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A 12-GHz Low-Cost Earth Terminal for Direct TV Reception from Broadcast Satellites

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Abstract—**A low-cost 12-GHz receiver for TV reception from high-power broadcast satellites is described. System designs using 0.6- and 1.2-m parabolic dishes with high-efficiency Cassegrain and prime-focus feed configurations have been studied. The front end consists of an MIC image-enhanced mixer, 1.2-GHz low-noise amplifier, and a Gunn diode LO. The signal is then fed to an indoor unit which has been designed using both surface acoustic wave (SAW) and lumped element 70-MHz bandpass filters and demodulators. Audio and video receiver outputs are fed directly to the baseband circuitry of a standard receiver.**

I. Introduction

HERMES communications technology satellite operating at 12 GHz and with an EIRP of 59 dBW ushered in the era of the high-power broadcast satellite [1]. In early 1978 Japan will be launching a similar satellite with an EIRP of up to 57 dBW [2]. The Nordic countries (Nordsat) [3] and West Germany [2] are seriously considering domestic high-power (60–67-dBW) broadcast satellite systems. Such systems permit the use of small TV receive-only terminals with values of G/T as low as 5 dB/K. This paper describes the development of such a terminal, discusses various alternatives which were considered, and comments on some of the advantages and disadvantages of each. The overall goal was to realize component designs which would lend themselves readily to low-cost medium-to-large-volume production.

II. System Configuration

Fig. 1 illustrates the system configuration which was adopted. Table I lists some of the target system parameters. Note that the design bandwidth of the terminal under construction was 85 MHz (although only 30 MHz per channel is actually required). However, some effort was made to anticipate future requirements for operation over the full 500-MHz allocated band. A double conversion approach was selected over single conversion. AFC on the second LO permits the use of super high frequency (SHF) LO's with relaxed long-term and temperature stability requirements. Also, future requirements for multichannel reception can be met by switching the second LO.

The single conversion approach would not only require the first LO to be remotely tunable to pre-set frequencies over the 500-MHz satellite band, but it would also be

TABLE I
Target System Parameters (1.2-m Dish)

Frequencies - Signal	12.038 - 12.123 GHz[1]
- SHF LO	10.8805 ± 0.005 GHz
- 1st IF	1.200 GHz
- UHF LO	1.130 ± .005 GHz
- 2nd IF	70.0 ± 0.2 MHz
Polarization	Linear
G_a/T_s	11.5 dB/°K (beam centre)
Outdoor Unit Gain	> 26.0 dB
UHF Cable (RG213U)	15m (<6 dB loss)
Indoor Unit Noise Figure	< 12 dB
Noise Bandwidth	22 MHz
Differential Phase	< 4°
Differential Gain	< 15%
Video Signal-to-Noise Ratio[2]	> 39 dB at threshold
Audio Signal-to-Noise Ratio[3]	> 45 dB at threshold
Audio Subcarrier Frequency	5.14 MHz
Outdoor Unit - Temperature Range	- 40°C to + 50°C
- Humidity	5% to 100%

[1] Actual band used is 12.0655-12.0955.
[2] Video peak to peak weighted (excluding sync. tip) to rms weighted noise.
[3] Audio rms TT/N weighted.

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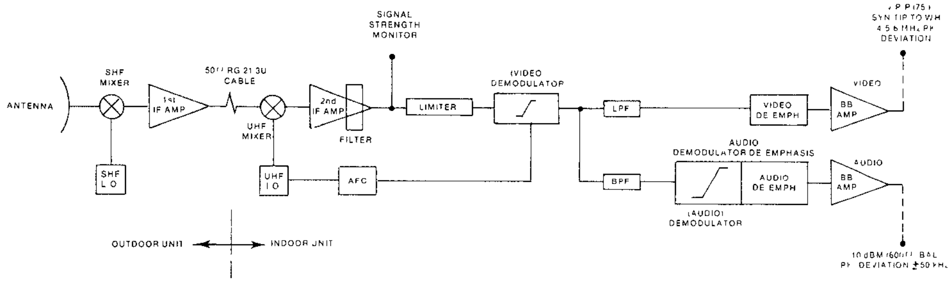


Fig. 1. Low-cost terminal system configuration.

required to be stable to approximately ±15-ppm total over the very exacting temperature and humidity requirements imposed on the outdoor unit (ODU). Furthermore, such a configuration would not lend itself readily to community reception. The advantages of commonality between the direct and community reception systems would not, therefore, be realized. Another advantage of the double conversion approach is the greater ease of realizing image and LO rejection. It appears that the only significant advantage of the single conversion system is the potentially lower system noise figure. However, with the latest developments in low-noise UHF bipolar transistor technology, this may not prove to be a significant advantage in the near future.

Once a double conversion configuration has been selected, it becomes necessary to define suitable first and second intermediate frequencies. The choice of IF's has been examined in some detail by the European Broadcasting Union [4], [5]. In general, for reasons of ease of filtering and avoidance of interference either to or from external equipment, a first IF in the range of 900–1400 MHz was recommended. Similarly, second IF's of 120 MHz or 400–500 MHz were recommended. However, with the general availability of equipment and components at 70 MHz for use in conventional terrestrial communication systems, the use of 70 MHz as the second IF was considered to be more desirable.

The system consists of three subassemblies, each of which will be described separately.

