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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.

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.

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TABLE I
SATELLITE GROUND-TERMINAL CHARACTERISTICS

Ground-terminal type	Receive only.
Antenna	10-ft parabolic dish; limited manual steering.
Antenna gain	49.98 dB at 12-GHz 54-percent efficiency.
Antenna polarization	Linear.
Antenna noise temperature [T_A]	200 K (maximum).
Low noise receiver	Double conversion—first IF at 1 GHz and the second at 80 MHz; channel separation after first IF.
Receiver noise temperature [T_R]	1200 K (maximum).
Receiver bandwidth	11.700–12.200 GHz.
Individual RF carrier bandwidth (per TV channel)	36 MHz.
Modulation	Frequency modulation.
Demodulator	Discriminator (12-dB threshold).
Receiving system noise temperature [$T_S = T_A + T_R$]	1400 K.
Receiving system [G/T] [antenna gain to system noise temperature ratio]	17.52 dB/K.

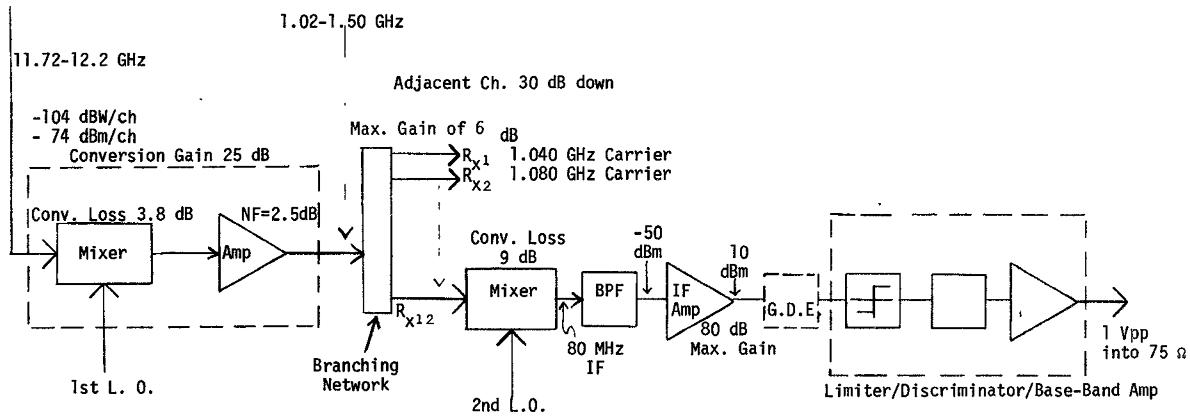

Fig. 1. Block diagram of 12-channel receiver.

After branching, the 12 individual channels are available at the inputs of the second mixers at a maximum level of about −40 dBm. A second conversion to an IF of 80 MHz is then carried out. This frequency was chosen to facilitate the design of the required LO frequencies. Appropriate BPF's and amplifier stages follow the second mixer to obtain an adequate level to drive the limiter and discriminator (3 V peak to peak across 300 Ω) at 80 MHz. An output amplifier provides a 1-V peak-to-peak composite video signal across 75 Ω, which is compatible with typical CATV remodulators for local distribution. Provision is made for group-delay equalizers (GDE's) to adjust the channel delay characteristics.

Notice that each channel receiver is identical beyond the channel-dropping filter. The particular channel is obtained by mixing with the appropriate second LO frequency. The passband characteristics of the branching-network filter and the IF-amplifier filters have been chosen to provide a total of 30-dB attenuation at adjacent channel band edges. The desired frequency response of the cascade of filters is shown in Fig. 2.

