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I. Design and Calibration of X-band Satellite Communications Ground Terminals

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Synopsis: This paper is directed towards the establishment of design parameters for X-band ground terminals, the emphasis being on technical efficiency, flexibility, and suitability for operation with present and foreseeable satellite transponders. By considering all the pertinent factors such as satellite parameters, environmental conditions, modulation and multiple-access techniques, communication availability, expected terminal lifetime with respect to that of a satellite, it is shown that the terminal parameters can be well defined.

The paper subsequently discusses in some detail how the parameters may be realized in a practical sense and how the terminal may be calibrated to measure accurately the quantities which enter into the power budget.

1. Introduction

Quite a volume of literature has been published describing satellite ground terminals. The greater part of these terminals has been designed for experimental purposes. Special design techniques for a variety of functional units, and the application of modulation as well as access techniques have been treated widely. It is not my intention to elaborate here on any particular feature or design in isolation, but rather to examine the present state-of-the-art in satellite communications technology as a whole, and to consider possible operational requirements with a view to developing design criteria for the ground terminals which are efficient and suitable for use in any present or predictable operational system. We shall, of course, pay particular attention to the antenna, its steering mechanism, and its calibration, because this is the most expensive item of the terminal (about half the price of a link terminal) which, once obtained, is also the most difficult to change and modify.

In order to arrive at specific conclusions we shall make certain restrictive assumptions. We shall only be concerned with X-band operation using hard-limiting heterodyne transponders, although many of the relationships and trade-offs between design parameters could well be applied more generally. The block diagram of such a satellite is shown in Fig. 1. Satellite transponders will be designed to work either in a hard-limited or in a 'linear' mode. Where bandwidth is not a limiting factor the former mode allows a more efficient usage of the transponder