III. Antenna

Cassegrain and prime focus configurations were both considered in order to permit the relative merits of each for small aperture antennas to be determined. The design efforts for the antenna centered on obtaining maximum gain while minimizing the size and complexity. Sidelobe levels were not considered to be a problem for this application.

A. Cassegrain

Initially, the use of a Cassegrain antenna using a shaped subreflector with a conical scalar feed was investigated. It was felt that a slightly lower antenna noise temperature might be achieved using this approach. Compared to the conventional prime focus system, this approach also eliminated the feeder waveguide loss prior to the ODU input. Furthermore, polarization alignment could easily be achieved by rotating the polarizer section of the feed from behind the dish.

The choice of a corrugated conical feed was made to provide E- and H-plane phase and amplitude symmetry while achieving low side-lobe spillover power at the subreflector. A cross section of the feed is illustrated in Fig. 2(a). The feed parameters were obtained using a computer program based on work by Mentzer and Peters [6]. Unlike the typical procedure, the corrugations were designed to be perpendicular to the feed axis and their depth maintained as shallow as possible to ease anticipated fabrication problems. The number of corrugations per wavelength and the corrugation width-to-space ratio were selected to optimize the tradeoff between feed loss and mismatch. Close agreement was achieved between the theoretical and measured feed patterns for a feed fabricated of aluminum. A feed readily molded from plastic and metallized gave performance only slightly degraded from that of the machined aluminum feed.

The shape of the subreflector was determined using a computer program based on work by Collins [7]. It was found analytically and later verified experimentally that some sacrifice in phase distribution across the aperture had to be accepted in order to achieve uniform amplitude distribution. The amplitude and phase patterns of the composite feed and subreflector were quite uniform across the reflector aperture. A prominent amplitude null realized on axis was obviously desirable to minimize the effects of subreflector aperture blockage. The overall aperture efficiency of a 1.2-m (4-ft) antenna using the above feed and subreflector was measured to be greater than 67 percent over the complete 500-MHz satellite band.

B. Prime Focus

Concurrent with the studies at the Communications Research Centre (CRC) of the Cassegrain system, a program was sponsored at the University of Manitoba to investigate the design and relative merits of high-efficiency prime focus systems. The study [8], [9] indicated that aperture efficiencies upwards of 80 percent could theoretically be achieved by use of simple feed designs similar to that illustrated in Fig. 2(b). An evaluation of the relative merits of the two approaches indicated that if the ODU could be made to fit behind the feed and therefore not increase the feed blockage, the prime focus system would be both simpler and potentially more economical. Furthermore it would likely be more tolerant to dimensional and alignment inaccuracies. The development and evaluation of such a configuration suitable for use at 12 GHz was therefore

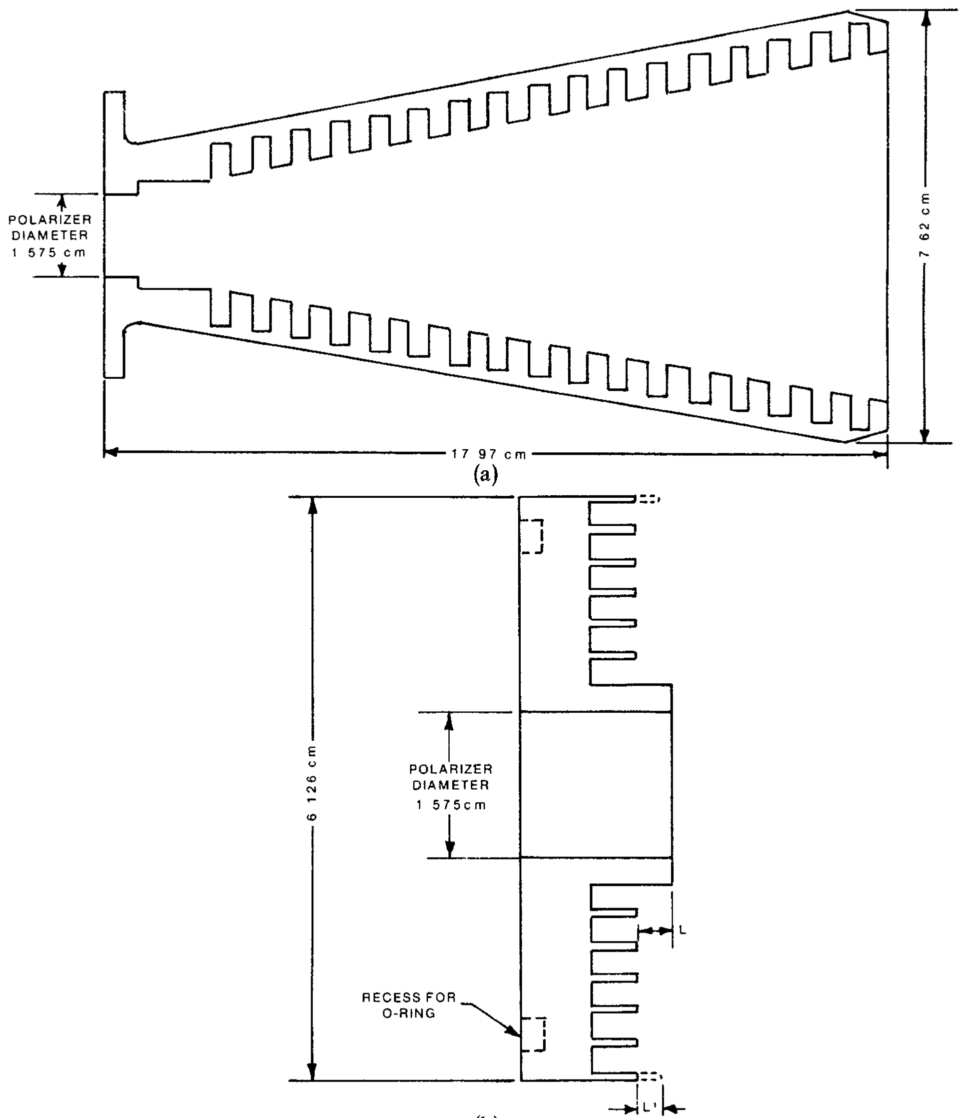


Fig. 2. (a) Conical scalar feed for Cassegrain antenna. (b) 90° scalar feed for prime focus antenna.