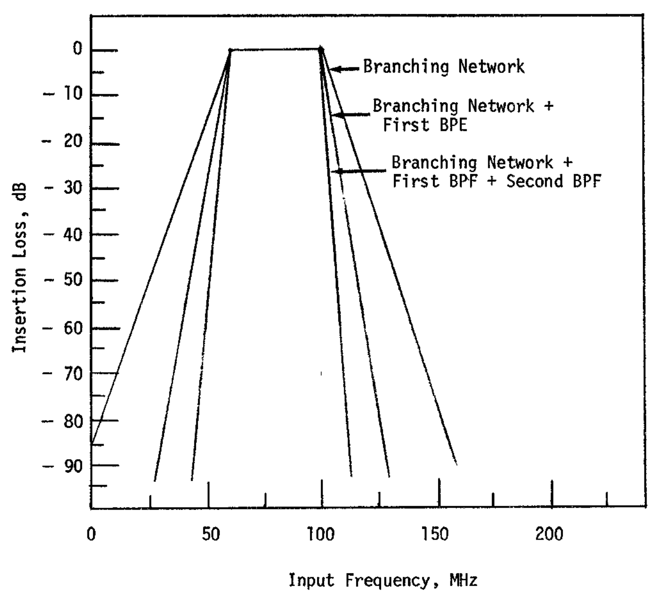

Fig. 2. Comparison of amplitude responses of branching network, first BPF and second BPF and resultant response.

Branching Network

The arrangement of the branching network is shown in Fig. 3. A 3-dB hybrid is used to separate the even and odd channels into two separate 50-Ω distribution lines in order to reduce the amount of adjacent-channel interference. Two isolators are used to provide a low-input VSWR for

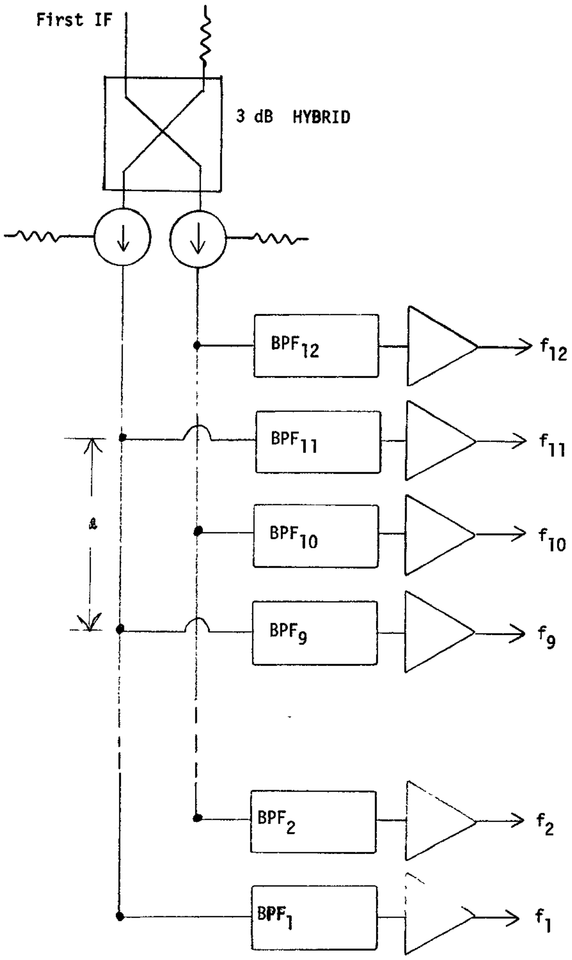

Fig. 3. Block diagram of 12-channel branching network.

the front-end converter output and to provide isolation for the even (or odd) channels rejected by the filter manifold. It may be possible to use only a single isolator at the input if the hybrid provides sufficient isolation between the two distribution lines. Since dielectric and conductor losses are greater at the higher frequencies, it is desirable to locate the higher frequency filters nearer the source.

The line lengths (l_{ij}) between the filters are adjusted such that at its center frequency each filter is presented with an open circuit produced by the remaining network beyond it.

The channel-dropping filters are 50-Ω doubly-terminated 0.1-dB Chebyshev filters designed to provide at least 30-dB attenuation at $f_i \pm 40$ MHz (the adjacent-channel bandcenter) [6]. They consist of five $\lambda/2$ (half-wavelength) parallel coupled resonators implemented in stripline. First-cut experimental designs show a ±0.5-dB ripple over the range $f_i \pm 18$ MHz with a 3-dB bandwidth of 40 MHz and midband insertion loss of 4.2 dB. The measured insertion loss at $f_i \pm 40$ MHz is >30 dB.

