References

[1] R. D. Campbell, "Radar interference to microwave communication services," *Elec. Engrg.*, vol. 77, pp. 916–921; October, 1958.

[2] A. H. Ryan, "Control of microwave interference," IRE TRANS. ON RADIO FREQUENCY INTERFERENCE, vol. RFI-1, pp. 1–10; May, 1959.

[3] M. P. Forrer and K. Tomiyasu, "Effects and measurements of harmonics in high-power waveguide systems," 1957 IRE NATIONAL CONVENTION RECORD, pt. 1, pp. 263–269.

[4] ——, "Determination of higher order propagating modes in waveguide systems," *J. Appl. Phys.*, vol. 29, pp. 1040–1045; July, 1958.

[5] D. J. Lewis, "Mode couplers and multimode measurement techniques," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 110–116; January, 1959.

[6] V. G. Price, "Measurement of harmonic power generated by microwave transmitters," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 116–120; January, 1959.

[7] C. C. Cutler, "Spurious modulation of electron beams," PROC. IRE, vol. 44, pp. 61–64; January, 1956.

[8] R. L. Jepsen, "Ion oscillations in electron beam tubes; ion motion and energy transfer," PROC. IRE, vol. 45, pp. 1069–1080; August, 1957.

[9] A. D. Sutherland, "Relaxation instabilities in high-perveance electron beams," IRE TRANS. ON ELECTRON DEVICES, vol. ED-7, pp. 268–273; October, 1960.

[10] Y. Koike and Y. Kumagai, "An experiment of ion relaxation oscillation in electron beams," PROC. IRE, vol. 49, pp. 525–526; February, 1961.

[11] V. Met, "Absorptive filters for microwave harmonic power," PROC. IRE, vol. 47, pp. 1762–1769; October, 1959.

[12] V. G. Price, *et al.*, "Harmonic suppression by leaky-wall waveguide filter," 1959 IRE WESCON CONVENTION RECORD, pt. 1, pp. 112–118.

[13] H. A. Wheeler and H. L. Bachman, "Evacuated waveguide filter for suppressing spurious transmission from high power *S*-band radar," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 154–162; January, 1959.

[14] O. Doehler, "Space charge effects in traveling-wave tubes using crossed electric and magnetic fields," *Proc. Symp. on Modern Advances in Microwave Techniques*, Brooklyn, N. Y., November 8–19, 1954. Polytech. Inst. of Brooklyn, N. Y., vol. 4, pp. 101–121; 1955. See especially pp. 115–119.

[15] R. W. Gould, "Space charge effects in beam-type magnetrons," *J. Appl. Phys.*, vol. 28, pp. 599–605; May, 1957.

[16] G. Novick and V. G. Price, "Measurement and Control of Harmonic and Spurious Microwave Energy," General Electric Microwave Lab., Palo Alto, Calif., Final Rept., Phase III, Contract AF30(602)1670, ASTIA Document No. AD-214430; May 15, 1959.

[17] L. A. MacKenzie, "Klystron Cavities for Minimum Spurious Output Power," Cornell University, School of Elec. Engrg., Ithaca, N. Y., Res. Rept. No. EE418; January 31, 1959.

[18] F. B. Llewellyn, "Electron-Inertia Effects," Cambridge University Press, London, England, 2nd ed., chs. 3, 5; 1943.

[19] K. Tomiyasu and M. P. Forrer, "Diode oscillation in high-voltage klystrons," IRE TRANS. ON ELECTRON DEVICES, vol. ED-8, pp. 381–386; September, 1961.

[20] T. D. Sege, "Electron beam forming apparatus," U. S. Patent No. 2,916,659; December 8, 1959.

[21] M. Chodorow, *et al.*, "Design and performance of a high-power pulsed klystron," PROC. IRE, vol. 41, pp. 1584–1602; November, 1953. See especially p. 1601.

[22] K. Tomiyasu, "On the possibility of drift-tunnel oscillations in high power klystrons," PROC. IRE, vol. 49, pp. 1207–1208; July, 1961.

[23] B. Fank, "Investigation of the Transverse-Field Klystron," Stanford Electronics Labs., Stanford University, Stanford, Calif., Tech. Rept. No. 305–1; March 10, 1958.

[24] K. Tomiyasu, "Table for determining cutoff frequencies for circular waveguide," to appear in *Microwave J.*

[25] D. R. Hamilton, "Summary of microwave tube types and functions," in "Klystrons and Microwave Triodes," M.I.T. Rad. Lab. Ser., D. R. Hamilton, Ed., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 7, ch. 2; 1948. See especially p. 30.

[26] J. Marcum, "Interchange of energy between an electron beam and an oscillating electric field," *J. Appl. Phys.*, vol. 17, pp. 4–11; January, 1946.

