresponding erroneous results. However, a considerable time for recovery must be allowed. In the application of this particular receiver, the recovery time of about 35 μsec was not considered objectionable.

CHAPTER 19

CRYSTAL-VIDEO RECEIVERS

By H. J. Lipkin

INTRODUCTION

19·1. Uses of a Crystal-video Receiver. *General Description.*—A crystal-video receiver is one on which the incoming r-f signal is detected immediately at the input from the antenna and the resulting video signal is amplified by a high-gain video amplifier. Because the crystal is superior to any other form of detector, it is almost always used for this purpose, thus giving this type of receiver its name.

The outstanding advantage of the receiver is the absence of r-f and i-f circuits, resulting in simplicity, small size, low cost, and broad r-f bandwidth. There are no r-f amplifiers requiring tuning, no local oscillator to be tuned and kept stable, and no i-f amplifier. At video frequencies, the desired gain can be obtained more economically in regard to power consumption and size of components than at radio frequencies or intermediate frequencies. The absence of the narrow-band i-f circuits necessary to obtain appreciable gain per stage at intermediate frequencies makes the receiver inherently a broad-band device, limited only by the selectivity of the input circuit.

The price paid for this simplicity and broad band is a large loss in sensitivity (about 40 to 50 db) compared with that obtainable with a good superheterodyne receiver. There are, however, applications where this low sensitivity is acceptable and where the small size or broad band is required. An additional disadvantage is the poor pulse shape obtained at the output of the receiver, because of certain unavoidable properties of such high-gain video amplifiers. Although high-gain video amplifiers have been developed that have moderately good pulse shape, the comparison of the signals obtained from these with the output of a good superheterodyne still leaves much to be desired. Furthermore, if good selectivity is required, cumbersome r-f filters must be inserted in the antenna line.

Thus the crystal-video receiver finds application as a complete receiver where a low sensitivity is adequate, where the requirements upon pulse shape are not too stringent (of particular interest are those cases where the output of the receiver is used merely as a trigger for a succeeding

unit and the shape of the output signal is of no consequence), and where little selectivity is required. It also finds application where these disadvantages are outweighed by its small size, low cost, simplicity, or broad band.

A crystal-video receiver may follow an r-f amplifier or even an i-f amplifier. The former case is the TRF receiver, usable at vhf, where satisfactory r-f amplifiers are available. The second case is the superheterodyne receiver with low i-f gain and high video gain. The main advantage of this type of receiver is the possible saving of size and power as a result of using video gain rather than i-f gain. Such a receiver would be economical in applications requiring a high ratio of i-f bandwidth to video bandwidth, as a much higher gain per stage would then be obtainable at video than at intermediate frequency. It should be noted, however, that the extra gain required to compensate for the inefficiency of operating a second detector at low levels makes it impractical to use such a receiver unless a considerable saving in gain is effected by amplifying at a narrow video bandwidth rather than at a wide i-f bandwidth. For common receivers, where both the i-f and video bandwidths are determined by the pulse response, the factor of 2 saved in using video (6 db per stage) is insufficient to justify the use of this type of receiver.

19·2. Performance of Crystal-video Receivers.—The performance of a crystal-video receiver can be specified by its sensitivity, r-f bandwidth (or frequency coverage), and video response. These quantities are somewhat modified in their meaning because of the effects of the square-law characteristic of the crystal detector.

Square-law Detector.—Because of the absence of an LO signal, the crystal detector operates at low levels where it is a square-law device; that is, the current (or voltage) of the output video signal is proportional to the r-f *power* fed into the crystal. For this reason, conventional methods for measuring and specifying sensitivity cannot be used without precaution. The chief differences between square-law and a linear receiver are as follows:

1. Because of the nonlinearity of the detector, there is no quantity analogous to conversion loss in a mixer that can be specified. In other words the loss is a function of the signal level.
2. If two signal levels are compared, the difference between them, expressed in decibels, will be twice as great if measured at video as if measured at radio frequency. Thus the difference between the tangential signal and a signal equal to noise is 8 db in the video but only 4 db in the radio frequency. Since it is the minimum usable r-f signal that determines the sensitivity of a receiver, any reference will be to the r-f level.

3. Because of Item 2, any change in the video circuit that improves the signal-to-noise ratio by a given number of decibels in the video will improve the sensitivity of the receiver by only half as many decibels. This renders invalid the property of linear receivers that a change in pulse length will necessitate a change in receiver bandwidth of the same ratio and will, therefore, change the receiver sensitivity by the same ratio, so that the energy of the minimum detectable pulse is a constant. In a crystal-video receiver, a change in bandwidth by a given factor will change the noise level, but this will appear as only half the change in sensitivity that would occur in a linear device. If the pulse width is decreased by a factor of 4, therefore, the receiver video bandwidth will be increased by a factor of 4, thus increasing the video noise level by 6 db. Because of the square-law action of the detector, the r-f signal need be increased only 3 db to account for the 6-db change in video noise level. Therefore, the energy of the minimum detectable pulse has been decreased by a factor of 2, or by the square root of the change in pulse width. Thus, the energy per pulse of the minimum detectable signal is not a constant but varies inversely with the square root of the pulse width.

For these reasons the term noise figure, frequently used in linear receivers, is meaningless when applied to crystal-video receivers, since it is a function of signal level and therefore of pulse width (see Sec. 1·1).

Sensitivity.—The sensitivity of a crystal-video receiver is almost entirely a function of the crystal, because at video frequencies the noise produced in the amplifier is generally negligible in comparison with the Johnson noise of the crystal. This can be assured by using crystals of sufficiently high video impedance.

The sensitivity of crystal-video receivers is generally specified in terms of the tangential signal. With crystals available at the present time, sensitivities of approximately 10^{-8} watt are obtainable at microwave frequencies (3000 to 10,000 Mc/sec) and sensitivities of about 10^{-9} watt at vhf (less than 1000 Mc/sec). The sensitivity remains constant at roughly 10^{-9} watt below 1000 Mc/sec and falls off with frequency above this point. The apparent constancy between 3000 and 10,000 Mc/sec is the result of extra effort put into the development of better video crystals for the 10,000-Mc/sec band.

R-f Bandwidth.—At microwave frequencies, where the r-f circuit is in the form of a transmission line or a cavity, the minimum frequency sensitivity is set by the inherent frequency sensitivity of the crystal. This can be represented in the 3000-Mc/sec band as a simple single-tuned circuit having a Q of 5 to 10 and a bandwidth, therefore, of 300 to 600

Mc/sec. At 10,000 Mc/sec the Q is considerably higher, being about 100, which corresponds to a bandwidth of 100 Mc/sec. At lower frequencies, where lumped-constant circuits must be used, the bandwidth is limited by the shunt capacity in the same manner as the i-f bandwidth in normal i-f amplifier stages. The r-f bandwidths obtainable are, therefore, of the same order of magnitude as the i-f bandwidths of common input circuits, and the usual types of circuits, single-tuned, double-tuned, etc., can be used.

For most applications, the use of the 3-db point to specify bandwidth is quite arbitrary, since these applications require a broad r-f bandwidth for covering a wide band of radio frequency but do not require the complete bandwidth for any one signal. Therefore, all that is important is that the sensitivity of the receiver be adequate throughout the band, and variations within the band are of no particular importance, except in so far as excessive sensitivity is detrimental. If the maximum sensitivity is 10 db above that required for the application, the usable bandwidth extends to the points that are 10 db down. However, 3-db bandwidth will always be specified, unless specifically noted otherwise, because of conventional usage.

Video Properties.—The video amplifier used for crystal-video receivers must of necessity produce considerably more pulse distortion than is found in superheterodyne receivers because of the high gain and the enormous dynamic range required. Because the receivers are often required to operate with a dynamic range of 60 db (from 10^{-8} to 10^{-2} watt), the amplifier, because of the square-law characteristic of the detector, must have a dynamic range of 120 db. This, together with the high-gain requirement (100 to 120 db), requires that there be pulse distortion in the amplifier. This distortion will show up in the form of one or more overshoots, as shown in Fig. 19·1. These can be reduced in time duration to the same order of magnitude as the pulse length and by suitable designs can be cut down to only a single overshoot with no secondary overshoot in the same direction as the original signal. The first overshoot, however, is always present and can never be completely eliminated.



FIG. 19·1.—Waveform of pulse obtained from several successive coupling-circuit stages.

In addition, the square-law characteristic of the detector affects the over-all response of the amplifier and the output video signal, making it impossible to use a crystal-video receiver for applications requiring linearity. Furthermore, the square-law characteristics reduce by a factor of 2 the range in decibels between the minimum detectable signal and the signal that will be just limited.

CRYSTAL AND CRYSTAL HOLDER

19·3. Video Crystals.—Inasmuch as the sensitivity of a video receiver is determined almost entirely by the characteristics of the detector, this detector must be one that will have the best possible signal-to-noise ratio. At video frequencies, the crystal behaves like an ordinary resistance in generating Johnson noise[1] and therefore is better than any tube that has excess noise. Furthermore, the rectification efficiency of the crystal is superior to that of most tubes.

The crystals used for low-level detectors are similar to those used for mixers as described in Sec. 2·3 and illustrated in Fig. 2·4*a* and *b*. The coaxial unit, shown in Fig. 2·4*b*, is used in the 10,000-Mc/sec band because the cartridge of Fig. 2·4*a* is too sensitive to temperature changes. For successful detector operation at 10,000 Mc/sec the contact area between the crystal and catwhisker must be small. To obtain this the mechanical force on the catwhisker must be small so that the contact is abnormally vulnerable to mechanical displacements of any sort, such as those produced by differential expansion between the metal and the ceramic. Such differential effects are reduced in the coaxial construction. The ceramic cartridge is satisfactory for use at 3000 Mc/sec and below.

Sensitivity.—The sensitivity of mixer crystals is specified in terms of conversion loss and noise temperature. Low-level detectors, because of their nonlinearity, cannot be specified in this manner. Instead it has been found convenient to define a quantity called the "figure of merit" which combines the rectification properties and the noise generation of the crystal in such a way as to produce a number that is proportional to the signal-to-noise ratio.

The rectification properties of a video crystal may be described in terms of the current sensitivity of the crystal as a detector. This quantity is the constant of proportionality between the r-f power into the crystal P and the direct current i which would be delivered to a zero-resistance load. It is expressed in microwatts per microampere and is the constant b in the relation

$$i = \frac{P}{b} \tag{1}$$

which describes the rectification process.

The crystal may be regarded as a current source shunted by its internal video resistance R, or it may be regarded as a voltage source

[1] It may be shown by quite general arguments that any two-terminal network containing no sources of power will have appearing at its terminals Johnson noise given by the temperature of the network and the resistive component of the impedance measured between these terminals.

having a voltage given by

$$e = iR = R\frac{P}{b}. \tag{2}$$

The rms noise voltage appearing on the grid of the first stage will be given by

$$n = \sqrt{4kTB(R + R_a)}, \tag{3}$$

where k is Boltzmann's constant, T is the absolute temperature, B is the noise bandwidth, and R_a is the equivalent noise resistance of the input tube. The quotient of Eqs. (2) and (3) gives the signal-to-noise ratio:

$$\frac{e}{n} = \frac{P}{\sqrt{4kTB}\, b} \frac{R}{\sqrt{R + R_a}}. \tag{4}$$

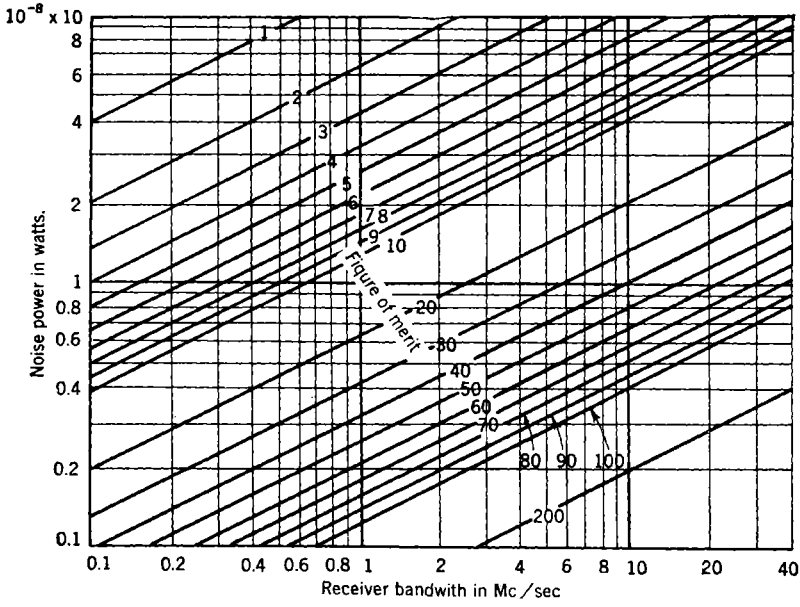


FIG. 19·2.—Noise power as a function of bandwidth and figure of merit.

It can be seen from Eq. (4) that the signal-to-noise ratio improves if R is increased. Hence, it is detrimental to the receiver sensitivity if the crystal is loaded with any parallel resistance. Therefore, the resistance R is taken as the crystal resistance. If a representative value is taken for R_a (1100 ohms is ordinarily used), the quantity $R/b(R + R_a)^{1/2}$ is a characteristic of the crystal and is called the figure of merit FM. Substitutions in Eq. (4) gives

$$\text{FM} = \sqrt{4kTB}\,\frac{\left(\frac{e}{n}\right)}{P}. \tag{5}$$

Thus, the figure of merit is the product of a constant $\sqrt{4kTB}$ and the ratio $\frac{(e/n)}{P}$. If P_1 is taken as the power required to produce a signal whose output power is equal to noise power, the relation becomes

$$P_1\text{FM} = \sqrt{4kTB}. \tag{5a}$$

Curves giving P_1 as a function of B for various values of the figure of merit are shown in Fig. 19·2. Other criteria for sensitivity correspond to different values for the ratio e/n, and the required power may be computed accordingly. For example, to find the power necessary to give a tangential signal, the values given by Fig. 19·2 should be increased by 4 db.

Video Resistance.—The video resistance R appears in the definition of the figure of merit, in a way indicating that higher sensitivities can be obtained with higher-impedance crystals. This is, in actuality, not entirely true, since the current sensitivity tends to be poorer for higher impedance crystals and therefore counteracts the effect of the higher impedance. In general, however, there is a trend toward better crystals at higher impedance.

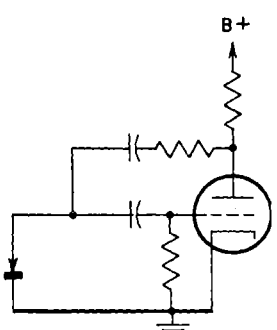


FIG. 19·3.—Inverse-feedback circuit for loading input circuit.

The video impedance of the crystal has an important effect upon the bandwidth of the video input circuit and, therefore, upon the entire amplifier. Shunting the crystal with a low resistance is detrimental to the signal-to-noise ratio and so is never done. The bandwidth of the input circuit is therefore determined entirely by the product of the video impedance of the crystal and the total shunt capacity. A high crystal impedance will result in a narrow bandwidth and a slow response to pulses. It is undesirable that the crystal have too high a video resistance, although sometimes loading by inverse feedback, as discussed in Chap. 4, will help. Figure 19·3 shows a typical circuit.

It is important that there is no direct current flowing through the crystal, since this produces a large amount of excess low-frequency noise. If direct coupling is used, the bias on the first stage must be sufficient to prevent any grid-current flow through the crystal. If this cannot be done, capacitive coupling must be used, as is shown in Fig. 19·4. (The

d-c return path for the crystal through the leakage resistance is satisfactory.) In this connection, particular notice should be taken of the feedback circuit of Fig. 19·3. This is the only method of connecting the feedback resistor that will not produce excess noise, as would be the case if direct coupling were used, or produce an overshoot because of the effect of feedback in lowering the input resistance of the feedback stage, as would be the case if the resistor were connected between the first two grids.

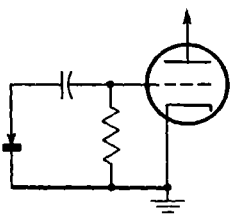


FIG. 19·4.—Capacitance-coupled input circuit.

Another effect of the video resistance of the crystal is to determine the absolute noise level, as can be seen from Eq. (3). Thus, higher-impedance crystals require less gain in the video amplifier. Furthermore, variations of the video impedance from crystal to crystal will cause variations in the noise output of the receiver. Note that this noise output is not connected with the figure of merit which merely specifies the r-f power necessary to produce a given signal-to-noise ratio. Thus, it is possible for two crystals to have the same figure of merit and, therefore, the same sensitivity but to differ widely in noise output.

High Level Properties.—As the power input to the crystal increases, the crystal impedance will eventually begin to decrease until it is down to around 50 ohms. Eventually the crystal will burn out, generally when the r-f power is of the order of magnitude of 1 watt. If input power levels can rise above this value, some sort of protection is usually required. The decrease in impedance with high signal level furnishes excellent self-protection for the crystal in TRF receivers. The current necessary to supply 1 watt into a 50-ohm load is about 140 ma. This is clearly in excess of the maximum current that any common r-f amplifier tube is capable of delivering. Hence, in a TRF receiver crystal burnout is no problem.

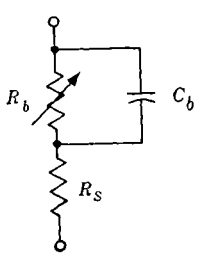


FIG. 19·5.—Equivalent circuit for crystal contact.

R-f Impedance.—The r-f impedance of the crystal contact can be represented by an equivalent circuit as shown in Fig. 19·5. The barrier resistance R_b is the nonlinear element responsible for the d-c volt-ampere characteristic of the crystal near zero bias. The barrier capacity C_b arises from the storage of charge in the rectifying barrier. The spreading resistance R_s results from the constriction of current flow lines in the semiconductor near the contact. At higher frequencies the capacitance C_b acts as a shunt across the nonlinear resistance R_b and reduces the rectification efficiency. Thus, the figure of merit for crystals is lower at micro-

wave frequencies than at vhf, giving rise to the decreased sensitivity discussed in Sec. 19·2. Because of the resistance R_s, it is impossible to tune out the capacitance C_b with an external transformer. To obtain maximum efficiency of rectification, however, the crystal must be driven by a source whose impedance is the complex conjugate of the crystal impedance, including the reactance introduced by the cartridge. To simplify the design problem for the crystal holder, the variation of r-f impedance from crystal to crystal should be reduced to a minimum.

At frequencies below 1000 Mc/sec the effect of the capacitance is negligible. Therefore, the rectification efficiency is independent of frequency in this range, and the crystal r-f impedance is a pure resistance having the same value as the video impedance.

Low-temperature Performance.—The nature of the rectification process in crystals leads to the theoretical prediction, which is verified experimentally, that its properties will vary radically with temperature. At low temperatures the video impedance R becomes very large, increasing much more rapidly than $1/T$, and the rectification constant b increases. This simultaneous variation tends to maintain a constant figure of merit, but the increase in resistance produces a corresponding increase in the output noise level, which can be serious in applications where the receiver is used unattended to drive a trigger circuit responding to all signals above a given threshold level. At low temperatures, the noise level may rise above this threshold and cause considerable extraneous triggering. This effect is an inherent property of the crystal. If the receiver is unattended, the only solution is to keep the temperature range within reason, if necessary by thermostating the crystal holder. In an attended receiver, the gain of the amplifier can be adjusted from time to time to keep the noise level constant.

Types of Crystals.—The common video crystals in use at present are the 1N27, 1N32, 1N30, and 1N31. Of these, the first two are designed for use in the 3000-Mc/sec band, and the last two for use in the 10,000-Mc/sec band. Any of these crystals can be used successfully at lower frequencies, since the r-f impedance match is not so critical as at microwave frequencies where the crystal impedance has an appreciable reactive component. Each of these four types was designed for a particular application demanding certain specifications which, however, give an idea of what can be done in the present state of the art.

The 1N27 crystal was designed for use in a receiver with stringent bandwidth requirements and therefore has an upper limit of 4000 ohms for the video impedance. The minimum figure of merit is 60.

The 1N32 crystal is a development of the type 1N27 for higher figure of merit. Because there was no bandwidth restriction in its application, the allowable impedance range is 5000 to 20,000 ohms. The minimum figure of merit is 100.

The 1N30 is a crystal in the ceramic cartridge for the 10,000-Mc/sec band. Because of the difficulties connected with the ceramic cartridge, this type of crystal is no longer being manufactured. The video resistance is between 7000 and 21,000 ohms, and the minimum figure of merit is 55.

The 1N31 is the successor to the 1N30 in the coaxial cartridge. It has a minimum figure of merit of 55 and a video resistance between 6000 and 23,000 ohms.

The r-f impedance of these crystals is held within reasonable limits by measuring the figure of merit and resistance in a standard crystal holder electrically similar to the one used in practice. It is desirable that the holder be fixed tuned, which is the case for most of these crystals at present.

19·4. Crystal Holders.—The crystal holder for a microwave crystal-video receiver is mechanically very similar to the mixers described in Sec. 2·4 and illustrated in Figs. 2·5 to 2·7. There is no LO power to be fed in; hence this extra complication is eliminated. The difference in r-f impedance between mixer and detector crystals resulting from the different working level requires some change in the r-f matching transformers.

R-f Properties.—A microwave crystal holder should be so designed as to match the input line to any crystal with a minimum of reflection loss and a minimum number of tuning adjustments. The degree to which this is possible is set by the spread of crystal impedances that must be matched. The crystal holder must, by a suitable combination of transformers and matching stubs, obtain the optimum result from a given distribution in crystal impedance.

The maximum bandwidth of the crystal holder is set by the crystal, although if a considerable transformation ratio is required for matching, the attainable bandwidth may be less than this. If the application requires that the bandwidth of the holder be narrower than the crystal bandwidth, some form of r-f filter may be used ahead of the crystal holder. In some special applications, where a loss of sensitivity is allowable, a relatively high Q cavity can be used as a crystal holder to save space and weight. Usually, the inherent bandwidth of the crystal is accepted as the r-f bandwidth of the receiver.

There are two types of tuning adjustments: one to allow for the spread of impedances from crystal to crystal and one to vary the frequency band over which the receiver is sensitive. The former type of adjustment is often undesirable, because the only method of adjustment is to maximize the receiver response to a given signal. In portable applications, where there may not be a signal source available, it is difficult to replace a crystal in the field and maintain proper sensitivity unless the holder is fixed tuned. Hence, development has tended toward fixed-tuned crystal

holders which will cause a small loss in sensitivity with some crystals. Tuning to change the frequency band is generally unobjectionable as long as the setting for a given band does not change when the crystal is changed.

Sign of Output.—An important property of a crystal holder is the sign of the video pulse obtained at the output, as this has great bearing on the design of the amplifier. The ceramic cartridge has been standardized

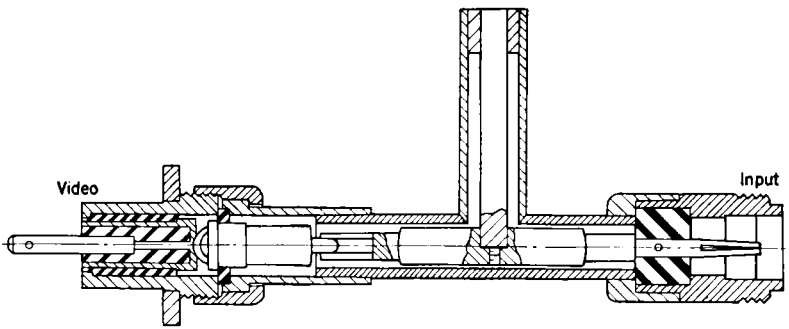


FIG. 19·6.—*S*-band crystal holder.

to give a positive video pulse at the small end of the cartridge or, conversely, a negative pulse at the large end. The coaxial type of cartridge gives a positive video pulse from the center conductor. Holders have been designed for both signs of output, but mechanical considerations have made some of these rather impractical.

Examples.—Figure 19·6 shows a fixed-tuned crystal holder for the 3000-Mc/sec band. It has a reflection loss of less than 1 db on 40 per

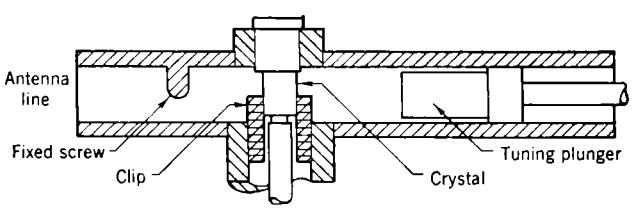


FIG. 19·7.—*X*-band crystal holder.

cent of the 1N27 crystals, less than 2 db on 70 per cent, and less than 3 db on 95 per cent. The output, taken from the large end of the cartridge, is negative. A variation of this type of holder to obtain positive output takes the output off the stub, with a suitable capacity coupling to prevent the video from being short-circuited by the plumbing.

Figure 19·7 shows a crystal holder for the 10,000-Mc/sec band, using the type 1N30 crystal with the ceramic cartridge. Because the 1N30 crystals vary widely in their r-f impedances, it is impossible to design a

satisfactory fixed-tuned holder; therefore a tuning adjustment is required. The video output from this type of holder is positive, since it is taken from the small end of the crystal. It is difficult to design a satisfactory holder for this band that will have negative output. If the crystal is placed across the waveguide, good contact is assured only when the large end of the crystal is clamped firmly in place and grounded, as in Fig. 19·7.

The holder for the coaxial crystal is similar to the mixer shown in Fig. 2·7. Here also a positive output is obtained. It is difficult to obtain negative output from the coaxial crystal, since this means taking the video output from the outer conductor.

For the 3000-Mc/sec band either sign of output is therefore possible, but for the 10,000-Mc/sec band only positive output can be readily obtained.

Crystal Holders for Vhf.—At frequencies below 1000 Mc/sec the reactive component of the crystal impedance becomes negligible, except for the stray capacity across the crystal as a whole. The problem of matching is, therefore, enormously simplified, since the r-f impedance of the crystal is equal to its video resistance.

At frequencies where lumped-constant circuits are used, the crystal holder is merely a mechanical device to hold the crystal, and the impedance matching can be achieved with a double-tuned circuit or a tapped coil. The circuit around the crystal is similar to that of an ordinary superheterodyne second detector, except that the impedance of the crystal is higher at low levels.

At frequencies where distributed constant circuits such as transmission lines are used, the crystal holder will hold the crystal across the line at the point of proper impedance. Matching is achieved by placing the crystal at the proper point in the line. Since the impedance of the crystal is purely resistive, the variation in r-f impedance can be cut down by using crystals with a small spread of video resistance. It is therefore possible to fix the crystal holder at a definite point on the line.

VIDEO AMPLIFIER

19·5. Special Problems of High-gain Video Amplifiers.—The high-gain video amplifier used in crystal-video receivers is quite different from the superheterodyne video amplifiers discussed in Chap. 8. In order to meet the stringent requirements on gain, dynamic range, and sometimes size, considerable sacrifice in performance must be made, particularly in regard to pulse response.[1]

To obtain maximum sensitivity, the amplifier should have sufficient

[1] See also Vol. 18, Chap. 10, Radiation Laboratory Series.

gain to bring pulses up from the noise level to an amplitude suitable for viewing on an indicator or for triggering an auxiliary circuit. Thermal noise level for a bandwidth of 1 Mc/sec is 4×10^{-15} watt. For an amplifier input impedance of 4000 ohms, this corresponds to a noise voltage input of 4 μv. To obtain an output of 4 volts, a gain of 120 db is required.

Signals as high as several volts, however, may also appear at the input. If no special care is taken to handle such large signals, they will drive successive stages of the amplifier to cutoff and into heavy grid cur-

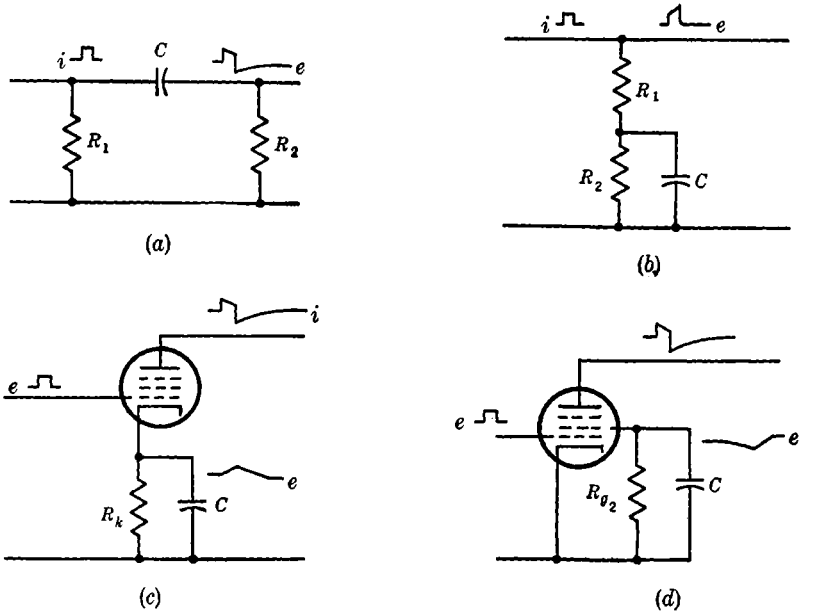


FIG. 19·8.—Overshoot waveforms. (*a*) Coupling circuit; (*b*) decoupling circuit; (*c*) cathode bypass; (*d*) screen bypass.

rent, blocking the amplifier for a considerable period after each strong pulse and rendering it completely useless for the amplification of weak signals. Some applications require that the insensitive period following signals be reduced below a specified minimum; others require further that the shape or the length of the strong pulse shall be preserved despite the limiting. In either case, the problem of overload becomes the essential feature of amplifier design, requiring an entirely different approach from that of Chap. 8.

Overshoots and Recovery.—In almost any video amplifier, many condensers are used for coupling and bypass purposes. If these condensers are not to cause distortion of pulses, they must behave as a perfect short

circuit for anything other than direct current. This means that the voltage across the condenser and therefore its charge must remain constant regardless of the presence of signals. In actuality, however, all signal voltages tend to charge or discharge each condenser during each pulse. After the end of the pulse, each condenser tends to restore itself to its normal charge, thereby producing transients that appear as signals following the pulse. These signals are referred to as (low-frequency) overshoots, because they arise from the poor low-frequency response of the amplifier, that is, the failure of the condensers to act as a perfect short circuit at low frequencies.

In a high-gain video amplifier, overshoots become a serious problem because of the large dynamic range. It is not sufficient that the overshoots be reduced to 10 or even 1 per cent of the signals producing them. One per cent of a very strong signal is still considerably stronger than the weak signals to which the amplifier should be sensitive. If the overshoot on a 1-volt signal is to be less than noise, which is 4 μv at the input, it must be less than 0.0004 per cent. It is for this reason that the problem of overshoots is given the extensive consideration that follows.

Figure 19·8 shows typical waveforms for coupling, decoupling, cathode-bypass, and screen-bypass circuits. In each case a charge is built up on the condenser during the pulse, causing the output pulse to have a slanting top. This charge remains upon the condenser at the end of the pulse and then leaks off exponentially, producing the overshoot. The calculated values of the magnitude of the overshoot for the various cases are given in Table 19·1.

TABLE 19·1.—FRACTIONAL OVERSHOOT AND RECOVERY TIME CONSTANT FOR COMMON CIRCUITS

Circuit	Fig. 19·8	Fractional overshoot ($\tau \ll T$)	Recovery time constant, T
Coupling	*a*	$\frac{\tau}{(R_1 + R_2)C} = \frac{\tau}{T}$ *	$(R_1 + R_2)C$
Decoupling	*b*	$\frac{\tau}{R_1 C} = \frac{R_2}{R_1}\frac{\tau}{T}$	$R_2 C$
Cathode-bypass	*c*	$\frac{g_m \tau}{C}$	$\frac{C}{g_m + \frac{1}{R_k}}$
Screen-bypass	*d*	$\frac{\tau}{T}$	$\frac{R_{g2} r_{g2} C}{R_{g2} + r_{g2}}$

* τ is the pulse length.