In order to make up the insertion loss of the filter and the distributed losses in the manifold and to help provide additional isolation against LO leakage from the second mixer into adjacent channels, a single-transistor (HP 35821E) amplifier is placed at the output of each filter. This amplifier has a nominal gain of 10 dB and reverse isolation greater than 17 dB over the 1–1.5-GHz band. Forward isolation is nominally 8 dB over the same range when the transistor is not biased.

Since the even- and odd-numbered filters are well isolated, the operation of the entire branching network can be determined by considering either set of filters. For the purpose of analysis the manifold with the odd-numbered channel filters was modeled on a computer and the insertion loss and return loss of each filter were examined. The results of this calculation, as shown in Fig. 4, indicate that little degradation is to be expected in individual filter performance due to interaction with other filters. Theoretical passband ripple is about 0.1 dB while the out-of-band insertion loss exceeds 30 dB at $f_i \pm 40$ MHz for each filter. The in-band return loss is reasonable with typical value >15 dB.

III. L-BAND-TO-BASEBAND RECEIVER

The second mixer is a commercial 50-Ω double-balanced flat pack made by Merrimac Research and Development, Inc., with LO and RF bandwidths from 50 to 1500 MHz and a conversion loss of 9 dB. A 50-Ω BPF (first BPF) follows the mixer to eliminate spurious out-of-band components and to provide improved channel selectivity. It consists of four *LC* resonators designed for 0.01-dB passband ripple and upper and lower cutoff frequencies of 98 and 62 MHz, respectively. The filter uses standard fixed capacitors and custom wound inductors. The measured insertion loss was 0.5 dB at 80 MHz and 1 dB at the cutoff frequencies; 30 dB of insertion loss was obtained at 40 and 120 MHz.

Following this filter is an 80-dB-gain amplifier centered at 80 MHz with 40 MHz of bandwidth. Its input impedance is 50 Ω and its output impedance is 300 Ω. A block diagram is shown in Fig. 5. Three MC 1590 high-frequency integrated circuits are capacitively coupled to provide 23-dB gain per stage. Provision is made for automatic gain control (AGC), but most testing is done at full gain. A second five-resonator BPF is located after the first stage. It is similar to the previous filter, but tuned to provide 30 dB of insertion loss at 50 and 110 MHz. The impedance level is 100 Ω. The measured response of this filter showed a passband ripple of 1 dB and an insertion loss of 0.5 dB at 62 and 98 MHz, respectively. The power amplifier consists of a feedback pair of 2N918 transistors capacitively coupled to the preceding stage. It provides 11 dB of gain and produces a 3-V-peak signal into 300 Ω.

Fig. 6 shows the frequency response of the entire IF amplifier (first and second BPF, preamplifier, and power amplifier) when driven by the second mixer. The input to the mixer was swept from 1180 to 1270 MHz at a level of −55 dBm. The second LO frequency was 1160 MHz at a level of 9 dBm. In order to obtain this response, the filters were tuned with one variable capacitor in the first BPF and three variable capacitors in the second BPF. A net midband gain of 80 dB was obtained with ±0.5-gain ripple and a bandwidth of 36 MHz. The trace below the response wave represents the network-analyzer detector characteristic.

