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Improved Microwave Repeaters for Hungarian All-Solid-State Communications Systems

T. BERCELI, G. HAMMER, F. RÁKOSI, G. REITER, AND S. SZÉNÁSI

Abstract—**Improvements in microwave repeaters are outlined utilizing new circuit concepts for the receiver, transmitter, and branching-filter systems. In this way, a 1.5-dB decrease in the receiver noise figure and a 1-dB reduction in the attenuation of branching-filter systems are obtained with a simultaneous 2-dB overall increase of transmitter-multiplier and upconverter efficiencies, resulting in a higher signal-to-noise ratio. An improvement is achieved in the AM-to-PM conversion and group-delay characteristics, too. The new circuits have been developed for Hungarian all-solid-state communications systems operating in the 4-, 6-, and 8-GHz frequency bands.**

I. INTRODUCTION

IMPROVEMENTS in microwave repeaters are outlined utilizing new circuit concepts for the receiver, transmitter, and branching-filter systems. In this way, an essential reduction in the receiver noise figure and in the attenuation of branching-filter systems is obtained with a simultaneous increase of transmitter-multiplier and upconverter efficiencies, resulting in a higher signal-to-noise ratio. The new circuits have been developed for Hungarian all-solid-state communications systems operating in the 4-, 6-, and 8-GHz frequency bands. Thus the paper also gives a report on present state-of-the-art techniques in Hungary.

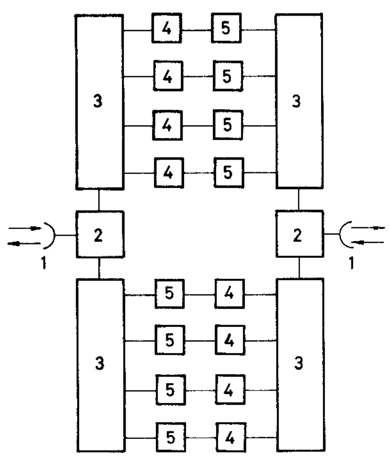

Fig. 1. Schema of a repeater. 1: Antenna. 2: Transmitter–receiver combiner. 3: Branching filter. 4: Receiver. 5: Transmitter.

The schematic of a repeater is shown in Fig. 1. At the repeater stations, one aerial is used in one direction, transmitting four transmitter signals and receiving four receiver signals simultaneously. The transmitter and receiver signals of the same polarization are combined or selected by transmitter–receiver combiners. The transmitter or receiver signals are combined or selected by branching filters.

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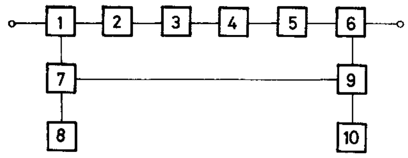

Fig. 2. Block diagram of receiver and transmitter. 1: Receiver mixer. 2: IF filter. 3: IF group-delay equalizer. 4: IF main amplifier. 5: IF limiter. 6: Upconverter. 7: Shift mixer. 8: Shift oscillator. 9: Power splitter. 10: Heterodyne generator.

The block diagram of a receiver and transmitter is shown in Fig. 2. The received signal is transposed to 70 MHz by the receiver mixer. This IF signal is filtered, group-delay equalized, and amplified in the receiver, then the signal is passed into the transmitter. There the IF signal is limited and transposed into the microwave band by an upconverter. The heterodyne signal for the upconverter is provided by a high-power heterodyne generator through a power splitter. A fraction of this power is led to the shift mixer where this signal is mixed with the signal of a shift oscillator and, in this way, the heterodyne signal for the receiver mixer is obtained.

The repeaters are applied in point-to-point telecommunications links. Most of the baseband noise in a multihop link originates in the repeaters. Thus the most important characteristic of a repeater is the noise contribution taking part in the overall baseband noise of the link. The noise contribution can be derived from the thermal and intermodulation noises.

The thermal-noise contribution is dependent on the radio-frequency signal-to-noise ratio obtained in a single-hop link. The thermal-noise contribution is decreased by increasing the transmitter power and by decreasing the receiver noise figure and the attenuation of the branching-filter systems, supposing that the other parameters are constant.

The intermodulation noise is mainly dependent on amplitude and group-delay variations and on AM-to-PM conversion. These characteristics are improved by the developed new circuits, too.