The effect of several circuits of this type in cascade can be understood by considering the effect of one of these circuits on a signal that already has an overshoot. During the pulse, the condenser will charge, as in

the simple case. At the end of the pulse, the overshoot produced by the charge on the condenser adds to the overshoot on the input signal, producing an overshoot whose initial amplitude is the sum of the two. The condenser then discharges and charges up in the opposite direction, since the input overshoot is now equivalent to a signal. A secondary overshoot is therefore produced, which is the "overshoot of the overshoot." Similarly if a signal passes through several circuits, there will be several overshoots, one for each circuit, as shown in Fig. 19·1.

The magnitude of a multiple overshoot can be determined qualitatively by considering it as the sum of the individual overshoots. The initial value of the overshoot is clearly the sum of the initial values of all the overshoots acting independently. However, since an overshoot of long time constant has the major part of its energy at low frequencies, it will be attenuated by any circuit having a short time constant. Thus the amplitude of any component overshoot is reduced in a circuit having a shorter time constant by a factor that is approximately the ratio of the two time constants. The multiple overshoot, therefore, can be considered as the sum of individual component overshoots, each having a magnitude given by Table 19·1 and modified by suitable factors to account for the reduction in amplitude by all circuits of shorter time constant.

Table 19·1 shows that the magnitude of the overshoot produced by a given circuit is inversely proportional to the time constant of the circuit. It is, therefore, theoretically possible to make all overshoots negligible simply by making all the time constants sufficiently large. There is, however, a practical limit to the maximum usable values of resistance and capacitance, and it is generally impossible to design a high-gain amplifier with no overshoots at all. Furthermore, even if it were possible, some form of low-frequency rejection filter to eliminate microphonics would still be needed. This filter would necessarily introduce an overshoot, since it would effectively introduce at least one short time constant.

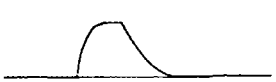


FIG. 19·9.—Effect of shunt capacity on waveforms of a pulse.

Effect of Shunt Capacity.—In the previous consideration of overshoots, the shunt capacity always present in coupling circuits has been neglected. In common *RC*-coupled circuits, the effect of shunt capacity is to slow up the leading and trailing edges of the pulse, because this capacity must be charged before any voltage change takes place. The normal effect of shunt capacity is that shown in Fig. 19·9.

With some circuits another effect of shunt capacity may be present which will cause apparently unexplainable overshoots. This can occur in all circuits where the full voltage across the coupling condenser does

not appear as a signal on the grid after the pulse but divides among several branches.

Consider the typical coupling circuit of Fig. 19·10, with shunt capacitances C_1 and C_2. At the end of the pulse there will be a brief flow of current charging up the capacitances C_1 and C_2. This will divide the voltage on the coupling condenser between the grid and plate circuits inversely as the associated shunt capacitances. Thus the magnitude of the overshoot depends upon the ratio of the capacitances rather than on the ratio of the resistances. After a short period of time, however, the charges on these shunt capacities change as current begins to flow through the resistors, and the conditions described by Table 19·1 take effect. The time required for this to take place depends upon the time constants of the shunt capacitance and the load resistors.

The initial magnitude of the overshoot as caused by this shunt capacitance can differ considerably from the values obtained from the previous analysis. For the case where R_1 is very much greater than R_2 or for any other circuits that depend for their effectiveness upon having only a small portion of the condenser voltage appearing on the grid, the actual initial overshoot may be considerably greater than would be expected from Table 19·1. This effect will be of comparatively short duration, as the ratio of its time constant to that of the expected overshoot is equal to the ratio of shunt capacitance to the coupling condenser. Thus, it will appear as a sharp spike at the beginning of the overshoot as shown in the waveforms in Fig. 19·10.

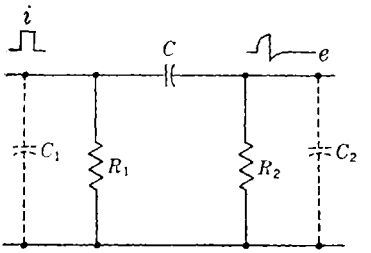


FIG. 19·10.—Transient effect of several shunt capacitances.

This effect may be troublesome in cases where the circuit has been designed to have a negligible overshoot according to previous considerations. The spike may be sufficiently larger than the overshoot to come through the amplifier as an appreciable signal. The effect is difficult to track down experimentally because it depends on the stray capacities which are affected by every change. One obvious cure is to transfer the spike to the plate circuit by loading the offending grid circuit with a small condenser, thus changing the distribution of the coupling-condenser voltage between the plate and grid circuits.

Microphonics and Low-frequency Interference.—In any amplifier that has a high gain in the audio range, extraneous signals may appear as the result of mechanical shock or vibration or of pickup from power supplies or stray fields due to adjacent power equipment. Each indi-

vidual amplifier presents its own problem in this respect, depending upon the special performance requirements of the amplifier and the particular conditions under which the amplifier will be used. There are, however, certain general precautions in the way of shielding and cushioning that are applicable in all cases.

In spite of these precautions, it is almost certain that extraneous signals will be present with sufficient amplitude to necessitate their removal by some sort of filtering in the amplifier. This filtering must be inserted at some point before the signals reach sufficient amplitude to cause variation in the transconductance of the tubes, or else the low-frequency dis-

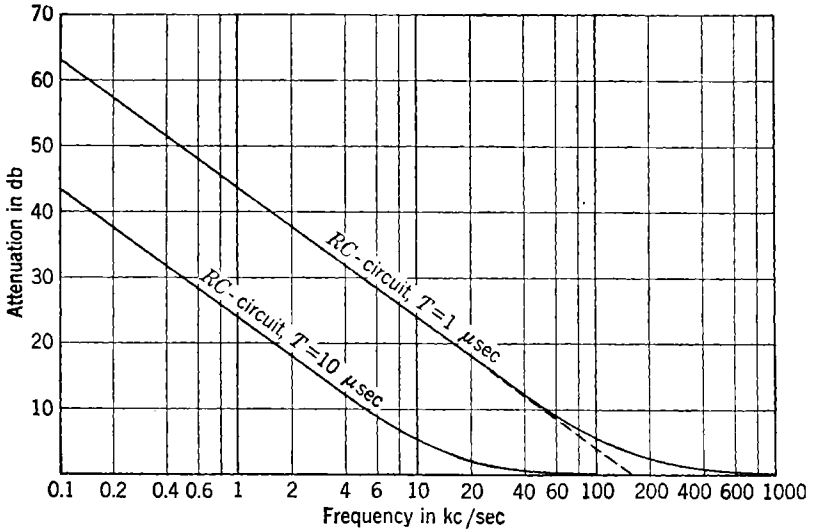


FIG. 19·11.—Attenuation introduced by a short time constant.

turbance will modulate the high-frequency noise that is always present and so be transmitted through the filter. The simplest form of such a filter is a short time constant. Figure 19·11 shows the attenuation vs. frequency for several values of the time constant. The shorter the time constant the better will be the low-frequency rejection. If, however, the time constant is too short relative to the pulse length, the pulse will drop sharply in amplitude during its length. This is generally undesirable. If several short time-constant circuits are present in an amplifier, the low-frequency rejection will be much better than with only one, but this will introduce more overshoots. In the design of an amplifier using *RC* coupling, the time constants are made as short as possible without making the pulse response unsatisfactory.

Pulse Stretching.—There are some applications for high-gain video amplifiers that require that the pulse length be preserved in passing

through the amplifier, regardless of signal strength. This presents a serious problem because of the large dynamic range of the amplifier. A signal that is several volts at the input will be repeatedly amplified and limited throughout the amplifier; thus, the signal appearing at the output will be only that portion of the input pulse which is below the saturation level of the amplifier, or about one-millionth of the original signal, amplified up to saturation level. The width of the output pulse will, therefore, be the width of the input pulse measured at a point 120 db down. Any slope in the trailing edge of the pulse due to shunt capacity, as shown in Fig. 19·9, will result in considerable stretching of a strong pulse after repeated amplification and limiting.

The amount of stretching produced is, of course, a property of the amplifier and is independent of the length of the input pulse. The amount of stretching or lengthening, measured in microseconds, is usually more or less proportional to the strength of the input signal in decibels. Therefore the performance of an amplifier in this regard is often specified by a "stretching coefficient" expressed in decibels per microsecond. Thus if an amplifier had a coefficient of 100 db/μsec, a pulse 1 μsec long and 100 db above the minimum input level would appear in the output as a 2-μsec pulse. A similar 5-μsec pulse would appear in the output with an apparent length of 6 μsec.

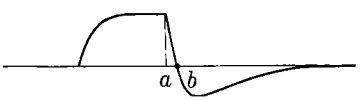


FIG. 19·12.—Limitation of pulse stretching by the introduction of an overshoot.

One obvious method of reducing pulse stretching is by shortening the time constants of the circuits causing it. This may involve a considerable sacrifice in gain per stage, since the minimum shunt capacity is determined by the interelectrode tube capacitances and by wiring capacitance. The only method of reducing the time constant is, therefore, the reduction of the load resistors with the consequent loss of gain. Another method of improving the pulse stretching is by the introduction of an overshoot early in the amplifier before appreciable limiting takes place. This has the effect of making the signal cross the baseline at a definite point, as shown in Fig. 19·12, instead of approaching the baseline asymptotically, as in the previous case. Therefore the maximum amount of pulse stretching that can occur in the amplifier is the distance *ab* in Fig. 19·12, regardless of the amount of amplification and limiting following this point.

19·6. Circuits Used in High-gain Video Amplifiers.—The choice of a circuit to be used for a given video receiver depends upon the desired recovery and overshoot performance more than upon anything else. Each type of circuit used is characterized by its overshoot performance and, in particular, by the manner in which the grid-current problem,

discussed in greater detail below, is handled. (Another important consideration, in the case of small amplifiers, is the weight and power consideration. Small amplifiers are discussed in detail in Sec. 19·7.)

Although it is impossible to eliminate overshoots completely, it is advantageous wherever possible to design circuits so that the overshoots produced by them will be negligible at the output of the receiver. This means that the overshoot following a very strong signal must be less than the magnitude of a very weak signal or that the fractional overshoot must be less than the ratio of the weakest signal to the strongest, which is the reciprocal of the dynamic range of the amplifier.

Because of limiting, the dynamic range of signals actually present will vary from point to point in the amplifier. At each point of limiting, all signals above a certain amplitude are clipped. The dynamic range beyond this point is thus reduced, and therefore the allowable fractional overshoot is increased. The first few circuits have the most severe requirements on the time constants in order to eliminate overshoots. The nearer a given circuit is to the output end the shorter its time constant can be without introducing appreciable overshoot.

If the dynamic range of an amplifier is one million, the fractional overshoot produced by the circuit preceding the first point of limiting must be less than one-millionth if the overshoot is to be negligible at the output end. This can generally be achieved only if there is a short time constant later on in the amplifier. If all time constants were long, the fractional overshoot produced by the first circuit would be approximately equal to the ratio of the pulse length to the time constant. To make this less than 10^{-6}, a time constant of the order of seconds would be required for a 1-μsec pulse. This is out of the question in most practical cases. The use of a short time constant later on in the amplifier will, as discussed in Sec. 19·5, reduce the magnitude of the overshoot in the first circuit by a factor approximately equal to the ratio of the time constants. Thus the fractional overshoot will be given by

$$a = \frac{\tau}{T_1}\frac{T_2}{T_1} = \frac{\tau T_2}{T_1^2}, \tag{6}$$

where τ is the pulse length, T_1 the first time constant, and T_2 the second time constant. This may be solved for T_1 to give

$$T_1 = \sqrt{\frac{\tau T_2}{a}}. \tag{7}$$

If it is assumed that the overshoot must be one-tenth of the weakest signal in order to be negligible, a is 10^{-7} for a dynamic range of 120 db. Then with typical values $\tau = 2$ μsec and $T_2 = 5$ μsec, T_1 is found to be equal to 10,000 μsec, which is not an unreasonable value for some circuits.

"Brute Force" Circuit.—An obvious approach to the design of a high-gain video amplifier is to make all time constants long except one, which is short for the rejection of microphonics. This will be possible if a time constant of 10,000 μsec can be achieved in all circuits except one.

For coupling circuits, the upper limit on grid resistors for most tubes is 1 megohm. If no grid current is drawn, a condenser of 0.01 μf will give a time constant of 10,000 μsec. If the value of the resistor is closer to 0.1 megohm, the condenser required is 0.1 μf. These are reasonable values for use in most amplifiers, except those where an extremely small package is required. However, the use of a postage-stamp-size 0.01-μf condenser still allows a fairly small amplifier to be built.

Decoupling circuits can generally be designed for negligible overshoot. The plate-load and decoupling resistors may usually be made as high as 10,000 ohms. This would require a capacity of 1 μf to get a time constant of 10,000 μsec. For a large-size receiver this is allowable, but for small receivers some other scheme must be used. Generally, there are several small time constants later in the receiver which alleviate the problem.

Screen-bypass circuits require components of the same order of magnitude as decoupling circuits and are therefore subject to the same limitations. For this reason, it is often advantageous, especially where small size is important, to use triodes and eliminate the screen-bypass circuit.

Table 19·1 shows that the resistance determining the time constant of the cathode circuit is the reciprocal of the transconductance of the tube. Enormous cathode-bypass condensers would be required to obtain a time constant of 10,000 μsec. A tube having $g_m = 3000$ μmhos would require a cathode-bypass condenser of 30 μf. For this reason, it is common practice to leave cathode resistors unbypassed. If the cathode resistor is small, the loss in gain due to degeneration is not appreciable. In some cases the cathode resistor is bypassed by a small condenser in order to improve the high-frequency response. In this case the time constant is so short that the overshoot is negligible and is obscured by the slow trailing edge of the pulse caused by poor high-frequency response.

All these considerations would indicate that it is quite feasible to build an amplifier in which all overshoots save one are negligible. However, an important difficulty is introduced when stages are driven into the grid-current region. If grid current is drawn, the grid resistor is shunted by the input conductance of the tube and it no longer affects the time constant, which is now dependent upon the plate-load resistor and the coupling condenser. Since the plate-load resistor must carry the plate current of the tube, values higher than about 50,000 ohms are not suitable, except in certain special applications in which triodes are

used. The plate resistance of the triode, therefore, effectively shunts the load resistor in these special cases, and the parallel combination is generally less than 50,000 ohms. The coupling condenser required for a time constant of 10,000 μsec is therefore at least 0.2 μf and is generally higher. Elimination of overshoots becomes impractical with simple *RC* coupling if grid current is drawn.[1]

In the "brute force" circuit, grid current is prevented by choosing the operating points of the stages so that the limited output of each stage having positive output will not drive the succeeding stage into grid current.

If the amplifier handles signals of only one polarity, the stages can be divided into two types, those which receive a negative signal and those which receive a positive one. The limited output of a negatively driven stage is the signal produced by cutting off the tube, which is equal to the voltage drop in the load resistor. To keep this low, a tube that draws a relatively low plate current for a given transconductance should be used and should be operated at a low bias. In order to obtain satisfactory amplification from the positively driven stage, while operating it at a high bias, the tube must have a reasonable value of transconductance at a large negative bias. This property is obtainable only in tubes of high power consumption such as the 6AG7. If the amplifier must handle signals of both polarities, high-current tubes must be used in each stage and must be operated with sufficient bias and low enough current to give a satisfactory limited output. Because increasing bias and decreasing plate current tend to decrease the transconductance, the gain obtainable per stage in an amplifier required to handle signals of both polarities will be considerably less than that of an amplifier handling only one polarity.

Inverse-feedback Pairs.—The inverse-feedback-pair circuit described in Chap. 6 can be used in a high-gain video amplifier to improve the overshoots. Because of two effects, which were not treated in Chap. 6 as they are of little interest to the designer of an i-f amplifier, considerably better overshoot performance can be obtained from a properly designed feedback pair. If the amplifier is designed so that negative signals appear at the input of the pair, the constants of the pair may be so adjusted that without excessive loss of gain, the positive signal appearing at the second

[1] Note, however, that the use later in the amplifier of several small time constants instead of only one introduces another factor in Eq. (6) reducing a. This allows the use of a smaller time constant than 10,000 μsec in the first circuit. Also, note that the grid resistor is shunted by a low resistance only during the pulse but not during the overshoot. The time constant of recovery for the overshoot therefore includes the grid resistor and can be made considerably larger than the time constant during the pulse. This will increase the factor by which the overshoot is reduced by following short-time-constant circuits.

grid will be limited to a value that does not cause grid current to flow. Also, feedback has the effect of reducing the overshoot of the interstage coupling circuit by a considerable factor from what it would be in the absence of feedback. A typical circuit is shown in Fig. 19·13.

It is shown in Chap. 6 that the relative values of the load resistors in a feedback pair can be adjusted without changing the over-all response of the pair, provided that the feedback resistor is properly varied. This has the effect of varying the relative gains of the two stages without changing the over-all response. The overloading of the second grid can be minimized by putting as much gain as possible in the second stage and as little as possible in the first. Thus, the maximum signal appearing at the second grid will be much less than in a circuit with the same over-all gain and no feedback, and the overshoot on maximum signal will not be so great. In some cases, it may even be possible to design the pair so that no grid current at all is drawn.

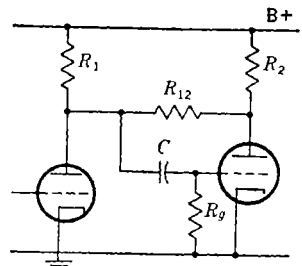


FIG. 19·13.—Inverse-feedback pair.

The theory of Chap. 6 cannot be applied to this problem, however, without reservations, because the large signals drive the tubes over a sufficiently wide range to make the assumptions of linearity invalid. Although a theory using a nonlinear characteristic or making certain approximations could probably be worked out, this type of amplifier has been designed by a cut-and-try procedure, using the published curves of tube characteristics.

The inverse-feedback-pair circuit allows the use of common amplifier tubes operating at comparatively low current. This constitutes a considerable saving in weight and power consumption. Note, however, that this circuit is subject to the spike type of overshoot due to shunt capacity, which is not true of the brute-force circuit, where all the voltage across the coupling condenser appears as an overshoot across the grid. The inverse-feedback pair is restricted to circuits where only one polarity of signal is used and a negative signal is applied to the first grid of the pair. For this type of amplifier, performance as good as that of the brute-force circuit can be obtained at a considerable saving of weight and power. For amplifiers required to handle both polarities of signals, the only approaches discussed here that can be used are direct coupling and the prevention of grid current by brute force.

Direct-coupled Inverse-feedback Pair.—An obvious method of elimination of the problem of charging of the coupling condenser by grid current is the elimination of the coupling condenser. Any conventional direct-coupled circuit can accomplish this, carrying along with it the usual

problems of a more complicated power supply. Use of an inverse-feedback pair in conjunction with direct coupling,[1] as shown in Fig. 19·14 allows the use of ordinary power supply.

In this circuit, the first tube of the pair is operated at a low plate voltage, at which the tube still performs satisfactorily as an amplifier. The large cathode resistor of the second tube furnishes sufficient cathode bias to bring the cathode potential above that of the grid by the desired amount. Because of the two types of d-c degeneration present in this circuit, it is unusually stable with respect to power supply and tube variations.

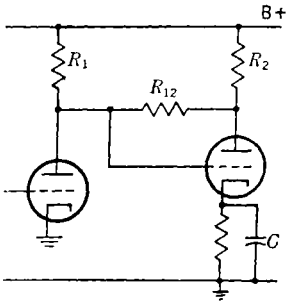


FIG. 19·14.—Direct-coupled inverse-feedback pair.

The unique feature is the effect of the feedback in reducing overshoot caused by the cathode bypass condenser, thus making it possible to bypass the large cathode resistor, which would otherwise cause considerable loss in gain. This circuit has the advantage of all others considered here, in that it has eliminated the problem of grid current. Note, however, that it is restricted to applications where only negative signals are impressed upon the first grid of the pair.

Double Short-time-constant Circuit for Single Overshoot.—Where there are two short time constants, there will, in general, be a secondary overshoot. By allowing grid current to be drawn in the stage following the second short time constant, the magnitude of the first overshoot may be increased to the point where the second overshoot is effectively eliminated. The waveform so produced is shown in Fig. 19·15. The secondary overshoot, although still recognizable, no longer crosses the baseline and so does not act like a second signal. This behavior may be understood in terms of the picture previously given in Sec. 19·5 of the formation of the secondary overshoot by noting that the time constant of the second circuit is short during the signal pulse because of grid current. Therefore, a large primary overshoot is produced. On the other hand, when the pulse ceases and the overshoot from the first circuit appears at the second circuit, the time constant of this latter circuit is long, and therefore the magnitude of the secondary overshoot is comparatively small.

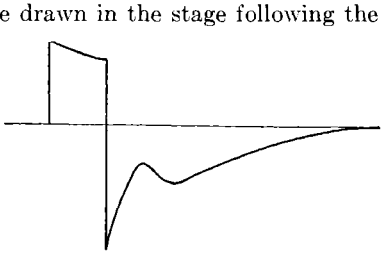


FIG. 19·15.—Elimination of secondary overshoot by drawing grid current.

[1] The unique advantages of the direct-coupled inverse-feedback pair were demonstrated by R. J. Grambsch at the Radiation Laboratory.

The recovery in this case will be poorer than in the absence of grid current; however, in some applications the elimination of the secondary overshoot is more desirable than the quick recovery.

The condition for no secondary overshoot may be shown to be $T_{2b} \geqq T_1 + T_{2a}$, where T_1 is the time constant of the first circuit, T_{2a} that of the second in the presence of grid current, and T_{2b} that of the second in the absence of grid current.

Multiple-overshoot RC.—For applications where multiple overshoots are allowable and the desired recovery time is moderately long in comparison with the pulse length, a multiple-overshoot amplifier is the simplest solution. Simple *RC*-coupled circuits can be used with moderately short time constants in the coupling circuits. Grid current is allowable *provided that grid resistors are comparable to or smaller than plate-load resistors*, thus making the time constants with and without grid current approximately equal. If this is not done, the overshoots produced will have a very long recovery. Bypass and decoupling circuits can use much shorter time constants than are allowable in a single-overshoot amplifier, since there are more short time constants in the amplifier to reduce the overshoots produced by these circuits.

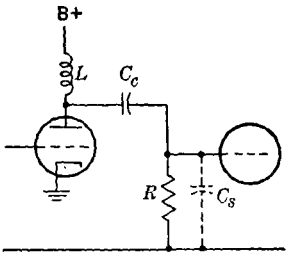


FIG. 19·16.—Choke-coupled amplifier stage.

Multiple-overshoot amplifiers are much less subject to interference from microphonics and extraneous low-frequency signals than are single-overshoot receivers because of the greater low-frequency rejection afforded by the larger number of short time constants.

Choke-coupled Circuit.—It has been shown that when grid current is drawn, the plate-load resistor must be made as high as possible to increase the time constant of charging of the coupling condenser. This produces a large voltage drop across the resistor and uses considerable power. A high power-supply voltage and a high-wattage resistor are necessary. To avoid these complications a choke may be used in place of the plate-load resistor. The choke acts as a high impedance during the charging of the coupling condenser but has a negligible d-c voltage drop. The circuit is shown in Fig. 19·16.

In this circuit, the inductance of the choke may resonate with the stray shunt capacity or with the coupling condenser, producing a train of damped oscillations following each pulse that are equivalent to a series of overshoots. To avoid this effect, both circuits must be at least critically damped by the grid resistor. The value of the grid resistor is determined by the amount of gain desired, and therefore the inductance

of the choke is determined by this condition of greater-than-critical damping. The conditions are

$$L \geqq 4C_sR^2 \tag{8a}$$

and

$$L \leqq \frac{C_cR^2}{4}. \tag{8b}$$

This is achieved generally by making the inductance as large as is practically possible (85 mh is a good value), thus satisfying condition (8*a*) and making the coupling condenser large enough to satisfy condition (8*b*).

If the inductance is sufficiently high, the fractional overshoot will be that of an *RC*-circuit having a resistance equal to the grid resistor regardless of the presence of grid current.

Output Circuits.—If the application for which a video receiver is used requires that the output signal be viewed on an indicator, standard output circuits, such as have been described in Chap. 8, can be used. There are many applications where the amplifier output is never viewed, however, but is merely used as a trigger for some other circuit. For these applications, it is often advantageous to make the amplifier output circuit a trigger circuit of some sort. The most common type of output trigger circuit is the multivibrator, one form of which is shown in Fig. 19·17. This circuit belongs to the class of *monostable circuits* as described in Vol. 19, Chap. 9, of this series.

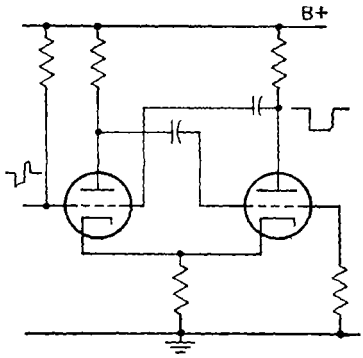


FIG. 19·17.—Multivibrator output circuit.

It is generally advantageous to trigger the multivibrator with a negative signal on the normally conducting tube. In this way the amplification of this tube is utilized. Furthermore, the multivibrator action can prevent the charging of the input coupling condenser by grid current. If there were no multivibrator action, the positive overshoot would drive the input grid into the positive region and charge the coupling condenser with the grid current, thus producing a long secondary overshoot. The multivibrator action, however, causes a large negative pulse having a duration determined by the constants of the circuit to appear on this grid. If this pulse is greater in amplitude than the positive overshoot on the input signal, the grid will remain negative and no grid current will be drawn. Care must be taken that the positive overshoot

is not greater than the negative multivibrator pulse; otherwise the overshoot will serve as a trigger to the multivibrator, causing the cycle to end too soon.

If a multivibrator output circuit is used with a multiple-overshoot amplifier, it is often advantageous to make the length of the multivibrator cycle slightly greater than the time required for all the overshoots following a strong signal to decay. There will then be only one output signal for each input signal to the amplifier, regardless of the number of overshoots produced in the amplifier. This will eliminate the possibility of overshoots appearing as spurious signals and providing false triggers to later circuits.

One variation of this circuit can be used in cases where the original pulse length is to be preserved, but all overshoots must be eliminated. In this case, the input signal to the multivibrator is a positive one, with a negative overshoot. The positive signal is amplified by the first tube producing a negative signal that has no effect on the second tube, which is already nonconducting. The negative overshoot, however, is in the proper direction to cause multivibrator action, and the resulting multivibrator pulse can be made sufficiently long to cover all the overshoots. If the output is taken from the plate of the first tube, the result will be a negative signal followed by a large positive overshoot, equal in length to the length of the multivibrator pulse. If this overshoot is objectionable, it can be clipped by a diode.

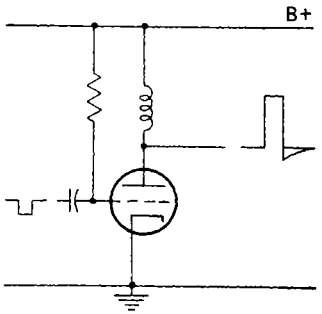


FIG. 19·18.—Output circuit for low-voltage amplifier.

Another special output circuit is shown in Fig. 19·18. It is used in amplifiers operating at a low voltage, where it is desirable to get a high-voltage output signal, usually considerably higher than the supply voltage. The output stage is a power-amplifier tube, operated to draw a high plate current through an inductive load. When a negative signal on the grid of this tube shuts off the current, there will be a high voltage generated across the inductance, which can be many times greater than the supply voltage. It is also possible to combine this circuit with the multivibrator by using an inductive plate load in the tube that is normally conducting.

19·7. Small Amplifiers.—One of the main advantages of the crystal-video receiver over the more sensitive and higher-fidelity superheterodyne is that of small size and lower power consumption. Therefore most applications of the video receiver are in cases where performance is

secondary to these considerations. The choice of tubes and circuits for small amplifiers is, therefore, of considerable importance in video-receiver design and is based upon criteria considerably different from usual amplifier design. Circuits are chosen not for their performance but for their ability to make use of small components and to require a minimum of power. Very often the circuits actually used are extremely inefficient according to normal criteria such as gain-bandwidth product or number of stages.

The various applications for small amplifiers can be grouped into several general classes according to the power available for the amplifier. There are applications where sufficient power is available to meet the needs of any circuit chosen, so that a lightweight compact amplifier using heater-type tubes with moderately high plate voltages and currents is suitable. There are other applications where the available power supply limits the plate current, the heater current, or both. Here, tubes with a-c heaters will again be used, but they may have to be types selected for low heater drain, operated at lower plate current. It is quite possible that this type of amplifier may use more tubes than one in which B-supply current is no consideration.

Where the only source of power is a storage battery or generator of 28 volts, it is advantageous to design an amplifier that will not require a higher plate supply voltage, thus eliminating the need for dynamotors, vibrators, or similar equipment. The primary concern is to obtain the desired gain and performance with a 28-volt B-supply, size and heater power being of secondary importance. B-supply current, negligible in comparison with heater current, is of no concern whatsoever.

For some portable applications, the only source of power available is dry batteries. Since the life of such equipment is inversely proportional to the power drain, it is important that this be kept as low as possible. Filament-type tubes must be used wherever possible because of the enormous saving in power over tubes using indirectly heated cathodes.

Choice of Tubes.—In small amplifiers, the prime consideration in choosing tubes and circuits is gain, which directly affects the size of the amplifier, rather than bandwidth, which affects only the pulse shape. The best tube for a given application is the one that will furnish the most gain for a given space or power-drain requirement. In many applications, the bandwidth or the pulse response is required only to be such that the output waveform indicates the presence of signals, without giving any information regarding the nature of the signal.

When this is the case, triodes are generally superior to pentodes for small amplifiers, despite the higher figure of merit of pentodes. The high amplification factor of pentodes cannot easily be utilized to obtain high gain unless large plate-load resistors are used. This produces a high

d-c voltage drop across the load resistor, thus necessitating the waste of considerable power and the use of high-wattage load and decoupling resistors. The use of the choke-coupled circuit to circumvent this problem introduces added complexity to the circuit and increases the size by adding the chokes which can easily be larger than the tubes used. If suitable compromise values are chosen, an amplifier can still be built using pentodes without excessive power requirements in which the gain per stage is higher than that obtainable with triodes. The use of twin-triode tubes, however, makes possible a considerably higher gain per envelope than that obtainable with pentodes. Furthermore, the elimination of the screen dropping resistor and bypass condenser may allow a considerable saving in size, especially where miniature and subminiature tubes are used, and the size of the amplifier may well be determined more by the number and size of the other components than by the tubes. The screen bypass condenser, for the first tube in particular, presents a serious problem, since it may be fairly large to ensure good recovery.

The main disadvantage of triodes for normal video-amplifier use, namely the Miller effect, is of little significance for this type of amplifier. Although the grid-plate capacitance appears effectively across the input circuit of each stage, multiplied by the gain of the stage, causing the input capacity of the stage to be unusually high, this merely has the effect of reducing the bandwidth and distorting the pulse and has a negligible effect on the gain.[1] The coupling condenser must be kept considerably larger than the equivalent input capacity, however, since the two capacitances are effectively in series across the plate load acting as a capacitance voltage divider. If the coupling condenser is too small, a considerable portion of the signal will be lost across it. This puts a limitation on the shortest possible time constant that can be used without loss of gain.

Choice of Circuit.—The choice of the circuit to be used for a small amplifier depends upon the size and power requirements and the desired recovery. Brute-force prevention of grid current, with large time constants for all circuits except one, is out of the question for a small amplifier because of the large physical size of the components required. The inverse-feedback pairs, direct or capacity coupled, can be used for amplifiers using heater-cathode-type tubes, with a high-voltage plate supply. Because of the absence of cathodes in filament-type tubes, direct coupling is out of the question for them. Neither type of inverse feedback can be used under low-voltage conditions without excessive loss of gain. Thus, the double short-time-constant is the only scheme outlined here that is usable for designing a single overshoot amplifier operating with a low plate-supply voltage.

[1] This is not true in the early stages of the amplifier before the signal has been distorted and lengthened.

For multiple overshoot amplifiers, straight *RC* coupling is simplest and best. Choke coupling can be used, but the advantages scarcely justify the added weight and size. In applications using a 28-volt power supply, more stages will be necessary for a given gain than with heater-cathode-type tubes operated with a high B-supply voltage, because of the reduced transconductance.