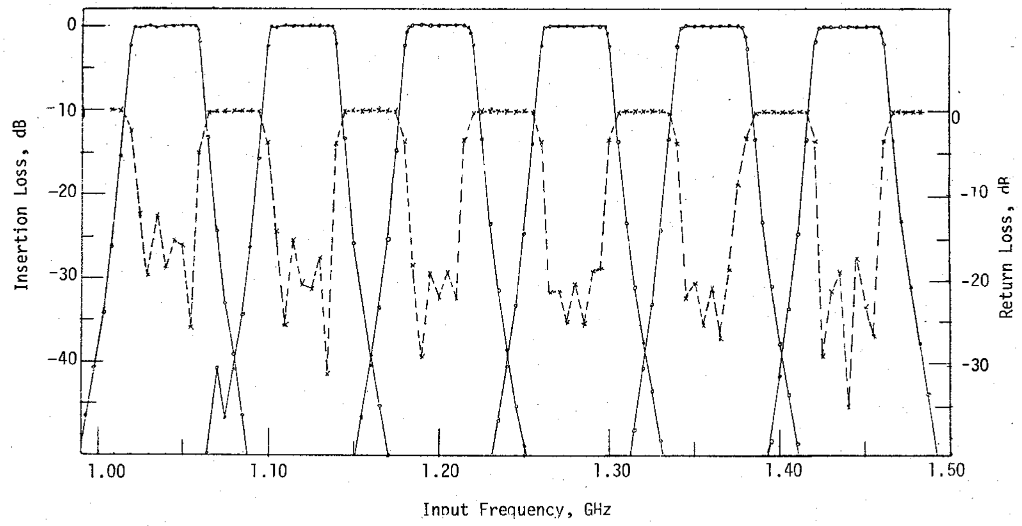

Fig. 4. Branching-network insertion loss (solid line) and return loss (dotted line) as a function of input frequency.

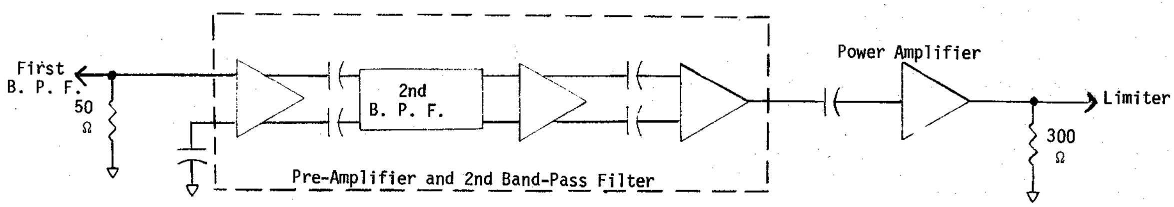

Fig. 5. Block diagram of 80-MHz IF amplifier.

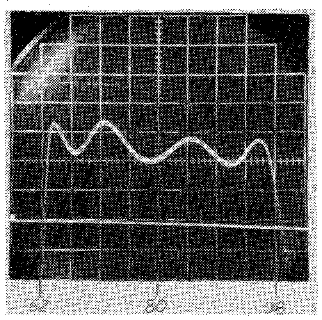

Fig. 6. IF-amplifier frequency response and response of diode detector. Vertical scale: 1 dB/cm. Horizontal scale: MHz as marked.

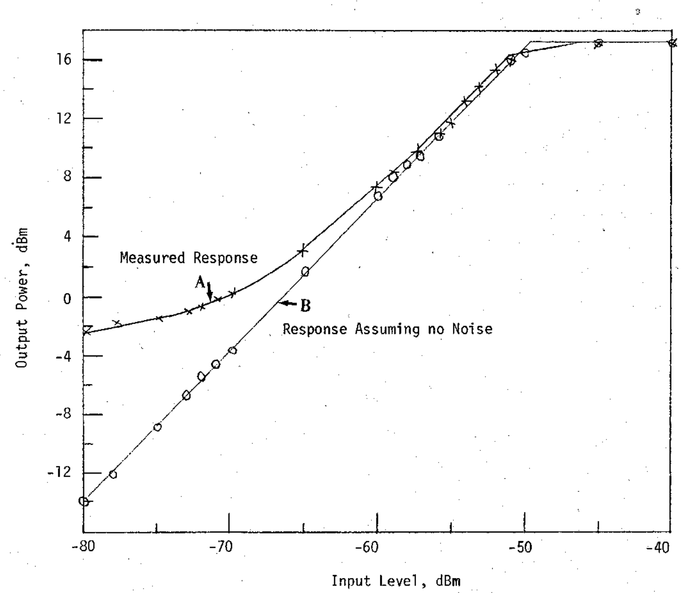

Fig. 7. IF-amplifier output power as a function of input power.

The RF input to the second mixer was matched experimentally to provide a return loss greater than 15 dB. In a similar manner the LO part was matched to provide a nominal return loss of 10 dB.

The minimum detectable signal at the mixer's input was found to be approximately −86 dBm. A plot of output power of the IF amplifier as a function of input power to the mixer is shown in Fig. 7. Curve *A* is the measured response. The output wide-band noise level is −2.67 dBm and affects the dynamic response at low levels. Curve *B* shows the equivalent linear dynamic range if no noise is present. From Curve *A* it is apparent that the range of linear operation is between −65 and −52 dBm. Operation into the nonlinear range, as will be seen later, results in higher intermodulation distortion than operation in the linear range.