undertaken. Further analytical and experimental investigations were undertaken to optimize the lengths of the outermost ring and polarizer extensions (L' and L in Fig. 2(b)). The polarizer diameter was selected to yield approximately 40 dB of attenuation at the LO frequency and also attenuate spurious inputs at the image frequency. It was found that the improvement by varying L' was small and therefore, for simplicity, L' was set to zero. The measured patterns of the feed design which was eventually chosen, and which used the optimized value of L, are shown in Fig. 3. This feed, when tested with a 1.2-m reflector with an $f/D = 0.375$ yielded an aperture efficiency of better than 75 percent with the secondary pattern shown in Fig. 4. With a 0.6-m dish ($f/D = 0.40$) and a new value of L, this same feed yielded an efficiency of about 65 percent. Much of the difference was accountable to increased blockage.

IV. The Outdoor Unit

The possibility of using a FET amplifier front end was considered. However, the expected noise figure improvement to be achieved over that of an image-enhanced mixer was not considered substantial enough to warrant the expected price increase. For community reception systems, where many subscribers are involved, the increased cost may be justifiable.

A. Mixer

The design used for the mixer is relatively straightforward. Two GaAs beam lead Schottky diodes (DC1306M[1]) were used in the arms of a rat-race hybrid. In a manner similar to that used by Degenford and Newman [10], the diodes were preceded by coupled line image bandstop filters which presented a short-circuit impedance to the diodes. The use of radial line stubs in the bandstop filters resulted in a wider bandstop characteristic than was achievable using conventional open-circuit stubs. The design selected yielded a 16-dB image rejection bandwidth of greater than 200 MHz. IF bandstop filters, transparent to the SHF frequencies were

[1] AEI Semiconductors, Lincoln, UK.

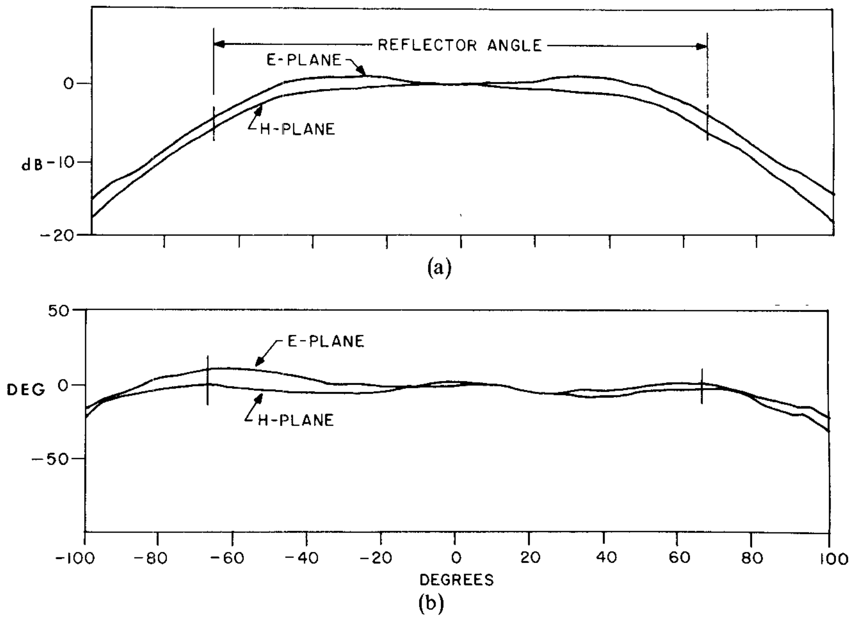


Fig. 3. Prime focus antenna scalar feed pattern. (a) Relative gain. (b) Phase.

also located in each arm of the rat-race hybrid. These filters were totally planar and consisted of a high-impedance line section $\lambda/4$ long at 12.08 GHz resonated with a planar capacitor at 1.2 GHz. The 15-dB bandwidth realized for these filters was approximately 250 MHz. Little effort was made to accurately measure the conversion loss of the mixer. Rather, efforts were concentrated on realizing the lowest noise figure for the composite mixer and a 1.2-GHz amplifier.

B. First IF Amplifier

The 1.2-GHz amplifier used three NEC 2107 bipolar transistors at a cost of $9 each in small quantities. A version which used two NEC 2135 devices at $3.50 each for the second two stages performed almost as well. Conventional MIC design techniques were used. Initially, device noise circles were obtained using an in-house developed MIC stub matching circuit. A single-stage circuit, designed and fabricated with the aid of computer optimization, yielded a noise figure of 2.2 dB with a 11.5-dB gain. Stub matching was used primarily, although current studies are directed towards realizing broader bandwidth transformer or lumped element matching approaches. The interstage lengths were optimized empirically. Amplifiers with 0.4–0.7-dB lower noise figures employing the Fujitsu FT1720AR, the NEC 645, and the HP HXTR-6104 are currently being developed as it is expected that the cost of these devices will drop to levels acceptable for this application in the near future.