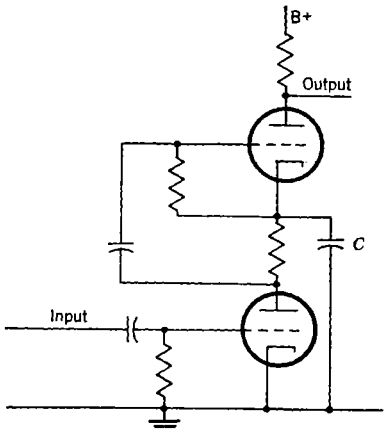


FIG. 19·19.—Series-pair connection of amplifier stages.

For some special applications where a high B-supply voltage of the order of 300 volts is available but the allowable current drain is low, a reduction of the current drain by a factor of 2 can be achieved by connecting the tubes in pairs, in series across the B supply, as shown in Fig. 19·19. For pulses, the pair represents an ordinary pair of amplifier stages, since the cathode of the second stage is effectively grounded by the large condenser *C*. For direct current the two tubes are in series

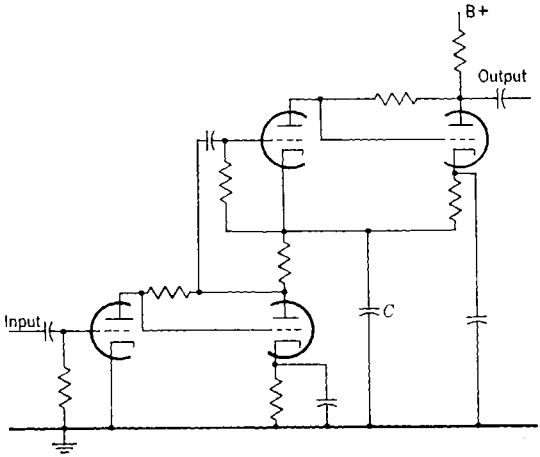


FIG. 19·20.—Series connection of direct-coupled inverse-feedback pairs.

across the B supply. Thus, the voltage across each tube is half the B supply (150 volts, if the total is 300), which is still enough for good amplification, whereas the current drawn from the supply for the two tubes is

only the plate current of one tube. This same idea can be extended to the direct-coupled inverse-feedback pair, as shown in Fig. 19·20.

The characteristics of the amplifier are not determined completely by the circuit diagram. As has been discussed previously, stray capacity can play an important part in the pulse response of the amplifier. Even more important in small amplifiers is the possibility of stray feedback which may have a peculiar effect on the pulse response of the amplifier and, if there is enough of it, may result in oscillation. These effects can be avoided by proper layout and proper grounding as discussed in Chap. 10. Briefly, the following points must be considered: The stages should be laid out in the form of a strip, from input to output; common ground points on the chassis should not be used for several stages; heaters should be grounded to the chassis at points where there are no other connections.

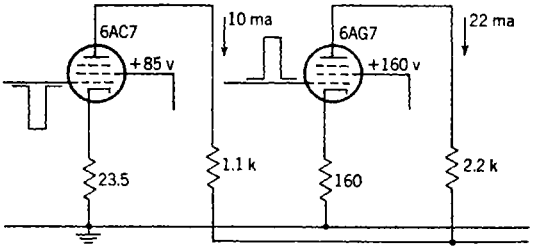


FIG. 19·21.—Illustration of the principle of design.

19·8. Examples.—Three examples of crystal-video receivers are discussed in this section. They have been chosen in an attempt to present a fair cross section of the types of receivers that might be built.

Receiver 1.—This receiver uses large tubes, 6AC7 and 6AG7, and was designed for a fixed installation. It is shown in Figs. 19·21 to 19·24. The amplifier is designed using the brute-force circuit to take a negative input pulse; thus a receiver for the 3000-Mc/sec band can be made by the addition of a crystal holder like that of Fig. 19·6, which is not shown in the photographs of the receiver. The receiver works with 2-μsec pulses and limits the stretching to 60 db per microsecond. This means that the amplifier must have its pulse stretching limited to 120 db per microsecond because of the doubling effect of the square-law detector.

The stages of the amplifier are designed in pairs, a typical one of which is shown in Fig. 19·21. In order to drive the 6AG7 into grid current, the grid voltage must rise above the cathode potential. At zero grid bias, the plate current of the 6AG7, as operated in the amplifier, is about 64 ma. Adding to this the zero-signal screen current of 6 ma, there will be a total of 70 ma flowing through the cathode resistor during

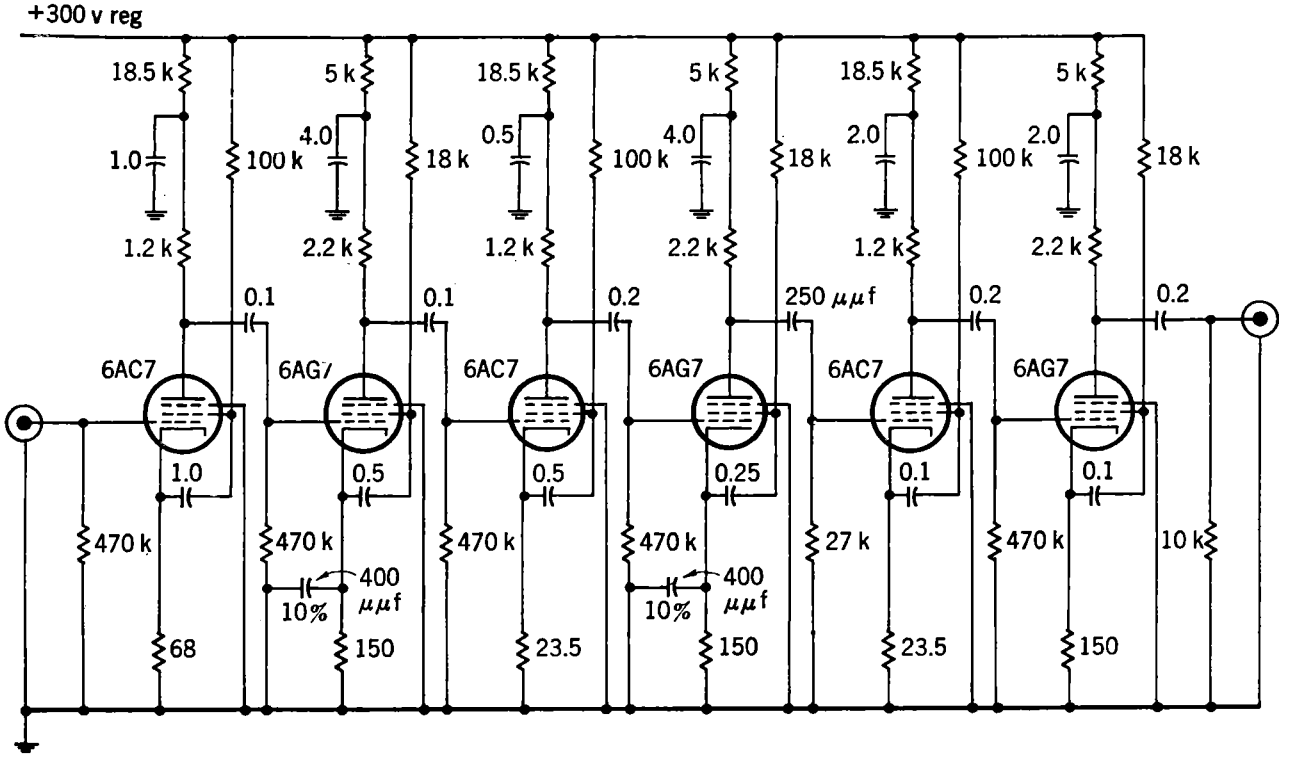


FIG. 19·22.—Circuit diagram of Receiver 1.

a pulse that barely drives the tube into grid current. The cathode voltage will therefore be $0.070 \times 160 = 11.2$ volts. If no grid current is to be drawn, the signal on the grid of this tube must, therefore, be limited to less than 11.2 volts. The limited output of the 6AC7 is

$$0.010 \times 1100 = 11 \text{ volts},$$

which is less than 11.2 volts and will not drive the 6AG7 into grid cur-

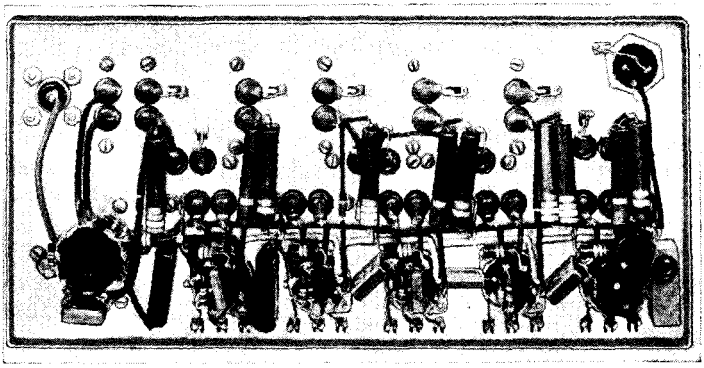


FIG. 19·23.—Bottom view of chassis, Receiver 1.

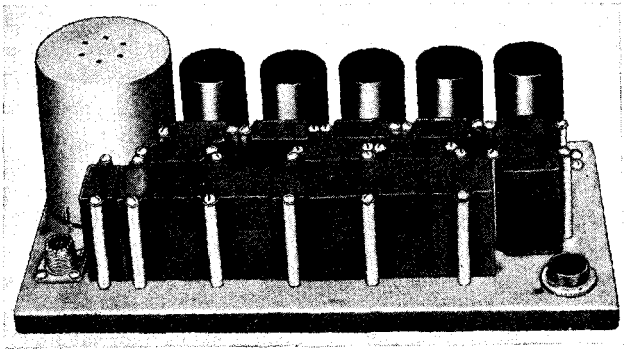


FIG. 19·24.—Top view of chassis, Receiver 1.

rent. In the circuit diagram of Fig. 19·22 the values are slightly different and would indicate on the basis of the previous calculation that grid current would occur. Actually, the tube characteristics are such that the circuit of Fig. 19·22 does not show any effects of grid current.

The time constants of all circuits are long except for the coupling circuit between the fourth and fifth stages, which is made short to introduce a single overshoot and to reduce microphonics. Because the

application requires that there be a sag of not more than 20 per cent in a 2-μsec pulse, a time constant of 10 μsec is used for this circuit. Since the signal on this grid is negative, the overshoot produced will drive the tube into grid current, thus shortening the time constant of recovery for this circuit.

Cathodes are left unbypassed, except for the small condensers used for peaking in the second and fourth stages. The screen bypass condensers are returned to the cathodes rather than to ground, so that the signal component of the screen current will not flow through the cathode resistor and produce added degeneration. The degeneration is kept as low as possible by the use of small cathode resistors.

The cathode resistor in the first stage is higher (68ω) than the other 6AC7 cathode resistors (23.5 ohms—actually two 47-ohm resistors in parallel) in order to ensure that no grid current flows through the direct-coupled crystal. Capacitive coupling is out of the question because of the enormous condenser that would be required to prevent overshoot. To ensure that the pulse stretching be held to 120 db/μsec, the bandwidth of the amplifier must be somewhat over 2 Mc/sec. The narrowest circuit is likely to be the input circuit. Hence the crystal used is the 1N27 whose video resistance is specified to be less than 4000 ohms.

The layout of the amplifier is quite straightforward and is shown in Fig. 19·23. The row of condensers nearer the tubes are the screen bypass condensers, the others being decoupling condensers. Of particular importance are the ground connections. There are three ground points for each tube, one for the shell, one for the heater, and the other for the suppressor, which carries also some circuit grounds. To ensure stability, all circuit grounds should be connected to the suppressor pin and none to the heater. In this case, it has been found permissible to violate this principle to a small extent without undue instability by returning the cathode circuit to the heater pin. As constructed, the amplifier is stable with a small cover over the circuits of the first tube and the remainder of the amplifier open. For convenience, a cover fitting over the entire amplifier is used.

To reduce microphonics, the first tube is shock-mounted. The leads to this tube, as shown in Fig. 19·23, do not go to the socket itself but to tie points mounted on a bakelite ring which are connected by flexible leads to the shock-mounted tube.

The amplifier requires an electronically regulated power supply furnishing 120 ma at 300 volts. Filtering in the heater line is necessary to prevent transients on the power line from being picked up within the amplifier and being amplified as signals.

Figure 19·24 shows the top view of the chassis of this amplifier with the large condensers used for screen bypass and decoupling.

Receiver 2.—This receiver was originally designed for an ultraportable application in the neighborhood of 3000 Mc/sec. It has a relatively narrow r-f bandwidth which, however, may be set anywhere within a broad band of frequencies. The original specifications allowed some

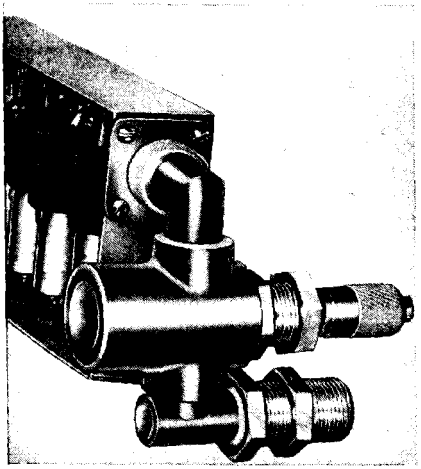


FIG. 19·25.—Crystal holder and tuned cavity.

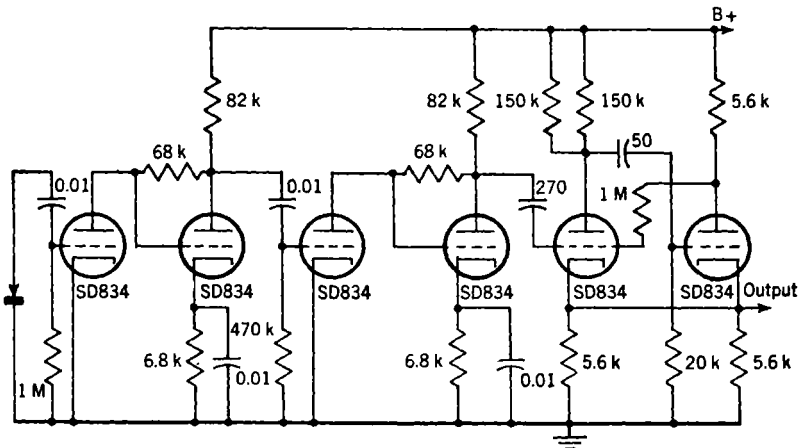


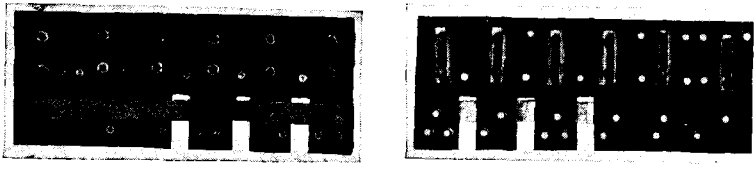
FIG. 19·26.—Circuit of amplifier, Receiver 2.

reduction in sensitivity, so the r-f filter and crystal holder are combined in one unit which is visible in Fig. 19·25.

The cavity is tuned by the micrometer which is not expected to hold a calibration. There is an insertion loss of several decibels in this cavity.

Because the receiver is tuned each time it is used, the problem of designing the holder to match a spread of crystal impedances is not serious.

The amplifier circuit uses direct-coupled inverse-feedback pairs with a multivibrator output circuit. The inverse-feedback pairs are designed as described in Sec. 19·6. The last two tubes form a multivibrator which



(*a*) (*b*)

FIG. 19·27.—Mounting strip of subminiature amplifier. (*a*) Bottom view; (*b*) top view.

is triggered by the negative signal appearing on the grid. The short time constant for the elimination of long overshoots and low-frequency interference is the time constant of coupling in the multivibrator, 50 $\mu\mu$f and 20,000 ohms, giving a time constant of 1 μsec. In addition, the time constant in the grid of the first tube of the multivibrator is

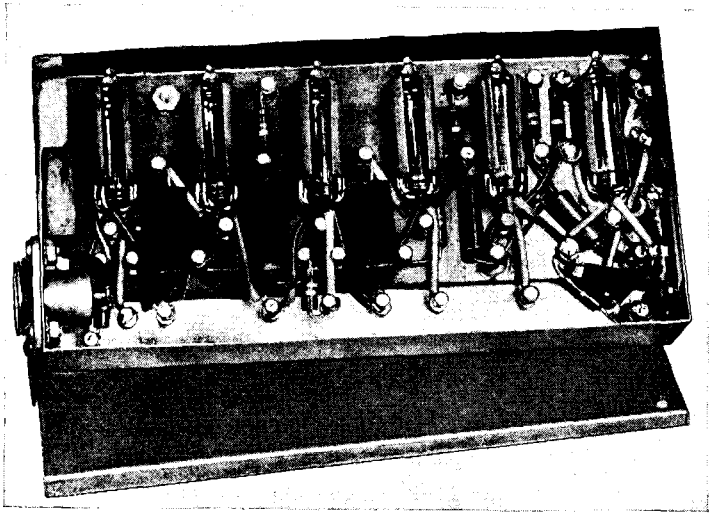


FIG. 19·28.—Bottom view of chassis, Receiver 2.

shorter than is apparent because the 1-megohm grid resistor is shunted by the conductance of the grid, which is positive except during the period of the multivibrator.

Overshoots from other sources are minimized by the use of direct coupling to the positively driven stages and of long time constants in

the negatively driven stages. Decoupling has been eliminated, as the amplifier is quite stable if the B-supply impedance is low. This can generally be achieved by using a condenser of moderate size (between 0.1 and 1 μf) across the supply. The circuit diagram for this amplifier is shown in Fig. 19·26.

The unique feature of this amplifier is the small size obtained by the use of type SD-834 subminiature tubes. The details of layout and construction are shown in Figs. 19·27 and 19·28. The tube clamps are mounted on a bakelite strip, together with turret lugs which are used for all connections. The slots are put into the strip to allow for the mounting of the postage-stamp-size 0.01-μf condensers. The power leads are strips of copper foil running along the back side of the strip, as shown in Fig. 19·27*a*, the upper strip being the B bus and the bottom strip being the heater bus. This serves the purpose of keeping the power leads close

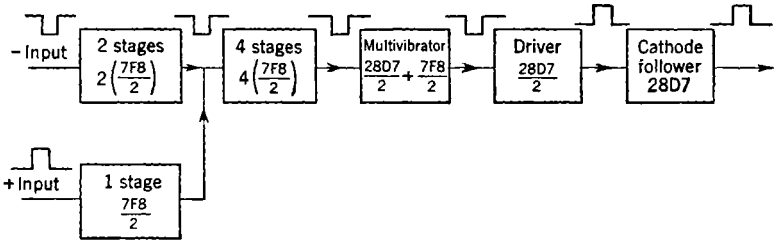


FIG. 19·29.—Block diagram of receiver.

to the chassis and out of the way, avoiding unnecessary congestion of leads. The strips, kept away from circuit components and having high capacity to ground and low inductance, are not so prone to cause feedback trouble as are ordinary wire leads.

The strip is mounted into the chassis, using another bakelite strip of similar dimensions to insulate the copper strips from the chassis. All connections are made to the turret-lug terminals on the strip, except for ground connections which are made to turret lugs mounted directly into the chassis below the strip. There is a separate ground lug for each stage, plus other ground lugs where the heaters are grounded. Some lugs serve as ground for two heaters, others for one, depending upon convenience, but no lug is used for both heaters and other circuits. With the precaution of separate grounds and the copper strip as the heater lead, there is no need for further filtering of the heaters.

Receiver 3.—This receiver, shown in Figs. 19·29 to 19·31, is designed for an airborne application, where it is desired to place the receiver at a point remote from the rest of the equipment, including the power supply.

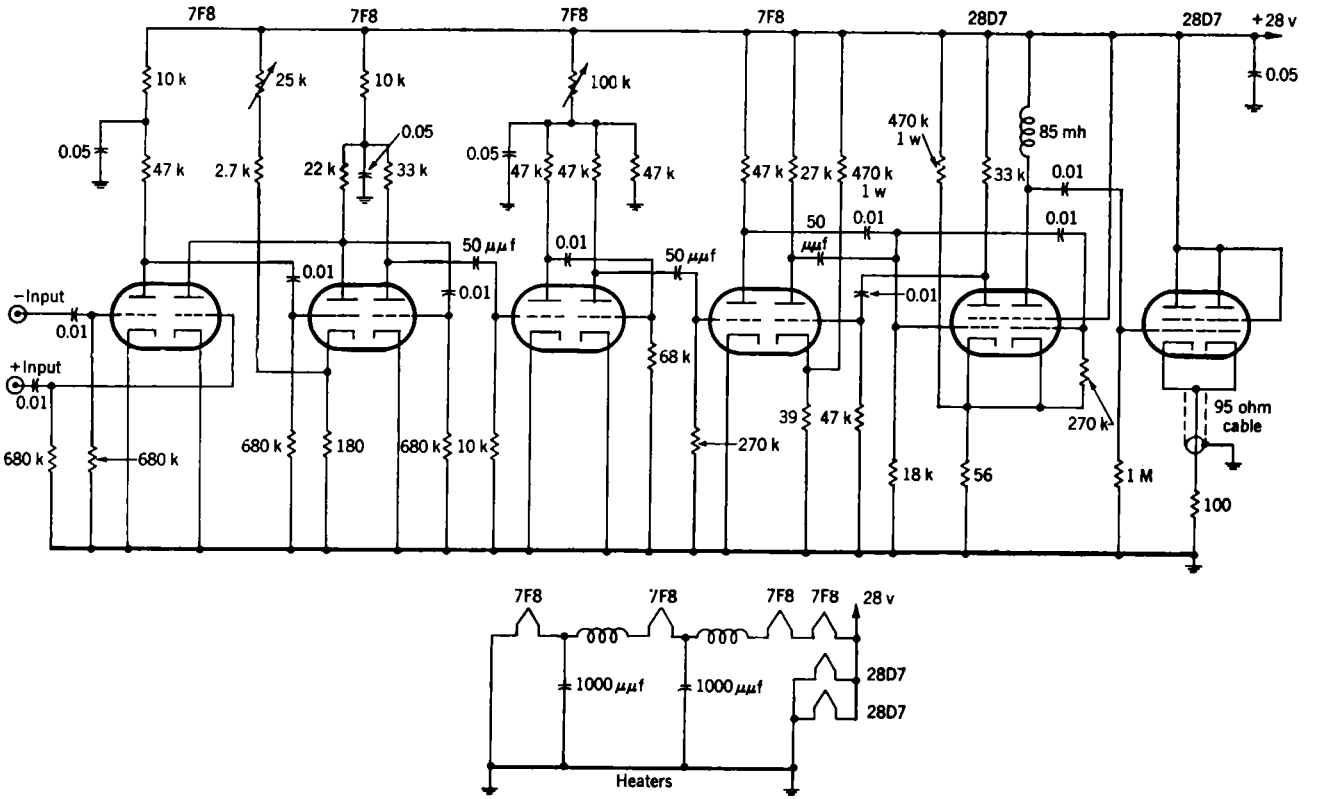
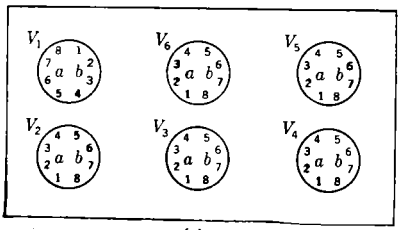
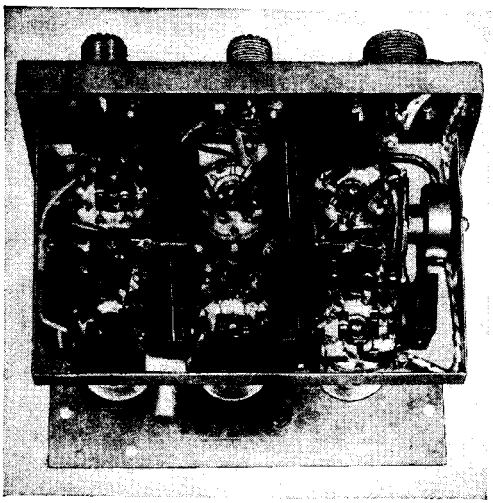


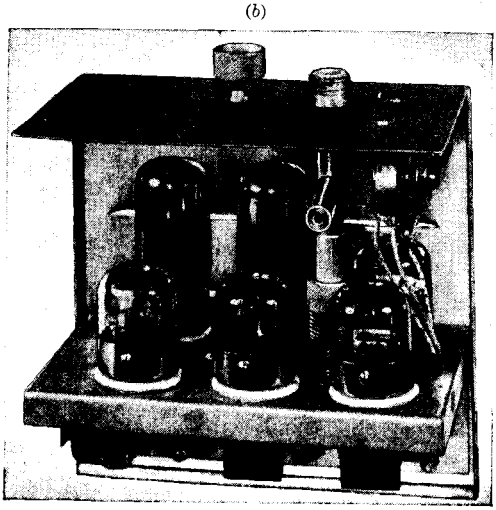
FIG. 19·30.—Circuit diagram of receiver.



(a)



(b)



(c)

FIG. 19·31.—(a) Tube layout; (b) bottom view of chassis; (c) top view of chassis.

It was also required that the receiver must use 28 volts and must be capable of delivering a 12-volt trigger into a 95-ohm cable. The receiver must furnish a trigger for any signal that it receives in either of two bands, one in the neighborhood of 10,000 Mc/sec and one in the neighborhood of 3000 Mc/sec. This necessitates two crystal holders feeding into a double-channel amplifier. Since the 3000-Mc/sec crystal holder has a negative video output and the 10,000 Mc/sec has a positive video output, the two amplifier channels must take signals of opposite sign.

To give maximum sensitivity, the trigger circuit must be adjusted so that it just fires on noise. The use of a separate trigger circuit for each channel would require too many tubes. Therefore the noise outputs of the two channels add, each reducing the sensitivity of the other channel. If satisfactory sensitivity is to be obtained on both bands, the gains of the two channels must be balanced. Two gain controls are necessary, one that affects only one channel and is therefore a balancing control and one that affects both channels and controls the output noise level.

A block diagram of the amplifier is shown in Fig. 19·29. The two input channels are mixed after one stage of amplification for the positive signal and two stages for the negative, in order to get the necessary relative inversion. This is followed by four stages of amplification and a multivibrator. In parallel with the normally conducting tube of the multivibrator, except for the plate circuit, is a driver stage using the high-voltage output circuit described in Sec. 19·6 under "output circuits." The high output pulse is used to drive a cathode-follower output stage. Considerable output is obtained by use of type 28D7 twin beam-power-amplifier tubes as the driver and the cathode follower. The 7F8 twin triode, although not designed for 28-volt operation, performs quite well as an amplifier under this condition and is used for all stages except the output.

The circuit shown in Fig. 19·30 is one of straight *RC*-coupling with multiple overshoot. The time constants are chosen to make the recovery as rapid as possible. Short time constants are introduced into the amplifier as late as possible, so that the signal will be limited before reaching these circuits and will produce smaller overshoots. It is important that long overshoots do not reach sufficient amplitude to drive tubes to grid current. Hence, the short time constants are used in the coupling circuits of the first negatively driven stages where the long positive overshoot would cause grid current.

The time constants in the early stages are made long. The problem of grid current is disregarded in the first positively driven stage (the maximum possible signal is still so small that any overshoot produced by grid current will be made negligible by the short time constants following). The following stage, however, although negatively driven, will be driven

into grid current by the long overshoot. Since this would produce an extremely long paralysis of the amplifier, a short time constant is used for coupling at this point. The following stage will be driven into grid current by the signal, hence the grid resistor is held to the same order of magnitude as the plate resistor to prevent a long recovery. The coupling condenser is made as large as possible (0.01 μf) to reduce the amplitude of the overshoot produced here to a minimum. At the output of this stage, the long overshoot has been amplified to a level where it is again troublesome, and another short time constant is inserted. The time constant of the coupling into the multivibrator is long, but the grid resistor is kept low because of the requirements of the multivibrator, which is a straightforward circuit. All decoupling-circuit time constants are large; thus there are only two short time constants in the circuit.

The layout of this amplifier presents a peculiar problem because the space requirements did not permit the use of a narrow strip. The layout as shown in Fig. 19·31 solved the problem.

Figure 19·31*b* is a photograph of the layout, and Fig. 19·31*a* is a diagram showing the pin numbers of the tubes and numbering them for identification. Figure 19·31*c* shows the top of the chassis and the 3000-Mc/sec crystal holder. The functions of the tubes, as numbered in Fig. 19·31*a*, are as follows:

Triodes V_{1a} and V_{2a} are the first two stages of amplification in the 3000-Mc/sec band channel, the input lead coming in from the crystal holder, which is on top of the chassis and located at the center of the group of tubes—V_1, V_2, V_3, and V_6 as shown in Fig. 19·31*c*. The connection to the holder is obscured in Fig. 19·31*b* by the parts covering it, but the input coupling condenser can be seen running beside tube V_1 next to the large decoupling condenser.

Triode V_{1b} is the input stage for the 10,000-Mc/sec channel. The crystal holder for this channel is not mounted in the chassis but on the antenna. A small length of cable is used between the holder and the connector on the front panel in front of tube V_1.

The two channels are mixed at the plate of V_{2a}, and the following four stages of amplification are triodes V_{2b}, V_{3a}, V_{3b}, and V_{4a}. Beam power tube V_{5a} and triode V_{4b} form the multivibrator, with tube V_{5a} being the normally conducting tube; tube V_{5b} is the driver; and tube V_6, both halves connected in parallel, is the cathode follower.

The input and high level amplifier stages are at opposite corners of the amplifier. Although the output stage is adjacent to the input, the feedback loop is broken by the normally nonconducting half of the multivibrator. Therefore there will be no transmission around the loop, except immediately after a signal, when the recovery time of the multivibrator prevents continuous ring-around. The stray feedback is

reduced still more by the 0.05-μf decoupling condensers visible in Fig. 19·31*b* which act as interstage shields, since they are at ground potential for pulses. To prevent excessive temperature rise resulting from the large heater power (11.2 watts) required by the 28D7 tubes, good ventilation must be provided. In the receiver the temperature rise at some points on the chassis would be as much as 90°C in the absence of cooling.

CHAPTER 20

SUPERREGENERATIVE RECEIVERS

By G. O. Hall

Introduction.—The superregenerative principle has been known since the early 1920's when it was described by E. H. Armstrong.[1] Radio amateurs have made wide use of superregenerative receivers in the intervening years. During the war the emphasis that was placed on lightweight portable equipment of small cost led to a study of the superregenerative principle, and some new techniques were evolved.

The limit of gain in an ordinary regenerative amplifier is reached when positive feedback is increased to a point where the tube oscillates. Superregeneration extends operation into the region of oscillation by allowing the circuit to oscillate for only a fraction of the time. This extension greatly increases the gain over ordinary regenerative amplifiers. Voltage gains of over a million have been secured from a one-tube superregenerative circuit. The circuit may be used as an extremely sensitive tuned detector or as a high-gain r-f amplifier.

20·1. Advantages and Disadvantages.—The merits of the superregenerative circuit are (1) simplicity, (2) high gain from a single tube, (3) light weight, (4) low cost, and (5) ease of combining transmitting and receiving functions.

The advantages of simplicity and light weight become increasingly important at frequencies above 100 Mc/sec, as other types of receivers (TRF, superheterodyne) tend to become more complicated at these higher frequencies. The feature of light weight has led to extensive use of superregenerative receivers for airborne and ground-transportable radar beacons. In several beacon applications a single tube has served the double function of reception and transmission.

There are several undesirable characteristics that are inherent with superregenerative receivers. These are (1) gain instability, (2) poor selectivity, (3) reradiation, and (4) high noise level. Fortunately, the majority of these undesirable features can be suppressed to a point where the over-all performance is satisfactory for a great many applications.

The maintenance of proper sensitivity is not a serious problem for applications where the receiver is supervised, because the operator can

[1] Edwin H. Armstrong, "Some Recent Developments of Regenerative Circuits," *Proc. IRE*, **10**, 244, August 1922.

regulate the receiver with manual controls. Automatic systems, such as radar beacons, which must operate reliably without attendance, present a problem in gain stability. This is particularly true for pulse-reception receivers, since such receivers operate in the *linear mode* which is critical in adjustment. This problem is solved by the incorporation of *automatic gain stabilization* circuits.