In order to study the effects of intermodulation distortion, two signals (tones) were fed simultaneously into the RF port of the mixer. One tone, f_{mb} at 1240 MHz, produced the desired 80-MHz signal while the other tone, f_{int}, was varied between 1220 and 1260 MHz to produce interfering signals in the 60–100-MHz IF band. A spectrum analyzer was then used to measure the resulting spurious signals. Fig. 8 shows the power levels of the resulting spurious signals relative to the level of f_{mb} plotted as a function of their frequency for three different levels of f_{int}. The desired signal f_{mb} is at a level of −50 dBm, the point at which the amplifier begins to saturate. The level of f_{mb} was then lowered to −55 dBm. It was found that spurious signals are not detectable if f_{int} is much below −55 dBm, as is shown in Fig. 8. As can be seen, operation

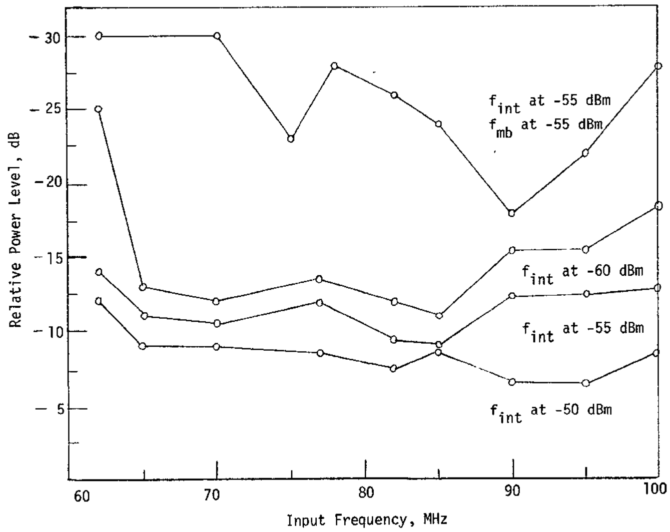

Fig. 8. Level of intermodulation products relative to level of f_{mb} as a function of frequency for different levels of f_{int} where f_{mb} = −50 dBm except as noted.

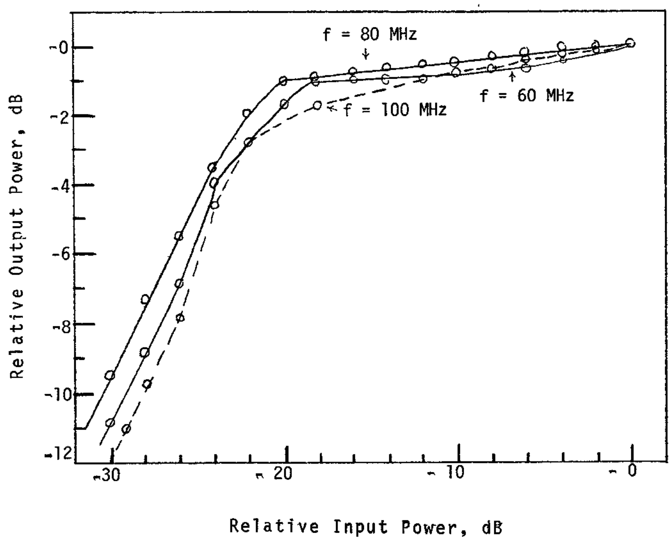

Fig. 9. Limiter output power as a function of input power for three different input frequencies.

of the amplifier in the nonlinear region produces considerable intermodulation product levels.

Provision is made for the insertion of GDE's as needed. Following the IF amplifier and GDE is the limiter which consists of two identical diode limiters separated by a 2N918 transistor amplifier which provides 6 dB of gain. The diodes are HP silicon Schottky-barrier devices. A nine-element low-pass filter is used to remove the harmonic content of the limited output. The filter has a passband ripple of 0.5 dB and an insertion loss of 1 dB at 100 MHz, and 18 dB at 120 MHz. When the dynamic range of the limiter and filter was measured at 80 MHz, it was found that the output power varied not more than 0.83 dB for a 20-dB change in input power as shown in Fig. 9. When driven by the IF amplifier, the limiter will be operated at a maximum input level of 12 dBm, corresponding to a relative power level of −8 dB in Fig. 9. Hence as seen there, the dynamic range is 12 dB. The frequency response was found to be quite flat from 60 to 100 MHz.