A preliminary single-stage design using the Fujitsu transistor has yielded a 1.8-dB noise figure with an 11-dB gain.

C. SHF Local Oscillators

Various approaches to the 10.880-GHz LO design were examined. The design goals included a power output of $+8.0$ dBm with a total frequency variation over the temperature range of less than ± 5 MHz. A Gunn diode was

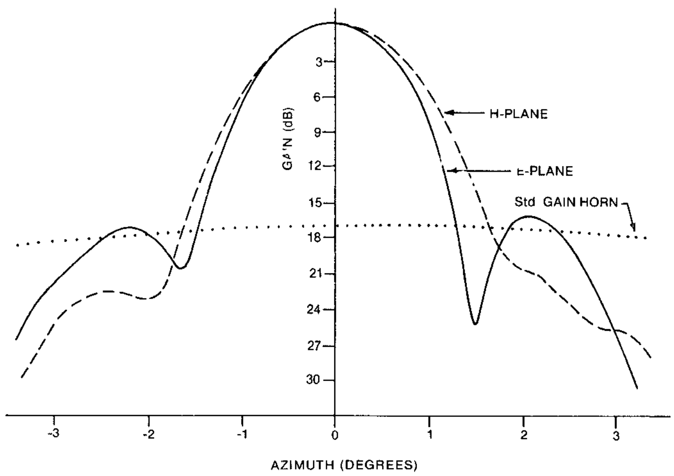


Fig. 4. Secondary pattern of a 1.2-m parabolic reflector with $f/D = 0.375$ and the optimized 90° scalar feed.

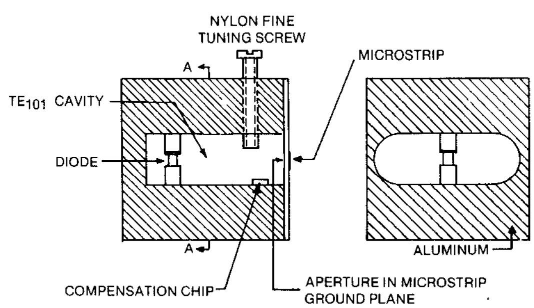


Fig. 5. Gunn diode oscillator using a TE_{101} temperature compensated cavity.

selected as the active element to take advantage of its potentially low-cost low-voltage requirement and low FM noise characteristics.

The first approach pursued and the one used in the ODU's described later is illustrated in Fig. 5 and uses a

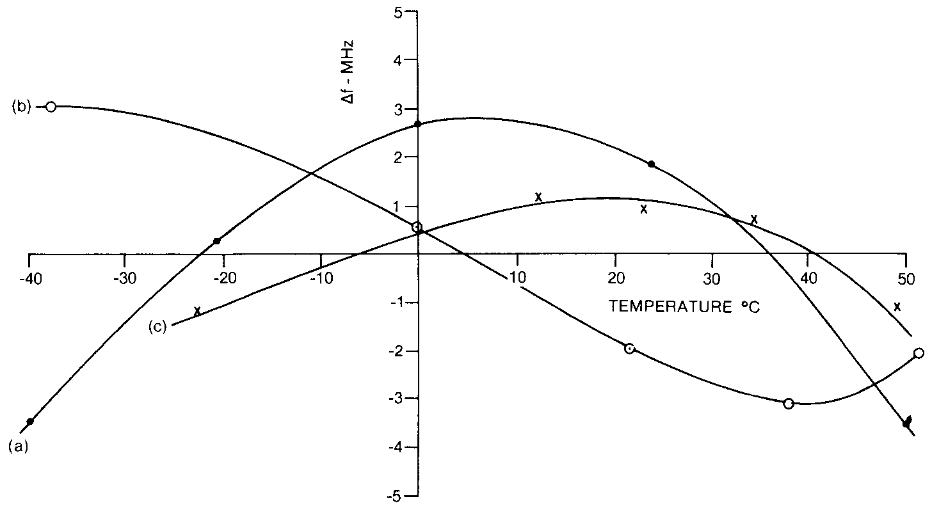


Fig. 6. Oscillator frequency temperature stability achieved using (a) a TE_{101} cavity temperature compensated using a negative T_{cc}, high ε_r dielectric chip; (b) a TE_{011} cavity compensated using a high coefficient of expansion material; and (c) a microstrip resonator temperature compensated using a high ε_r dielectric with a negative T_{cc}.

conventional TE_{101} cavity temperature compensated using a high ε_r chip with a negative coefficient of permittivity. The oscillator output is coupled through an aperture in the ground plane of the MIC output circuit [11]. This approach yielded the temperature stability curve *a* in Fig. 6. The compensation was felt to be quite sensitive to the dimensions of the dielectric chip and might not, therefore, be acceptably repeatable for this application.