II. RECEIVER

In order to achieve a low noise figure, reactive terminations are needed at the image, sum, and harmonic frequencies in the receiver mixer [1]–[3]. Further, to obtain low-amplitude and group-delay response fluctuations, the reactive terminations should be nearly frequency independent. Another important parameter is the small size requirement in order to reduce overall equipment size.

Receiver mixers utilizing a circulator or a hybrid arrangement are well known [4]. Both of these arrangements have a relatively high space requirement, and furthermore the electrical distance between the filters and the diodes is long, resulting in frequency-dependent reactive terminations at the image, sum, and harmonic frequencies. Besides, in the former arrangement, the circulator loss will increase the receiver noise figure.

A new receiver-mixer arrangement [5] was developed which offers an optimum solution of the previously stated problems. The simplified cross section of the new receiver mixer is shown in Fig. 3. In this arrangement, the receiver and heterodyne signal filters are combined by a common-cavity resonator and the circuit comprising the mixer diode is connected to this common-cavity resonator. If a suitable design is applied [6], the receiver branch of this combined filter will show a maximum flat or a Chebyshev-type response in a wide passband, while obtaining a narrow passband in the heterodyne branch of the filter, as can be seen in Fig. 4. This figure shows the VSWR characteristics of a combined filter at a port formed on the common cavity.

In order to obtain these characteristics, the resonant frequency of the common cavity is adjusted to the reciever-band center frequency f_{s0}. The cavity coupled directly to the common cavity in the heterodyne branch is tuned to the heterodyne frequency f_h. The resonant frequency of the next cavity is equal to $2f_h - f_{s0}$. The common cavity and the latter two cavities form a three-resonator filter in which two resonators are detuned in opposite directions from the heterodyne frequency with the same value. Consequently, the heterodyne branch exhibits a narrow passband around f_h.

The filter-cavity resonators are formed by inductive irises. The length of the cavities is nearly $\lambda_g/2$ except the common cavity which has a length of λ_g approximately. At the center part of the common cavity, a coupling slot is cut into one of the waveguide broad sides. The common cavity is coupled to a reduced-height waveguide by this slot, as can be seen in Fig. 5.

The mixer diode is placed in the reduced-height waveguide parallel to the electric lines. One end of the diode is connected to a low-pass filter preventing RF leakage. The other end of the diode is connected to a short coaxial line used as a series-matching element. The length of this line can be changed by a step-by-step method. A movable waveguide short circuit serves as a parallel-matching element.

At the image frequency, both branches of the combined filter have a stopband; therefore, total reflection from the filter is observed in the reduced-height waveguide. With a suitable distance between the diode and coupling slot, a nearly short-circuit termination for the image frequency is obtained at the diode junction. As this distance can be made quite short in our arrangement, the image termination is approximately frequency independent. Reactive terminations for the sum and harmonic frequency signals are provided by a waffle-iron-type harmonic filter inserted at a convenient place between the mixer diode and coupling slot.

The impedance of the reduced-height waveguide is equal to the optimum generator impedance, yielding a minimum noise factor. The IF signal is led from the diode through the low-pass filter to an IF preamplifier mounted on the mixer to get a direct connection to the diode. The preamplifier noise figure is 1.8 dB, gain equals 30 dB, and input resistance can be varied between 100 and 150 Ω.

In the 4-GHz mixer, a GaAs Schottky-barrier diode is

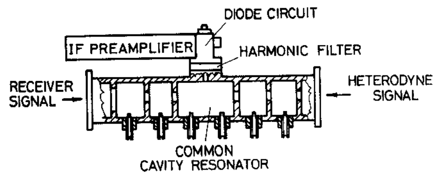

Fig. 3. Cross section of receiver mixer.

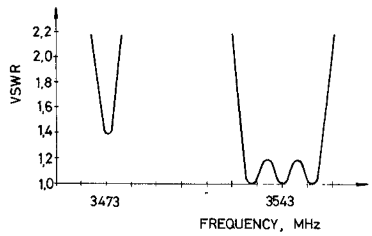

Fig. 4. VSWR characteristics of combined filter.

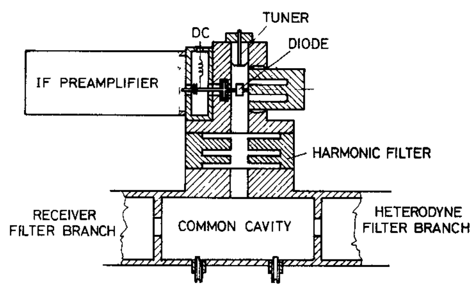

Fig. 5. Cross section of mixer circuit.