[27] T. Wessel-Berg, "A General Theory of Klystrons with Arbitrary Extended Interaction Fields," Microwave Lab., Stanford University, Stanford, Calif., M.L. Rept. No. 376; ONR Contract N6 onr 25123 (NR 373 361); March, 1957.

[28] ——, "Large-Signal Analysis of Monotron Oscillators with Constant RF Field," Norwegian Def. Res. Establ., Bergen, Norway, Tech. Note No. 1, June, 1960; Rome Air Dev. Ctr., Contract No. AF 61(052)-264 (RADC TN 60–179), June, 1960.

[29] K. Bløtekjaer and B. Grung, "Optimum RF Field Distribution of Monotron Cavities," Norwegian Def. Res. Establ., Bergen, Norway, Tech. Note No. 2, September, 1960; Rome Air Dev. Ctr., Contract No. AF 61(052)-264 (RADC TN 61-14), September, 1960.

[30] J. A. Ruetz and W. H. Yocom, "High-power traveling-wave tubes for radar systems," IRE TRANS. ON MILITARY ELECTRONICS, vol. MIL-5, pp. 39–45; April, 1961.

[31] D. G. Dow, "Behavior of traveling-wave tubes near circuit cutoff," IRE TRANS. ON ELECTRON DEVICES, vol. ED-7, pp. 123–131; July, 1960.

[32] R. M. Bevensee, "A unified theory of electron beam interaction with slow wave structures, with application to cutoff conditions," *J. Electronics and Control*, vol. 9, pp. 401–437; December, 1960.

Practical Design and Performance of Nearly Optimum Wide-Band Degenerate Parametric Amplifiers*

M. GILDEN†, MEMBER, IRE, AND G. L. MATTHAEI‡, MEMBER, IRE

Summary—The design of a two-resonator single-diode degenerate parametric amplifier is described, which incorporates features that give it nearly optimum wide-band performance. These features include the use of almost lumped circuit elements, a separate pump resonator which is very lightly coupled to the diode and pump circuits, and a diode resonated in series rather than in shunt, from which several advantages accrue. A bandwidth of 21 per cent with 15-db midband gain (double channel) is obtained at 1 Gc using two resonators, as compared with 8 per cent using one resonator. Both measured responses are found to be in excellent agreement with theoretical responses obtained with a digital computer. The measured double-channel noise figure was 1 db. Theoretical and experimental results are presented which show this type of amplifier to be remarkably insensitive to tuning errors. Good results were also obtained using two identical amplifiers in balanced operation with a 3-db coupler so as to eliminate the need for a circulator.

* Received by the PGMTT, May 1, 1961; revised manuscript received, July 24, 1961. The research was sponsored by the Wright Air Dev. Div., Wright-Patterson AFB, under Contract No. AF 33(616)-5803.

† Microwave Associates, Burlington, Mass. Formerly with Stanford Research Institute, Menlo Park, Calif.

‡ Stanford Research Institute, Menlo Park, Calif.

Introduction

Single-diode parametric amplifiers with increased bandwidth can be obtained by using properly designed multiresonator coupling circuits instead of single-resonator circuits. Such increases in bandwidth were first demonstrated by Seidel and Herrmann[1] for the case of degenerate parametric amplifiers. In their analysis the parameters of the resonators are fixed by forcing the frequency derivatives of the gain function to be zero at midband. More recently Matthaei[2,3] has shown that multiresonator coupling circuits can be designed to give relatively large bandwidths for either degenerate or nondegenerate parametric amplifiers as well as for up-converters. His analysis is based on a filter-theory viewpoint.

A single-diode degenerate parametric amplifier was designed and constructed to verify part of the design theory of Matthaei and to ascertain what practical difficulties might arise. Included in the device were several special features which helped to fulfill the objectives of the theory for optimum design.[2,3] A brief discussion of an experimental verification has previously been given by the authors.[4] In this paper (and in two SRI reports[5]) the single-diode degenerate amplifier is discussed in more detail, including the sensitivity of such amplifiers to mistuning and their operation in balanced and cascaded configurations using 3-db directional couplers.

Description of the Single-Diode Amplifier

The equivalent circuit of the two-resonator amplifier (as operated with a circulator) is shown in Fig. 1. The variable-capacitance diode is represented by a series circuit, where C_0^s is the series average value of the diode capacitance, C_1^s is the series time-varying component of capacitance, L_d is the diode parasitic inductance, and R_s is the diode loss resistor.[2,3] Capacitance C_0^s and L_d along with some additional inductance make up the series resonator indicated by X_1 in the figure. The second resonator in shunt, indicated by B_2, is included in order to obtain greater bandwidth. The pump power is supplied to the diode through the pump resonator, which is loosely coupled to the series resonator and pump generator.