A second approach has been tested and used an MIC Gunn oscillator, transmission stabilized to a TE_{011} cavity. The oscillator was coupled to the cavity at both the input and output through apertures in the ground plane of the MIC substrate in a manner similar to that described by James *et al.* [12]. The oscillator frequency was temperature compensated using the techniques described by Atia and Williams [13]. In this technique the dimensional variation of the aluminum cavity was offset by the high thermal expansion rate of the slab of compensation material.[2] Curve *b* in Fig. 6 shows the results achieved with this method. Besides its greater simplicity, this approach has the advantage that the frequency compensation is much more insensitive to dimensional inaccuracies. A modified ODU which uses this technique is currently under development.

Fig. 6 also includes the stability of a microstrip resonator, temperature compensated using a dielectric overlay (curve *c*). The dielectric overlay, besides having a high ε_r, also has a negative coefficient of permittivity. The thickness of the overlay was adjusted to obtain optimum compensation over the total temperature range. This approach is very promising for applications such as required here since it permits an essentially planar design to be employed.

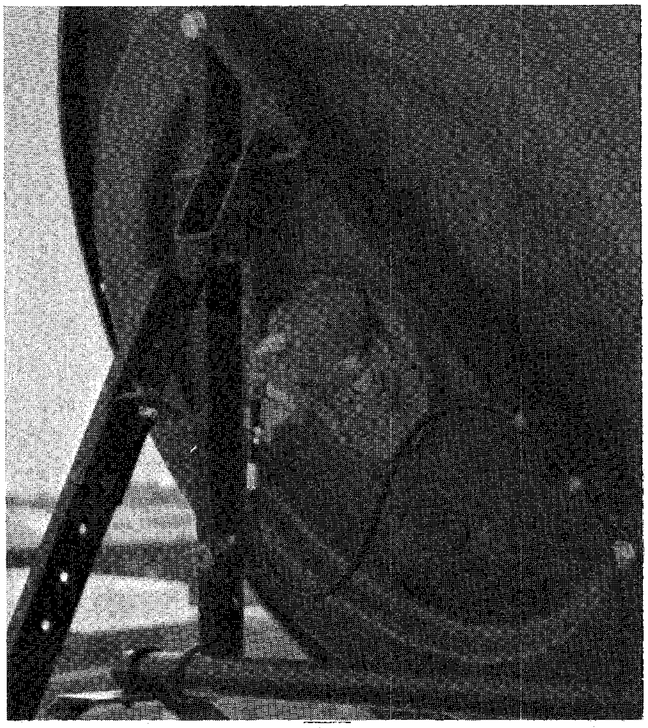
D. ODU Assembly

The different packaging approaches used with the Cassegrain (ODU 1) and prime focus (ODU 2) systems are illustrated in Fig. 7(a) and (b). Both use essentially identical RF circuit designs. ODU 1 uses an SMA to microstrip transition at the signal port with the LO coupled via a coax line through a hole in the mixer substrate. In ODU 2 the signal is probe coupled directly from the polarizer unit while the LO uses a straight MIC-to-MIC interconnect. ODU 1, for reasons of flexibility, consists of separate polarizer (probe coupled) and RF sections, although one integral unit could easily be realized. In both cases, polarization alignment is achieved by rotating the complete ODU with respect to the dish. However, ODU 1 has the advantage of ease of adjustment since all adjustment is made from behind the dish. Weather protection was achieved with ODU 1 by encapsulating the entire unit in a polymeric material. In ODU 2, the same function was performed by use of O-rings.

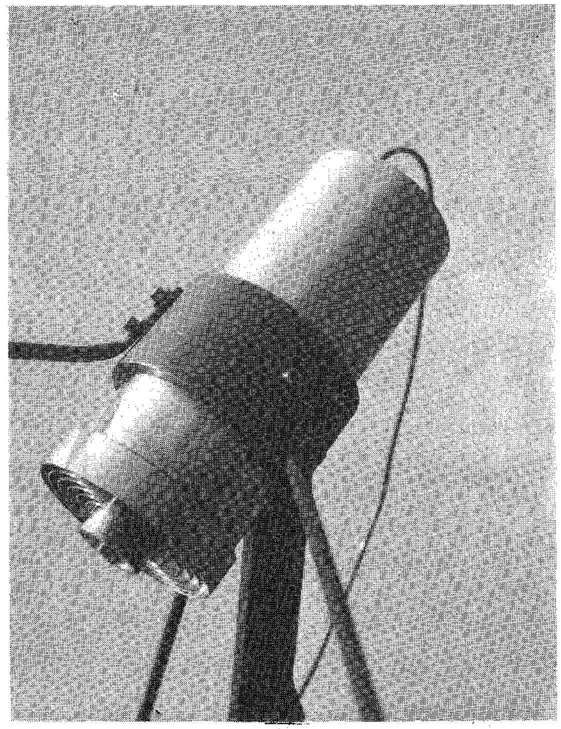
The 500-MHz noise figure (SSB corrected for image) and image and signal gain responses of ODU 1 are shown in Fig. 8. The performance of ODU 2 was generally similar although a noise figure degradation of about 0.4 dB was suffered. Over the design bandwidth of 85 MHz, the noise figure of ODU 1 is ~ 6.1 dB and, in fact, remains < 6.2 dB for almost the full 500-MHz band. In principle, the gain response, although not flat, might also be acceptable over the full 500-MHz band. However, an increase in the range of the ODU output signal level (~ 12 dB) as well as the steep gain slopes which occur in the upper and lower portions of the band (~ 2 dB/20 MHz) would have to be accommodated. This would place much more stringent requirements on the indoor unit (IDU) demodulator in order to prevent severe AM/PM conversion. Much of this gain slope is undoubtedly associated with the narrow-band amplifier design approach used. Fig. 9 illustrates the variation in the noise figure of ODU 2 at 12.08 GHz over the ambient temperature range of -40 to $+50$°C. Note that the worst case noise figure of 6.7 dB corresponds to a system G/T_s degradation of 0.8 dB.