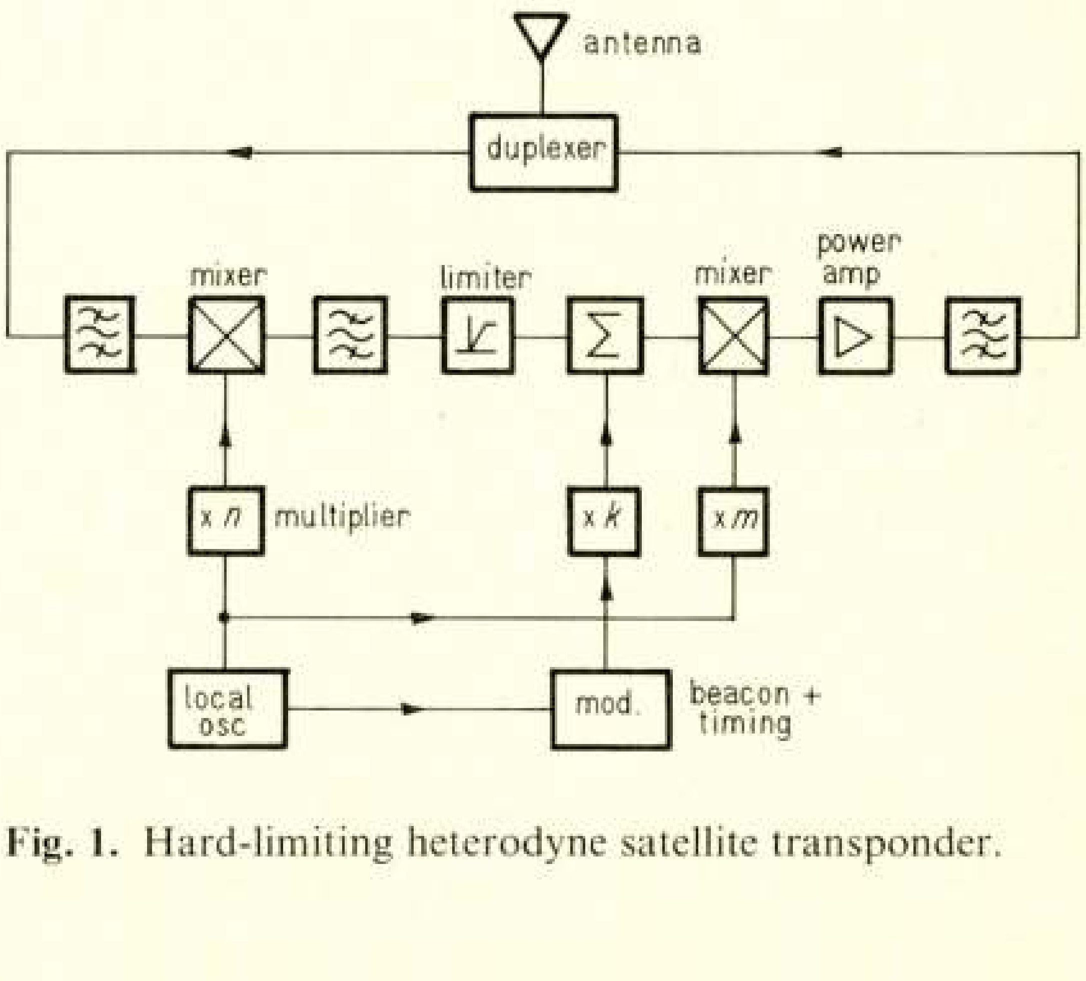

Fig. 1. Hard-limiting heterodyne satellite transponder.

Lecture delivered at the Symposium on antennas organized by the Nederlands Elektronica en Radiogenootschap on 18 May 1971 at Delft University of Technology, Department of Electrical Engineering. Voor de aankondiging zie 'De Ingenieur' 1971, nr. 19, blz. A 336.

power, while the linear or quasi-linear operation is preferred where frequency spectrum is at a premium and it is unlikely that the transponder could be driven into saturation by intentional interference.

Furthermore, we shall consider terminals with antennas of about 20 ft and greater in diameter, requiring some form of automatic tracking facility, and capable of providing multiple voice and data channels. A typical terminal may be considered as comprising four distinct parts:
- link terminal;
- telephone and telegraph multiplex equipment;
- primary power supply and distribution;
- switching equipment.

The main subject of the present paper is the link terminal which, almost independently of size, would be composed of the following sub-systems:
- antenna and feeds;
- antenna mount and control;
- communications transmitter;
- communications receiver;
- frequency generation;
- test equipment.

Fig. 2 gives a diagrammatical representation of the internationally agreed operational frequency bands [1]. It should be noted that two 50 MHz bands are exclusively allocated for fixed-satellite communications. The ground terminal should, however, be designed to permit operation anywhere within the two complete 500 MHz bands, and, where applicable, make use of the full bandwidth.

2. Influence of Satellite Parameters on Terminal Design

Many X-band satellites have already been built and successfully orbited and it may therefore be useful to examine their characteristics and try to determine the extent to which they define the ground terminal which is to be used in future operational systems. The more important parameters of some of these satellites are summarized in Table 1 which also shows predicted design trends for future satellites.

We shall now consider the relationships between the satellite characteristics and ground terminal parameters.

2.1. *Transmitter Power*

The minimum ground terminal effective radiated power (ERP) may be determined from items 4, 6, and 8 of Table 1, and from a knowledge of the number of carriers accessing the satellite.

The receiver input noise temperature of the satellite will include a contribution from the earth of up to 300 K. Thus, assuming a total noise temperature of 400 K, and a 50 MHz bandwidth with a gain of 14 dB for the receiving antenna, the minimum ground terminal ERP is found to be about + 105 dBm.