[1] H. Seidel and G. F. Herrmann, "Circuit aspects of parametric amplifiers," 1959 IRE WESCON Convention Record, pt. 2, pp. 83–90.

[2] C. W. Barnes, G. L. Matthaei, and R. C. Honey, "Applications of New Techniques to Low-Noise Reception," Stanford Res. Inst., Menlo Park, Calif., Quart. Progress Rept. No. 7, Sec. 3-A, SRI Project No. 2550, Contract No. 33(616)-5803; March, 1960. See also Quart. Progress Rept. No. 8.

[3] G. L. Matthaei, "A study of the optimum design of wideband parametric amplifiers and up-converters," IRE Trans. on Microwave Theory and Techniques, vol. MTT-9, pp. 23–28; January, 1961.

[4] M. Gilden and G. L. Matthaei, "A nearly optimum wideband degenerate parametric amplifier," Proc. IRE (Correspondence), vol. 49, pp. 833–834; April, 1961.

[5] M. C. Pease, *et al.*, "Application of New Techniques to Low Noise Reception," Stanford Res. Inst., Menlo Park, Calif., Quart. Progress Rept. No. 10, Sec. 3-A, SRI Project No. 2550, Contract No. AF 33(616)-5803; December, 1960. See also Final Rept. of February, 1961.

A drawing of the strip-line realization of the equivalent circuit in Fig. 1 is shown in Fig. 2, and a photograph of the completed amplifier with its cover plate removed is shown in Fig. 3. A Hughes 1N896 diode is used which has a computer-type package. This diode has 0.020-inch-diameter wire leads which, because of their high characteristic impedance in the structure, provide the additional series inductance required to make the series circuit resonant at the signal frequency $f_0 = 1$ kMc. The shunt resonator is formed using a small, short-circuited stub to provide the inductance, and a metal block insulated with dielectric material to form the capacitance. The capacitor block was designed so that it would become resonant at the pump frequency

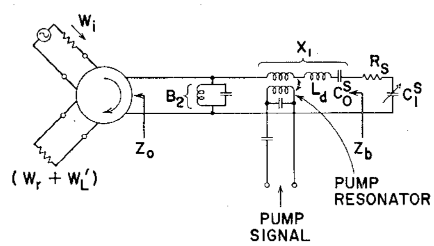

Fig. 1—Equivalent circuit of double-resonator degenerate parametric amplifier.

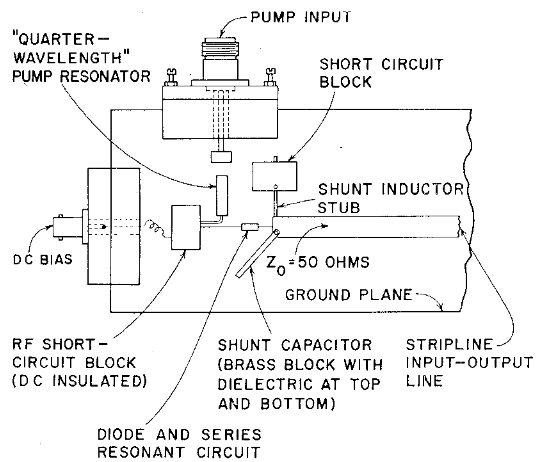

Fig. 2—Construction details of double-resonator degenerate parametric amplifier.

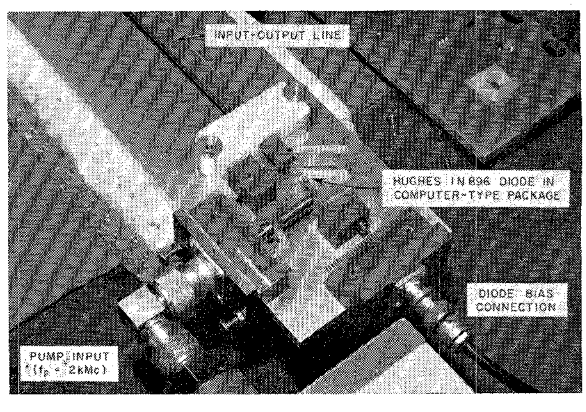

Fig. 3—Photograph of double-resonator degenerate parametric amplifier (cover plate removed).

$f_p = 2f_0$ and thus provide a short circuit across the input-output line at that frequency. The pump resonator consists of a nominally quarter-wavelength resonator which is inductively coupled to the diode resonator and capacitively coupled to the pump generator line. The short-circuit end of the pump resonator also provides its mechanical support. DC bias is applied to the diode using an external battery.