A simple power regulation and transistor bias circuit was included in each ODU. An unregulated 17-V dc was sent to the ODU from the IDU via the RF interconnect cable. This

[2] Johnston Industrial Plastics Jaytrex 1000.



(a)



(b)

Fig. 7. The outdoor units (a) for the Cassegrain system (shown mounted on the back of the 1.2-m dish) and (b) for the prime focus system.

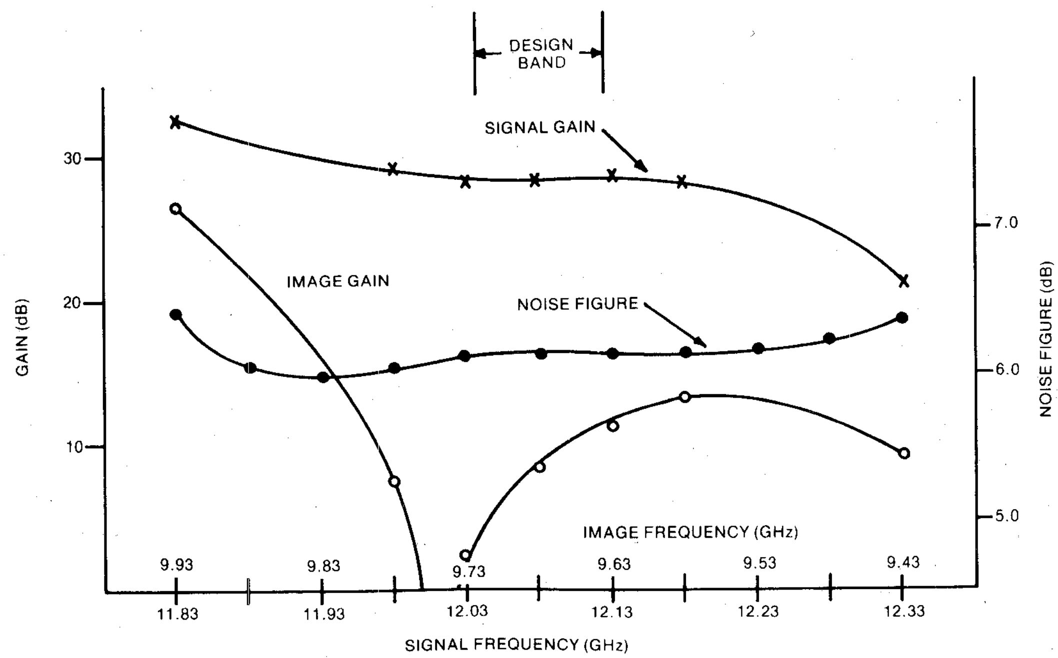


Fig. 8. The outdoor unit (ODU 1) SSB noise figure and signal and image gain responses over a 500-MHz bandwidth.

circuit then regulated the 17–12 V $\pm$ 0.1 V, biased the oscillator and 1.2-GHz amplifier, and supplied the desired base biases to each of the transistor stages.

V. The Indoor Unit (IDU)

Since most circuits in the IDU generally used fairly conventional designs, only those areas where significant decisions had to be made or problems were encountered will be mentioned.

A. UHF Mixer/LO

For the sake of expediency, the design adopted used a commercially available double balanced mixer in a flat-pack multipin configuration and a tunable LO in a TO-8 package.

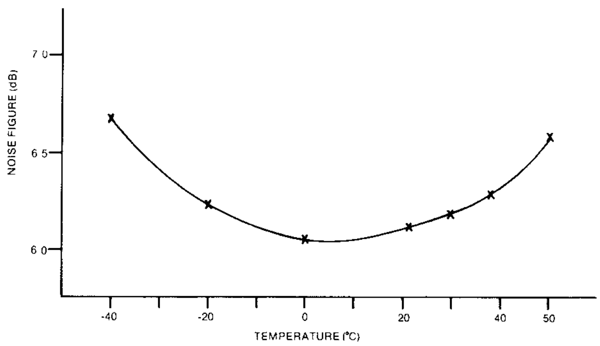


Fig. 9 Variation with temperature of the outdoor unit (ODU 1) noise figure.

To reject the UHF image, a simple $\lambda/4$ coax open-circuit stub was placed in shunt at the input to the indoor unit. This approach met the requirements for the development system although it would obviously be unacceptable for a multichannel system study. In parallel with the work proceeding at CRC, an IDU feasibility study funded by CRC was also performed by Electrohome Limited.[3] Their conclusions indicated that a tunable 1.2-GHz mixer/LO complete with integral image filter and preamplifier could be designed using techniques similar to those currently in use in conventional UHF TV tuners and would cost less than $10.

B. IF Amplifier, AGC, and Limiting

The main difficulty encountered here was achieving the required performance over the large percentage bandwidth required (28.5 percent). No suitable commercial IC amplifier units could be located.

Ultimately, a cascade of wideband emitter compensated transistor stages was used preceding and following the IF filter. Problems then encountered in maintaining stability were overcome mainly by use of extensive interstage shielding. No AGC was incorporated into the design as the expected maximum input level variation to be accommodated was only 10 dB. It was felt this range could be easily accommodated by the limiter. However, a three-stage limiter was eventually required to achieve the degree of limiting necessary to avoid AM to PM conversion.