The poor selectivity of the superregenerative receiver can cause considerable difficulty if active r-f channels are close together in frequency. Fortunately, in the frequency ranges where superregenerative receivers become practical, r-f channels are fairly well separated, and there is little difficulty from interference. Reradiation becomes a problem only when two or more sets are located close together (within 1 mile).

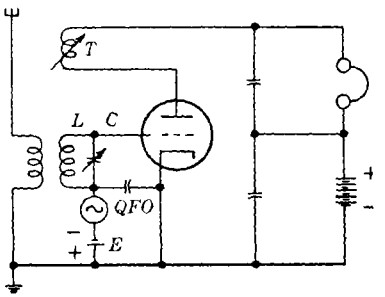


FIG. 20·1.—Circuit of superregenerative detector.

The noise level will be higher than that of a well-designed superheterodyne receiver, but for many applications the advantages of light weight, simplicity, etc., will outweigh this disadvantage.

20·2. Theory of Operation.—The superregenerative principle can be applied to any type of oscillator by the use of the same basic theory. The tubes usually employed are the small low-power high-frequency triodes. In the region of 3000 to 10,000 Mc/sec velocity-modulated oscillators or low-power magnetrons may be used with some sacrifice in sensitivity.

A simple superregenerative detector for communications work is shown in Fig. 20·1. In this circuit, positive feedback is secured by coupling the plate back to the grid through the coil T. The amount of feedback is controlled by the orientation of this coil with respect to L. The grid circuit has a negative bias supply in series with a sine-wave quench oscillator *QFO* so that the circuit is placed in an oscillating condition on each positive peak of the quench voltage. On each negative swing of quench voltage the tube is cut off. The frequency of the quench oscillator is well above the audio range so that it will not interfere with the modulation frequencies. The plate circuit will have pulses of current occurring at the quench frequency and producing a certain average plate current when no signals are present. When signals are present, the areas under the envelopes of the individual bursts of oscillation increase, the amount of increase being dependent upon the signal amplitude. By this action the average plate current varies in amplitude at an audio rate that represents the demodulation of the received carrier.

The phones are bypassed for the quench frequency and respond to the audio rate of change in plate current.

20·3. Growth and Modes of Oscillation.—The fundamental action in a superregenerative circuit is centered around the growth of oscillations in an oscillator. To study the characteristics of this growth, a rectangular quench voltage will be assumed.

The variation in grid voltage is shown in Fig. 20·2*a*. Between positive pulses, the grid is at the negative potential $-E$ which biases the tube beyond E_{co}, the cutoff voltage. During the positive pulses the grid potential is raised to a point where the circuit oscillates as indicated by projection into the shaded oscillation region. Thus oscillations grow during the A periods and decay during the B periods.

There are two modes of operation defined for superregeneration. They are linear mode (Fig. 20·2*b*) and logarithmic mode (Fig. 20·2*c*).

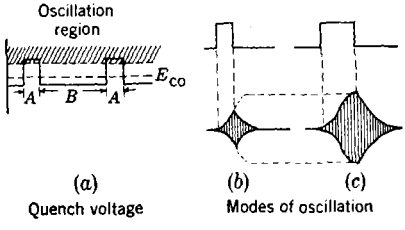


FIG. 20·2.—Quench voltage and relation to modes of oscillation.

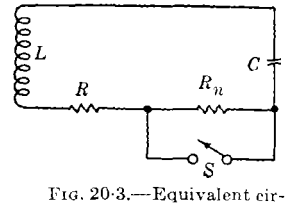


FIG. 20·3.—Equivalent circuit for superregenerative tank circuit.

The linear mode results when the positive quench period A is so short that the oscillations do not have time to build to full saturation amplitude. The logarithmic mode occurs when the A period is sufficiently long to allow oscillations to build to full amplitude before the end of the A period.

It is not essential that the quench voltage be applied to the grid circuit as the on-oscillation off-decay cycle can also be obtained by applying a suitable quench voltage to the cathode or plate. However, plate and grid quenching are more common than cathode quenching.

In constructing an equivalent circuit for an oscillator, the feedback can be represented as a negative resistance. With this concept, the tank circuit of the superregenerative receiver becomes the circuit shown in Fig. 20·3 where R is the physical loss resistance, R_n the negative resistance representing feedback, and the switch S represents the action of the quench voltage. Switch S is open during the A periods and closed during the B periods.

The general equation for the circuit is

$$L\frac{di}{dt} + R_T i + \frac{1}{C}\int i\,dt = 0, \tag{1}$$

where R_T is the net tank resistance. This equation may be solved for e_c, the voltage across the condenser;

$$e_c = ke^{-\alpha t} \cos(\omega t + \theta), \tag{2}$$

where

$$\alpha = \frac{R_T}{2L}, \qquad \omega = \sqrt{\frac{1}{LC} - \frac{R_T^2}{4L^2}}.$$

For the types of circuits to be considered $1/LC \gg R_T^2/4L^2$, and it will be assumed that $\omega = \omega_0 = 1/\sqrt{LC}$.

When there are no externally applied signals, noise voltage is the only source of voltage when period A begins. Then for the no-signal condition, where V_n is the noise voltage appearing across the condenser at the instant S is opened,

$$e_c = V_n e^{-\frac{R_T}{2L}t} \cos \omega_0 t. \tag{3}$$

Since the circuit oscillates, R_n is larger than R; the net tank resistance R_T is negative, and the envelope of the oscillations is

$$e_c = V_n e^{(R_T/2L)t} \tag{4}$$

This equation[1] indicates that oscillations will increase continuously with time if the net tank resistance remains negative. However, the negative portion of the tank resistance R_n is dependent upon two factors: (1) the percentage of feedback and (2) the amplification of the circuit.

As oscillation starts, the operating point for the grid is in the negative region, and the amplification will be practically constant until the oscillations drive the grid either positive or beyond cutoff.

Amplification will diminish as the grid is driven positive or beyond cutoff, and the magnitude of R_n will decrease. As the oscillations continue to grow, R_n continues to decrease until $R_n = R$ and the so-called "saturation" oscillation condition exists.

20·4. Gain in the Logarithmic Mode.—The growth of oscillations under no-signal conditions is shown in Fig. 20·4, where V_n is the noise voltage present at the start of oscillations and E_{max} is the saturation amplitude. Since at $t = 0$, $e^{(R_T/2L)t} = 1$, the oscillations start with an amplitude V_n. Since noise is random in amplitude, the area under the oscillation envelopes varies from pulse to pulse in a random manner. This results in a modulation of plate current that causes a rushing sound in the phones. This rushing sound is a criterion for superregenerative action in a communication receiver and is often used as a rough check in adjusting a receiver. When signals are received, the bursts of oscillation become regular and the noise disappears.

[1] From this point on, R_T is defined as the absolute value of $(R - R_n)$. Thus the sign of the exponent will show directly whether a function is expanding or decaying.

If a signal voltage V_s is present, the voltage on the condenser at the start of the oscillations will be $V_s + V_n$ and Eq. (4) becomes

$$e_c = (V_s + V_n)e^{+\frac{R_T}{2L}t} \tag{5}$$

The envelope for oscillations with a signal present is the upper curve of Fig. 20·4. Oscillation has started at V_s volts higher than V_n, and saturation amplitude is reached t_a sec (time of advance) sooner than when only the noise voltage was present.

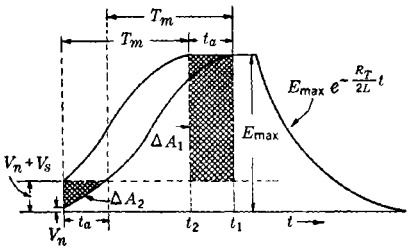


FIG. 20·4.—Growth and decay of oscillations for the logarithmic mode.

By referring to Fig. 20·4 the change in area with signal voltage can be calculated for the logarithmic mode. Since the growth of oscillations will follow the same law on each cycle, the time for oscillations to grow from some fixed amplitude to E_{max} will always be the same. Specifically, the time required for growth from the $V_n + V_s$ voltage level to E_{max} is the same for both curves and is represented by T_m.

Consider first the difference in areas above the level $V_n + V_s$. Since the upper curve started from $(V_n + V_s)$ at a time t_a sec before the lower curve, it reached E_{max} earlier by t_a sec and enclosed the additional area $t_a(E_{max} - V_n - V_s)$ represented by the large shaded area ΔA_1. Below the voltage level $V_n + V_s$, it is evident that the additional area is represented by the small shaded section ΔA_2.

For convenience $(V_n + V_s)$ will be replaced by V_2, and V_n by V_1. The time of advance t_a is obtained by solving

$$V_2 = V_1 e^{(R_T/2L)t_a}, \tag{6}$$

giving

$$t_a = \frac{2L}{R_T} \ln \frac{V_2}{V_1}. \tag{7}$$

Figure 20·4 has been drawn out of scale to show the effect of initial conditions on the growth of oscillations. Ordinarily E_{max} is measured in volts, whereas signals are of the order of microvolts. Since the voltage rise being considered in Eq. (6) is only a small fraction (about 1/10,000) of the total rise to E_{max}, R_T can be considered constant and Eq. (7) is valid over the entire range of signal voltages encountered in ordinary receptions.

The increase in area above V_2 becomes

$$\Delta A_1 = t_a(E_{max} - V_2) = (E_{max} - V_2)\frac{2L}{R_T} \ln \frac{V_2}{V_1}. \tag{8}$$

The area ΔA_2 below V_2 is approximately half of the rectangle $t_a V_2$. Since the entire rectangle has an area of about 1/10,000 of ΔA_1, the error will be negligible if the entire change in area is considered to be $E_{max} t_a$. Then

$$\Delta A = E_{max} \frac{2L}{R_T} \ln \frac{V_2}{V_1}. \tag{9}$$

Since the decay of oscillations in the logarithmic mode is always from the same level (E_{max}) and starts at a time determined only by the quench voltage, the area under the decaying portion of the cycle is constant. Thus Eq. (9) gives the entire change in area per quench cycle. The change in average voltage of the detected output, assuming linear detection, is

$$V = f_q \Delta A = f_q E_{max} \frac{2L}{R_T} \ln \frac{V_2}{V_1}, \tag{10}$$

where f_q is the quench frequency in cycles per second.

This equation indicates the tremendous gain obtainable by superregeneration. The factor $f_q E_{max}(2L/R_T)$ is of the order of volts, and $\ln (V_2/V_1)$ is a factor dependent on voltage ratios and not absolute amplitudes. For example, if a signal is received of amplitude equal to V_n, then the increase in output over no signal will be

$$V = f_q E_{max} \frac{2L}{R_T} \ln 2. \tag{11}$$

This can easily mean a gain of a million, since noise voltage V_n is only a few microvolts.

20·5. Output Characteristics for the Logarithmic Mode.—Since for any particular circuit the quantity $f_q E_{max}(2L/R_T)$ is a constant, a plot of detector output vs. signal ratio for linear detection becomes the logarithmic curve of V_2/V_1 shown in Fig. 20·5. The flattening of this curve with increase in signal ratio shows that the logarithmic mode has a limiting action similar to AVC. This is borne out in practice by a marked reduction in interference from noise impulses (such as ignition noise) that are many times stronger than the signal. This curvature also explains the disappearance of the rushing noise when signals are received. The slight addition of V_n to a moderately large signal voltage produces a

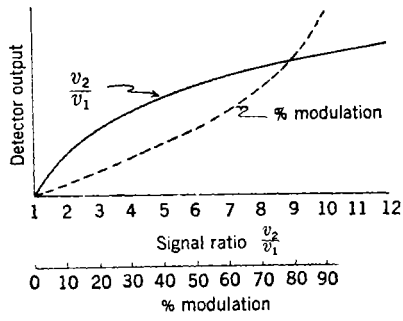


FIG. 20·5.—Variation of linear detector output for logarithmic mode.

negligible addition in the output, since the effect on the voltage ratio is small.

The curve of detector output vs. percentage modulation is fairly linear up to about 60 per cent modulation but shows a "volume expansion" characteristic above 60 per cent which grows rapidly worse as the modulation approaches 100 per cent. Thus the logarithmic mode is not suitable for broadcast reception. Deep peaks of modulation in voice or music produce violent volume changes.

20·6. Determination of Maximum Usable Quench Frequency.—It is evident that maximum gain will be obtained when the quantity

$$f_q E_{\max} \frac{2L}{R_T}$$

is a maximum. Assume first that $E_{\max}(2L/R_T)$ has already been established and that the maximum usable quench frequency is to be determined. The optimum A period (Fig. 20·2) for the logarithmic mode would be one just long enough for the oscillations to reach $E_{\max}$ from the weakest signal, since extending the period of oscillation at full amplitude gains nothing in detected output. Assume that a 100-Mc/sec receiver, operating in the logarithmic mode, has a growth time of 5 μsec for the weakest signal to be received. This will establish the minimum usable A period as 5 μsec as indicated in Fig. 20·6. At the end of the A period oscillations decay exponentially, the equation for voltage being

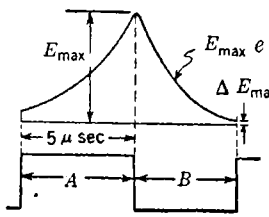


FIG. 20·6.—Decay of oscillation in the logarithmic mode.

$$e_c = E_{\max} e^{-\frac{R}{2L}t} = E_{\max} e^{-\frac{\omega t}{2Q}}, \tag{12}$$

where t is measured from the start of the B period.

Since the tube is cut off (equivalent to a closed switch in Fig. 20·3) the Q in Eq. (12) is for the tank circuit alone (no feedback).

The B period must be sufficiently long so that the amplitude of the oscillation envelope $\Delta E_{\max}$ remaining at the end of the period is less than the amplitude of the received signal, or the circuit will operate on its own signals to the exclusion of received signals. If a Q of 100 is assumed at 100 Mc/sec, the time constant will be $2Q/\omega = 0.315$ μsec. Table 20·1 lists the relation between $\Delta E_{\max}$ and the duration of the B period for the range that will be of interest. Thus a duration of 3.15 μsec (10 time constants) for the B period will reduce the envelope to $0.00005E_{\max}$ for this hypothetical 100-Mc/sec receiver. The total period is 8.15 μsec,

TABLE 20·1.—RELATION BETWEEN ΔE_{max} AND DURATION OF B PERIOD

Duration of B in time constants	$\frac{\Delta E_{max}}{E_{max}}$
5	0.00670
6	0.00250
7	0.00090
8	0.00040
9	0.00010
10	0.00005

or f_q is roughly 123 kc/sec. By this method a quench cycle can be selected for maximum sensitivity under given conditions of E_{max} and $2L/R_T$.

20·7. Considerations for Maximum Gain.—To secure the optimum condition, all three factors E_{max}, $2L/R_T$, and f_q must be considered simultaneously. If f_q is maintained at the maximum usable frequency as the factor $2L/R_T$ is increased, considerations are as follows:

1. As the factor is increased, the rate of growth will be decreased, the A period lengthened, and f_q decreased.
2. If the rate of growth is reduced too much, oscillation will be unstable.
3. R_T can be reduced by reducing the Q of the tank and by reducing feedback, since $R_T = |-R_n + R|$. (a) Reduction of Q will accelerate decay and reduce the required length of the B period. (b) Reduction of feedback will not affect the rate of decay.
4. L can be increased at low frequencies, but it may be impossible at high frequencies.
5. There will be some reduction in E_{max} as the factor is increased.
6. If f_q is reduced too much as a result of increased $2L/R_T$, trouble may be encountered in separating f_q from the modulation frequency.

20·8. Sinusoidal Quench Voltages.—Up to this point a rectangular quench voltage has been assumed. Since most circuits employ sinusoidal quench voltages, the foregoing analysis must be reexamined to determine whether or not a change to a sinusoidal quench voltage will materially affect the final results. Furthermore, selectivity and the factors that determine it must be investigated before final conclusions can be drawn regarding the design of a superregenerative circuit.

When the grid voltage varies sinusoidally, the net tank resistance R_T varies with time between the limits of R and $R - R_{n\,max}$. For the usual triode (not remote cutoff) μ increases rapidly as the grid voltage is raised above cutoff and then remains practically constant. The general characteristic will be similar to the μ vs. e_g plot of Fig. 20·7a. Figure 20·7b shows the variation in the net tank resistance R_T with sine-wave quench. As the quench voltage e_q crosses cutoff, the net tank resistance

decreases from R at a to 0 at b on to the maximum negative value $R - R_n$ at c. From a to b the net tank resistance is still positive, oscillation does not occur, and the system operates as an ordinary regenerative circuit. As R_T crosses zero, oscillation starts and the superregenerative cycle begins.

With rectangular quench the shape of the envelope of oscillations is given by

$$e_c = Ve^{(R_T/2L)t} \tag{20·4}$$

where R_T is assumed constant. The slope of the curve is given by

$$\frac{de_c}{dt} = V\frac{R_T}{2L}e^{(R_T/2L)t} \tag{13}$$

which gives an initial slope of $V(R_T/2L)$.

With the variation of R_T shown in Fig. 20·7*b*, R_T is zero at b, ($t = 0$) and the initial slope is zero. In the interval bc the value of R_T is increas-

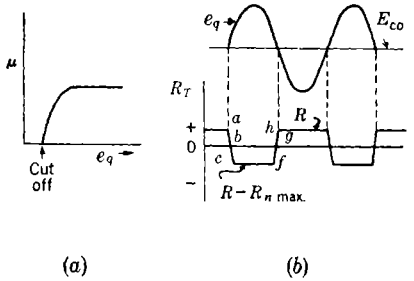


FIG. 20·7.—Variation in net tank resistance with quench voltage.

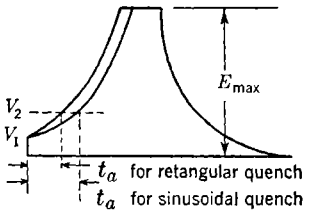


FIG. 20·8.—Comparison of oscillation growth with rectangular and sinusoidal quench voltages.

ing negatively, and the condition of R_T assumed for the rectangular case is not reached until c.

Figure 20·8 shows a comparison between growth of oscillations with a rectangular quench voltage and a sinusoidal quench voltage. Both start from the initial voltage V_1, but the sine-wave-quenched curve starts with zero slope and always remains below the curve for rectangular quench. As soon as $R_T = |R - R_{n\ max.}|$ (point c of Fig. 20·7), the two curves are parallel, since the exponent of e is identical for both curves. The time required for the lower curve (sine-wave quench) to reach a higher voltage (V_2) is greater than for the upper curve (rectangular quench), so the time of advance t_a is greater for the sine-wave quench, and the gain, other things being equal, is greater for sine-wave quench than for rectangular quench. Since the quench voltage enters the oscillation region obliquely (with sine-wave quench), the presence of a signal adding to the quench voltage will cause oscillation to start earlier. The

amount of this advance in the start of oscillations will increase with signal strength and add to the over-all gain. On the decay cycle, as R_T recedes from $R - R_{N\,max}$ to R (f to g to h, Fig. 20·7), maximum rate of decay will not be attained until point h is reached where $R_T = R$. From this point on, the law for decay will be the same as that for the rectangular quench.

The results of the change to sine-wave quench may be summarized as follows:

1. The shape of the R_T variation has changed from a rectangle to a trapezoid, the slope of the sides being dependent upon the slope of the sine wave as it crosses cutoff.
2. The cycle now includes a period of regenerative operation.
3. The gain has been increased, but so has the minimum usable quench period.
4. The general law for gain, in the logarithmic mode, is essentially unaltered.

In selecting the quench frequency for optimum gain unequal A and B periods were indicated. Figure 20·9 illustrates how the periods can be altered by changing the bias and amplitude of the quench voltage.

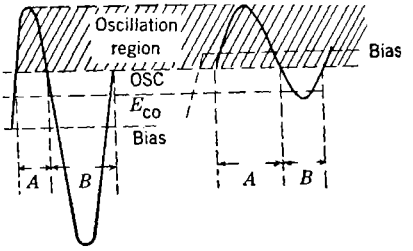


FIG. 20·9.—Control of A and B periods by bias and quench amplitude.

20·9. Selectivity.—Since the bandwidth of the superregenerative receiver is determined by the response curve of a single tuned circuit and not by a series of cascaded tuned circuits, the selectivity cannot be made so great as that of a TRF or superheterodyne receiver. Since the application of superregeneration is usually at the higher frequencies (where other systems tend to become more complicated), interference problems are generally not serious.

In the preceding section it was noted that the circuit passed through a regenerative period before reaching the superregenerative condition of oscillation. This regenerative period improves the selectivity over the selectivity that would be attained by the tank circuit alone, since feedback reduces R_T and increases Q'. (Q' refers to the oscillator tank circuit with feedback present, $Q' = \omega L/R_T$.)

If the circuit is brought suddenly into the oscillating condition (rectangular quench), it is logical to assume that the initial voltage at the start of oscillations is a voltage determined by the response curve of the tank alone. On the other hand, if the circuit is slowly brought into oscillation, there will be a progressive amplification during the regenerative

period, Q' will be increasing, and there will be a progressive increase in the input selectivity up to the point of oscillation. This indicates that a low-frequency sine-wave quench will extend the period of regeneration and improve selectivity, particularly if the tube is so biased that the slope of the quench voltage is small throughout the regenerative period.

The amplitude modulation of oscillations (rise and decay) produces sidebands of the resonant frequency. These new frequencies can be determined by analyzing the envelope of the oscillations, since the sidebands will consist of the resonant frequency plus and minus the frequency components of the envelope. Figure 20·10 shows the line spectrum of a succession of pulses at a recurrence frequency f_q. This spectrum approximates the d stribution of frequencies encountered in the superregenerative receiver because it is for pulses having an exponential rise and decay. It is noted that only the discrete frequencies f_q, $2f_q$, $3f_q$, . . . , nf_q are produced, the amplitude of the individual lines of the spectrum decreasing as the order of the harmonic of f_q increases.

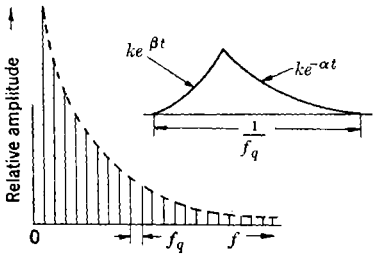


FIG. 20·10.—Line spectrum of a series of pulses at a recurrence frequency f_q.

If the superregenerative tube is operating as an amplifier (no detection), the sidebands have little or no effect on the selectivity, since beat frequencies (between the sidebands and incoming signals) fall far outside the response curve for the tank. If, however, the superregenerative tube performs detection, the frequencies spaced at f_q, $2f_q$, etc., can beat with incoming signals to produce frequencies that fall within the range of the circuits following the detector. A signal off resonance by nf_q will beat with the corresponding frequency in the spectrum to produce a sum or difference frequency that is passed. Thus, the circuit exhibits multiple resonance, having resonance points at intervals of f_q. This multiple resonance effect was pointed out by H. Ataka.[1] He suggested using multiple resonance as a means for accurately calibrating a wavemeter.

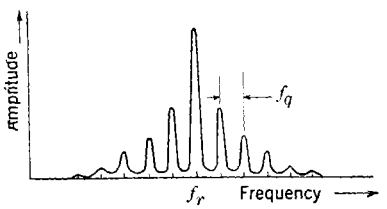


FIG. 20·11.—Multiple resonance curve.

The general shape of a multiple resonance curve is shown in Fig. 20·11. The amplitude of the individual resonance points will depend

[1] Hikosaburo Ataka, "On Superregeneration of an Ultra-short-wave Receiver," *Proc. IRE*, **23**, 851, August 1935.

upon the Q of the circuit, the conversion gain, and the amplitude of the particular frequency (nf_q) in the spectrum.

As the quench frequency is reduced, the frequency separation between the lines in the spectrum is decreased and any particular harmonic differs from the center frequency by a smaller amount. If, for example, the bandwidth of a tank circuit is 1 Mc/sec, the tenth harmonic of a 50-kc/sec quench frequency falls at the half-power point whereas it would require the twentieth harmonic of a 25-kc/sec quench to reach the same point. Thus, a low-frequency quench will reduce the effect of multiple resonance and improve selectivity. A general rule for preventing multiple resonance difficulties is to maintain a ratio of at least 100 between the radio frequency and the quench frequency. Figure 20·12 shows the relationship between selectivity and quench frequency. The data plotted is obtained from a 30-Mc/sec superregenerative receiver having a tank circuit with a Q of 100. A reduction in quench frequency from 50 to 8 kc/sec has reduced the bandwidth (at 20 db) by a factor of 3. It is to be noted that even at a 50-kc/sec quench frequency the bandwidth (about 1 Mc/sec) is much narrower than for a simple tuned circuit with a Q of 100, which would have a bandwidth over 3 Mc/sec wide at 20 db down. The selectivity of a superregenerative receiver then can be made much better than a simple tuned circuit. Comparative tests have shown that a properly designed superregenerative circuit has considerably better selectivity than the best regenerative receiver.

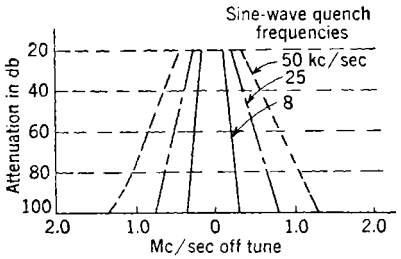


FIG. 20·12.—Selectivity of a 30-Mc/sec receiver.

20·10. Noise.—In an ordinary TRF receiver (assuming linear detection), the noise[1] that appears in the output is determined by the bandwidth of the output circuits. If, for example, the detector output pass band is restricted to 5 kc/sec, the noise appearing in the output will be from only those noise components which were present in a 10-kc/sec band of the tuned r-f input circuit even though the r-f circuit has a bandwidth many times wider.

In the case of a superregenerative circuit the noise in the output is from an r-f band considerably wider than twice the output bandwidth. An examination of the detected output with increasing modulation frequency will show how this is true. The variation in detector output

[1] Much of the material for this section was derived from Hazeltine Electronics Corp. Report No. 1506W.

frequency as the modulation frequency is increased is shown in Fig. 20·13. This figure is based on a detector output pass band wider than $f_q/2$.

The superregenerative process consists of sampling the r-f voltage in the tank circuit once each quench cycle and producing an envelope of oscillations for each sampling. If the modulation frequency f_m is low compared with the quench frequency f_q, there will be many samplings per modulation cycle and the average voltage of the detector output will vary in synchronism with the modulation. As the modulation frequency is increased, the number of samplings per modulation cycle decreases; and when f_m is greater than $\frac{1}{2}f_q$, the rise and fall of voltage in the tank as viewed by the sampling is at a frequency less than $\frac{1}{2}f_q$. As a matter of fact the frequency in the detected output will be the difference between f_q and f_m. As f_m increases toward f_q, the frequency in the detected output decreases as the difference between f_q and f_m becomes less and less, until at $f_m = f_q$ samplings occur once each modulation cycle and always at the same point on each modulation cycle, so the detector output frequency is zero.

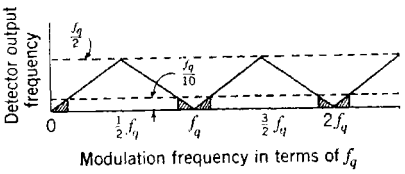


FIG. 20·13.—Detector output with increasing modulation frequency.

As indicated in Fig. 20·13, this process continues as f_m increases, having maximum detector output frequency of $\frac{1}{2}f_q$ when f_m is $\frac{1}{2}f_q$, $\frac{3}{2}f_q$, etc. This process will continue on out until the sidebands of the modulation frequencies are eliminated by the response of the r-f tuned circuit. Thus it is found that all modulation frequencies (including noise) which can appear in the pass band of the r-f tank circuit will be present in the output in a band of frequencies from 0 to $\frac{1}{2}f_q$.

Consider now that the pass band of the detector is reduced to $\frac{1}{10}f_q$ as indicated by the line on Fig. 20·13. The noise in the output will be the sum of the shaded contributions at $f_m = 0$, $f_m = f_q$, etc., and not just the one contribution at 0 as with an ordinary r-f amplifier.

The fact that the noise is considerably higher for superregenerative receivers than for other conventional types such as TRF and superheterodyne must be taken into account when considering the application of superregenerative receivers. It is to be noted that the most effective means for reducing noise in superregenerative circuits is reduction of the r-f bandwidth and not the output bandwidth. There is some evidence that tube noises, particularly shot effect, are increased by the intermittent tube operation associated with the superregenerative action.

20·11. Reradiation.—A superregenerative circuit must, of necessity, radiate energy if the tank circuit is coupled directly to the antenna. The

band of energy is broad because of the wide selectivity curve and may cause interference in near-by receivers operating on adjacent channels. The amount of reradiated energy can be reduced by decreasing the coupling to the antenna, but this reduces the sensitivity. A better method is to reduce the amplitude of E_{max}, since this has little effect on the ultimate sensitivity. The gain of the superregenerative circuit is reduced, but this can be restored by increasing the amount of video gain. More elaborate schemes are

1. Insert an r-f amplifier between the antenna and the superregenerative circuit. With present tubes and techniques, this is practical up to frequencies of about 1000 Mc/sec.
2. Convert the signal frequency to a lower frequency by superheterodyne methods, and operate the superregenerative circuit at the intermediate frequency.

20·12. Conclusions Regarding Logarithmic-mode Superregenerative Receivers.—It has been pointed out that maximum gain is attained when $f_q E_{max}(2L/R_T)$ is a maximum. The maximum usable value of f_q is determined by the sum of the A and B periods. As $E_{max}(2L/R_T)$ is increased, the sum of the A and B periods increases and the maximum usable quench frequency is reduced.

Selectivity is increased by reducing f_q to accentuate the regenerative period. Narrowing the r-f bandwidth improves the signal-to-noise ratio.

The first consideration in design is the maximizing of $E_{max}(2L/R_T)$. Since the value of L is usually determined by other considerations and its usable range is limited, attention is focused on reduction in R_T. It can be controlled by (1) the Q of the tank circuit and (2) the feedback. The Q of the tank circuit will be reduced as coupling to the antenna circuit is increased and more impedance is reflected into the tank circuit. This provides a means for reducing Q [to increase $E_{max}(2L/R_T)$] while improving the energy transfer from the antenna.

Consequently the Q of the tank should be made as high as possible, and the operating Q adjusted to an optimum point by varying the antenna coupling. The final desired performance can be attained by regulating (1) antenna coupling, (2) quench frequency, (3) quench-voltage amplitude, and (4) grid bias.

Since these adjustments interact, the final settings will be arrived at after some juggling. If the antenna is coupled too tightly, the oscillator will become unstable; but if feedback is increased by increasing the quench voltage, it can again be stabilized.

In most communications receivers the quench frequency is not

adjusted to the maximum usable frequency, since the gain is ample and selectivity is better at a lower quench frequency.

There is generally no necessity for automatic gain stabilization for the sets operating in the logarithmic mode, since most systems are supervised. Gain control is usually accomplished either by a grid-bias control or by a quench-voltage amplitude control.

A separate detector is not essential for communications receivers provided the quench frequency is kept sufficiently low to avoid multiple resonance effects.

In constructing a superregenerative receiver, it is important that good uhf techniques be employed with particular emphasis on shielding and r-f bypassing.

20·13. Self-quenching Circuits.—The superregenerative circuit may be made to operate self-quenched by using a short-time-constant *RC*-circuit in the grid.

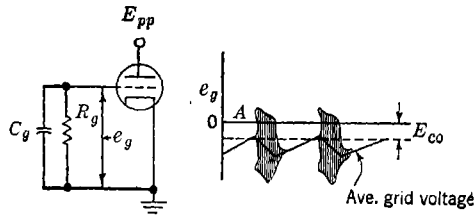


FIG. 20·14.—Self-quenching oscillator (r-f components omitted).

The fundamental circuit and resulting waveforms are shown in Fig. 20·14. Assume that the tube has started oscillating (point *A*). The positive swings drive the grid positive, cause grid-to-cathode conduction, and charge the condenser C_g. This causes the average grid voltage to decrease as shown by the heavy line. As the grid voltage decreases, a point is reached where oscillation cannot be sustained, and the oscillations die out leaving the grid at a negative voltage beyond E_{co}, the cutoff voltage. The condenser discharges through R_g, and the grid voltage rises to the "start oscillations" potential, and the cycle is repeated. The repetition frequency will depend on the time constant R_gC_g and can be increased or decreased by changing the value of C_g and R_g. With this type of quench the oscillations always reach the same amplitude and enclose the same area.