The design of the circuit elements included a number of design features which made possible nearly optimum performance of a two-resonator amplifier. The diode was resonated in series so that its internal inductance L_d merely contributes to the total inductance required for the series resonant circuit and does not degrade the performance (when operating below the self-resonant frequency of the diode). In an amplifier with the diode resonated in shunt, diode self resonance can have a very serious effect on bandwidth if the diode self-resonant frequency is at all near the operating frequency. The circuit elements were also made as nearly lumped as possible so that the multiple resonances which occur in any distributed elements would be far removed from the signal frequency. Such multiple resonances tend to increase the reactance or susceptance slope of the diode resonance at the operating frequency and as a result tend to narrow the bandwidth. The pump resonator was lightly coupled at both input and output so that it had little effect on the diode circuit at other frequencies, but it would still give efficient power transfer to the diode at the pump frequency. About 6 mw of incident power was required to drive the diode over its full range of capacitance.

There is another advantage in resonating the diode in series in this particular amplifier. For the desired input frequency and 50-ohm input impedance level, the required diode is conveniently of a size (2.2 $\mu\mu$f zero-bias capacitance) which allows the use of higher Q diodes. If the diode had been resonated in shunt and the same input impedance level was used, a diode with much larger capacitance would have been required, and the Q's for such diodes are considerably lower. The series diode arrangement used also keeps stray capacitance across the diode barrier junction to a minimum and thus prevents degradation of the effective diode C_1/C_0 ratio.[6] (Keeping this ratio large is very important for obtaining large bandwidth.)[2,3] Also, to keep C_1/C_0 large the diode was pumped as hard as possible consistent with good noise figure.

By making the distance from the RF short-circuit block (see Fig. 2) to the diode about a quarter wavelength at the upper-sideband frequency $f_0+f_p=3$ kMc, the diode is made to see a large reactance at that frequency and also at the second harmonic of pump, $2f_p=4$ kMc. In this manner, dissipation of power at these frequencies is prevented.

Experimental Results

Single-Diode Degenerate Parametric Amplifier

A negative-resistance amplifier requires a circulator in order to obtain best performance with a single amplifier unit.[7] Since currently no 1-Gc circulators with adequate bandwidth appear to be available, the amplifier was tested using a 3-db directional coupler to give separate input and output ports as shown in the block diagram (Fig. 4.) The precision directional coupler was fabricated using an interleaving printed circuit construction developed at SRI on another program.[8] It had a residual VSWR less than 1.05 and an isolation in excess of 30 db across the frequency band of interest. The frequency response, corresponding to operation with an ideal circulator, was obtained using the circuit of Fig. 4 in which a reference output level was established with a short circuit in place of the amplifier. In this circuit a short circuit would correspond to an amplifier with a gain of unity. Since a broad-band detector was used and the input signal was modulated, the output power represents the sum of both signal and idler power.

The measured and computed response of the amplifier with and without the shunt resonator is shown in Fig. 5. The value of C_0^s and R_s used in computing the response was established by measurements, the L_d value was as suggested by the diode manufacturer, and the C_1^s value was the value required to give 15-db midband gain (as determined by computation). The other circuit-element values were fixed by the prototype filter used in the amplifier. The agreement between the computed and measured results gives very encouraging verification of the previously developed theory.[2,3] The computed 3-db bandwidth of the single-resonator design is seen to be 81 Mc while the measured bandwidth is practically the same. The 3-db bandwidth of the computed double-resonator response is about 221 Mc, while the corresponding measured bandwidth is about 210 Mc giving a fractional bandwidth of 21 per cent. These values of bandwidth correspond to a C_1/C_0 value of about 0.32.

The noise figure of the parametric amplifier was measured using a circuit shown in the block diagram of Fig. 6. The pads and the tuners were included to insure that the amplifier would be well matched and would see identical impedance when the noise source was on and

[6] The C_0 and C_1 referred to here are the conventional parallel-equivalent average and time-varying capacitance coefficients as defined by the Fourier series $C(t)=C_0+2C_1 \cos(2\pi f_p t)+\cdots$. The computation of the series coefficients, C_0^s and C_1^s in Fig. 1, is discussed elsewhere,[2,3] but $C_0^s/C_1^s=C_1/C_0$ where C_0^s is roughly equal to C_0.

[7] Of course, using two identical amplifier units in a balanced circuit with a 3-db directional coupler can give an equivalent performance, as is shown by results given later in this section.

[8] W. J. Getsinger, S. B. Cohn, and J. K. Shimizu, "Design Criteria for Microwave Filters and Coupling Structures," Stanford Res. Inst., Menlo Park, Calif., Tech. Rept. No. 10, Sec. 2, SRI Project No. 2326, Contract No. DA 36-039 SC-74862; July, 1960.