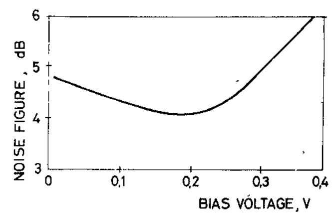

Fig. 6. Noise figure as a function of bias voltage.

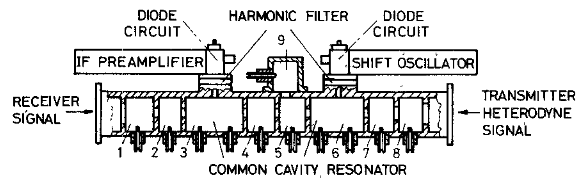

Fig. 7. Cross section of combined receiver–shift mixer.

applied. This diode exhibits a noise figure of 5.6 dB in a conventional mixer with resistive terminations at the image, sum, and harmonic frequencies. In our arrangement, the noise figure is reduced due to the reactive terminations at the useless frequencies. The best noise figure is 4.1 dB including filter loss of 0.4 dB. The noise figure is dependent on bias voltage as shown in Fig. 6. The minimum noise figure is obtained at 0.2-V bias voltage with a heterodyne signal level of 5 mW.

Efforts were made to reduce the intermodulation noise contribution of receivers by decreasing the AM-to-PM conversion. The source of AM-to-PM conversion is generally in the IF main amplifier. By applying only *RC* coupling networks all over the IF amplifiers, AM-to-PM conversion is diminished to 0.2–0.3°/dB for the complete IF section [7].

In our microwave equipments, the receiver heterodyne signal is derived from the transmitter heterodyne signal by a shift mixer. An arrangement similar to the receiver mixer is used for the shift mixer too. By combination of the filters of the receiver and shift mixers, a combined receiver–shift mixer has been developed as given by the simplified cross section shown in Fig. 7. The receiver signal is given on one end of the combined 3-section filter, while the transmitter heterodyne signal is coupled to the other end of this filter. The mixer diode circuits are connected to the reduced-height waveguides coupled to the common-cavity resonators.

The design of the 3-section filter is similar to that of the combined filter in the receiver mixer. Cavities 1, 2, 3 are tuned to the receiver-band center frequency f_{s0}. The resonant frequency of other cavities is adjusted as follows: cavity 4 to f_{rh}; cavity 5 to $2f_{rh} - f_{s0}$; cavity 6 to f_{rh}; cavity 7 to f_{th}; and cavity 8 to $2f_{th} - f_{rh}$, where f_{rh} is the receiver heterodyne frequency and f_{th} is the transmitter heterodyne frequency. Auxiliary cavity 9 is used to increase the attenuation between the shift and receiver mixers. This attenuation is more than 90 dB.

III. TRANSMITTER

The transmitter signal is provided by a high-level upconverter which serves for heterodyning the microwave heterodyne signal and the modulated IF signal. The efficiency of the high-level upconverter is of primary importance, especially in all solid-state systems where transmitter power is given directly by the upconverter. Thus any increase in upconverter efficiency will result in an increased signal-to-noise ratio.

The known arrangements used for high-level upconverters contain a circulator or a hybrid [8]. Both arrangements have a relatively low efficiency as some resistive load will be present at the useless combination frequencies. Further, the long electrical distance between the filters and the diodes is also a disadvantage.

A high efficiency requires a definite electrical condition at the ports, substituting the diode junction at different frequencies [9]. At these ports, a specified impedance should be present at the intermediate, heterodyne, and useful sideband frequencies; on the other hand, an open circuit is required at the unwanted sideband and harmonic frequencies. From the latter, only the first unwanted sideband and the second harmonic frequency is of importance; this means that for reaching a high efficiency, the open circuit at the diode junction has only to be exhibited at these frequencies.

The electrical condition for high efficiency can be

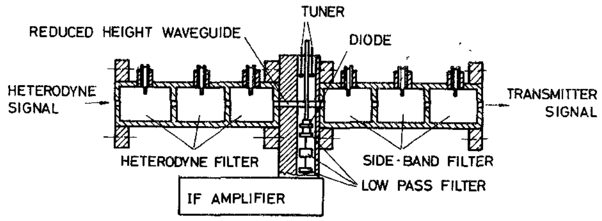

Fig. 8. Cross-sectional view of upconverter.

established by a new upconverter arrangement [10] which has a simplified cross section, as shown in Fig. 8. The varactor diode is placed into a coaxial transmission line; at one end of this line, a tuning element, and at the other end a low-pass filter is used. An IF amplifier mounted on the upconverter is coupled directly to the low-pass filter. The coaxial line containing the varactor is coupled to a reduced-height waveguide forming a connection with the heterodyne and sideband filters. These bandpass filters establish ports for the heterodyne and transmitter signals.