The Electrohome study concluded that a somewhat different approach employing AGC, two stages of limiting, and a reduced amplifier bandwidth would be satisfactory, although it is expected the performance would be somewhat degraded from that of the CRC unit.

C. Bandpass Filter

The IF noise filter is a 5-pole 0.1-dB ripple Chebychev design. For simplicity, ease of tuning, and lowered cost, a design using five parallel resonant sections and simple capacitive coupling was employed. Although not the optimum approach for such large bandwidths, the performance achieved was found to be more than adequate for this application (22 MHz-noise bandwidth).

A SAW bandpass filter designed for this application is shown in Fig. 10. Care in packaging was required to maximize out-of-band rejection and minimize passband ripple and insertion loss. Further amplification was required to overcome the insertion loss of the filter. The noise bandwidth of this 70-MHz IF filter was 20 MHz. At present it is not clear if the SAW approach will be cost effective because of the requirement for additional amplification. The obvious advantage of the SAW is the elimination of all requirement for final tuning of each IF filter.

D. Discriminator and AFC Loop

The discriminator used was a lumped element version of the transmission-line bridge type. The required linearity was ± 22 percent over ± 8 MHz and ± 10 percent over ± 13 MHz. One of the main difficulties encountered was maintaining the center frequency output sufficiently close to zero volts to ensure, via the AFC loop, the required ± 200-kHz IF stability. Channel selection may be achieved simply by inserting a dc offset into the second LO AFC voltage.

A discriminator was developed using a SAW filter structure to provide a frequency dependent time delay. Unfortunately, to reconstruct the modulation, this approach required that the input signal be sampled at or greater than the modulation signal Nyquist rate. Due to the long delays encountered in the basic SAW structure, this resulted in performance which was useful only to derive the AFC voltage. At present more conventional designs using both filter and differential time delay schemes are being examined.

VI. The System Performance

The complete prime focus system using a 0.6-m dish is shown in Fig. 11. The IDU may be seen to the left of the dish. Table II lists the performance goals and results achieved by the small terminals when used with Hermes. Results for the 1.2-m Cassegrain antenna system and the 1.2- and 0.6-m prime focus systems are listed. The performance shown in this table is that at the center of the satellite beam and with no significant atmospheric attenuation. It is worth noting that with an FET amplifier front end, with 4.5-dB noise figure, a G/T_s improvement of 2–3 dB is possible. The Cassegrain system with ODU 1 was operated successfully throughout the winter of 1976–1977 with temperatures falling as low as -25°C. Except for a temporary failure due to severe icing of the antenna, service was uninterrupted. It should be appreciated that propagation fading, particularly due to precipitation, is not generally as severe as encountered by terrestrial systems. Furthermore, the potential user of such terminals would likely be much more tolerant of occasional fading.

The advantages of the portability of the 0.6-m system as opposed to the 1.2-m system are quite dramatic. For example, the complete earth terminal including the IDU (but not the TV) can easily be packaged to enable it to be hand carried and erected by a single individual.

VII. Cost Study and Conclusions

This study concluded that a wholesale price (in quantity $> 10\,000$) for the IDU (with multichannel capability) would

[3] Department of Communications contract number 12ST.36001-6-2598.

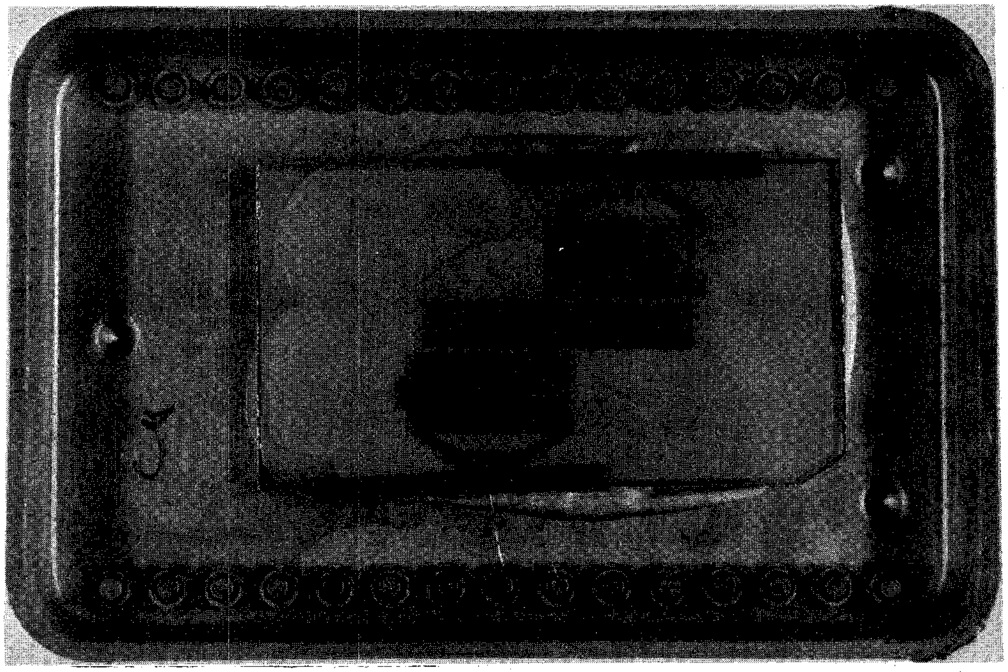


Fig. 10. The 70-MHz SAW bandpass filter. The noise bandwidth of this unit was 20 MHz.