The presence of a signal voltage in the tank causes oscillations to start earlier, since the signal voltage adds to the rising bias voltage. Once oscillations are started, the envelope of oscillations includes the same area as for the case of no signals. Consequently the repetition rate of these bursts of oscillations increases with the amplitude of received signals, and

the change in output voltage is the result of this increased rate and not of increased area per burst of oscillations.

It is also possible to use a single tube as a combined superregenerative r-f oscillator and sine-wave-quench oscillator. This is accomplished by adding a suitable low-frequency tank circuit and feedback loop for the quench frequency. One circuit of this type is shown in Fig. 20·15. This method of self-quenching is considered to be superior to the use of a short-time-constant grid circuit.

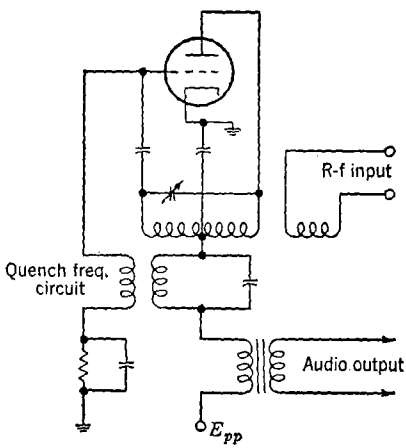


FIG. 20·15.—Combined superregeneration detector and sinusoidal quench-frequency oscillator.

20·14. Linear Mode—General. In applications of superregenerative circuits to pulse reception, most of the developmental work was on linear-mode superregenerative receivers. Many of the circuits evolved were for operation at frequencies considerably higher

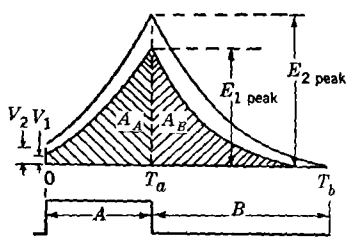


FIG. 20·16.—Growth and decay of oscillations for the linear mode.

than are encountered in ordinary communication work. Although this discussion on linear-mode superregenerative receivers will be concerned with high-frequency pulse reception circuits, the basic principles will apply equally well to communication circuits.

Gain for the Linear Mode.—Since oscillations never reach saturation in the linear mode, the amplitude of the envelope of oscillations varies with the strength of the received signal. The growth and decay of oscillations will follow the same exponential law as the logarithmic mode. Figure 20·16 shows the envelopes of linear-mode oscillations for two different signal amplitudes. A rectangular quench voltage is assumed, and the equations previously given for the logarithmic mode apply. For an initial voltage V_1, the voltage across the condenser becomes

$$e_c = V_1 e^{(R_T/2L)t} \tag{20·4}$$

for growth and

$$e_c = E_{1\text{ peak}} e^{\frac{-R}{2L}t} \tag{20·12}$$

for decay, where $R_T = (R - R_n)$ as before and $E_{1\,\text{peak}}$ is the peak voltage of the envelope as shown in Fig. 20·16.[1]

The area under the growth curve (A period) is

$$A_A = \int_0^{T_a} V_1 e^{(R_T/2L)t}\,dt = V_1 \frac{2L}{R_T}\left[e^{(R_T/2L)T_a} - 1\right] \tag{14}$$

and the peak voltage is

$$E_{1\,\text{peak}} = V_1 e^{(R_T/2L)T_a}. \tag{15}$$

The area under the decay curve (B period) becomes

$$A_B = V_1 e^{(R_T/2L)T_a} \int_{T_a}^{T_b} e^{\frac{-R}{2L}(t-T_a)}\,dt = V_1 \frac{2L}{R} e^{(R_T/2L)T_a}\left[1 - e^{\frac{-R}{2L}(T_b - T_a)}\right]. \tag{16}$$

The total area under the envelope is

$$A = A_A + A_B = V_1\left[\left(\frac{2L}{R} + \frac{2L}{R_T}\right)e^{\frac{R_T}{2L}T_a} - \left(\frac{2L}{R_T} + \frac{2L}{R} e^{\frac{R+R_T}{2L}T_a - \frac{R}{2L}T_b}\right)\right]. \tag{17}$$

Since the quantities within the bracket are constants, the total area can be written as

$$A = V_1(k_1 - k_2). \tag{17a}$$

For a quench frequency f_q the average linearly detected output is then

$$E_1 = f_q V_1(k_1 - k_2). \tag{18}$$

If the initial voltage is increased to V_2 (Fig. 20·16), the oscillation envelope starts at V_2 and reaches the higher peak value $E_{2\,\text{peak}}$ by the end of the A period, and the average linearly detected output becomes

$$E_2 = V_2 f_q(k_1 - k_2), \tag{19}$$

and

$$\text{Gain} = \frac{E_2 - E_1}{V_2 - V_1} = f_q(k_1 - k_2). \tag{20}$$

For pulse reception the detector often has a sufficiently long time constant so that the peak value of the oscillation envelope is obtained as an output. Under these conditions the gain becomes

$$\text{Gain} = \frac{E_{2\,\text{peak}} - E_{1\,\text{peak}}}{V_2 - V_1} = e^{(R_T/2L)T_a}. \tag{21}$$

[1] This derivation is based on the simplifying assumption that R_T is constant. Since the change in area under the oscillation envelope for the linear mode involves the entire growth of oscillations, the error introduced by this simplification is greater than is the case for the logarithmic mode.

Although the quench frequency does not enter directly into the gain formula for peak detection, it must be sufficiently high so that the detector output does not fall off too much between pulses.

It is noted that the quantity $e^{(R_T/2L)T_a}$ appears as the important factor in gain. Consequently, the condition for maximum gain involves a large value of R_T. This is opposite to the conditions for maximum gain in the logarithmic mode. It is important, however, that the relation between $R_T/2L$ and T_a be such that the oscillations do not reach saturation amplitude. If a receiver has been designed for linear-mode operation (high quench frequency and large negative resistance), a transition into the logarithmic mode by saturation oscillation leads to an extreme reduction in gain from that point on.

20·15. Output Characteristics for the Linear Mode.—Since the output voltage for linear detection is equal to the signal voltage times a constant $f_q(k_1 - k_2)$, the output is linear with respect to signal voltage. A plot of detector output voltage vs. input signal voltage is shown in Fig. 20·17. Since the output voltage is linear with respect to signal voltage, the linear mode of operation does not produce the undesirable "volume expansion" characteristic exhibited by the logarithmic mode. Consequently, the circuit will respond properly to communication signals with deep modulation.

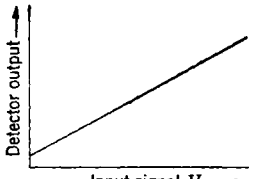


FIG. 20·17.—Variation of linear detector output for linear mode.

Quench Frequency.—In order to secure proper linear-mode operation, the quench frequency must be considerably higher than for the logarithmic mode. A rapid growth (large R_T) and a high quench frequency lead to maximum gain.

With the linear mode, the maximum usable quench frequency will depend upon the strongest signal to be received and the decay time constant $2Q/\omega$. It is essential that there be sufficient attenuation in the B period to dispose satisfactorily of the preceding oscillations before the next A period begins, or the output may not follow the modulation. To avoid the possibility of this difficulty, it is good design practice to allow a factor of safety in the B period by providing a decay time longer than would be indicated by calculations involving the maximum signal and $2Q/\omega$. Selectivity, as with the logarithmic mode, will be improved by extending the period of regeneration. If the bias is such that the A period is produced by the positive tips of the sine-wave quench voltage, use will be made of the reduced slope near the peak of the sine wave. In this way the regenerative period can be lengthened for the high quench frequencies employed in linear-mode receivers. Use of the sine-wave tips for the A period will, of course, result in a longer B than A period,

but this condition is generally required for linear-mode operation anyway.

For pulse reception the tube is so biased that the oscillation envelope is small under no-signal conditions. The use of the reduced slope portion of the sine-wave quench for the A period allows some time for the signal to be amplified by regeneration, so that the signal voltage, added to the rising quench voltage, can cause a greater advance in the starting time of oscillations and produce a larger amplitude of oscillations.

The minimum usable quench frequency for pulse reception is determined by the duration of the shortest r-f pulse to be received and the permissible "jitter" in the output. If, as in Fig. 20·18*a*, the period of the quench is less than the duration of the r-f pulse, an output is assured for every received pulse. On the other hand, if the period is longer than the r-f pulse, as in Fig. 20·18*b*, output is dependent upon the phase of the quench voltage with respect to the received r-f pulse.

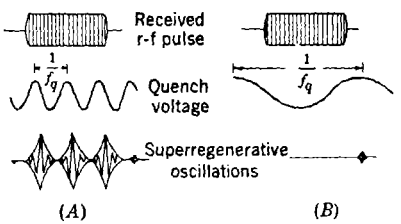


FIG. 20·18.—Effect of quench frequency on pulse reception.

Since this phase relation is entirely random, the leading edge of the output pulse will vary from pulse to pulse by approximately $1/f_q$ and the output will consist of a group of overlapping pulses. Although it is true that the most forward pulse in the group starts at the leading edge of the received pulse, this condition is not desirable, and jitter should be held to the minimum, consistent with other design considerations.

Although a high quench frequency is very desirable from the standpoint of jitter, there are other considerations that limit the maximum quench frequency. As the quench frequency is raised, the A period is shortened, and oscillations must grow more rapidly if ample sensitivity is to be attained. This means that $R_T/2L$ must be increased by use of either a higher Q tank or increased feedback. The circuit becomes more critical as the rate of growth is increased. Slight changes in operating conditions, such as voltages or loading, can cause the receiver to be insensitive, so that weak signals will be missed. If the change in conditions is in a direction to make the receiver more sensitive, saturation oscillations may result. In beacon systems where the receiver output is used to trigger a transmitter, false replies may be produced. The selection of a quench frequency for pulse reception becomes a compromise among sensitivity, stability, and jitter. Table 20·2 (page 571) gives the operating frequency, pulse duration, and quench frequency for a number of superregenerative pulse receivers.

Detection.—Detection may be performed by the superregenerative tube as in logarithmic-mode receivers. In applications where the output is used to trigger a transmitter and an automatic gain stabilization circuit is not required, plate-circuit detection has proved very satisfactory. The sudden increase in current engendered by the receipt of a signal provides a negative pulse in the plate circuit that can be coupled to the transmitter-triggering circuit.

A separate detector is employed when automatic gain stabilization is required and also when the output pulse is to be displayed on an indicator.

The most common type of separate detector is shown in Fig. 20·19. The coupling condenser C is small (usually a few micromicrofarads), to reduce the loading on the tank circuit and to provide a short-time-constant charge circuit. The diode is a low-resistance path when its plate is

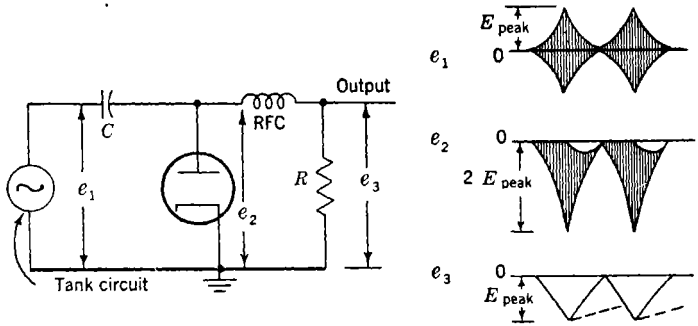


FIG. 20·19.—Separate detector for superregenerative amplifier.

positive and so prevents the right-hand side of the condenser from going positive. Consequently, the positive excursions of the r-f oscillations e_1 charge C to the full positive peak value of voltage. As the oscillations leave their positive peaks, the plate of the diode is at a lower voltage than its cathode, and any discharge must occur through the load resistor R. Since the time constant RC is long compared with the period of the r-f oscillations, there is a negligible discharge between positive peaks. The voltage across the diode is shown by e_2, and the average d-c voltage at the plate of the diode is a negative voltage equal to the peak of the r-f wave. The choke removes the radio frequency, and the average d-c voltage e_3 appears across the load resistor. Once the r-f envelope has passed its peak, the positive peaks of the r-f oscillations are smaller than the voltage on the condenser and there is a continuous discharge through R. If the discharge time constant RC is sufficiently short, the condenser will completely discharge between bursts of oscillations, as shown by the solid line in e_3. Filtering can be accomplished by increasing the time constant,

as shown by the broken line. If the discharge time constant is made too long, a single output signal may be secured for two pulses that are close together in time. It may also distort the trailing edge of the output pulse to such an extent that it is difficult for the operator to distinguish between different width pulses displayed on an indicator. Sometimes a bias voltage is placed in the cathode to prevent output from the small oscillations initiated by noise.

20·16. Automatic Gain Stabilization (AGS) Circuits.—The maintenance of uniform sensitivity becomes a problem when superregenerative receivers are operated without supervision. In airborne installations the problem is aggravated by the wide variation in operating conditions. Consequently, AGS circuits are incorporated in practically all unsupervised airborne superregenerative receivers.

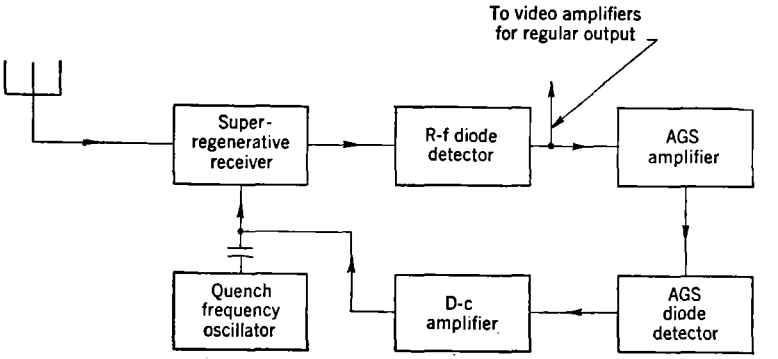


FIG. 20·20.—Block diagram of automatic gain stabilization.

The criterion for sensitivity in a linear-mode superregenerative receiver is the amplitude of the noise-initiated oscillations which are present under no-signal conditions. The AGS circuit regulates the amplitude of these oscillations by controlling either the bias or the quench-voltage amplitude. Figure 20·20 is a block diagram for a typical AGS circuit. Noise-initiated oscillations are detected by the r-f diode and applied to the AGS amplifier. The amplified pulses are detected, filtered, and converted into a d-c grid voltage for the superregenerative receiver tube. Polarities are such that an increase in amplitude of the noise pulses increases the grid bias on the superregenerative tube and vice versa. Quench voltage is applied to the grid through a condenser, so that the d-c grid voltage is determined entirely by the AGS circuit.

The receiver-AGS portion of a 175-Mc/sec radar beacon is shown in Fig. 20·21. Noise-initiated pulses of radio frequency occurring at a 300-kc/sec (f_q) repetition rate are detected by the diode detector that produces negative output pulses. These negative pulses are applied to

the grid of the AGS amplifier through a short-time-constant coupling circuit. The transformer in the plate of the AGS amplifier is tuned to the quench frequency, 300 kc/sec, by means of movable iron slugs in the primary and secondary windings. Since the pulses occur at 300 kc/sec, the amplification is greatly increased by this tuned circuit. The AGS diode produces a negative output across the 0.22-megohm load resistor. The diode output is then filtered by the low-pass filter network (consisting of the 0.10-megohm resistor and associated condensers) before

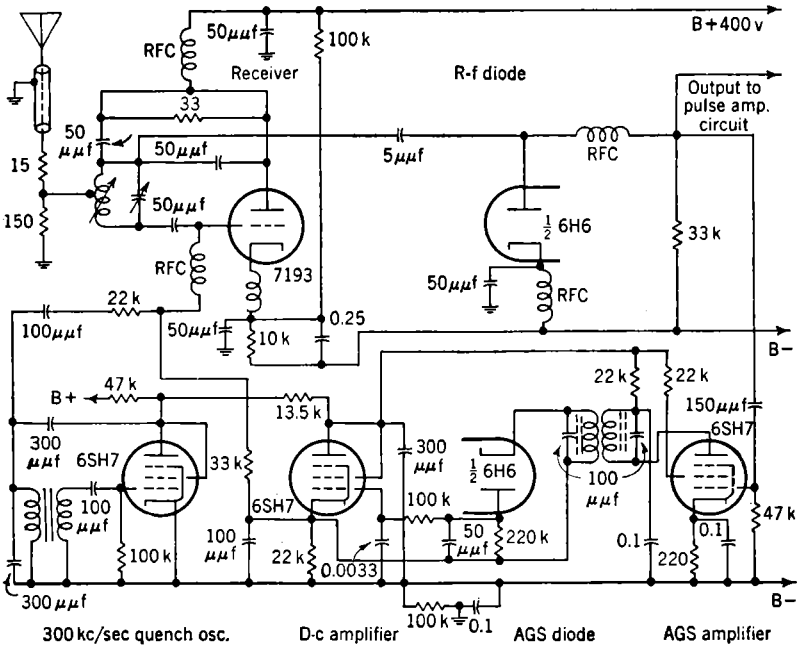


FIG. 20·21.—175-Mc/sec superregenerative receiver with automatic gain stabilization circuits.

application between grid and cathode of the d-c amplifier. By applying the signal between grid and cathode, an amplified signal can be taken from the cathode, which is at a more reasonable voltage level than the plate. (If the signal had been applied between grid and B−, the d-c amplifier would be a cathode follower with a gain less than unity.) With this connection (often called "bootstrap") the AGS diode and associated circuits move up and down with the level of the output and so must be isolated from B−. The 100-μμf condenser across the 22,000-ohm cathode resistor provides additional filtering, so that the voltage applied to the grid of the superregenerative receiver is practically pure direct

current. The 100-$\mu\mu$f coupling condenser serves as a d-c block between the 300-kc/sec quench oscillator and the grid of the 7193 receiver tube. The AGS circuit is very sensitive. A change of 1 mv in the amplitude of the input pulses to the AGS amplifier will produce a change in AGS output voltage of approximately 5 volts. The time constants in the circuit are such that there is no tendency toward hunting.

The tuned circuit associated with the receiver tube also serves as the tank circuit for the transmitter in the beacon. When a beacon-interrogating pulse is received and the transmitter fires, large amplitude r-f oscillations exist in the tank and are detected and applied to the AGS amplifier circuit. These transmitted pulses have a maximum duration of approximately 20 μsec; the recurrence rate (rate of interrogation) is less than 500 cps. Accordingly, under the maximum conditions, a 20-μsec transmitted pulse occurs every 2000 μsec. For the other 1980 μsec between transmitted pulses, 600 noise pulses, spaced at intervals of $3\frac{1}{3}$ μsec, occur. Since there is only one transmitted pulse of large amplitude to 600 noise pulses, the effect on the AGS circuit is negligible. In addition, the short-time-constant coupling circuit in the grid of the AGS amplifier differentiates the long pulse to reduce its duration.

20·17. Single-cycle Superregenerative Receivers.—The single-cycle receiver, as its name implies, has a single period of sensitivity instead of recurrent sensitive periods. The receiver is made sensitive by a pulse from a control circuit and so will not respond to an r-f pulse unless it has the proper time relation to this control pulse.

A single-cycle superregenerative receiver which was used in an airborne beacon is shown in Fig. 20·22. The superregenerative receiver, formed by V_2 and its associated circuits, is tuned to a frequency of 200 Mc/sec. Positive sensitizing pulses are secured from the multivibrator circuit of $V_{1a\text{-}b}$.

V_{1a} is normally conducting, while V_{1b} is normally cut off by the positive bias applied to its cathode from the voltage divider made up of the 470,000- and 27,000-ohm resistors. When a control pulse is received, a negative pulse (waveform *a*) is applied to the grid of V_{1a}. This cuts V_{1a} off, and an increase in voltage appears at its plate. This increase in voltage elevates the grid of V_{1b}, causing it to conduct and driving its plate negative as shown by waveform *b*. The 50-$\mu\mu$f condenser couples the change in plate voltage of V_{1b} back to the grid of V_{1a} and holds V_{1a} cut off until the condenser discharges. As soon as the condenser has discharged sufficiently to allow V_{1a} to conduct, its plate voltage falls and cuts off V_{1b}. The cycle occupies approximately 5 μsec. Waveform *b* is differentiated by the circuit between the plate of V_{1b} and ground. A portion of this differentiated waveform is secured at the arm of the 1000-ohm sensitivity control and is applied to the grid of V_2 as a sensitiz-

ing pulse. Waveform *c* is the voltage at the arm of the sensitivity control when no 200-Mc/sec signal is present to trip V_2. Since this waveform does not go positive until approximately 4 μsec after the start of the waveform, there is this much delay between the leading edge of the control pulse and sensitivity in the 200-Mc/sec receiver.

The 39,000-ohm cathode resistor self-biases V_2 so that its quiescent gain is less than unity. The sensitivity control is adjusted so that the sensitizing pulses alone will not cause the circuit to oscillate. If a 200-Mc/sec signal of sufficient amplitude is in the tank, the tube will burst into oscillation near the most positive point of the sensitizing pulse.

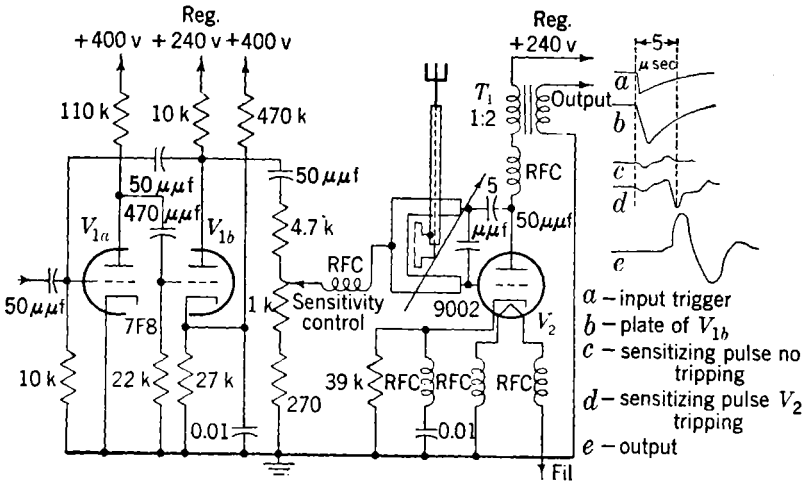


FIG. 20·22.—Single-cycle superregenerative receiver.

Once the circuit starts oscillating, it continues to oscillate until the cathode voltage rises to a point where the gain is insufficient to sustain oscillations. After oscillation ceases, the 0.01-μf cathode-bypass condenser must discharge before the circuit can be tripped again. The voltage at the arm of the sensitivity control with the circuit tripping is shown by waveform *d*. The negative swing during oscillation is due to grid current. The sudden increase in plate current, which occurs when the circuit oscillates, develops the output voltage shown by waveform *e*. Since the amplitude and duration of the oscillations is controlled by the circuit and not by the level of the input 200-Mc/sec signal, the output amplitude is the same for any signal strong enough to trip the circuit.

The amplitude of the sensitizing pulse must be such that it alone will not cause oscillations; yet it must bring the circuit close to oscillation if the receiver is to have proper sensitivity. This means that circuit conditions must be carefully controlled. Since AGS, as previously

described, cannot be applied to this system, control is accomplished by the use of well-regulated voltages.

The superregenerative circuit is in the form of a Colpitts oscillator using a capacity-shortened quarter-wave line as a tuned circuit. Figures

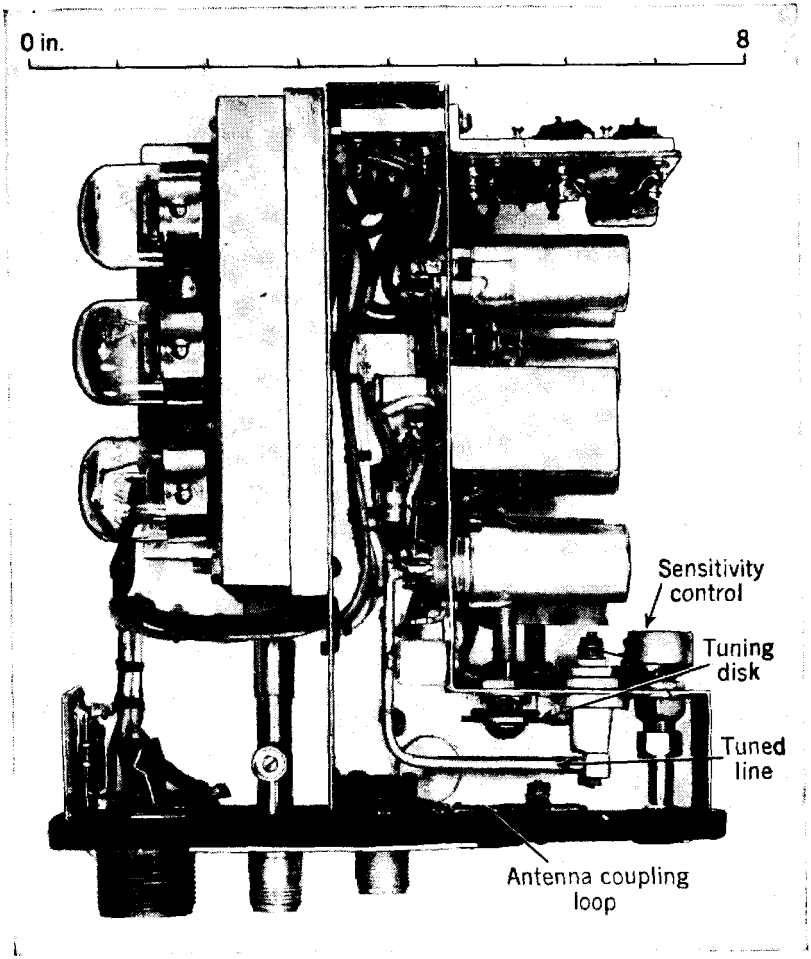


FIG. 20·23.—Complete beacon incorporating single-cycle superregenerative receiver.

20·23 and 20·24 show the general construction of the tuned line and arrangement of circuit elements. Tuning is accomplished by the inductive disk. The entire beacon assembly (exclusive of antennas) is only 8 by 7 by $2\frac{7}{8}$ in.

The sensitivity of this receiver is equal to that of sine-wave-quench superregenerative receivers operating at the same frequency. The features of simplicity and controlled time of sensitivity make this receiver adaptable to many pulse reception applications.

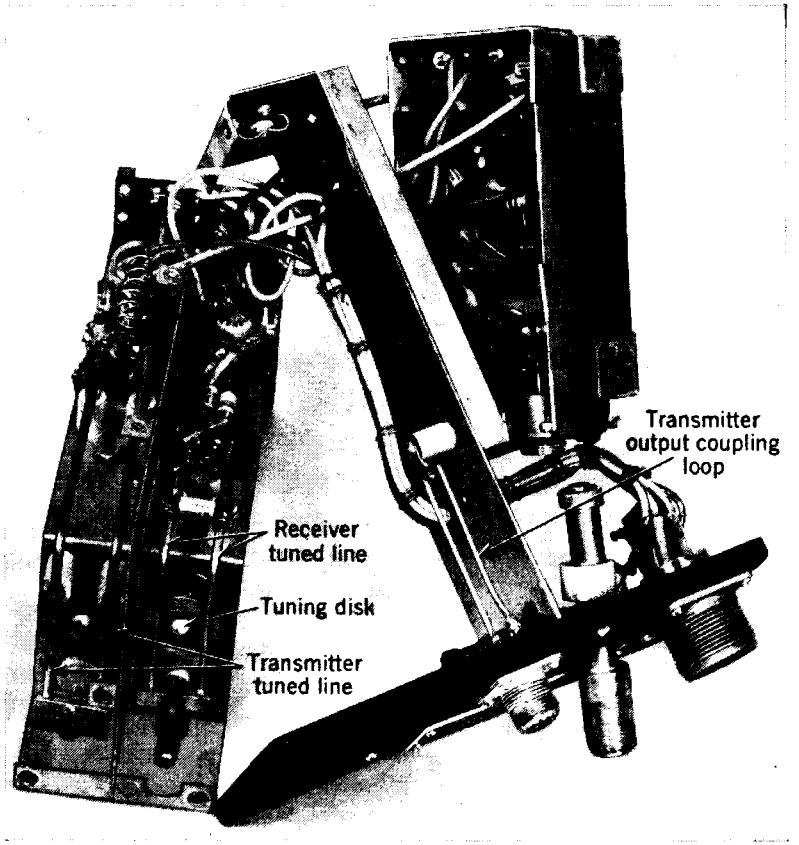


FIG. 20·24.—Exploded view of receiver.

20·18. Actual Receivers.—Table 20·2 lists six receivers that were employed in radar beacon systems. Although there were a great many more types in use than the six listed, this group covers the frequencies for which superregenerative receivers were employed, and the circuitry is fairly representative of all receivers operating at these frequencies. The sensitivities are not the ultimate obtainable at the listed frequencies. They were dictated by the specifications for the particular application.

Table 20·3 lists a group of experimental receivers that were con-

structed and tested to determine the characteristics of superregenerative receivers at higher frequencies. These were never utilized in any beacon systems.

TABLE 20·2.—REPRESENTATIVE RECEIVERS USED IN BEACON SYSTEMS

Approx. freq., Mc/sec	Sensitivity, watts	Tube type	Tuned circuit	Quench frequency, kc/sec	Duration of received pulses, μsec	Circuit diagram
175	5×10^{-11}	7193 triode	Lumped *LC*	300	9.0	Fig. 20·21
200	5×10^{-11}	9002 triode	Line	Single cycle	5.0	Fig. 20·22
700	2×10^{-11}	6F4 triode	Cavity	625	1.5	Fig. 20·25
700	2×10^{-11}	446B triode	Cavity	1000	1.5	Fig. 20·26
950	8×10^{-10}	6F4 triode	Cavity	900	1.2	Fig. 20·25

TABLE 20·3.—EXPERIMENTAL RECEIVERS

Freq., Mc/sec	Sensitivity, watts	Tube type	Tuned circuit	Quench frequency, kc/sec	Circuit diagram
650	8×10^{-13}	446B triode	Cavity	1800*	Fig. 20·30
1,000	1×10^{-12}	446B triode	Cavity	4500*	Fig. 20·30
3,000	1×10^{-10}	446B triode	Cavity	3400*	Fig. 20·30
3,000	5×10^{-10}	707A vel. mod.	Cavity Part of tube	1000	Not shown
10,000	5×10^{-10}	723A/B vel. mod.	Cavity Part of tube	1000	Fig. 20·33

* These quench frequencies were the maximum usable frequencies.

The receiver shown in Fig. 20·25 was employed in a lightweight portable beacon. The complete assembly including batteries weighed only 40 lb. The cavity was developed for this particular receiver and with slight modification was usable at the two required operating frequencies of approximately 700 and 950 Mc/sec. Construction details of the cavity are shown in Fig. 5·28. The cavity is essentially a short-circuited section of coaxial line as far as radio frequency is concerned, and the circuit can be represented as a Colpitts oscillator if the interelectrode capacitances are taken into account. The Q of the tuned circuit is approximately 250.

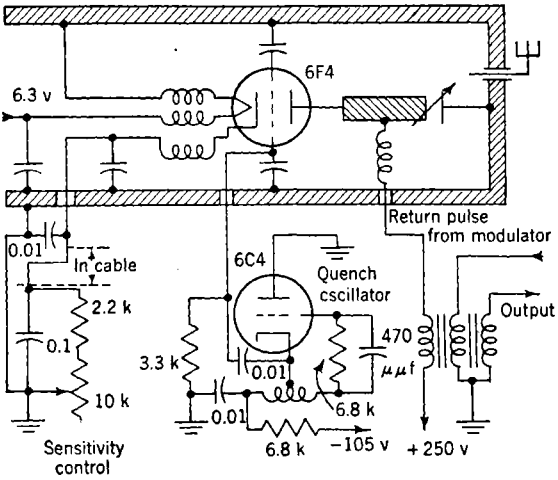


FIG. 20·25.—650- to 950-Mc/sec superregenerative receiver.