A design method for varactor-diode upconverters was developed based on a large-signal approximation. An inverse square-root function is applied to describe the capacitance versus voltage relationship of the varactor diode, and charge storage is neglected. Three ports are established, namely for the heterodyne, IF, and upper sideband (sum) frequencies. At any other frequency, open circuits are supposed to be terminations. Thus three nonlinear equations are set up giving the relationships between the currents. Based on this system of equations, all data for high efficiency are obtained: impedances and levels at three frequencies, bias voltage, dissipation, etc.

The arrangement shown in Fig. 8 makes circuit adjustment possible according to the preceding method. The impedances at the heterodyne and upper sideband frequencies are rather low (10–30 Ω) and the reduced-height waveguide serves matching. The heterodyne and sideband filters are used not only for filtering but also for impedance transformation. For fine adjustment, the position of the diode in the coaxial line can be varied by moving the inner conductor.

By the new arrangement, a 40-percent efficiency has been attained with an output power of 400 mW at 8 GHz, and a 35-percent efficiency was recorded with an output power of 1.2 W at 4 GHz.

An important task is to achieve a low AM-to-PM conversion. With decreasing IF voltage across the diode, AM-to-PM conversion is decreased; however, output power is decreased too, as can be seen in Fig. 9. A good compromise is obtained at an IF voltage of 9.5 V giving an output power of 1.1 W and an AM-to-PM conversion of 3°/dB. In the case of Fig. 9 the input power is 3.3 W at 4 GHz. The intermediate frequency equals 70 MHz. The IF impedance of the diode is about 400 Ω. Bias voltage is −10 V. Junction capacitance at −6 V is between 1 and 2 pF, cutoff frequency is 70 GHz at −6 V, and breakdown voltage is −60 V.

The upconverter and receiver mixer have been cascaded, and between the IF input and output of this setup, the amplitude and group delay versus frequency characteristics were measured with the results shown in Fig. 10. It can be seen that the 0.2-dB bandwidth is more than 24 MHz, and group-delay variation in this band is less than 3 ns.

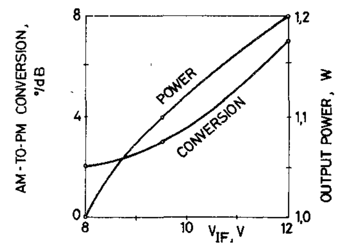

Fig. 9. AM-to-PM conversion and output power as functions of IF voltage across the diode.

The heterodyne signal is derived from a crystal-controlled oscillator by amplification, filtering, and frequency multiplication, as can be seen in Fig. 11. The low level of heterodyne noise is assured by a 125-MHz crystal oscillator and a narrow-band filter. At 250 MHz, 20-W power is obtained at the amplifier output and this is fed to a frequency-multiplier chain utilizing step-recovery diodes. The multiplication factor is 16, 24, or 32 depending on the output frequency.

High-level frequency-multiplier chains only comprise doublers and triplers because of dissipation problems. In order to increase the efficiency, combined multipliers have been developed unifying two doubler or tripler stages [11]. The multiplication factor of these combined units may thus be 4, 6, or 9.

In the case of a multiplication by 4, the combined multiplier comprises two diodes operating as doublers. The diodes are directly coupled and this increases efficiency by decreasing the coupling losses between the two diodes. The arrangement of a combined multiplier is shown in Fig. 12. The input-signal frequency is 250 MHz; therefore, this quadrupler contains mostly lumped elements, and only the output resonator is of the coaxial type. A network comprising three inductors and three capacitors is inserted between the input and first diode D_1. This network yields impedance transformation with low-pass transmission properties. C_3 is variable in order to provide a series-tuning possibility for diode D_1. Furthermore, a parallel tuning is presented by capacitor C_4. Between the two diodes, a series-resonant circuit is inserted, being tuned to the second harmonic of input signal by C_5. The coupling between the second diode D_2 and the output cavity is variable by screw S_1. The resonant frequency of this cavity is varied by screw S_2. The output coupler is a loop.