off. Care was also taken to reduce the leakage of local-oscillator power and excess noise from the receiver to the parametric amplifier. The image response of the receiver was also suppressed. With the filters and tuners required for a reliable measurement it was convenient to make a noise measurement only at the midband frequency. In the measurement the signal and idler frequencies were adjusted to be close enough together so that both passed through the receiver IF amplifier. The procedure included obtaining a noise-figure measurement F_A with the circuit as shown in Fig. 6 and also obtaining a measurement F_B with the noise source placed at Point B and a matched load at Point A. With this noise measurement the actual noise figure of the parametric amplifier is given by the expression

$$F = \frac{F_A}{L_{AB}} - \frac{F_B - 1}{G},$$

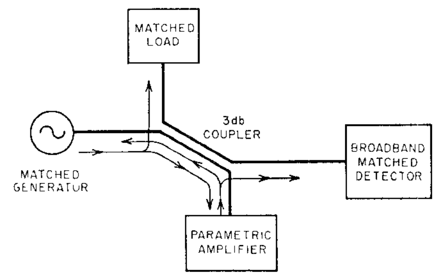

Fig. 4—Block diagram of circuit used in measuring the amplifier frequency response.

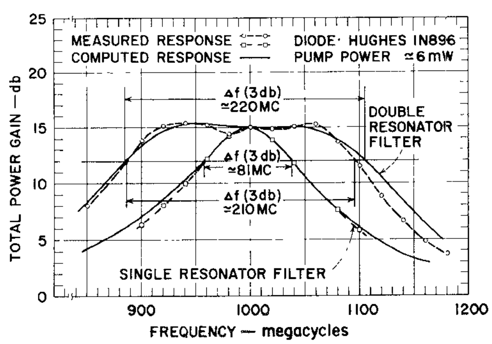

Fig. 5—Frequency response of the single- and double-resonator parametric amplifier.

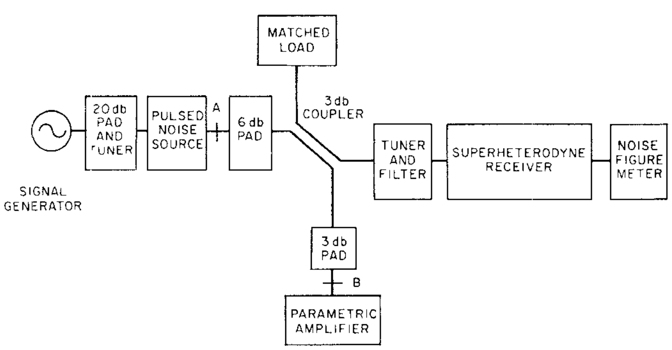

Fig. 6—Block diagram of apparatus for measuring noise figure.

where

- F is the parametric-amplifier double-channel noise figure.
- G is the power-gain *ratio* of the parametric amplifier.
- L_{AB} is the power-loss *ratio* from A to B ($L_{AB}>1$) including the 3-db loss of the directional coupler.

The noise figure F obtained by this procedure is a double-channel noise figure and is the noise figure which would be obtained if the device were operated with an ideal circulator. The computed value of double-channel noise figure for this amplifier was about 0.5 db as compared with measured values which ranged closely about 1.0 db. Considering the manufacturer's tolerances on the excess noise of the noise tube, the tolerances on the noise-figure meter accuracy, and the possible experimental error, the total possible error would be in excess of ± 0.5 db. Under these conditions the 1.0-db measured noise figure appears to be in satisfactory agreement with the 0.5-db theoretical figure.

Uncritical Nature of Degenerate-Amplifier Tuning

In the course of these experiments, it was found that the amplifier-tuning adjustment was not as critical as expected. This was subsequently further verified by examining the analytical expressions and by computing responses for improperly tuned amplifiers. The midband frequency of the amplifier, regardless of the tuning of the two resonators, is found to be always coincident with one-half the pump frequency. This is shown by the single-resonator computed responses in Fig. 7(a) for the same amplifier parameters used for Fig. 5, but with several different pump frequencies. The gain varied because the level of pumping was held fixed (*i.e.*, C_1/C_0 was fixed at 0.316); however, moving the pump frequency from 2.000 Gc to 1.820 Gc, the change in gain was small in the case of the single-resonator amplifier. In the case of the two-resonator amplifier responses shown in Fig. 7(b), the midband gain went up as a result of the effect of the second resonator on the impedance presented to the diode. If the pump power had been reduced (*i.e.*, if C_1/C_0 had been made smaller) for this case the response would probably have had a flatter top.

A related effect is the uncritical nature of the adjustment of the series resonant circuit. As shown by the single-resonator amplifier computed responses in Fig. 8, the midband frequency is not altered by rather large changes in the length of the inductive element and consequently the resonant frequency f_0 of the series circuit. However, to maintain the same value of midband gain, the C_1/C_0 ratio for the diode was adjusted in each case as indicated in the figure. It is interesting to note in the figure that in order to maintain 15-db gain the diode must be pumped somewhat harder (*i.e.*, C_1/C_0 must be larger) when the diode resonator is detuned. The detuned cases requiring larger C_1/C_0 values are seen to be accompanied by a somewhat larger bandwidth.