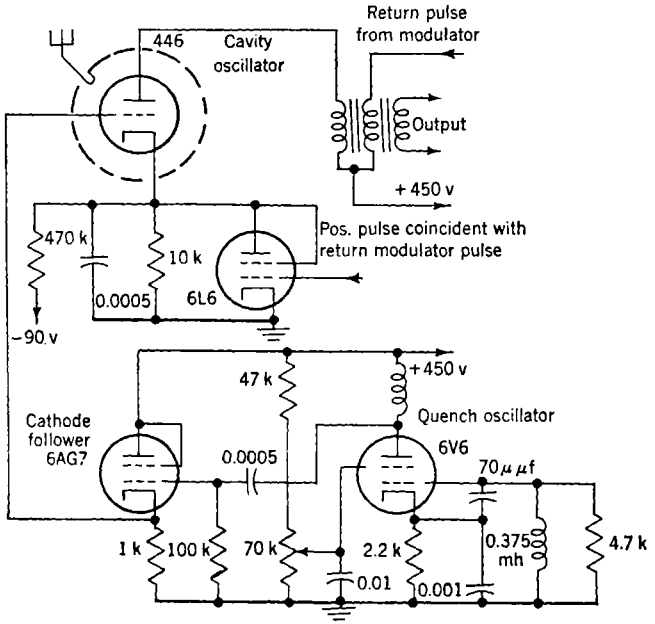


FIG. 20·26.—700-Mc/sec superregenerative receiver.

The quench oscillator is a Hartley circuit and delivers 25-volt peak-to-peak quench voltage to the grid of the 6F4. The quench frequencies employed are 625 kc/sec for the 700-Mc/sec receiver and 900 kc/sec for the 950-Mc/sec receiver. Since the quench-voltage amplitude is constant, sensitivity is adjusted by varying the cathode self-bias of the 6F4 receiver tube. The pass band of the receiver is 6 Mc/sec.

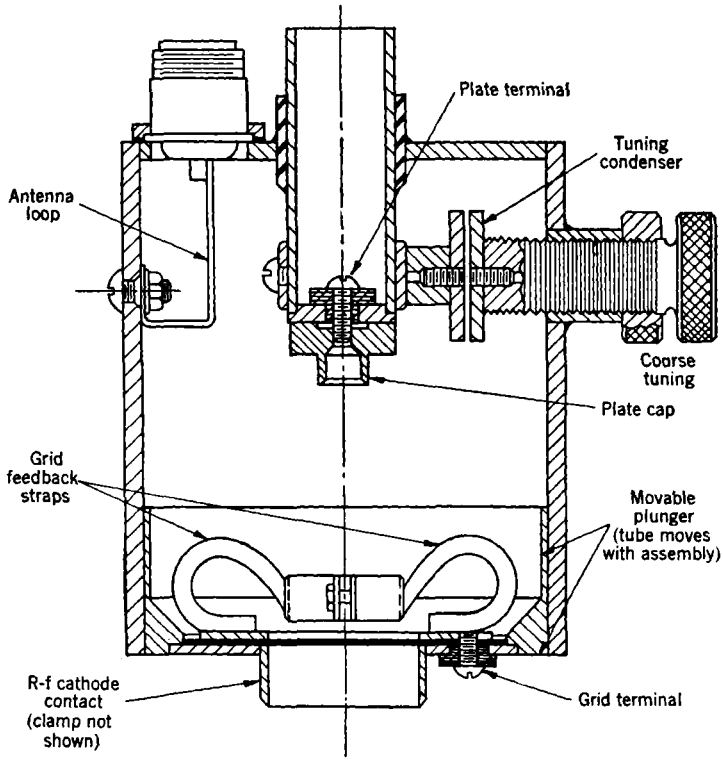


FIG. 20·27.—Construction of 700-Mc/sec cavity.

It is of particular interest that the 6F4 triode is used as both a receiver and a transmitter. The received r-f pulses produce positive output pulses at the secondary of the transformer in the plate of the 6F4. These are amplified and used to trigger a gas-tube modulator circuit which delivers 0.75-μsec voltage pulses back to the plate transformer. In this way a positive pulse of 600 volts lasting for 0.75 μsec is applied to the plate of the 6F4, which bursts into strong oscillation to produce the r-f reply pulse.

The receiver shown in Fig. 20·26 was also used as a lightweight

portable beacon. The transmitter-receiver section of the complete beacon is shown in Fig. 20·29. The over-all weight including batteries was 105 lb, and this model was superseded by the lighter-weight equipment of Fig. 20·25.

It is noted that the circuit is very similar to that shown in Fig. 20·25. The GL446B (lighthouse tube) is grid-quenched by a 1-Mc/sec sinusoidal quench voltage produced by an electron-coupled oscillator. The cathode follower between the quench oscillator and the GL446B provides

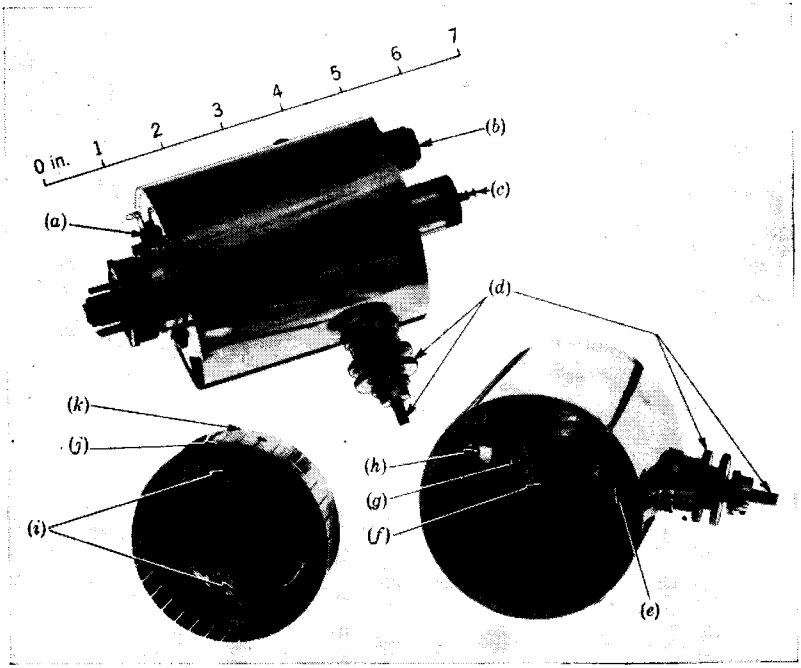


FIG. 20·28.—Cavity for 700-Mc/see beacon.

a low-impedance path for grid current while transmitting. Sensitivity is controlled by regulating the amplitude of the quench voltage. This is accomplished by varying the screen voltage of the quench oscillator.

The cavity oscillator uses grid straps for feedback as shown by the sectional drawing of Fig. 20·27. The GL446B tube, grid straps, and clamp are clearly visible in the photograph of the plunger assembly. The tube is clamped into the plunger assembly which is then slid into the chamber until proper contact is made with the plate cap. Plate and grid bypass condensers are built into the cavity, mica being used as a dielectric. The d-c connection to the grid is made by a soldering lug and conductor through an insulating bushing. The high-frequency cathode

connection is made by contact between the plunger and the shell. Since the GL446B tube incorporates a built-in condenser between the shell and the cathode, this connection does not provide a d-c path. The d-c cathode connection is made to a pin in the base.

When a pulse is received, the modulator circuit supplies a 1.5-μsec 600-volt pulse to the plate of the GL446B which is added to the normal

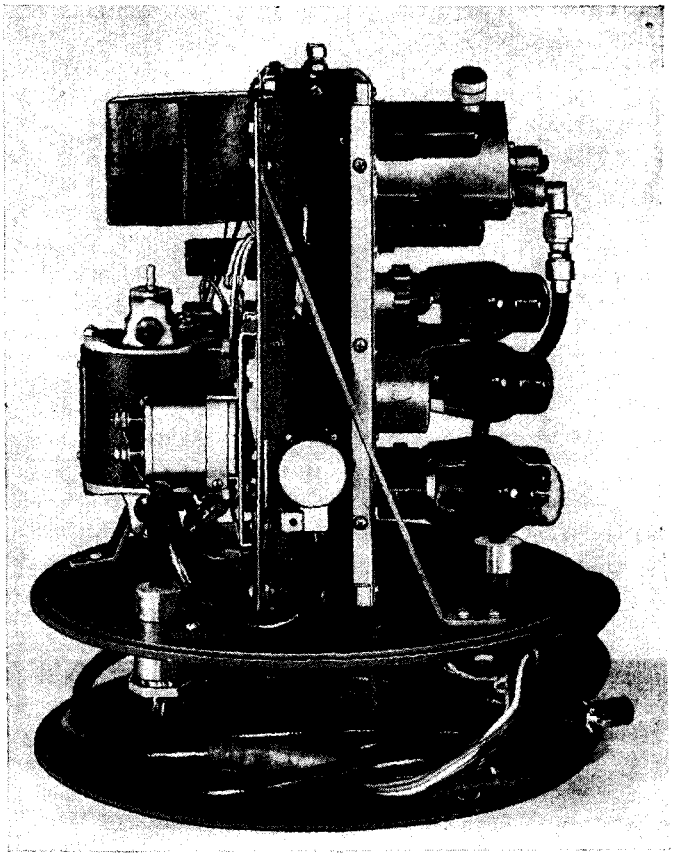


FIG. 20·29.—Complete 700-Mc/sec beacon.

plate-supply voltage of 460 volts to produce the reply pulses. In order that the tube should not develop an excessive cathode bias during transmission, the cathode-pulser tube (which is biased beyond cutoff during reception) is made conductive by a simultaneous 1.5-μsec pulse. The pass band of the receiver is 6 Mc/sec. Figure 20·29 is a photograph of the receiver.

The receiver shown in Fig. 20·30 was built to investigate the operation of the lighthouse tube (GL446B) in superregenerative circuits for wave-

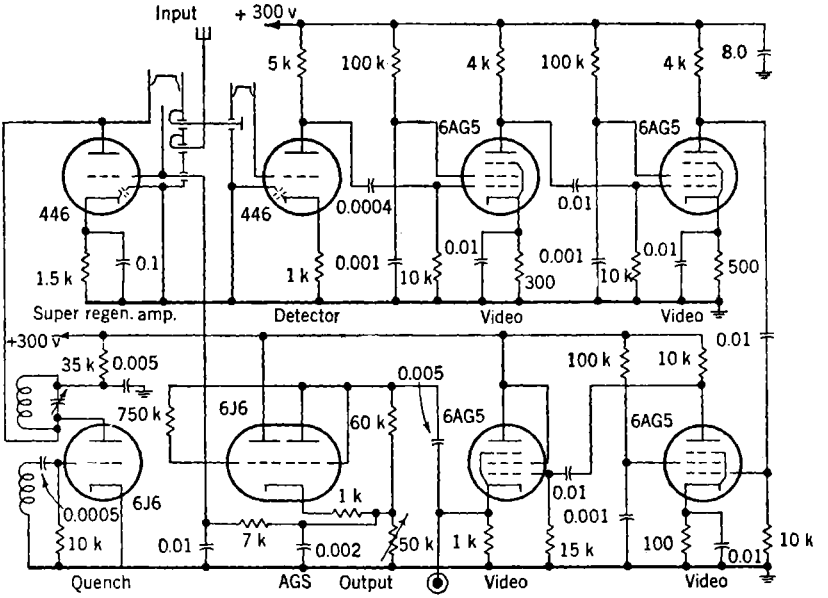


FIG. 20·30.—3000-Mc/sec superregenerative receiver.

lengths of 10, 30, and 45 cm and to determine the highest possible quench frequency at each wavelength. Measurements of sensitivity were made with c-w signals.

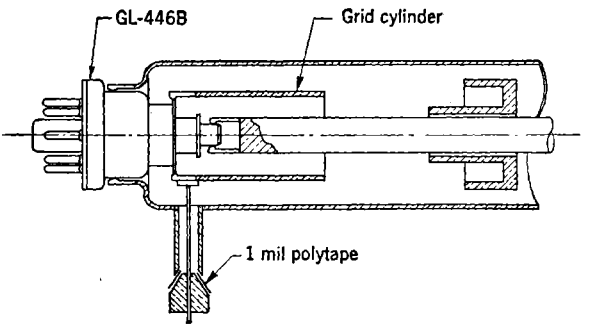


FIG. 20·31.—Cavity for 3000-Mc/sec receiver.

The cavity shown in Fig. 20·31 was used for the 10-cm receiver. Tuning was accomplished by moving the plate fingers on or off the plate contact of the tube. The plate choke was then moved into a position where the tube oscillated. The second cavity, consisting only of a grid-

cathode cavity, is the detector. A crystal may be used in place of the lighthouse tube detector. This method of detection (separate detector) was used in preference to self-detection because it is better for removing the quench frequency.

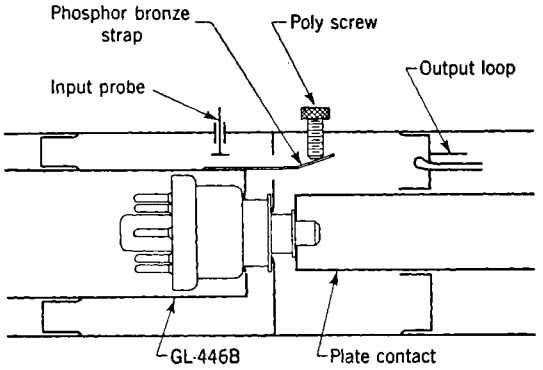


FIG. 20·32.—Cavity for 650- to 1000-Mc/sec receiver.

Plate quenching was found to be superior to grid quenching. It gave steadier operation and made possible the use of higher quench frequencies. Bias for the superregenerative tube was secured by the combination of cathode self-bias and grid bias from the AGS system. Although one stage of video amplification and one cathode-follower stage provided enough gain to secure output signals of 25 volts, it was necessary to add the two additional video stages to secure sufficient noise amplification to operate the AGS circuit properly.

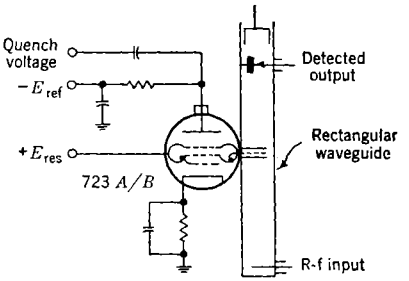


FIG. 20·33.—Schematic diagram of experimental 10,000-Mc/sec superregenerative receiver.

The AGS circuit operates in the manner previously described, one half of the 6J6 serving as an AGS detector and the other as a d-c amplifier. The variable 50,000-ohm resistor provides a means for adjusting the AGS output level and thus the sensitivity.

The cavity assembly for the 30- to 45-cm receiver is shown in Fig. 20·32. It consists of two coaxial cavities, one for the plate-grid line, the other for the grid-cathode line. Capacitive feedback is secured by the adjustable strap extending through the hole in the grid fingers to the plate cavity. The detector cavity is similar to the one used at 10 cm except that it is larger in size. This receiver was tested without AGS. Plate quenching and cathode biasing were employed.

Sensitivity was measured by determining the minimum detectable signal. This accounts for the apparent increased sensitivity of this receiver over others listed in Table 20·2 where sensitivities were determined by pulsed signals tangential to the noise.

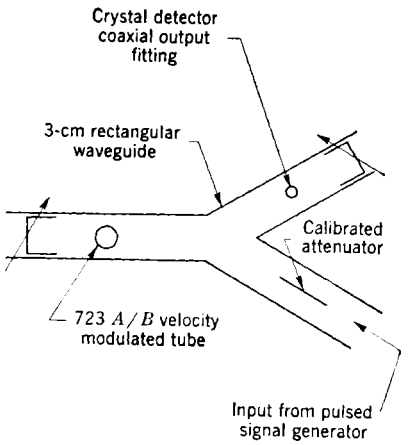


FIG. 20·34.—Diagram of experimental 10,000-Mc/sec superregenerative receiver.

The maximum usable quench frequency was determined by observing the quenching pattern at different quench frequencies. It was possible to determine the point of maximum quench frequency accurately as the sensitivity dropped off rapidly above this frequency. Quench voltages as high as 200 volts rms were tried to see if this would increase the maximum usable quench frequency, but this had no effect. The high quench frequency attainable at 30 cm was attributed to a greater range of feedback secured at this frequency with the particular cavity employed.

The experimental setup shown in Figs. 20·33 and 20·34 was made to ascertain the usability of velocity-modulated tubes for high-frequency superregenerative receivers. Sensitivity did not seem to be dependent upon the exact mode of oscillation. Some representative test data for 10,000 Mc/sec are given below. Quenching was with small-amplitude

TABLE 20·4.—TEST DATA FOR 10,000-MC/SEC RECEIVER

e Resonator, volts	e Reflector, volts	i Cathode, ma	Sensitivity, watts
+245	−115	17	5×10^{-10}
+200	−180	18	6×10^{-10}
+100	−100	7	7.7×10^{-10}

positive pulses (less than 10 volts) of durations between 0.25 and 1 μsec. Repetition rates were varied from approximately 1 to 2.5 Mc/sec. A 15,000-ohm cathode resistor was employed.

Similar measurements have been made with velocity-modulated tubes at 10 cm (3000 Mc/sec), and the sensitivities obtained were approximately the same as at 10,000 Mc/sec. It was thought that this was due to the fact that the principal noise contribution in velocity-modulated tube circuits is from the tube itself and that the noise level of the 10-cm tube was approximately equal to the noise from those operating at 3 cm.

CHAPTER 21

MTI RECEIVERS

By W. Selove

21·1. Introduction.—This chapter contains a description of receivers for use in coherent-phase MTI (moving target indication) systems.[1] Such systems, as their name implies, are used for the detection of moving objects by radar means, particularly under circumstances where the presence of a large number of stationary scattering objects near the moving object of interest makes its detection difficult. As ordinarily applied to pulse radar systems, this process involves the comparison of the echo pattern for a given area on successive transmitted pulses. This comparison is usually accomplished by delaying the echo-signal output of the receiver for a time equal to the interval between pulses and then subtracting this delayed signal from the undelayed output of the receiver resulting from the next pulse. Any parts of the echo signal, such as those due to fixed objects, that are the same from one pulse to the next will be eliminated or canceled. The whole comparison process has accordingly come to be known as "cancellation." Circuits and apparatus for accomplishing this will not be treated here, this discussion being limited to the special features that must be incorporated into the receiver proper to give best results in such service.

Special Features of MTI Receivers.—MTI receivers usually differ from normal radar receivers in three respects: (1) The local oscillator has much greater stability; (2) suitably phased c-w oscillations—*coh*erent *o*scillator (coho) signals—are introduced in the receiver to provide sensitivity to the phase of echo signals; and (3) the amplitude response of the receiver is specially designed to give maximum discrimination between signals from moving targets and similar signals from other causes.

In some MTI applications not all of these features are necessary. In fact, it may be possible to use an ordinary receiver with only slight modifications, such as the addition of shielding and power-supply filtering for the local oscillator and the injection of coho signal in some simple manner. In general, however, best MTI performance will be obtained if the MTI receiver is designed for its job.

Coho.—In a coherent-phase MTI system, moving targets are detected by measuring the change in phase difference between the transmitted

[1] MTI systems are discussed in Vol. 1, Chap. 16, of the Radiation Laboratory Series.

and received pulse in successive repetition intervals. The phase difference is measured by comparing the phase of the echo signal with the phase of a c-w signal that is locked in phase with the transmitter by each transmitted pulse. This locking is called "phase-coherence," and the oscillator that is so locked is called the "coherent" oscillator, or coho. Coho circuits and other details are discussed in Sec. 21·8.

In the process of frequency conversion in the mixer of a superheterodyne receiver, relative phase is preserved. Therefore if an oscillator operating at intermediate frequency is brought into phase coherence with the beat signal between the transmitter and the local oscillator, subsequent echo signals, after beating with the local oscillator, may be compared with the coho to give the same result that would be obtained if the coho operated at the transmitter frequency. This procedure is almost universally used for microwave MTI systems because it permits the complicated circuits required for locking to operate at a more convenient frequency than would be the case if a microwave oscillator had to be locked.

The signals from stationary targets, "permanent echoes," have the same phase relative to the transmitter pulse in successive repetition periods. In this respect permanent echoes, for example, echoes from ground "clutter," are similar to the coho signal. Even without a coho, therefore, a moving target can be detected by an MTI system if signals are received simultaneously from the moving target and from fixed targets. Such simultaneous signals may quite easily be obtained because the radar beamwidth is not zero. Thus if a large part of the indicator scope face represents regions from which ground clutter signals are received, satisfactory MTI operation may be possible without a coho (and also without an unusually stable local oscillator). This condition may exist, for example, in the case of an airborne system.

Use of Separate Preamplifier.—Most of the MTI receivers used at the Radiation Laboratory have had separate preamplifiers, the main part of the receiver having signals applied to it at intermediate frequency rather than radio frequency. This construction offers the quickest and most convenient way to attach MTI to a system in such a way as to permit either normal or MTI operation. The use of a separate preamplifier also eases the shielding problems in the MTI receiver inasmuch as less gain is required in the main part of the receiver. These shielding problems are generally more severe than in an ordinary receiver because of the greater complexity of circuits for MTI.

Gain-bandwidth Requirements.—An ordinary receiver requires enough i-f gain to amplify weak signals to a level at which they produce approximately linear response in the second detector. When a c-w coho signal is applied, however, the detector's incremental response is linear even for

infinitesimal signals, and the required i-f gain is determined by a different consideration, namely, to what fraction of the i-f limit level it is desired to amplify weak signals. The maximum value of the desired fraction is usually between one-half and one; the gain required is generally somewhat greater than for an ordinary receiver.

If in an ordinary radar receiver the i-f bandwidth is too great, the signal-to-noise ratio will be impaired even though the video bandwidth is suitably narrow. When a c-w signal of large amplitude compared with noise is present at the second detector,[1] however, this is no longer true, and bandwidth-narrowing in the amplifiers following the second detector has the same effect on signal-to-noise ratio as in the i-f amplifiers preceding the second detector. Hence, whereas in an ordinary receiver if the i-f bandwidth is chosen for maximum sensitivity, the following amplifiers must then have a greater effective bandwidth than that of the i-f amplifier in order to obtain good pulse definition, in an MTI system the bandwidth can be distributed more economically by making all stages in the entire system have the same effective bandwidth. This is particularly advantageous, because an MTI system contains many more cascaded stages than an ordinary radar system. Thus the i-f amplifier of an MTI receiver generally has greater bandwidth than a normal receiver for the same pulse length.

Different Types of MTI Receiver.—Three different types of receiver have been used on MTI systems at the Radiation Laboratory, the *lin-log, limiting,* and *IAGC* types. All have given apparently satisfactory (although not quantitatively evaluated) performance. The following sections contain a description of the theoretical and practical considerations that led to the use of the various receiver types and an evaluation of their characteristics.

21·2. General Requirements and Limitations. *Sensitivity to Moving Targets.*—Receivers for MTI service must have high sensitivity to moving targets in two distinctly different sets of circumstances—to signals in the clear and to signals in clutter. Sensitivity to weak signals in noise (that is, in the clear) will be the same for all receivers that respond similarly to noise and weak signals; the particular receivers discussed here all fall in this category. The maximum usable sensitivity to signals in clutter will in general be limited by fluctuations[2] of the clutter itself due to clutter fading, scanning, noise, and equipment instability.

[1] For this type of operation this circuit element might more properly be called a "converter." However, in this discussion the more familiar term will be retained.

[2] Two kinds of fluctuation are of importance: (1) fluctuations from pulse width to pulse width, that is, as a function of range, for a particular transmitter pulse, and (2) fluctuations from repetition period to repetition period for the echo signal from a particular range. The first kind will be referred to as *r*-fluctuations (range fluctua-

Clutter Fluctuations.—Theoretical and experimental investigation has shown[1] that clutter t-fluctuations produce a moving-target effect that may be represented in terms of a *fluctuation vector* varying randomly with respect to a *fixed clutter vector*, and that for a given type of terrain the rms amplitude of this fluctuation vector (exclusive of noise) is directly proportional to rms clutter amplitude. For ground clutter under typical conditions the fluctuation may be 40 to 50 db below rms clutter amplitude. For airborne systems the predominant fluctuations of ground clutter are likely to be due to "beating" among various targets which give simultaneous return because of the nonzero beamwidth and are likely to be relatively large compared with the ground-clutter fluctuations obtained with a stationary system. The fluctuations from sea return and clouds are usually large compared with ground-clutter fluctuations. In other words the echo pattern from a stable collection of scattering objects, such as those making up ground clutter, is very nearly the same from one pulse to the next, whereas a collection of relatively mobile scattering centers gives a more variable echo pattern.

Fluctuation Compression for Strong Signals.—It will often be true that fluctuations on strong clutter will be larger at the receiver input than noise fluctuations. For maximum sensitivity to signals both in the clear and in clutter, it is necessary that these large clutter fluctuations be compressed in the receiver so that at the receiver output they produce about the same size variations as noise. This requirement is imposed by the fact that, to be detected, a desired signal must cause an output variation at least as great as the largest undesired variations. Thus, if clutter fluctuations larger than noise are allowed to produce correspondingly large output fluctuations, either they will show up as spurious moving targets or else indicator sensitivity must be reduced to the point that the clutter fluctuations do not show, in which case weak signals (of the order of magnitude of noise) also will not show. Similarly, if clutter fluctuations are excessively compressed, there will be undue loss of sensitivity to signals in clutter.

For large signals, therefore, output fluctuation should be constant for a constant *percentage* input fluctuation, regardless of rms input; such a response is logarithmic in nature. However, compression is undesirable for weak signals. In fact, it should not be applied to clutter and signals unless the clutter is at least so large that clutter fluctuations (at the receiver input) are the same size as fluctuations due to noise. Thus the

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tions); the second as t-fluctuations (time fluctuations). The fluctuations referred to most often in this chapter are t-fluctuations, and the "t-" will therefore not be used when these are meant.

[1] See Chap. 15, Vol. 1, Radiation Laboratory Series.

optimum type of response[1] is one that is linear for signals below a certain level, called "crossover," and logarithmic for signals above this level. A receiver having such an amplitude response is known as a "lin-log" (linear-logarithmic) receiver.

21·3. The Linear-logarithmic (in Amplitude) Receiver. *Requirements.*—The desired characteristic is shown in Fig. 21·1. It can be described by the relations

$$\frac{dE_{out}}{dE_{in}} = \begin{cases} K_1 \text{ for } E_{in} < E_{in}^0 \\ \dfrac{K_1 E_{in}^0}{E_{in}} \text{ for } E_{in} > E_{in}^0. \end{cases} \tag{1}$$

When no coherent c-w signal is used in the MTI system, the lin-log characteristic may be applied to signals either before or after detection, although, of course, before cancellation. When a coherent c-w signal is used, however, the lin-log response must be obtained at intermediate frequency, before the coho signal is added; for the amplitude of a signal after addition of the coho signal depends on the phase difference between the two and hence cannot provide proper information to a circuit that is to compress fluctuations according to input signal amplitudes.

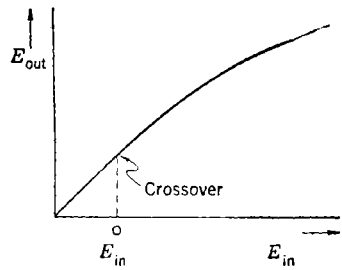


FIG. 21·1.—Ideal lin-log characteristic.

Methods.—Three different techniques have been used to obtain a lin-log characteristic. One employs feedback; the second makes use of the logarithmic response obtained when a negative voltage is applied to a variable-μ (remote-cutoff) tube; and the third involves a combination of the outputs of several channels with different gains, each having a response that is approximately linear up to a certain output and limited thereafter. The feedback method and the multichannel method will be discussed further. The second technique described above has sometimes been incorporated into a multichannel circuit,[2] but will be considered no further here.

Feedback bias for AGC has long been used in broadcast receivers. The ideal characteristic for such receivers, as shown in Fig. 21·2, is one that is limiting rather than linear-logarithmic, but by suitable choice of constants an essentially lin-log response can be obtained. This technique

[1] This was first suggested within the Radiation Laboratory by R. A. McConnell.

[2] Such a device was used by F. V. Hunt for a logarithmic vacuum-tube voltmeter. *Rev. Sci. Instruments*, **4**, 672–675 (1933).

is suitable for the purposes considered here only if the time required for the AGC to operate is short compared with the pulse length of the system. Thus the method is perfectly feasible for pulse lengths greater than several microseconds, but for pulse lengths of 1 μsec or less the AGC response time cannot generally be made sufficiently small because of the difficulty of avoiding regeneration or oscillation (see Chap. 9). It is because of this impossibility of obtaining sufficiently fast response that the lin-log receivers used at the Radiation Laboratory have not been of the AGC type.

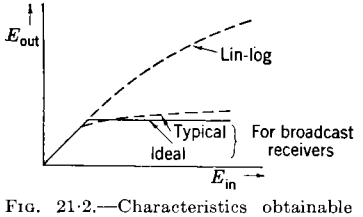


FIG. 21·2.—Characteristics obtainable with feedback gain-control bias.

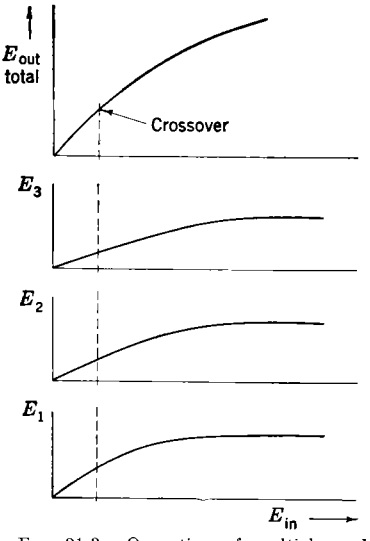


FIG. 21·3.—Operation of multichannel lin-log amplifier.

The principle of the multichannel type of lin-log network is illustrated in Fig. 21·3, and a basic block diagram is shown in Fig. 21·4. (E_1, E_2, and E_3 may be the outputs of V_1, V_2, and V_3 respectively.)

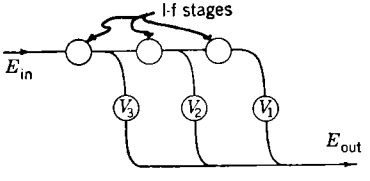


FIG. 21·4.—Simplified block diagram of lin-log amplifier.

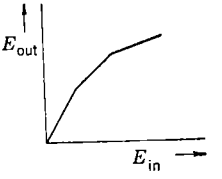


FIG. 21·5.—Lin-log amplifier response for linear sharply limiting individual-channel response.

The characteristic of each channel is represented in Fig. 21·3 as being approximately linear. If each stage had strictly linear response up to a sharply defined limit level, the total response curve would, of course, not be so smooth as that shown in Fig. 21·3 but would instead have the form shown in Fig. 21·5. In fact, even if the limit levels are not sharply

defined, the total response curve may still have this segmented form if the gain difference between successive channels is so great that any particular channel is not yet delivering appreciable output when the next more sensitive channel is nearly limiting. It has been found that with the circuits used, sufficiently good logarithmic response could be obtained for successive voltage-gain differences as high as ten times. Greater gain differences could be employed if the response of each channel were more nearly logarithmic than linear, but additional complications would be required to obtain such a response, and the resulting economy of channels would, for the applications encountered, consist only of a reduction from three channels to two. Hence, although more nearly logarithmic response might be desirable in some cases, the circuits actually used have had approximately linear response.

Even when the circuit constants and operating conditions of a lin-log network have been properly determined, the over-all response characteristic may suffer distortion if any of the following occurs in any channel:

1. Improper gain.
2. Improper limit level.
3. Nonflat limiting or overloading.

The likelihood of items 1 and 2 can often be reduced by using degeneration in each channel, although such degeneration generally tends to introduce some distortion in the over-all response by making the characteristic of each channel more linear. Problem 3 does not exist when the limiting is obtained in a video amplifier by driving a tube to cutoff with a negative signal, but such video limiting cannot be employed when a coherent c-w signal is used, because, as pointed out above, the lin-log response must be obtained at intermediate frequency before the addition of the coho signal.

I-f Limiting.—It has been found[1] that the high figure-of-merit tube types commonly used in high-gain, wide-band amplifiers have overload characteristics of the kind shown in Fig. 21·6. The various dashed lines represent response curves obtained with different tubes of a given type or under different operating conditions. The effect is apparently associated with the geometrical characteristics of the tubes involved. The 6AC7, for example, has a plate consisting of two comparatively narrow electrodes (which are so made to keep plate-to-ground capacitance low), to which cathode current is effectively beamed by grid support rods when the control grid is negative with respect to the cathode but which are easily missed by the electron paths existing when the control grid is positive. The cathode current then goes largely to the screen grid.