The first diode has a cutoff frequency of 40 GHz, a junction capacitance of 10 pF, and a breakdown voltage of −90 V. The data of the second diode are 60 GHz, 5 pF, and −80 V, respectively. With these diodes, the quadrupler provides 11.2-W output power at 1 GHz from an input signal of 20 W. Thus the efficiency is 56 percent.

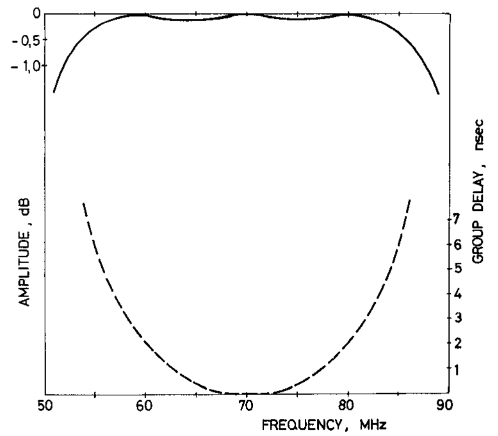

Fig. 10. Overall amplitude and group-delay characteristics of up-converter and receiver mixer.

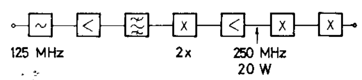

Fig. 11. Block diagram of heterodyne generator.

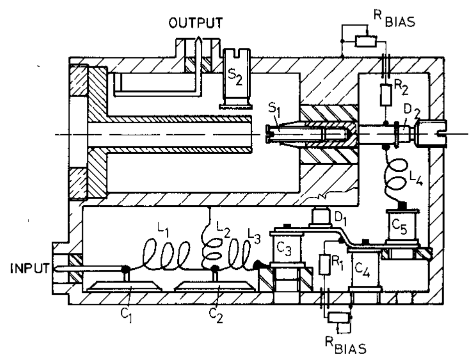

Fig. 12. Arrangement of combined quadrupler with 250-MHz input frequency.

The same principle was applied to develop a combined quadrupler with an input frequency of 1 GHz. The simplified cross section of this circuit is shown in Fig. 13. The first diode has a cutoff frequency of 60 GHz, a junction capacitance of 5 pF, and a breakdown voltage of -80 V. The data of the second diode are 80 GHz, 2 pF, and -45 V, respectively. With these diodes, the circuit provides 4.2-W output power at 4 GHz from an input signal of 11.2 W. Thus the efficiency is 37.5 percent.

The combined quadruplers were cascaded. Then the input frequency is 250 MHz and the output frequency is 4 GHz. For this case, the output power versus input power characteristics are shown in Fig. 14. At 20-W input power, the output power is 4.2 W, so the overall efficiency is 21 percent. The characteristics tend to saturate at an input power of 30 W. The relative bandwidth was also measured and found to be 1.5 percent with an input power of 20 W. The temperature dependence of the whole multiplier chain is 1 dB between 0°C and 50°C.

In order to compare the results, four separate doublers were constructed with the diodes applied in the combined circuits. This conventional multiplier chain delivers 2.7-W

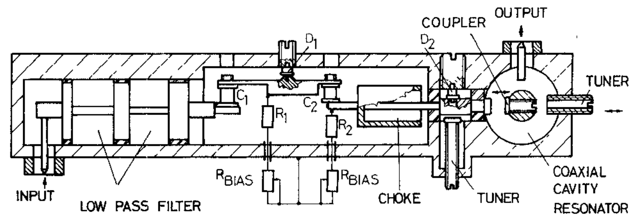

Fig. 13. Cross section of combined quadrupler with 1-GHz input frequency.

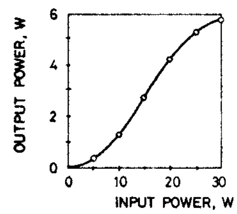

Fig. 14. Output-power versus input-power characteristics of two cascaded combined quadruplers.

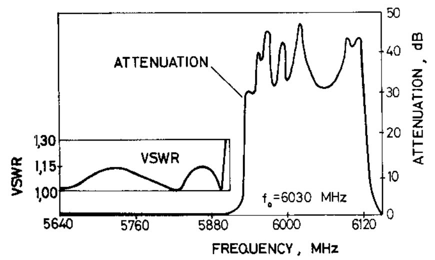

Fig. 15. Attenuation and VSWR response of bandstop filter.

output power at 4 GHz when the input-signal power is 20 W at 250 MHz. Thus the overall efficiency is 13.5 percent. Similar results are published [12].