Of practical interest, too, in these amplifiers is the degree to which the diodes, with their spread of values, can be interchanged in a particular unit. To verify this under controlled conditions a set of frequency responses of a two-resonator amplifier (the same as for Fig. 5) were computed for a range of values of average capacitances. As shown in Fig. 9, only the bandwidth and the ripple amplitude of the response vary; the midband frequency remains fixed. The change in bandwidth is not large, and should not be a serious problem. Note that if C_0 is smaller than its design value, the amplifier requires less pump power (*i.e.*, a smaller C_1/C_0 value) for the same midband gain than when C_0 has its proper value. However, if C_0 is larger, C_1/C_0 must be increased to maintain the same gain.

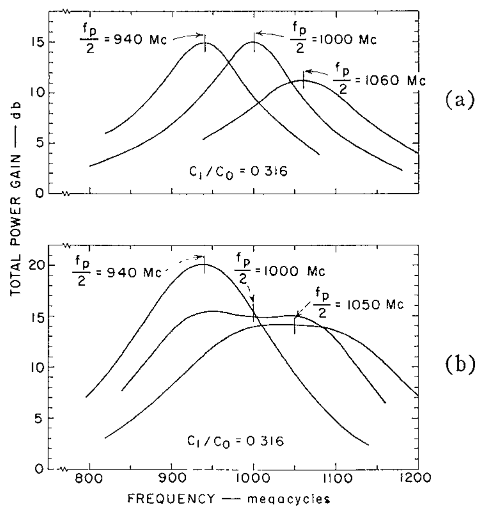

Fig. 7—Frequency-response dependence on pump frequency (computed results). (a) Single-resonator amplifier. (b) Double-resonator amplifier.

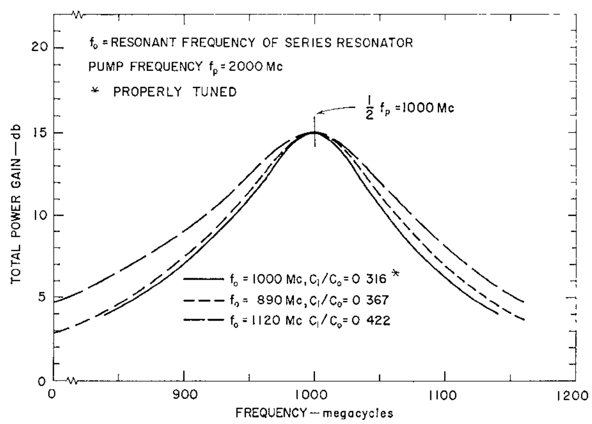

Fig. 8—Frequency-response dependence on the inductive element in the series resonant circuit (computed results).

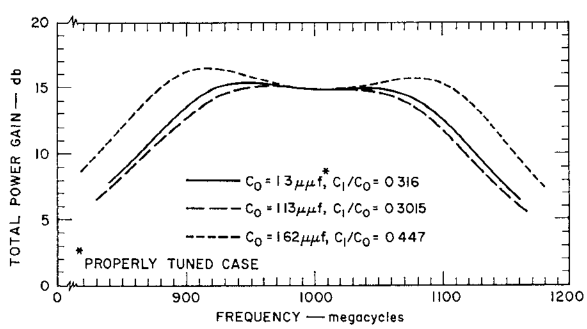

Fig. 9—Frequency-response dependence on the average diode capacitance in two-resonator amplifier (computed results).

The reason for the relatively uncritical nature of the degenerate parametric amplifier tuning can be seen by examining the power-gain expression. For this discussion, diode loss resistance R_s may be neglected, so the expression for power gain is simply[2,3]

$$|\Gamma|^2 = \frac{|Z_b Z_b' + X_{12}X_{21}|^2}{|Z_b Z_b'^* - X_{12}X_{21}|^2}, \tag{1}$$

where Z_b is the impedance seen looking back into the input filter as shown in Fig. 1 and Γ is the voltage reflection coefficient between the circulator and the amplifier. The asterisk indicates that the conjugate is to be taken. In (1), Z_b is evaluated at the signal frequency f, while Z_b' is the same impedance evaluated at the idler frequency $f' = f_p - f$, where f_p is the pump frequency. The quantity $X_{12}X_{21}$ in (1) is $1/[(2\pi C_1^s)^2 ff']$.

The power gain $|\Gamma|^2$ is most sensitive to variations in the difference expression in the denominator of (1). Study of the denominator shows that to a first approximation the imaginary part of the quantity within the magnitude signs will be unaffected by moderate mistuning of the diode resonator, while the real part will be only slightly affected.[5] Further, the expression obtained shows that the small change in the real part can be compensated for by a small increase in C_1/C_0. These results are consistent with the computed results in Figs. 7 to 9 and with our experimental observations.