[1] See Sec. 6·11.

Unfortunately, although the overload characteristic of a tube is improved by raising its plate voltage, its maximum output is also increased, so that the following tube is more easily overloaded. Conversely if the screen voltage is reduced, the maximum output is reduced, but the tube overloads at a lower signal voltage. Since the nature of the effect is

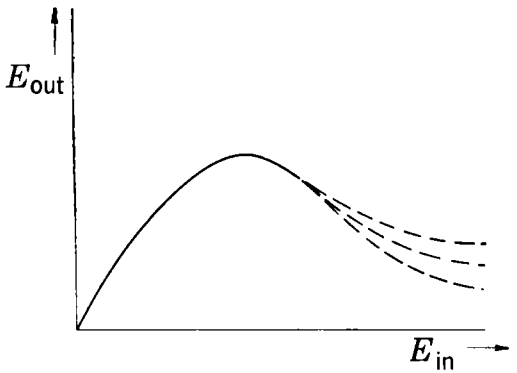


FIG. 21·6.—Typical i-f amplifier overload characteristic.

such that the maximum signal output current is limited, overloading of successive stages can be reduced or eliminated by use of suitably small load resistances, although this "suitable" value is generally somewhat smaller than that which would be used if only gain-bandwidth considerations were involved.

The methods that have been used to obtain fairly flat and uniform limiting have included combinations of those just stated, with the addition of one more, the use of grid-current limiting resistors or grid leaks. A section of the circuit of a typical amplifier using a grid leak is shown in Fig. 21·7. This method of reducing overloading effects is feasible only if the time constant RC can be made short compared with a pulse length and yet long compared with an i-f cycle, that is, only if the i-f pulse contains several cycles. Even if this condition is satisfied, the effectiveness of the grid leak cannot be made very great for pulses shorter than $\frac{1}{2}$ μsec or so, because C should be several times C_g in order that there be little attenuation from the capacitive voltage divider made up of C and C_g, and R should be large compared with the effective grid-cathode resistance when grid current flows. These conditions determine a minimum value of RC for effective grid-leak action.

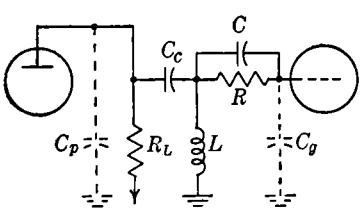


FIG. 21·7.—Simplified circuit diagram of i-f amplifier with grid leak.

Buffering.—In a multichannel lin-log amplifier, there may be the possibility of feedback from the output terminal through the least sensitive channel into one of the other channels. For the purpose of further discussion, a block diagram of a typical i-f lin-log amplifier is shown in Fig. 21·8. The arrows show the desired direction of signal transmission. Networks A attenuate signals, and the 180° phase-shift network is required to make the output of Channel 2 in phase with the outputs of Channels 1 and 3.

Let m = gain of V_1, V_2, or V_3, db,
n = gain of V_4, V_5, or V_6, db,
p = attenuation of A, in direction of arrow, db,
q = attenuation of A, in direction opposite to arrow, db,
r = attenuation of feedback through V_4, db

("db" is used here as $20 \log_{10} \frac{\text{output voltage}}{\text{input voltage}}$, without regard to impedance levels.)

In order that the circuit shall not oscillate, it is necessary that

$$p + q + r > 2m + n \tag{2}$$

at any frequency at which the phase shift around the loop from the output terminal through V_4, A, V_2, V_3, A, and V_6 is an integral multiple of 2π. The principal function of V_4 is to provide a means for obtaining as unilateral transmission as possible, that is, to provide buffering action between the output of V_6 and the input of V_2. In terms of the quantities listed above, it may be said that V_4 is included to make r as high as possible compared with n.

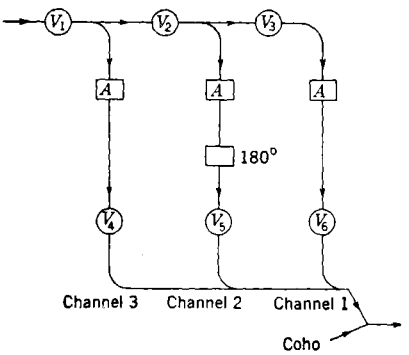


FIG. 21·8.—Block diagram of typical i-f lin-log amplifier.

The inequality stated above in Eq. (2) provides only that the amplifier will not oscillate. For two other reasons as well as for the necessity of having a safety factor, feedback through Channel 3 should be far less than the critical value indicated by Eq. (2). (1) Feedback considerably below the critical value for oscillation can still cause bandpass distortion severe enough to distort signals undesirably. (2) Some coho signal may feed back through Channel 3 in sufficient strength to cause V_6 to deliver an appreciable fraction of its maximum output. If this condition exists, the lin-log characteristic will effectively be applied to signals not accord-

ing to their amplitudes but according to the resultant amplitudes of signals plus coho. Thus the amount of compression suffered by a signal (and by its fluctuations) will depend on the phase difference between the signal and coho. For these reasons, a more suitable inequality for actual circuits than Eq. (2) is

$$p + q + r > 2m + n + 20\left(1 + \log_{10}\frac{\text{coho amplitude}}{\text{crossover amplitude}}\right) \quad (3)$$

where crossover amplitude means the output amplitude of crossover signals. Equation (3) applies only when coho amplitude is not appreciably less than crossover amplitude. This condition is usually desirable in the case of an i-f lin-log amplifier, as discussed below.

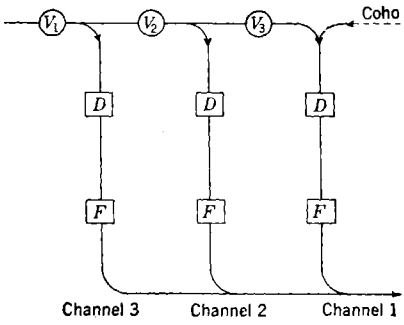


FIG. 21·9.—Block diagram of typical video-output lin-log amplifier. Circuits *D* are detectors; circuits *F* are filters.

If video output is desired rather than i-f output, either the circuit represented in Fig. 21·8 with the addition of a detector or the circuit of Fig. 21·9 may be used. The circuit of Fig. 21·9 provides lin-log response with a minimum number of tubes. This circuit will be discussed in more detail later; it is shown here because of its bearing on the buffering problem. It can be seen that if each of the channels includes an i-f filter (as will be the case if each channel has a detector), feedback through Channel 3 can easily be made sufficiently small.

Addition of Coho Signal.—If echo signals are to be compressed according to their amplitudes, the coho signal, if used, must be added *after* the echo signals have passed through the compressing network, since the amplitude of coho plus signal will depend on the phase difference between the two.

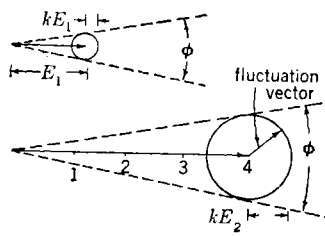


FIG. 21·10.—Response of linear receiver.

Coho Amplitude.—To see how the coho amplitude affects the response of a lin-log receiver, it is instructive to consider first the nature of the response of a lin-log amplifier to signals having various average amplitudes but constant percentage fluctuations.

Figure 21·10 is a vector representation of the output of a linear amplifier for two signals of relative amplitudes E_1 and $4E_1$, with the same per-

centage fluctuation k. (The fluctuation is represented in terms of a fluctuation vector as mentioned in Sec. 21·2). The dotted lines represent the limits of the phase variation possible in the output (or input) signal. Note that if the percentage fluctuation is the same for the two signals, the maximum phase variation ϕ is also the same.

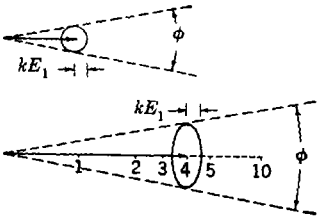


FIG. 21·11.—Response of lin-log receiver.

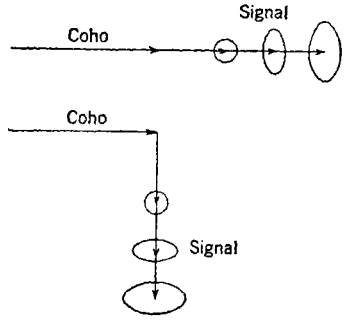


FIG. 21·12.—Lin-log receiver with large coho.

Figure 21·11 represents the output of a lin-log amplifier for the same two signals, assuming that E_1 is a signal which just reaches crossover. The numbers along the axis in the case of the larger signal denote relative *input* amplitudes. Thus Fig. 21·11 is essentially a conformal map of the response characteristic of a lin-log amplifier. Note that for constant percentage fluctuation of the input signal, the *magnitude* of the *amplitude* fluctuation of the output signal is the same for any signal above crossover as for one at crossover. Note, however, that the lin-log characteristic does not affect the phase fluctuation of signals.

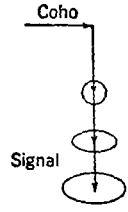


FIG. 21·13.—Lin-log receiver with small coho.

This lack of restriction of phase excursion may result in excessive output fluctuation (that is, fluctuation greater than that desired or expected from the lin-log amplifier) when the coho signal is added to output signals if coho amplitude is appreciably greater than crossover amplitude. This possibility is illustrated in Fig. 21·12 from which it can be seen that, for signal amplitudes greater than crossover but smaller than coho, the resulting fluctuation when coho and signal are in phase is the amplitude fluctuation of the signal, but the resulting fluctuation when coho and signal are approximately 90° out of phase is determined principally by the phase fluctuation of the signal and is greater than for the in-phase case.

From Fig. 21·13 it is clear that if coho amplitude is made no larger than crossover amplitude, the resulting fluctuation for a signal appreci-

ably greater than crossover will be essentially the properly (logarithmically) compressed amplitude fluctuation of that signal, regardless of the phase difference between the signal and coho.

The foregoing discussion shows that coho amplitude should not be appreciably *larger* than crossover amplitude. That it should also not be appreciably *smaller* than crossover amplitude is indicated by the following three considerations, which are listed in the order of their importance.

1. For isolated (not in clutter) moving target signals that produce output amplitudes larger than coho, the amplitude of the beat resulting when coho and signal are added will be that of coho. For good probability of detection of the moving target, this amplitude should be several times rms noise amplitude. Since under some operating conditions rms noise amplitude may be as great as one-tenth to one-third crossover amplitude, coho amplitude (which generally is not and need not be readily variable) should not be appreciably smaller than crossover amplitude.
2. In cancellation systems using a supersonic delay line, it is often convenient to apply coho and i-f output from the receiver directly to the delay line. In this case, coho may be considered to be a "carrier," modulated by the i-f signals. In the cancellation circuits commonly used, it is desirable that the carrier level not be too small a fraction (not less than one-half to one-eighth) of the maximum signal level.[1]
3. If coho amplitude is not larger than the largest output signal, then some clutter echoes will have such amplitudes and phases as to almost cancel coho; in the region of such clutter, the receiver will be relatively insensitive to moving targets for the reason given in Paragraph 1 above. The probability of occurrence of a given amplitude clutter echo decreases with increasing amplitude, so that the probability that coho will be approximately canceled by clutter increases with decreasing coho amplitude.

Methods.—The only method of adding coho signal to i-f output signals that has been considered thus far is simple addition at the output of an i-f lin-log amplifier. This is the most suitable technique if i-f output is desired from the receiver, but there is a somewhat simpler and almost equivalent method that can be used if video output is desired. Figure 21·9 shows this simpler method, which consists of adding coho only to Channel 1 of a lin-log amplifier. In terms of receiver types described later in the chapter, the action of this circuit may be likened to the combination of a limiting receiver with coho plus a lin-log receiver without

[1] See Sec. 14a·4, Vol. 20.

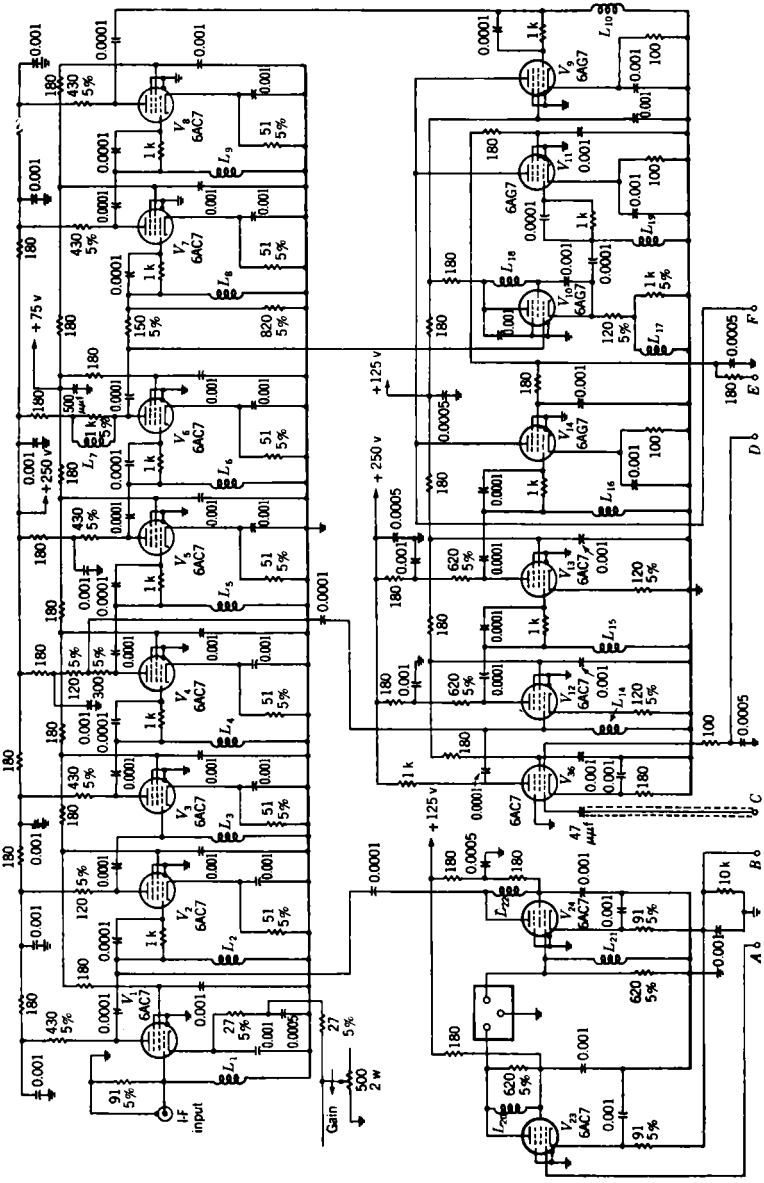
coho. Obviously coho can be added to one channel without being added to the others only if the output of that channel is detected separately from that of the others. This method is simpler because the phase relations between the signals in the various channels are unimportant and because the nature of the response is approximately independent of coho amplitude provided that the latter is greater than crossover amplitude. The response characteristic is less satisfactory than that of the circuit of Fig. 21·8 in one minor respect; namely, the response to fluctuations of a large (larger than crossover) signal depends on the phase of the signal with respect to coho. The response may be as much as 3 db greater for signals orthogonal to coho than for signals in phase with coho. (The nature and magnitude of this effect can be shown by diagrams similar to those of Figs. 21·10 to 21·13.)

Specific Circuits.—The general principles on which the design of a multichannel lin-log amplifier is based may be summarized as follows:

1. Suitable gain separations must exist between the channels and their limiters.
2. All channels should deliver the same limit-level output. An amplifier in which this is true can be more easily maintained and generally more easily designed.
3. Coho amplitude must have the proper relation to crossover amplitude.
4. Suitable buffering must be provided.
5. The side channels must be connected to the main amplifier in such a way as not to reduce undesirably the efficiency of the interstage coupling circuits in the main amplifier. The use of a voltage divider across the tuned circuit of the main amplifier to which the side-channel connection is made or the use of germanium crystals having low capacitance (see Chap. 7) for the detectors in the side channels aids in reducing the band-narrowing effect.

The actual design is determined by these principles and by the gain, bandwidth, logarithmic range, and nature of the output (that is, i-f signals or video signals). Of these latter factors, only the logarithmic range requires further discussion here.

The term "logarithmic range" is used here to describe the dynamic range of input signals between crossover and complete limiting. It is approximately true that a crossover signal is just strong enough to cause limiting in the most sensitive channel of a lin-log amplifier. If this were exactly the case, the logarithmic range would be the difference in gain between the most sensitive and least sensitive channels. Actually, however, for increasing input the gain of the most sensitive channel begins to decrease for signals considerably weaker than those just strong enough


+ 75 v
+ 250 v
+ 125 v
+ 250 v
+ 125 v
H-F input
Gain
A
B
C
D
E
F

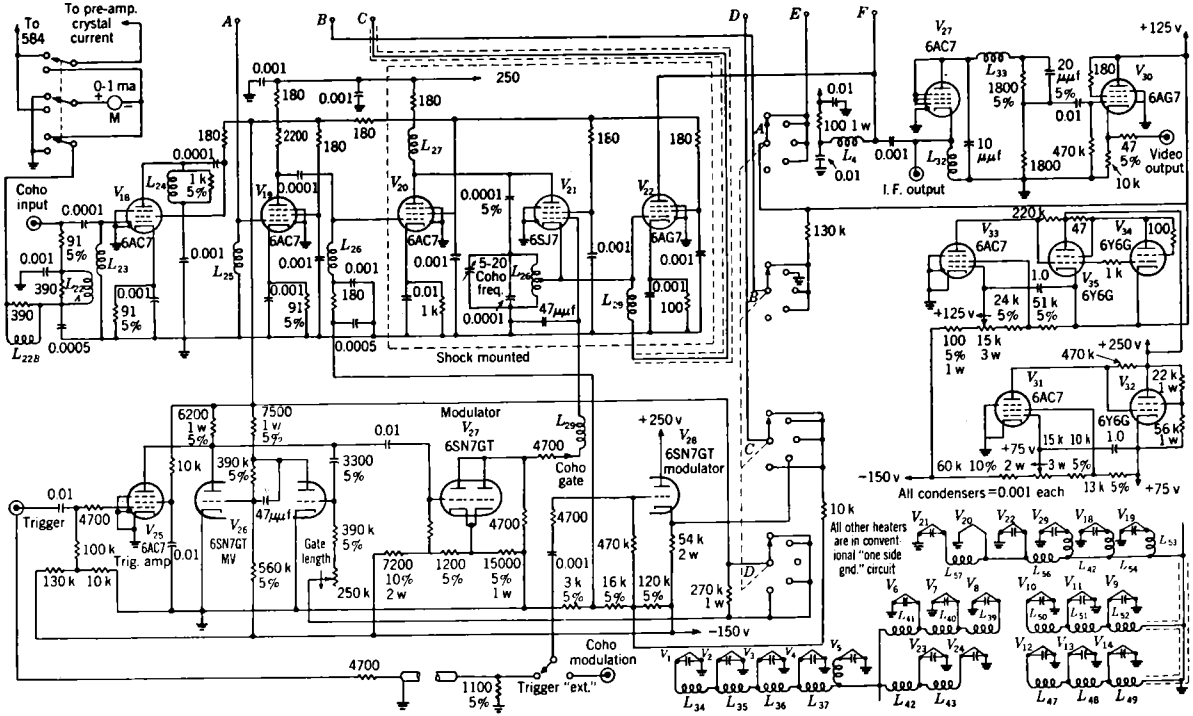
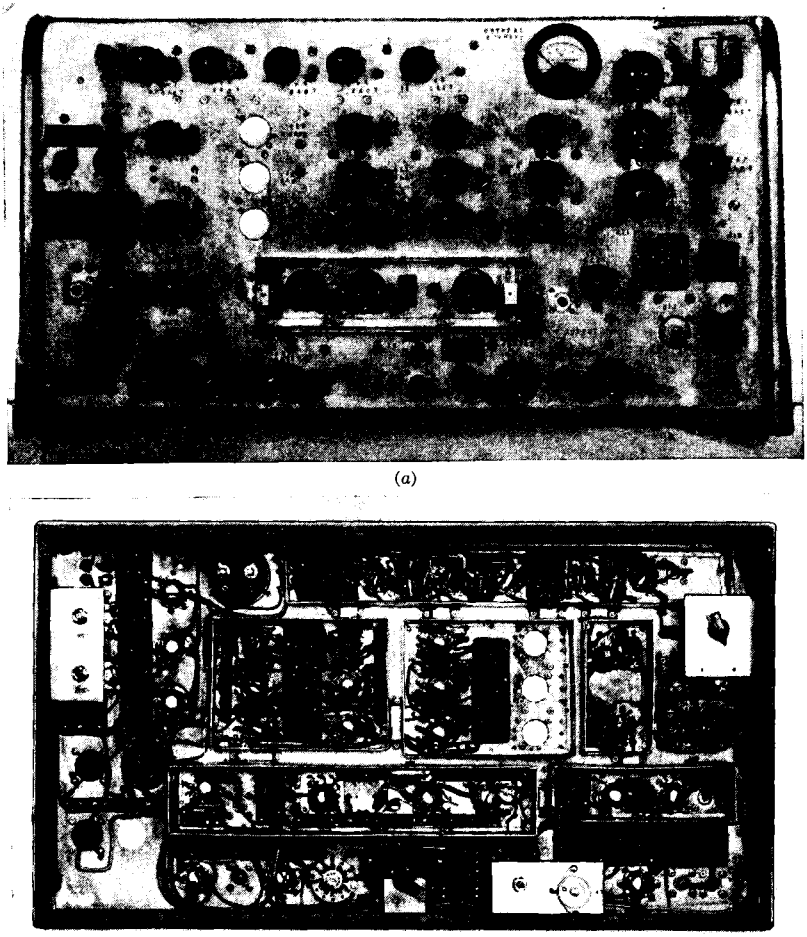


FIG. 21·14.—Circuit diagram of MTI receiver.

to cause limiting; in other words limiting in this channel is not caused by a crossover signal but by a signal at least 8 to 12 db stronger (in typical cases) than crossover. Thus, a typical lin-log amplifier having three channels with successive gain separations of 15 db might have a loga-



(*a*)

(*b*)

FIG. 21·15.—Main chassis of MTI receiver. (*a*) Back view; (*b*) internal view.

rithmic range of 40 db—30 db by virtue of the multichannel construction and 10 db between crossover and limiting in the most sensitive channel.

Figure 21·14 is the circuit diagram of an MTI receiver that was used in the field. This receiver has i-f output, coho amplitude slightly greater

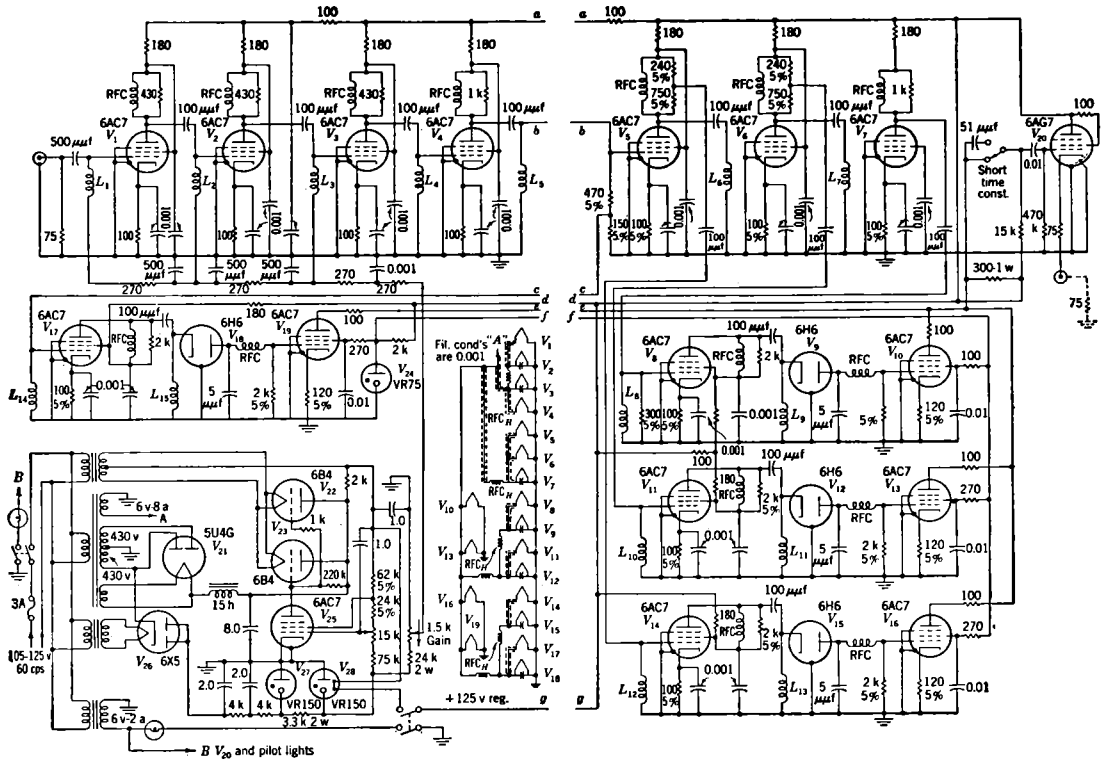


FIG. 21·16.—Circuit diagram of lin-log receiver.

than crossover amplitude, and logarithmic range of about 50 db. The bandwidth, exclusive of a preamplifier, is 3.5 to 4 Mc/sec.[1] Two views of the main chassis of this receiver are shown in Fig. 21·15*a* and *b*.

Figure 21·16 is the circuit diagram of a lin-log receiver that was designed for miscellaneous applications not requiring coho. Video limiting was used because it is more easily made to give consistent performance than i-f limiting. The logarithmic range is about 65 db, and the bandwidth, exclusive of preamplifier, is about 3 Mc/sec.

Figures 21·8 and 21·9 represent the arrangements that have come to be considered to be optimum for lin-log amplifiers delivering i-f or video output respectively. It will be seen that the numbers of tubes involved are considerably smaller than in the designs of Figs. 21·14 and 21·16. It is, of course, desirable to reduce to a minimum the number of those tubes which affect the gain differences between channels and the limit-level output of each channel. The circuits of Figs. 21·8 and 21·9 can be used to obtain up to about 45-db logarithmic range, which is sufficient for all MTI applications contemplated at present. For the logarithmic range of 45 db both circuits can be made to have bandwidths of at least 6 Mc/sec for the interstage circuits between V_1 and V_2 and between V_2 and V_3, thus permitting over-all bandwidths for the receiver as high as 2.5 or 3 Mc/sec to be obtained. (If less logarithmic range is required, or if more channels are used, the over-all bandwidth can be made greater.)

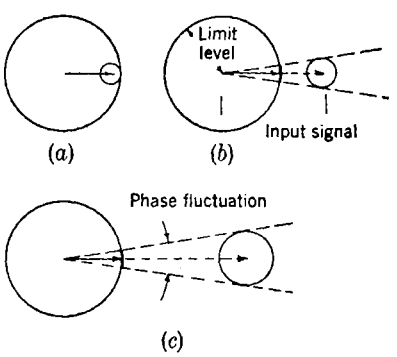


FIG. 21·17.—Logarithmic phase response in a limiting i-f amplifier.

21·4. The Limiting Receiver. *Theory of Operation.*—It has been shown that because clutter fluctuations (exclusive of noise) are proportional to rms clutter amplitude, the ideal receiver for MTI is one that compresses fluctuations of large clutter logarithmically. Section 21·3 described a method of obtaining logarithmic amplitude-fluctuation response. This section describes a method of obtaining logarithmic phase-fluctuation response.

It was mentioned above that signals of any amplitude having the same percentage fluctuation have exactly the same phase fluctuation. A receiver in which large signals limit in the i-f amplifier will thus give an i-f output having phase fluctuations depending only on the *percentage*

[1] See RL Report No. M-218, June 1945, for a more complete description.

fluctuation of a large signal[1] and not on the rms amplitude of the signal. This compressing action is illustrated in Fig. 21·17. It appeared that a limiting receiver would have certain advantages over a lin-log receiver (see Sec. 21·6).

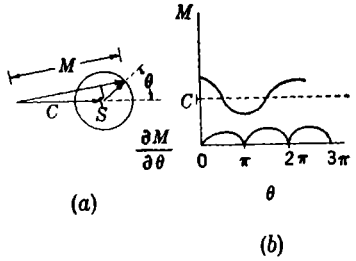


FIG. 21·18.—Limiting receiver with large coho; C = coho (vector), S = signal (vector), $M = |C + S|$.

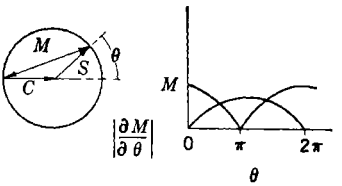


FIG. 21·19.—Limiting receiver with limit-level coho.

Coho Signal Addition: The "Balanced Detector."—There were two serious difficulties that delayed the application of a limiting receiver to MTI systems. The first one was to obtain flat i-f limiting instead of the condition shown in Fig. 21·6. The second and more serious difficulty was that with any simple method of adding coho signal, the receiver would be insensitive to fluctuations for certain phases of clutter. Figure 21·18 shows the manner in which the resultant of coho plus echo signal varies with the phase angle between the two for a large (limiting) signal and coho amplitude greater than limit level (output amplitude of a limiting signal). A moving target in clutter causes a small fluctuation in the total signal;

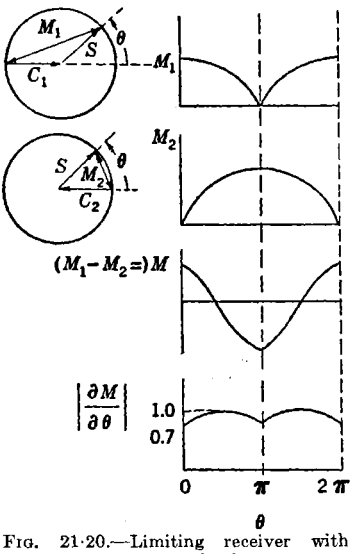


FIG. 21·20.—Limiting receiver with dual limit-level coho.

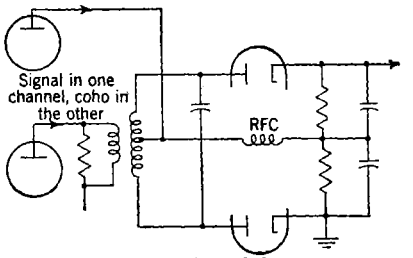


FIG. 21·21.—Balanced detector.

[1] This was pointed out at the Radiation Laboratory by A. G. Emslie.