The doublers were designed by using computer programs. The programs take into account the nonlinear junction capacitance, nonlinear series resistance, charge storage, and step-recovery effect. A period of input signal is divided into sections, and in every section analytical functions are used to describe diode behavior [13].

IV. BRANCHING FILTER SYSTEMS

The transmitter and receiver combiner contains a circulator, two bandstop filters, and isolators. The bandstop filters are low-loss types having a wide passband with abrupt cutoff and elliptical attenuation characteristics. The width of the transition band between passband and stopband is 26 MHz. In the 300-MHz-wide passband, highest VSWR is 1.1, and in the 190-MHz-wide stopband, lowest attenuation is 30 dB. The measured attenuation and VSWR response is shown in Fig. 15.

The transmitter and receiver signals of four microwave channels are combined and selected by branching filters. An important equipment requirement is the reduction of branching-filter loss as this loss would increase the attenuation between transmitter and receiver and thus decrease the signal-to-noise ratio in a hop. Another important requirement is the low group-delay fluctuation in the passband while maintaining the specified stopband attenuation. This will decrease the intermodulation noise.

In well-known branching-filter arrangements, the microwave channels are separated by bandpass filters connected to circulators [14]. The attenuation of this arrangement is substantially increased by the circulator losses, as a signal will pass eight circulator paths between transmitter and receiver. Arrangements not utilizing circulators for channel separation have also been published [15], where a microwave channel is selected by a combination of a bandstop filter and a bandpass filter.

A new branching-filter arrangement was developed comprising only a few circulators and thus having a low attenuation. Instead of bandpass filters, two diplexers, each separating two channels, are utilized. Two diplexers and one circulator are needed for separating four microwave channels, as is shown in Fig. 16. The microwave-channel frequencies are designated by f_1, f_2, f_3, and f_4.

The diplexer consists of two waveguide filters that contain inductive irises forming the cavity resonators [16]. The diplexer cross section in a symmetry plane parallel to the broad sides of waveguides is shown in Fig. 17. The first resonator of the filters is common. The other cavity resonators of the filters are connected to this common-cavity resonator by means of the iris a and the coupling hole b, respectively. The coupling hole b is formed on one of the narrow sides of the common-cavity resonator. In addition, port 1 of the diplexer and an auxiliary-cavity resonator are coupled to the common-cavity resonator through the iris c and coupling hole d, respectively. The latter is formed on the other narrow side of the common-cavity resonator.

The equivalent circuit of the diplexer is shown in Fig. 18. Here the irises and coupling holes are represented by inverters, and the cavity resonators by series-resonant circuits. The common-cavity resonator and the auxiliary-cavity resonator are represented by a 1-port between points A–A. Two Foster equivalent circuits can be derived for this 1-port [Fig. 19(a) and (b)] using the pole-zero pattern of Fig. 19(c). It can be seen that the common cavity has two resonant frequencies (f_{01} and f_{02}) due to the auxiliary-cavity resonator coupled to it. The design of the diplexer is based on the equivalent circuit [17].

Both microwave channels will pass the common cavity at port 1. Both resonant frequencies of this cavity correspond to the channel center frequencies. The other cavities of the diplexer are tuned to the proper channel center frequencies. The microwave channel signals to be separated are given on port 1. One of the channels will be dropped at port 2, and another channel at port 3, while any further channels are reflected at port 1. When combining two microwave channels, the signal flows are opposite.

The auxiliary-cavity resonator also has the effect of producing an attenuation pole at the resonant frequency (f_p) of the auxiliary resonator, being between the two channels to be separated. Applying additional auxiliary cavities at ports 2 and 3, attenuation poles at both sides of the passband are obtained. The additional auxiliary cavities in a very simple coaxial construction have been built.

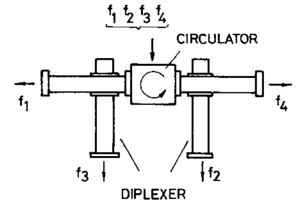

Fig. 16. Branching filter system.

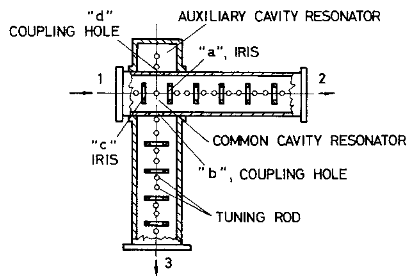

Fig. 17. Cross section of diplexer.