Balanced Amplifiers

The single-diode parametric amplifier becomes a use ful device when separate input and output ports can be provided. Amplifier bandwidths of the order of 20 per cent exclude the use of currently available circulators operating in the vicinity of 1 Gc, but an alternative is to use a pair of single-diode amplifiers with a 3-db directional coupler as shown in Fig. 10. This circuit, unlike one using a circulator, is bilateral, since it can amplify equally well in both directions. To prevent distortion of the frequency response it is required that the individual amplifier units have nearly identical impedance characteristics and that each work into a well matched load. This requirement means that the directional coupler and the terminations at *both* input and output ports must not produce an appreciable VSWR at the amplifier unit. The signal-frequency components will automatically add at the output port and cancel at the input port of the 3-db directional coupler if the input impedances of the two amplifiers are identical; however, for the idler-frequency signals the relative phases of the pump signals must be adjusted to make the idler outputs add at the output port. Making this adjustment at one frequency will provide proper idler phases at all frequencies.

The feasibility of the balanced operation is demonstrated by the experimental frequency response in Fig. 11 for the balanced amplifier shown in Fig. 10. A second amplifier unit identical to the one discussed above was used together with the precision 3-db directional coupler mentioned earlier. Both the transmission-power gain, which is the useful output, and the reflection-power gain are shown. The reflection-power gain (the db ratio of power reflected to power incident at the input of the balanced amplifier) is indicative of the balance between the two single-diode units. The transmission-power gain with a bandwidth of 210 Mc (*i.e.*, 21-per cent bandwidth) is a good replica of the single-amplifier frequency response (see Fig. 5), and the reflection power gain, except within a narrow band of frequencies, is always less than 0 db, indicating a satisfactory balance between amplifiers. The power gain includes both signal and idler outputs, since both carry the signal modulation.

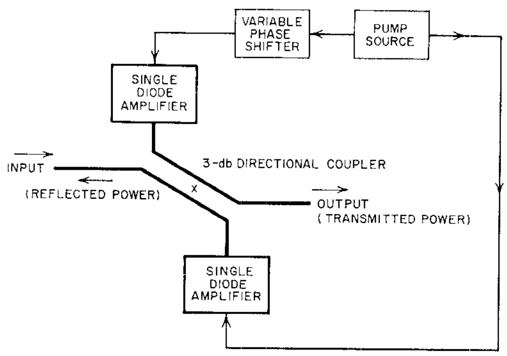

Fig. 10—Block diagram of two-port balanced amplifier.

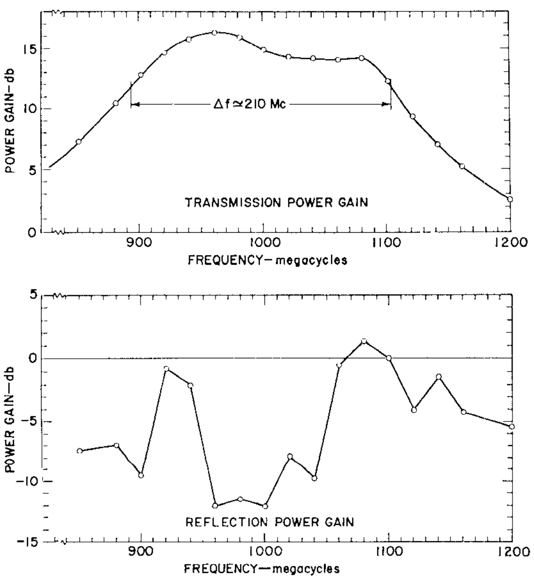

Fig. 11—Frequency response of balanced amplifier.

Cascaded Balanced Amplifier

The operating characteristics of the balanced amplifier being satisfactory, it seemed worth while to try to obtain even greater bandwidth by cascading several balanced amplifiers which are adjusted so that the individual pass bands are contiguous as shown in Fig. 12. To understand this circuit, note that the minimum value of gain of a degenerate amplifier is unity outside the pass band because the reflection from the amplifier corresponds to that of a pure reactance. Thus, when the first balanced amplifier doesn't amplify it merely passes the power on to the second amplifier. To obtain stable operation, the design gain must not be too large in order to provide a safety margin against distorting the response of the individual units and against reflections which could lead to oscillations.