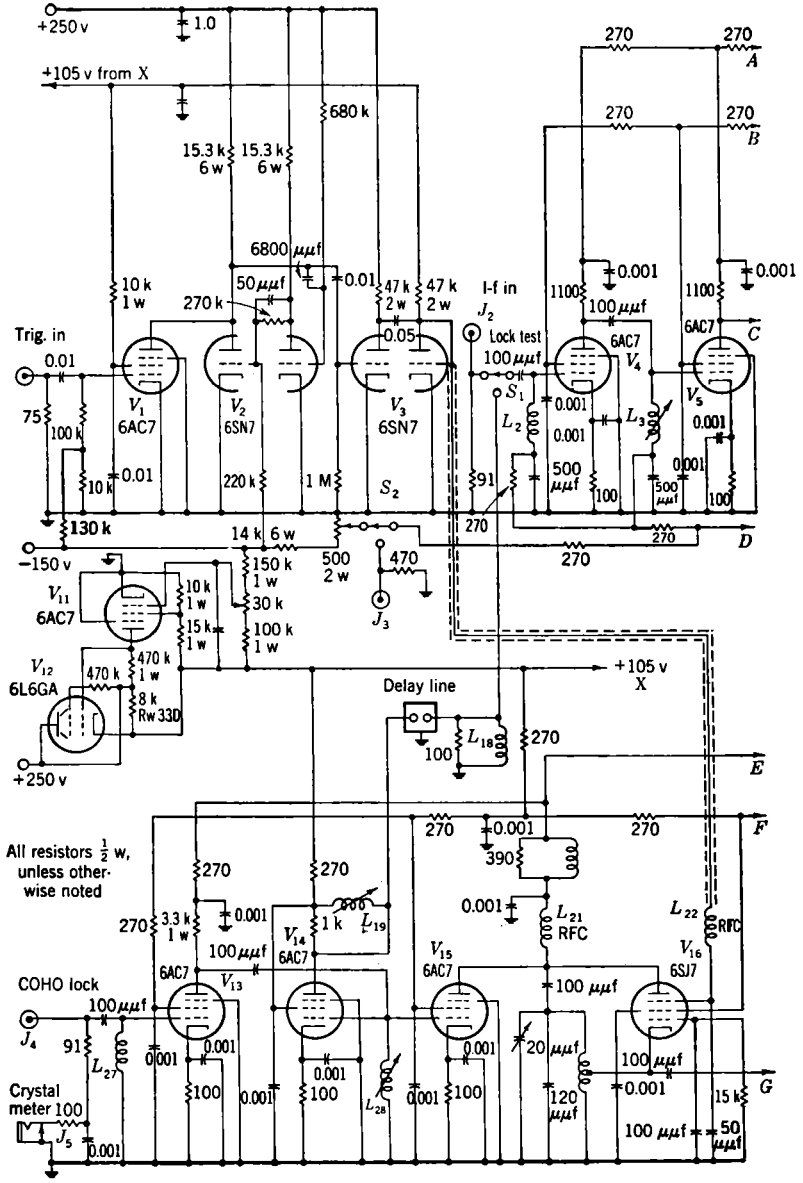
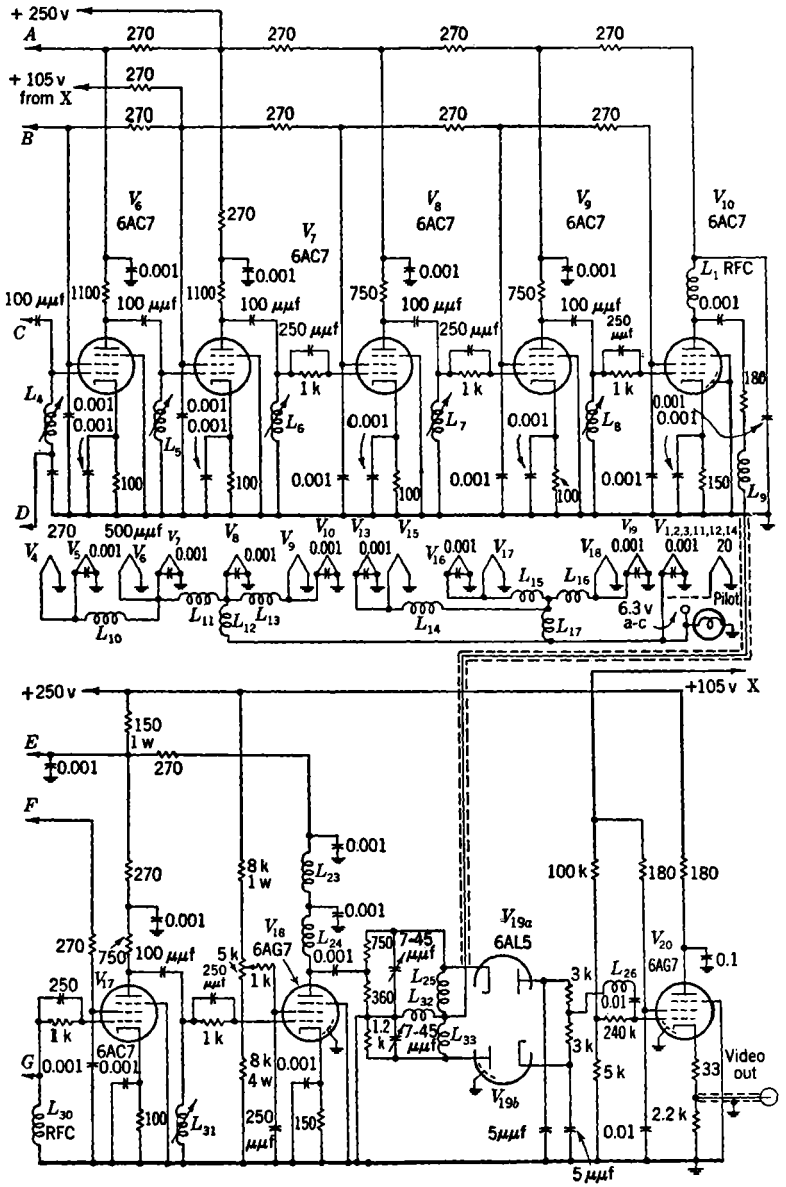
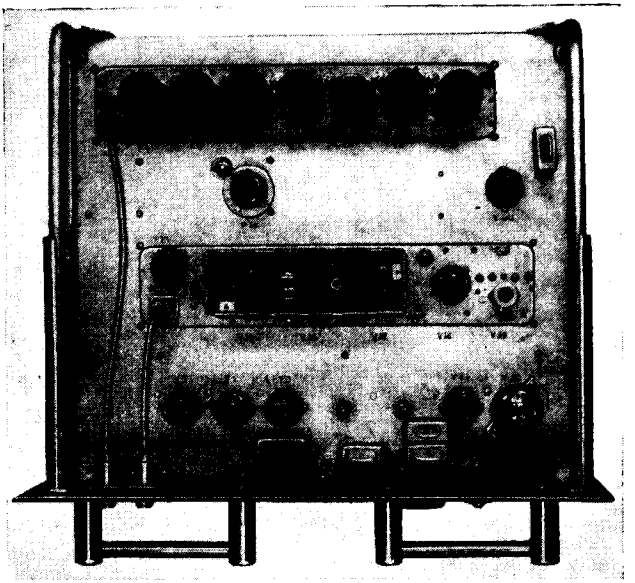


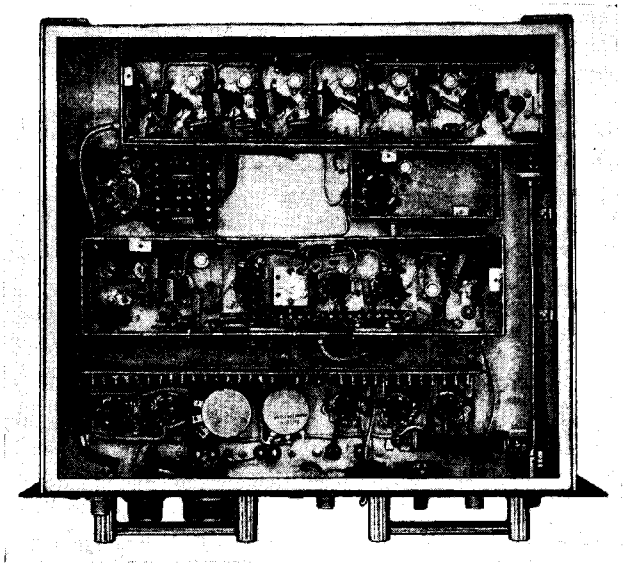
FIG. 21·22.—Circuit of a



limiting MTI receiver.



(*a*)



(*b*)

FIG. 21·23.—Limiting receiver. (*a*) Top view; (*b*) interior view.

it is the resulting phase fluctuation that is used to detect the moving target. Clearly, it is desirable that for a limiting signal the receiver be equally sensitive to small phase fluctuations for any value of θ; that is, $|\partial M/\partial\theta|$ should ideally be independent of θ. It is evident from Fig. 21·18 that for the simple method of coho addition represented, the receiver's sensitivity to (small) phase fluctuations varies considerably with θ and is, in fact, zero for signals in phase or out of phase with coho.

One of the "blind phases" (values of θ for which $|\partial M/\partial\theta| = 0$) can be eliminated by making coho level equal to limit level (Fig. 21·19). A method of eliminating *both* blind phases by using two coho signals (in separate detectors) 180° out of phase with each other is illustrated in Fig. 21·20.[1] The use of this dual-coho detection method with limit-level coho has made a limiting receiver practical.

The detector circuit used to obtain the operation represented in Fig. 21·20 is commonly called a "balanced detector," by analogy with a "balanced modulator." The essential features of such a circuit are that

1. In each of the two detectors, coho level is the same as limit level.
2. The phase angle between coho and signal at one detector is 180° different from that at the other.

A typical circuit is shown in Fig. 21·21. The detector circuit shown in Fig. 21·22 has also been used and has the advantages of ease of balance and lack of magnetic coupling.

Coho level must be fairly closely the same as limit level for most efficient operation. Maintenance of this equality can be aided by operating the coho-channel output tube under the same conditions as the signal-channel output tube. Automatic-level-control circuits can also be used; at this time there is not enough information available to determine how well equality of levels can be maintained without any control other than factory adjustment.

A Limiting Receiver.—Figure 21·22 shows the circuit diagram of a limiting receiver. Two views of this receiver are shown in Fig. 21·23*a* and *b*.

21·5. The IAGC Receiver. *Brief Description of IAGC.*—The term IAGC[2]—instantaneous automatic gain control—is commonly used to describe back-bias circuits designed to prevent i-f limiting of strong signals. Figure 21·24 illustrates the action of back bias (only the envelope of i-f signals is shown); Fig. 21·25 is a simplified diagram of a typical IAGC circuit. The correct back bias to keep the incremental gain of V_2 approximately constant is applied by the detector-amplifier circuit through the time constant RC, where C is the total capacity from the

[1] This scheme was proposed by A. G. Emslie and F. Cunningham.

[2] IAGC is treated in more detail in Chap. 9.

grid of V_2 to ground. The length of this time constant determines the exact response characteristics of the IAGC circuit. For MTI work, the time constant has generally been made approximately equal to the pulse length used in the system. (In some circuits that have been used, the effective time constant has depended on the magnitude and polarity of the incremental signal.) The action of a circuit having such a time constant and having some voltage delay in the detector part of the IAGC circuit is represented in Fig. 21·26.

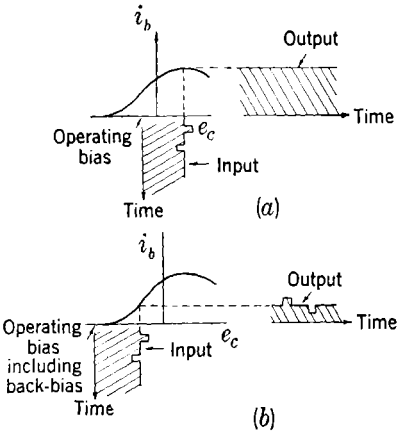


FIG. 21·24.—Operation of back bias.

Note that IAGC is quite different from the type of automatic volume control used in broadcast receivers. The latter is designed to reduce gain without distortion in the presence of strong signals, whereas IAGC, which is intended to preserve incremental gain, effectively removes the average component of strong signals and retains only the r-fluctuations.

It may be mentioned here that IAGC circuits do not perfectly preserve incremental gain and that some fluctuation compression is therefore

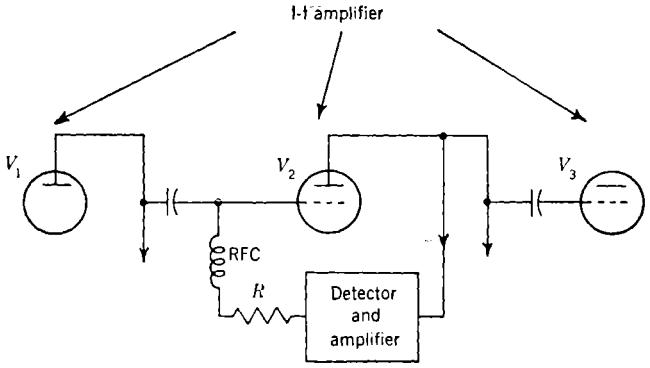


FIG. 21·25.—Simplified IAGG circuit.

applied to strong signals. In fact it seems possible to construct an IAGC circuit having an effect approximating that of a lin-log receiver followed by a differentiator circuit. As stated above, however, IAGC cannot readily be made "instantaneous" enough to provide lin-log response for pulses shorter than about 1 μsec.

The Use of IAGC Receivers in MTI.—Two reasons have been advanced for the use of IAGC receivers for MTI. (1) It has been claimed that IAGC operates as a "precanceler" of permanent echoes—a device that partially removes fixed echoes but does not appreciably affect *t*-fluctuating echoes. (2) IAGC is a desirable feature of receivers for *normal* radar systems—it is generally desirable to have "normal" presentation available even in a system used primarily for MTI—and simplification of production and maintenance problems will result if the MTI and normal receivers can be made approximately the same. This second reason is indisputably a good one, but close examination of the first leads to the following conclusions:

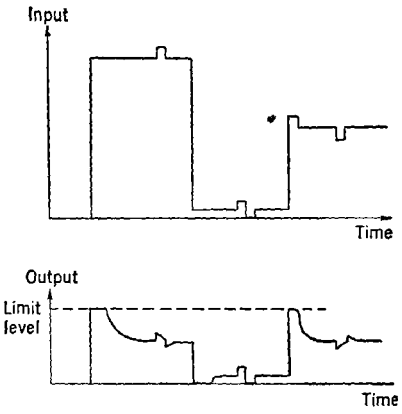


FIG. 21·26.—Operation of short-time-constant voltage-delayed IAGC.

1. The usefulness of IAGC as a precanceler depends on the *r*-fluctuations of normal clutter echoes and decreases as these fluctuations increase.[1] Ideally at least, IAGC does not reduce the incremental gain of the receiver—the gain to *r*-fluctuations— and it can be seen that after a large *r*-fluctuation the receiver is insensitive for a short time (Fig. 21·26). The receiver will thus be insensitive to amplitude *t*-fluctuations and partially insensitive to phase *t*-fluctuations, where clutter has large *r*-fluctuations. The harmfulness of this sensitivity loss is a statistical matter that has not been examined quantitatively; it can be said that *r*-fluctuations are indeed large in clutter and also that IAGC receivers have given impressive (although, to date, not quantitatively evaluated) performance.
2. The response of an IAGC receiver to a signal at a particular range depends on the amplitudes of the preceding signals; not only are *t*-fluctuations of strong signals generally not compressed, but *t*-fluctuations may even be expanded by the IAGC action. In view of the discussion in Sec. 21·2, one would expect an IAGC receiver to be unsatisfactory when clutter signals were so strong and the percentage *t*-fluctuation so high that clutter *t*-fluctuations were larger than noise. These IAGC receivers have been used for MTI only on relatively low-power ground systems where

[1] The subject of precancellation by IAGC and other methods is discussed further by W. Selove in RL Report No. 1016, March 1946.

clutter t-fluctuations are seldom larger than noise. In the case of an airborne MTI system using an IAGC receiver, however, it was found that the output due to clutter fluctuations was greater than that due to noise even through the system power was comparatively low (and the clutter signals therefore comparatively weak). The difference between ground and airborne systems is that for typical systems the percentage t-fluctuation is much greater in the airborne case.

Coho Addition.—The output of an IAGC receiver is restricted in amplitude; but unlike the limiting receiver, the output even for strong input signals contains amplitude fluctuations. Thus no blind phase exists such as would be the case for a limiting receiver and single coho. A satisfactory method of adding coho signal in an IAGC receiver is, therefore, simple addition. Coho level should be at least as large as the maximum signal output, because the latter in this type of receiver may be only a few times as large as noise. Simple addition of large amplitude coho has the advantage that for signals orthogonal to coho, output fluctuations depend essentially only on the output signal's phase fluctuations—there is thus for these signals the same kind of logarithmic compression that is obtained in a limiting receiver (see Fig. 21·17).

21·6. Comparison of Receiver Types. *Sensitivity to Weak Signals.*—For noise and hence weak signals, small compared with limit level, the signal-to-noise response of a "balanced detector" is the same as that of a simple single-coho detector. The IAGC receiver generally includes voltage delay in the IAGC circuit, so that for weak signals this receiver has a simple linear response. Thus the sensitivity to weak signals is the same for all three receivers.

Fluctuation Compression.—The lin-log receiver with small coho (small compared with maximum signal output) and the limiting receiver both give logarithmic compression of fluctuations of large signals. The IAGC receiver generally does not compress fluctuations and may even exaggerate them.

Response to Moving Targets in Clutter of Various Phases (with Respect to Coho).—Figure 21·19 shows the relative response to fluctuations on a large signal of a limiting receiver with a balanced detector. The (video output) lin-log receiver with (large) coho in one channel shown in Fig. 21·9 has exactly the same response curve. A lin-log i-f receiver with small amplitude coho or a lin-log receiver with no coho has a response independent of clutter signal phase—the ideal nature of this response. The response of an IAGC receiver may vary considerably with the signal phase, but the response will always be smaller for clutter signals orthogonal to coho than for clutter signals in phase or out of phase with coho.

Ability to Handle Very Large Signals (without Overloading).—The limiting receiver is best in this respect—it will handle the largest signals that can possibly be obtained. The IAGC receiver and lin-log receivers have about equal abilities—the largest signals that can be handled without overloading are those which limit the last unprotected tube or the last side channel. It is quite easy for either type to secure enough range for all presently contemplated applications.

Dynamic Range of Output Signals.—This dynamic range is of importance because of its effect on cancellation-circuit requirements. Thus, a receiver from which the largest output signal is only ten times rms noise, as is a typical case for an IAGC receiver in MTI service, imposes much less stringent requirements on cancellation than does one in which the largest output signal is one hundred times noise, as it is in a typical lin-log case. The effective dynamic-range compression of an IAGC receiver is perhaps its most important advantage.

A simple calculation shows that in a lin-log receiver which satisfies Eq. (1) of Sec. 21·3, the output for signals above crossover increases by an amount equal to the crossover output when the input increases by 8.7 db. Thus a lin-log receiver with 40-db log range has a maximum output 4.6 times (13 db greater than) crossover.

A limiting receiver is better than a lin-log receiver by just the factor of the logarithmic range—the maximum output of the former corresponds to a crossover signal.

Frequency (Carrier or Video) of Output Signals.—Any of the receivers described can be used to deliver video output. If i-f output is desired, however, the limiting receiver cannot be used unless the complete loss in sensitivity that exists for "blind phases" can be tolerated. That is, for efficient operation of a limiting receiver, a balanced detector should be used, and a balanced detector delivers only video output.

Difficulty of Maintaining Proper Operation.—Among video output receivers, the lin-log receiver shown in Fig. 21·11 is undoubtedly the easiest to maintain—it contains no tubes other than those in an ordinary receiver strip. The limiting receiver requires a critical coho level and a special detector circuit for optimum operation. The IAGC receiver contains two or three extra tubes but requires no operational adjustments—the extra complication is only one of design and maintenance.

Between i-f output receivers, the lin-log is probably slightly more difficult to maintain than the IAGC, since the former imposes certain requirements on the comparative gain of the various tubes involved.

Use in MTI with No Coho.—In some cases (for example, in airborne MTI) it may be desirable not to use a coho and to detect moving targets by their beating with clutter. In this case the receiver must be sensitive to amplitude fluctuations (since it will not be sensitive to phase fluctua-

tions); hence the limiting receiver will not do. Either the lin-log or IAGC receiver can be used.

21·7. Best Receiver Type for Various Applications. *General MTI Work.*—In view of the material of Sec. 21·6, it is believed that the limiting receiver is the best type for general MTI work. It has the theoretically optimum form of compression of fluctuations of strong signals, fairly uniform sensitivity in any phase of clutter, and an output dynamic range that under ordinary conditions is sufficiently small to be handled comfortably by present cancellation circuits.

For comparatively low-power ground systems in locations where high-percentage clutter fluctuation is not encountered, an IAGC receiver may be satisfactory, but quantitative information on the effect of IAGC on fluctuations of typical clutter must be obtained before this type of receiver can be definitely recommended. It seems evident that an IAGC receiver will be less effective for MTI operation in the presence of sea clutter and clouds than either of the other types.

Airborne MTI.—Theoretical analysis and experimental evidence show the ordinary type of IAGC receiver to be unsatisfactory for airborne MTI because of failure to compress suitably fluctuations of large signals. For airborne MTI systems with a coho, the best receiver is the limiting type; for systems with no coho, the lin-log is the only suitable receiver.

I-f Output.—The lin-log receiver is the most suitable type for general MTI work with i-f output. Theoretical analysis indicates that there is some hope of compressing the output dynamic range of this type of receiver to a range corresponding roughly to crossover—approximately the range of a limiting-type receiver.[1]

Other Possible Receiver Methods.—An interesting new method of operation of a limiting receiver, which would make the use of a coho unnecessary, has recently been proposed. This method consists of establishing phase coherence in the local oscillator. The receiver output would be *coherent i-f signals* (instead of the usual *coherent video signals*), which could be fed directly to a cancellation circuit. For use with delay-line cancellation systems, in the present state of the art, this technique is suitable only for pulse lengths greater than about 2 μsec; for storage-tube cancellation, however, shorter pulse lengths could be used.

21·8. Coherent Oscillator Circuits and Considerations. *Coho Locking.*—There are two conflicting requirements on the coherent oscillator; it must be possible to set its phase correctly with a short *locking pulse*, and once locked it must oscillate very stably (detailed stability requirements are discussed below). One method that has been widely used to permit these requirements to be easily satisfied is to turn off the coho a

[1] W. Selove, RL Report No. 1016, March 1946.

short time (about 50 μsec) before each locking pulse and to sensitize it at just the time of the locking pulse or slightly later. This gating method is illustrated in the circuit of Fig. 21·14, in which the coho, V_{21}, is gated by the operation of V_{25}, V_{26}, and V_{27}. The exact time of application of the enabling edge of the gate is not critical as long as it is within a microsecond or two following the locking pulse. The phase of the oscillator will be determined by the oscillations that the locking pulse establishes in the tuned circuit and that, because of the high Q of this circuit, persist in appreciable amplitude for an appreciable time (the decay time constant is Q/π cycles). It may be mentioned that the enabling gate must be applied in such a manner as not to set up transient oscillations which may affect the phase of the oscillator; in the circuit of Fig. 21·14 this condition is ensured by not having too steep a front on the gate.

It has also been found possible to phase-lock a continuously running stable oscillator by using a sufficiently strong locking pulse and by adjusting the oscillator operating conditions so that the oscillator is just operating stably, that is, not oscillating too strongly.

The locking pulse can be introduced into the oscillator circuit in any simple way—it may be applied to an extra modulation grid in the oscillator tube or, as in Figs. 21·14 and 21·22, may be simply added to the tuned circuit in parallel with the signal from the oscillator tube.

It has been found that a coho can be satisfactorily locked in phase even if the natural frequency of the oscillator is quite different from that of the locking pulse—20 per cent different, for example. The coho is usually operated at signal frequency, however, because otherwise the signals resulting when coho is added to echo signals will have a spectrum centered at some frequency other than zero, and the design of the video amplifier for these signals is thereby complicated.

Locking-pulse Considerations.—The i-f pulse to lock the coho is obtained by converting the transmitter pulse with the same (stable) local oscillator used to convert echo signals to intermediate frequency. The technique and problems are quite similar to those involved in obtaining an i-f pulse for AFC purposes (see Chap. 3).

The locking pulse has generally been obtained from a separate mixer fed by the transmitter through considerable attenuation. It is desirable to obtain as much signal output as possible from the mixer; but as in the case of AFC, increasingly high output is accompanied by a faster increase in spurious components. These spurious components are produced by two causes. One is the fact that various frequency components of the transmitter pulse beat with each other to produce mixer output at intermediate frequency. Since this spurious output varies as the square of the applied r-f voltage whereas the desired i-f output varies linearly with

the applied r-f voltage if the LO signal is sufficiently strong, this effect can be kept to a satisfactorily small value by keeping the applied r-f signal voltage suitably small. A level of about 1 mw has been found to be satisfactory.

Spurious components in the i-f locking pulse channel can also be produced by undesired signals in the LO line. In particular, if the local oscillator is coupled too tightly to the signal and locking-pulse mixers, signals and TR-tube spike may be introduced into the locking-pulse mixer. It is advantageous in this connection to have as great isolation as possible between the two outputs of the local oscillator (see Chaps. 2 and 3). It has been found practical to obtain the greatest possible attenuation between the local oscillator and the locking-pulse mixer by using a signal level of about 1 mw at this mixer and adjusting the LO coupling (rather than the signal coupling) to give the required signal output.

The gain required in the locking-pulse i-f amplifier depends on the necessary signal amplitude for stable locking. It has been customary to drive the final locking-pulse amplifier sufficiently hard to make it deliver its maximum possible output. The maximum i-f voltage available from the locking-pulse mixer, if spurious components are to be avoided, is perhaps 0.05 to 0.1 volt rms into a 100-ohm line, and the voltage required to drive the final locking-pulse amplifier to maximum output is usually about 1 to 5 volts. Hence, if a terminated line is used to couple the locking-pulse mixer to the amplifier, three stages are required for a pulse length of $\frac{1}{2}$ or 1 μsec. In the circuit of Fig. 21·14, it was possible to obtain enough output from V_{19} to permit biasing the final amplifier V_{20} to approximately Class B operation, thus suppressing small spurious signals in the locking-pulse channel.

The bandwidth of the locking-pulse amplifier should be about the same as that of the echo signal amplifier, namely, about the reciprocal of the pulse length. If the locking-pulse amplifier pass band is too narrow, two undesirable effects will result: (1) The locking pulse will not reach full amplitude, and (2) the gain of the amplifier will vary rapidly with the frequency of the i-f locking pulse.

Locking Test.—An MTI system is considerably more complicated than an ordinary radar system, and it is very desirable to have convenient methods of testing the MTI components. The accuracy with which the coho is locked can be investigated in a fairly simple manner, illustrated in the receivers of Figs. 21·14 and 21·22. In the latter circuit, for example, the locking pulse at the plate of the first locking amplifier V_{13} is applied not only to V_{15} but also to V_{14} which drives a short i-f delay line (of about 6-μsec delay). When switch S_1, at the grid of V_4, is thrown to the "locking test" position, the locking pulse is transmitted through the

signal channel a few microseconds after it is applied to the coho. If the coho is locking properly, its phase will be coherent with that of the delayed locking pulse; thus the latter will appear to be a "permanent echo."

The locking test pulse provides a convenient method of adjusting the LO or coho frequency so that the signal intermediate frequency and the coho frequency are identical. The delayed pulse is a beat note between the transmitter and the local oscillator. By the time it has arrived at the point where the coho signal is added, the coherent oscillator will have settled down to its steady frequency. If this frequency is not the same as that of the signal, another beat note will be produced which will show up superimposed on the pulse in the receiver output. Accordingly the procedure is to adjust the LO frequency until this video output pulse has a flat top. For this procedure to be successful, the over-all bandwidth of the circuits through which the locking pulse passes in traversing the signal channel should be great enough to deliver an appreciably flat-topped output signal.

Coho Stability.—In order that nonchanging input echo signals be nonchanging at the output, the coho must oscillate at very closely the same frequency during successive repetition periods. Analysis shows that the criterion of stability is the time rate of change of frequency.[1] The required stability in the coho can easily be obtained by the following measures. The coho and associated elements should be shock-mounted, a high-Q tank circuit with a high ratio of capacitance to inductance should be used, and the oscillator feedback should be adjusted to make the oscillator operate fairly weakly. The receivers of Figs. 21·14 and 21·22 use a 6SJ7 oscillator because of its comparative freedom from microphonics.

A method of checking roughly on the stability of the coho is illustrated in the circuit of Fig. 21·14. When switch S_2 is in the "stability test" position, the gate length of the coho gate is increased to more than a repetition period. If the transmitter is shut off, the coho will then oscillate freely, and some coho signal will be transmitted down the MTI delay line and compared a repetition period later with the signal then being produced by the coho. If the coho is oscillating stably, the phase difference between the delayed and undelayed coho signals will be the same in alternate successive repetition periods, and this fact can be indicated by suitable methods in the "comparison amplifier" part of the MTI system.[2]

Coho Output.—Methods of adding coho signal to echo signals are discussed in the preceding sections. The circuit in Fig. 21·22 shows a coho

[1] A method of rating and measuring the stability of oscillators for coherent-phase operation is described by S. Roberts, RL Report No. 819, October 1945.

[2] RL Report No. M-218, June 1945.

output amplifier operated under conditions similar to that of the signal output amplifiers, so as to deliver a coho signal level approximately equal to crossover. In the circuit of Fig. 21·14, the coho output amplifier is operated under conditions similar to those of the signal output amplifier, so that coho level and limit level can be maintained approximately constant.

The receiver of Fig. 21·14 also includes a tube V_{36} that can be used as a modulator to inject some coho signal into the echo-signal channel.

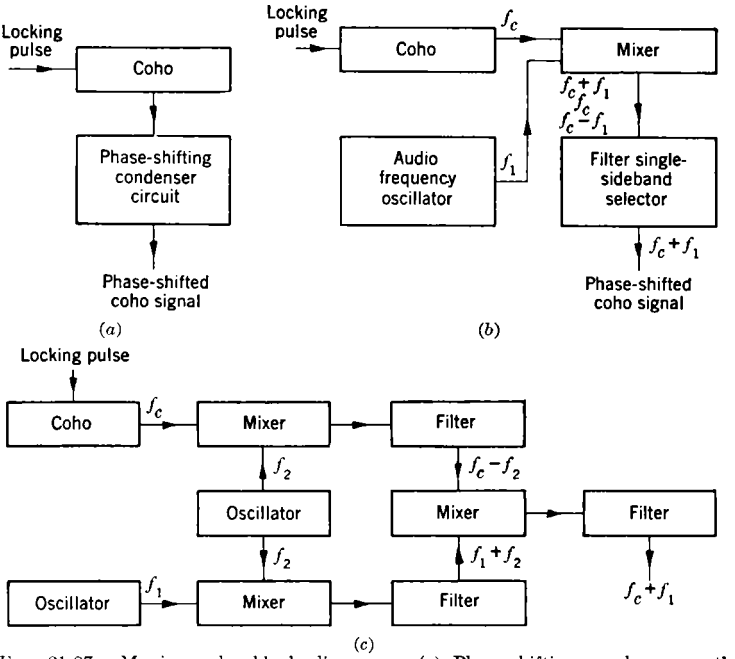


FIG. 21·27.—Moving coho block diagrams. (*a*) Phase-shifting condenser method. Rate at which condenser is rotated varies linearly with relative velocity to be compensated. (*b*) Single-sideband method. Frequency varies linearly with relative velocity to be compensated. (*c*) Single-sideband method with intermediate step.

This test signal is, of course, coherent with the coho and can be used to adjust the cancellation circuit.

"Moving" Coho.—For MTI systems in motion relative to ground or sea clutter, such as airborne or shipborne systems, signals from stationary clutter targets do not have the same phase relative to the transmitter pulse in successive repetition periods. It is possible, however, to adjust the phase of the coho signal so that signals from stationary targets will have the same phase relative to coho phase in successive repetition periods. Such a phase-adjusted coho is called a "moving" coho.

The necessary phase correction is a linear function of time for constant relative velocity of the system and the "permanent echo" target. Two methods of applying such a phase correction are shown in Fig. 21·27. The first uses a phase-shifting condenser of the type described in Vol. 17, Chap. 9. In the second, the phase of the coho signal is shifted in accordance with the instantaneous phase of an electronic oscillator, the proper signal being obtained by single-sideband filtering. The filter required for the method of Fig. 21·27*b* is generally impractically sharp—not only can suitably sharp filters not be obtained, but it is not desirable to have a narrow-band filter, because a narrow-band filter will have a slow response to the transient that occurs at each relocking of the coho. Moreover, a sharp filter requires that coho frequency be much more accurately controlled than would otherwise be necessary.

The objectionable requirement of a sharp filter can be relaxed by the use of one or more intermediate stages of frequency conversion, as shown in Fig. 21·27*c*. The oscillator of frequency f_2 need not be particularly stable, since its frequency does not affect the final coho signal output at all. Typical frequency values for such a unit might be $f_1 = 50$ kc/sec, $f_2 = 1$ Mc/sec, $f_3 = 30$ Mc/sec (f_1 can have some value different from zero even for compensation for targets with zero relative velocity if f_1 for such targets is made equal to an integral multiple of the repetition frequency).

A unit similar to that of Fig. 21·27*c*, but slightly more complicated, has been built and has given successful experimental results.[1]

[1] RL Report No. 975. A mechanical computer for varying f_1 in accordance with the velocity of the system and the direction of the beam is described in RL Report No. 900.

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