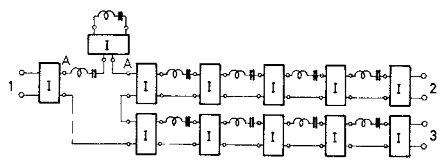
Fig. 18. Equivalent circuit of diplexer (I means inverter).

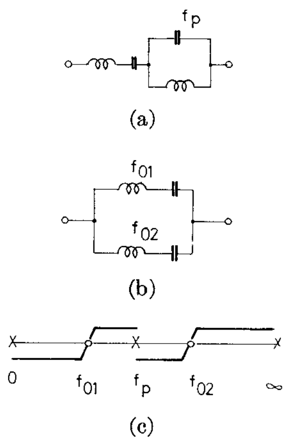

Fig. 19. Foster equivalents of common and auxiliary cavities.

The arrangement, as shown in Fig. 17, has been developed for the 4- and 6-GHz frequency ranges. For instance, a diplexer operating in the 4-GHz range exhibits 50–55-MHz-wide passbands and the VSWR in these bands has a fifth-degree Chebyshev response with a maximum value of 1.2. The measured diplexer attenuation responses are shown in Fig. 20. The average group-delay fluctuation within a single microwave channel is around 2 ns.

The same principle was used to construct simple rectangular-waveguide filters having attenuation poles at both sides of the passband [18]. In this case, the same stopband attenuation requirements are satisfied, even with a wider passband, and the group-delay fluctuation within

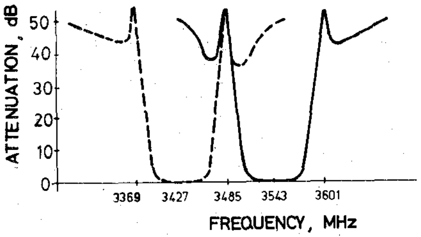

Fig. 20. Attenuation responses of diplexer.

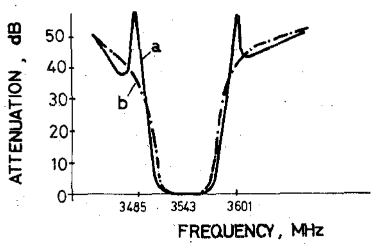

Fig. 21. Attenuation response of filters. *a*: With attenuation poles at the sides of passband. *b*: Without attenuation poles at the sides of passband.

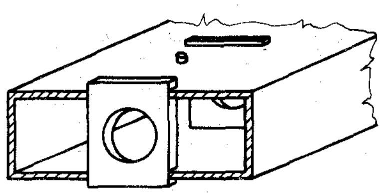

Fig. 22. A cavity formed by irises in a rectangular waveguide.

the microwave channel is thus lowered. The attenuation response of a filter with attenuation poles is plotted in Fig. 21. For comparison, there is also shown the attenuation response of a filter without attenuation poles. In both cases, the number of cavities used in the filters is the same.

Every filter is constructed in rectangular waveguides. The cavities are formed by inductive irises with uniform width, as can be seen in Fig. 22. In the center of the irises, round holes with different diameters are drilled which determine the iris susceptances. The filters are made of Invar The irises are soldered into slots cut in the broad sides of the waveguide. The inner surfaces are plated by silver and palladium.

V. CONCLUSIONS

Improvements in microwave repeaters were outlined utilizing new circuit concepts for the receiver, transmitter, and branching-filter systems. In this way, a reduction of 1.5 dB in the receiver noise figure and a decrease of 1 dB in the attenuation of branching-filter systems were obtained with a simultaneous 2-dB overall increase of transmitter multiplier and upconverter efficiencies. An improvement was achieved in the AM-to-PM conversion and group-delay characteristics, too. Circuit schemes, cross sections, and characteristics were shown for the developed arrangements. Finally, the developed circuits are also shown in photos. The combined receiver-shift mixer is shown in Fig. 23, the upconverter in Fig. 24, and the diplexer in Fig. 25.

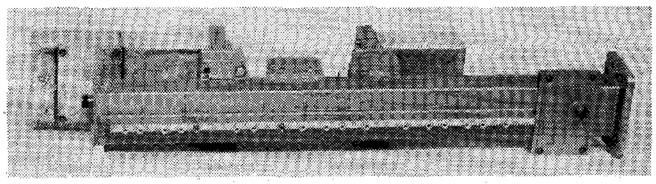

Fig. 23. Combined receiver-shift mixer.