The wide-band discriminator used is a variation of the transmission-line bridge type described by Seo and Lee [7], [8]. However, rather than using two λ/8 transmission lines, one open circuited and the other short circuited, these were replaced by their lumped-element equivalent circuits as shown in Fig. 10. The diode detectors used were the same as those used in the limiter. It was found during development of this circuit that the diodes presented sufficient capacitance so that the elements C_1 were not needed in the final configuration. A variable capacitor was used for C_2 to set the crossover frequency at 80 MHz. A linear response (~1-percent deviation from linearity) was achieved with this approach over the band from 60 to 100 MHz as is shown in Fig. 11. A final dc coupled output-differential amplifier is incorporated to provide an unbalanced output level of 1 V peak to peak into 75 Ω.

The entire L-band-to-baseband receiver is housed in a $1 \times 3 \times 12\frac{3}{16}$-in aluminum box. The complete package is shown in Fig. 12. Its total weight is $1\frac{3}{8}$ lb. All cir-

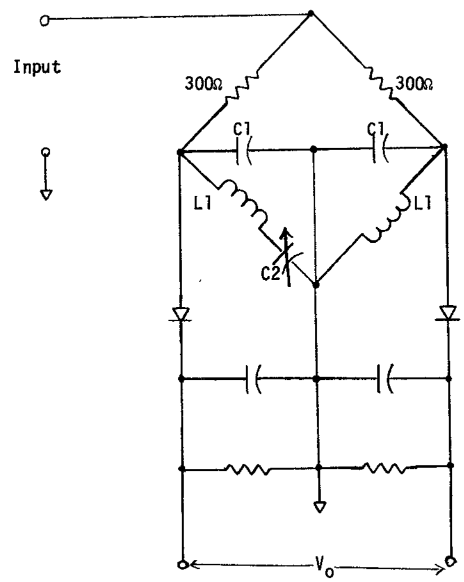

Fig. 10. Wide-band discriminator schematic.

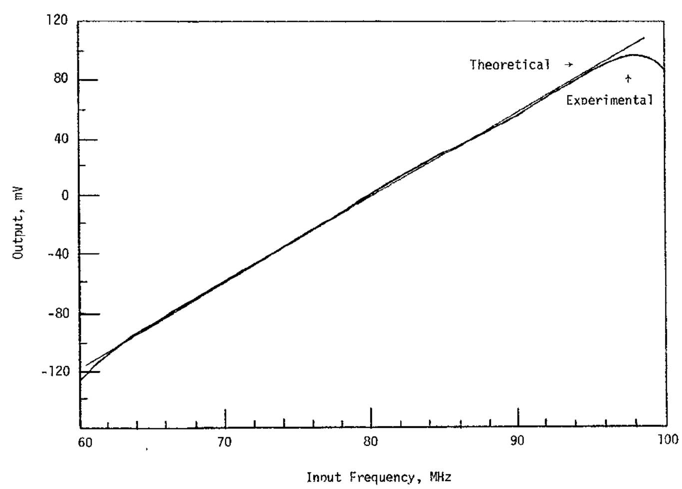

Fig. 11. Discriminator output as a function of frequency when connected to limiter output.

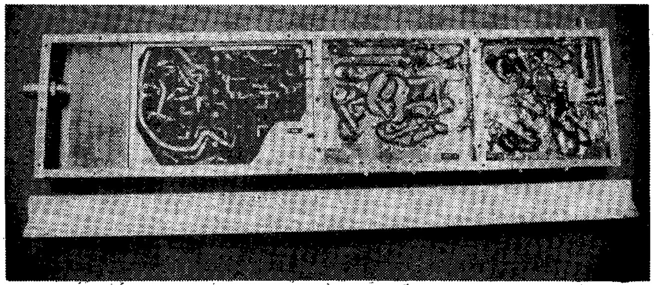

Fig. 12. Photograph of packaged single-channel receiver.