A two-stage cascaded balanced amplifier was assembled and tested. The first stage was the balanced amplifier described earlier with a midband frequency of 1 Gc and the second stage was of similar construction but scaled for operation at 1320 Mc. Unfortunately, the 3-db directional couplers had not originally been in-

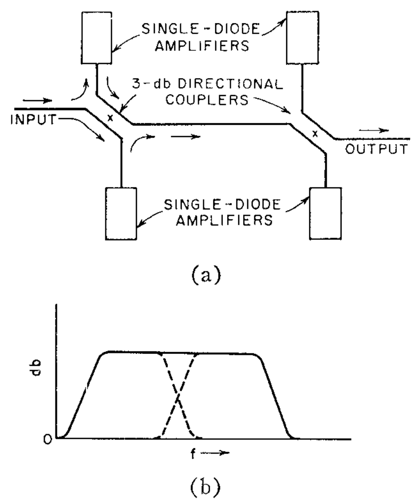

Fig. 12—Cascaded balanced amplifiers. (a) Balanced amplifier and tandem connection. (b) Composite frequency response.

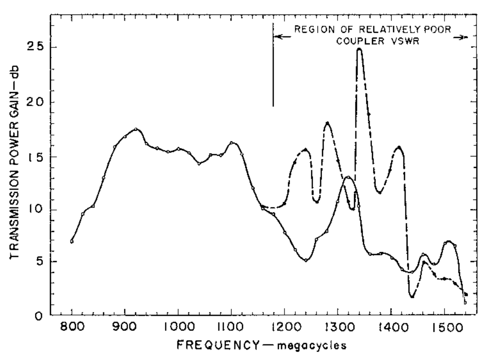

Fig. 13—Frequency response of cascaded balanced amplifiers.

tended for use up into the frequency range covered by the second amplifier. Therefore, because of reflections from the directional couplers, the frequency response of the second amplifier did not attain the flat broad-band shape of the first amplifier.

The measured frequency response of the cascaded amplifier is shown in Fig. 13. The heavy solid line represents the initial response after adjusting each balanced amplifier separately. Note that the shape of the response in the range of the well-matched 1000-Mc stage is a fair replica of the single-balanced-amplifier response in Fig. 5. The response in the range of the second stage is erratic because of the reflections from the directional couplers. An attempt to compensate for these reflections in the 1320-Mc second stage by adjusting the diode bias and the pump amplitude and phase finally resulted in the response at the higher end shown by the dashed line. The usable bandwidth, say for which the gain is over 10 db, extends from 850 Mc to 1420 Mc, which is approximately a 50-per cent bandwidth.

Discussion

This work has shown that practical, nearly optimum, single-diode degenerate amplifiers can be designed according to previously developed theory.[2,3] The construction used for the particular example had a number of significant features which are particularly advantageous for frequencies up to the self-resonant frequency of the diode. The excellent agreement between theory and experiment gave strong evidence of the validity and practical usefulness of the design methods which were used.

It was found that the C_1/C_0 ratio for practical diodes can be 0.32 or slightly larger. The measured and computed responses corresponded to a C_1/C_0 ratio of this value. Pumping the diode harder (which makes C_1/C_0 larger), made it possible to raise the midband gain as high as 20 db; however, above about 16-db midband gain the noise figure began to increase. Thus, $C_1/C_0 = 0.32$ appears to be about the maximum practical value for low-noise operation, at least for the case of Hughes 1N896 diodes. As was previously mentioned, making C_1/C_0 as large as permissible is necessary in order to obtain maximum bandwidth.

The investigations made of the performance of improperly tuned degenerate amplifiers showed that there is a tendency for detuning effects to cancel out. Because of this, if such amplifiers are designed so that the prototype design does not require quite all of the pump power that the diode can take, some latitude in diode parameters can be tolerated which would only require a readjustment of the pump power in order to obtain the desired midband gain. For a related reason, since the amplifier response always tends to center at exactly half the pump frequency, small adjustments in the midband frequency can be made by shifting the pump frequency.

Balanced operation of amplifiers giving separate input and output terminals was shown to be practical, although most commercial 3-db couplers would have too large a VSWR to be satisfactory for such purposes. It is, nevertheless, possible to build couplers with an adequately low VSWR, as is proved by the one fabricated for this particular application. Balanced operation using two amplifiers and a well matched 3-db coupler was seen to give wideband performance comparable to what would be obtained by a single amplifier with a very well matched, wideband circulator having very low loss. Until such circulators are available for all frequency bands of interest, the balanced amplifier appears to provide an attractive alternative. The chief limitation of this form of amplifier is that only well matched terminations can be used at both its input and output terminal. This condition is required not only to prevent oscillations due to multiple reflections but also to insure that each amplifier unit itself sees a well matched load so that its frequency response will not become distorted.

The results from the cascaded balanced amplifiers showed that this scheme does present a useful way for obtaining extremely wide bandwidth. Although the response at the higher end of the band was erratic due to the coupler VSWR with couplers which are well matched over the entire frequency band the response should be as desired. This type of amplifier has some advantages over the traveling-wave types of amplifiers in that it will give reasonably good gain with very wide bandwidth while using relatively few diodes. With this type of amplifier care must be taken, however, that the reflections of the terminations and the couplers are not too large, since this will distort the response and might cause oscillation.