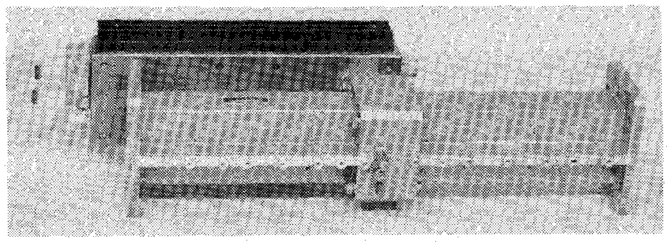

Fig. 24. Upconverter.

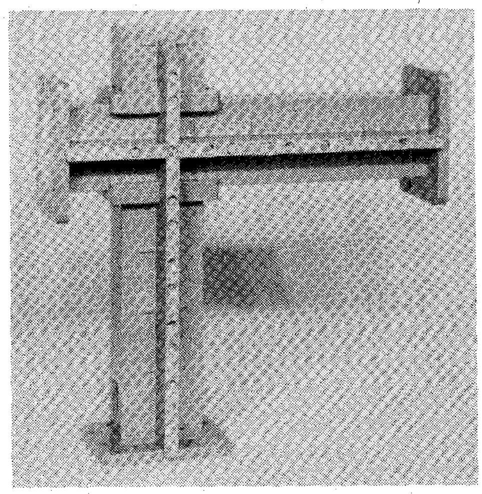

Fig. 25. Diplexer.

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A Low-Cost Multiple-Channel 12-GHz Receiver for Satellite Television Broadcasting

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Abstract—**The design of a low-cost FM-microwave satellite-ground-station receiver is described. It is capable of accepting 12 contiguous color-television equivalent-bandwidth channels in the 11.72–12.2-GHz band using a wide-band FM format and frequency division multiplexing (FDM) of the channels. Each channel has 36 MHz of usable bandwidth with a 4-MHz guard band and provides a CATV compatible output. The overall system specifications are first discussed. Then consideration is given to the design, fabrication, and evaluation of the different subsystems in the receiver.**

I. INTRODUCTION

DURING the past few years the use of artificial satellites located in geostationary equatorial orbit for direct television broadcast to augmented TV sets or limited rediffusion installations on earth has become a matter of increasing interest. If a multicarrier satellite transponder having a large number of TV equivalent-bandwidth channels were used in conjunction with local CATV distribution networks it would be possible to economically disseminate large amounts of information to many subscribers. One potential area of application is in the field of education [1]. In order to implement systems with hundreds of points of reception, low-cost ground-station receivers will be required. In this paper the design construction and performance of a candidate receiver is considered.

Based on a report of the results of the 1971 World Administrative Radio Conference [2] and the existing usage of the various services discussed there, operation in the 12-GHz region is indicated for television broadcasting from satellites. The receiver we have designed will accept 12 contiguous color-television equivalent-bandwidth channels in the 11.72–12.2-GHz band using a wide-band FM format and frequency division multiplexing (FDM) of the channels. Each individual channel is to have a usable bandwidth of 36 MHz and a guard band of 4 MHz. The receiver provides an output for each channel compatible with standard video signals suitable for use with a CATV headend.

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II. RECEIVER DESCRIPTION

The desired characteristics of the ground terminal based on an earlier study [3] are summarized in Table I. The block diagram of the receiver is shown in Fig. 1. The power per channel available at the input to the receiver is at a level of −74 dBm. Allowance is made for a 12-dB fade margin. The system noise temperature specification is 1400 K with 200 K being allocated to the antenna. In order to facilitate the separation of the 12 channels and to set the system noise temperature, the entire 500 MHz of information bandwidth is first down converted from 12 GHz to the band from 1 to 1.5 GHz. This is accomplished with a broad-band low-noise mixer amplifier with a conversion gain of 25 dB [4]. The individual channels are then routed to the appropriate *L*-band-to-baseband receivers by means of a branching network.

To provide gain to the incoming signal and to isolate leakage from the second local oscillator (LO) into adjacent channels, each channel-dropping filter utilizes a single transistor amplifier at the output of a five-resonator band-pass filter (BPF) [5]. In this manner each channel shows a maximum forward gain of 10 dB, and a minimum reverse isolation of 20 dB with 30-dB suppression of signals at the adjacent band centers. A minimum of 70-dB attenuation to LO leakage from the second mixer is provided by each filter amplifier.