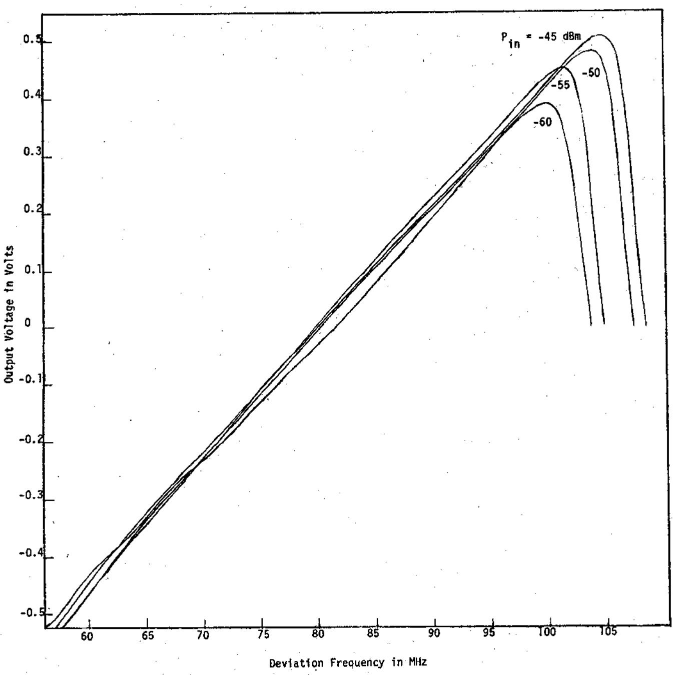

Fig. 13. Discriminated output of receiver as a function of frequency for several microwave input-power levels.

cuits were constructed in standard glass-epoxy boards. To prevent oscillation, shielding was required between the first and second preamplifier stages and also between the IF-amplifier output and limiter input. To prevent interference from spurious signals, the box is made RF tight overall. Two dc feedthrough connectors are used for connection to a ±32-V power supply.

The overall swept performance of this receiver from 1.2 GHz ± 20 MHz to baseband is shown in Fig. 13 for several values of input-signal level to the second mixer. Table II gives the overall performance characteristics for the receiver.

TABLE II
PERFORMANCE SPECIFICATION FOR *L*-BAND-TO-BASEBAND CHANNEL

Input level	−40 dBm max into 50 Ω
Minimum detectable signal	−86 dBm
Input return loss	>15 dB
Output	1 V peak to peak into 75 Ω
IF	80 MHz
Overall bandwidth	1.0–1.5 GHz
Bandwidth per channel	36 MHz ± 0.5 dB
Channel selectivity	30-dB suppression of adjacent channel band edges
dc power supply	±32 V 10 W
Dimensions	1 × 3 × 12 3/16 in
Weight	1 2/3 lb
Mixer—	
LO input level	+10 dBm
LO input return loss	>10 dB
IF amplifier—	
Output level	+16 dBm max into 300 Ω
Output noise level	−2.67 dBm
Dynamic range	15 dB
Limiter—	
Nominal input level	+10 dBm
Dynamic range	10 dB
Input impedance	300 Ω
Discriminator—	
Input impedance	300 Ω
Output linearity	<1-percent deviation from linearity

IV. DISCUSSION

As it is our intention that these receivers find widespread usage, both in educational and commercial systems, a low cost is a major design consideration. For example, double conversion is used which allows construction of the channel-dropping filters at *L* band, utilizing stripline techniques which are more economical than the use of waveguide or coaxial filters as required at higher frequencies. Beyond the channel-separation filters, the design of each channel is identical. Straightforward, yet innovative, designs were attempted which use standard components. Identical components have been used as much as possible to obtain economies of scale. The use of tunable elements has been avoided where ever possible to eliminate the need for complicated testing and alignment procedures.

A bill of material for a single channel receiver in lots of 1000 has been estimated at $52.00. The branching network (manifold and filters) could be implemented for an estimated $1500. The LO comb has not been studied so that no accurate cost figures are currently available. It is felt that a 12-channel *X*-band receiver of the type discussed here could be sold for under $10 000. Hopefully, the development of such receivers will speed the time when direct-broadcast satellite networks could be deployed.

ACKNOWLEDGMENT

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