Fig. 11. The complete low-cost earth terminal using the 0.6-m reflector with the 90° scalar feed outdoor unit. The indoor unit is also visible at the left.

be less than \$100. The cost of the reflector plus support structure for the 1.2-m dish has been estimated at \$60–\$90. Cost estimates for the ODU are much more variable and difficult to define, particularly in view of the rapidly varying prices for suitable microwave components. Depending largely on the fabrication procedure adopted, the wholesale cost of the ODU may be expected to vary from as low as \$150 to over \$300. In summary, therefore, the total cost of the terminal may be expected to be from \$310 to \$490. These

TABLE II
PERFORMANCE SUMMARY (BEAM CENTER AND NO PRECIPITATION ATTENUATION)

PARAMETER	TARGET (1.2m)	CASSEGRAIN (1.2m)	PRIME FOCUS (1.2m)	PRIME FOCUS (0.6m)
Received Power Level	-116.1 dBm	-116.1 dBm	-116.1 dBm	-116.1 dBm
Antenna Gain	41.5 dB	41.8 dB	42.3 dB	35.8 dB
Carrier Power Receiver Input	- 74.6 dBm	- 74.3 dBm	- 73.8 dBm	- 80.3 dBm
Noise KT_sB (B=22 MHz)	- 94.4 dBm	- 94.9 dBm	- 94.4 dBm	- 94.4 dBm
G/T_s	11.5 dB/°K	12.3 dB/°K	12.3 dB/°K	5.8 dB/°K
C/N (including uplink contribution)	18.9 dB	19.5 dB	19.5 dB	13.0 dB
Video SNR[1] - MAX	49.3 dB	49.9 dB	49.9 dB	43.4 dB
- Threshold	39 dB	38.3 dB	38.3 dB	38.3 dB

[1] Video peak-to-peak (excluding sync. tip) to rms weighted noise.

figures are in close agreement with the estimates arrived at by Sellberg [5] in his study of terminals for the potentially forthcoming Nordsat system.

ACKNOWLEDGMENT

Many people were responsible for major contributions to the development of this terminal. I wish specifically to acknowledge the effort of T. Nishizaki, G. St. Amand, O. Berolo, A. Kong, L. Shafai, B. Clarke, R. Hahn, and M. Cuhaci. Also, the support of the microelectronics group is greatly appreciated.

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New n-Way Hybrid Power Dividers

NOBUO NAGAI, MEMBER, IEEE, EIJI MAEKAWA, AND KOUJIRO ONO

Abstract—**A new n-way planar hybrid power divider (henceforth HPD) and an n-way coaxial-type HPD are proposed, and the synthesis methods of three-way and four-way planar HPD's are shown. As for the three-way planar HPD, the isolation characteristics among three ports show more than 20 dB, and the VSWR's of four ports show less than 1.4 in 2 : 1 bandwidth. The experimental characteristics are in good agreement with the theoretical analysis.**

I. Introduction

THE OBJECTIVE of this paper is to propose some new n-way hybrid power dividers (henceforth HPD) constructed by a new synthesis method. The n-way HPD's described by Wilkinson [1] and Yee *et al.* [4] are synthesized by using M sections of n uncoupled transmission lines of equal length with isolation resistors of the Y connection which are connected from the end of the n transmission lines to a common junction. The isolation resistors cannot be designed to be a planar structure. The n-way HPD's presented in this paper are synthesized by using M sections of n-wire coupled (or uncoupled) lines of equal length with isolation resistors which are connected by the ends of the neighboring wires.

The analysis of the n-way HPD is done by getting the eigenvalues and the corresponding eigenvectors of the characteristic admittance matrices of M sections for n-wire coupled (or uncoupled) lines and of M admittance matrices for the isolation resistors, and then by getting the equivalent circuit representation of the n-way HPD [8], [9]. The equivalent circuit representation is presented by n circuits which consist of a two-port for the even-mode circuit and $n-1$ one-ports for the odd-mode circuits.

It can be shown that the $(n+1)$-port made with a coupled n-wire line with isolation resistors of the Y-connection acts as an n-way HPD at narrow-band frequencies [6], [8], [9]. If we use isolation resistors different from the Y-connection, it needs some sections of isolation resistors and n-wire segments to perform matching and isolation among output ports at the required frequency.

This paper proposes an n-way HPD of a planar structure and a coaxial type n-way HPD, constructed by way of new connections of isolation resistors. As for the examples of the planar HPD, we show the circuits of three-way and four-way HPD's, and their VSWR's and isolation responses are shown in the figures. The isolation characteristics in three output ports of the three-way HPD show more than 20 dB, and the VSWR's of four ports show less than 1.4 in 2 : 1 bandwidth. The isolation characteristics in four ports of the four-way HPD show more than 24 dB, and the VSWR's of five ports show less than 1.2 in 2 : 1 bandwidth.

II. The n-Way Planar Hybrid Power Divider

Since we get isolation resistors of a planar structure by connecting $n-1$ resistors by the ends of the neighboring wires of the n-wire lines constituting an n-way HPD, we consider a $(1,n)$-port, shown in Fig. 1(a), which is designed by using M sections of n uncoupled transmission lines of equal length with the planar isolation resistors. The analysis

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