2.2. *RF Bandwidth of the Ground Terminal*

The bandwidth of the output stages of the transmitting channel

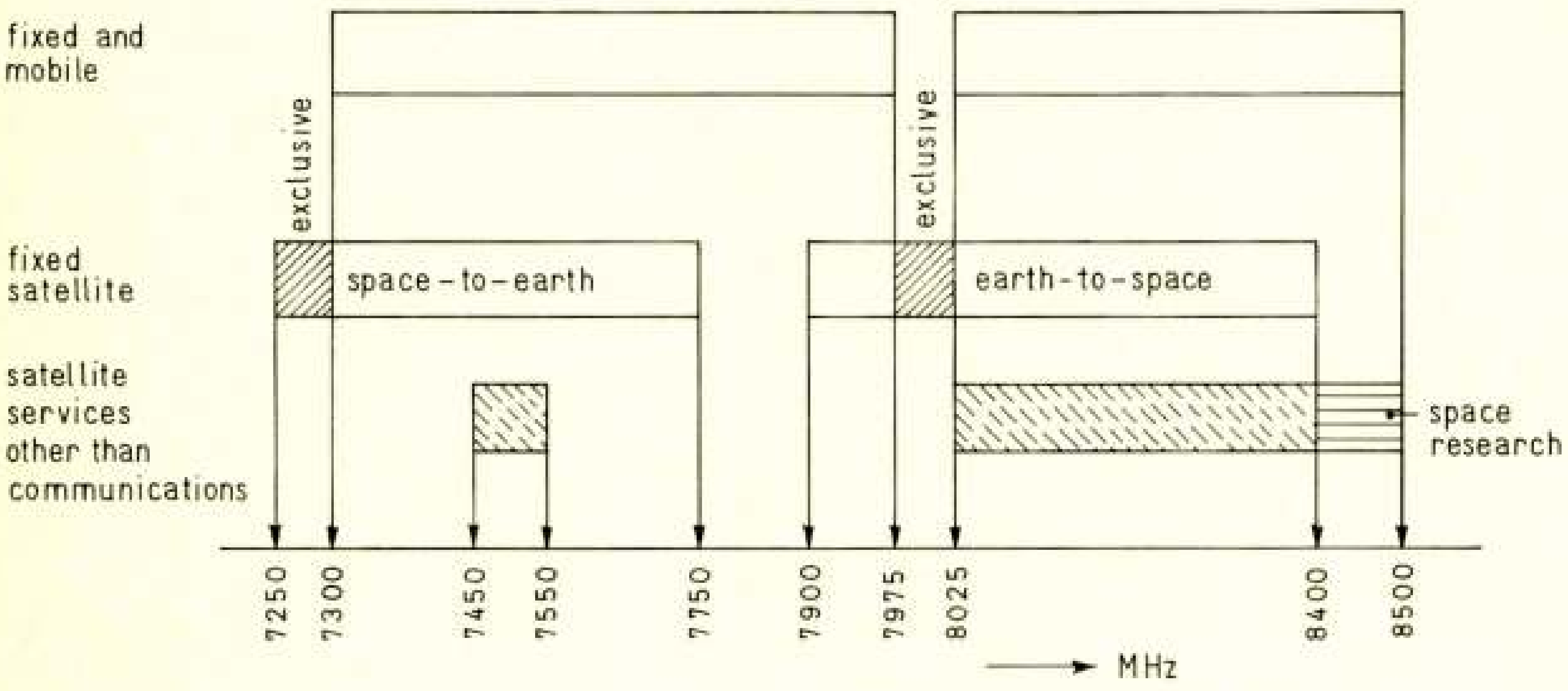

Fig. 2. ITU frequency assignment.

Table 1. Characteristics of Some X-Band Satellites.

	LES-2 1)	LES-4	IDCSP 2)	1970-1975
1. Orbit	elliptical, medium altitude	highly elliptical	near synchronous	synchronous
2. Transmit frequency (GHz)	7.75	7.75	7.25	7.25-7.75 +
3. Receive frequency (GHz)	8.35	8.35	8.00	7.90-8.40 +
4. Bandwidth (MHz)	20	20	20	> 20
5. Transmit EIRP (W)	0.2	2.0	8.0	$< 12\,n^{+*}$
6. Approximate antenna gain (dB)	0	0	4	> 14
7. Polarization	circular	circular	circular	circular
8. Noise figure (dB)	16	9	9	9
9. Number of transponders	1	1	1	multiple
10. Beacon frequency (GHz)	7.74	7.74	7.300	–
11. Beacon ERP (W)	0.01	0.05	0.5	> 1

+ CCIR Limit, +* n = number of 4 kHz channels.

1) Lincoln (Laboratory) Experimental Satellite. 2) Interim Defence Communication Satellite Programme.

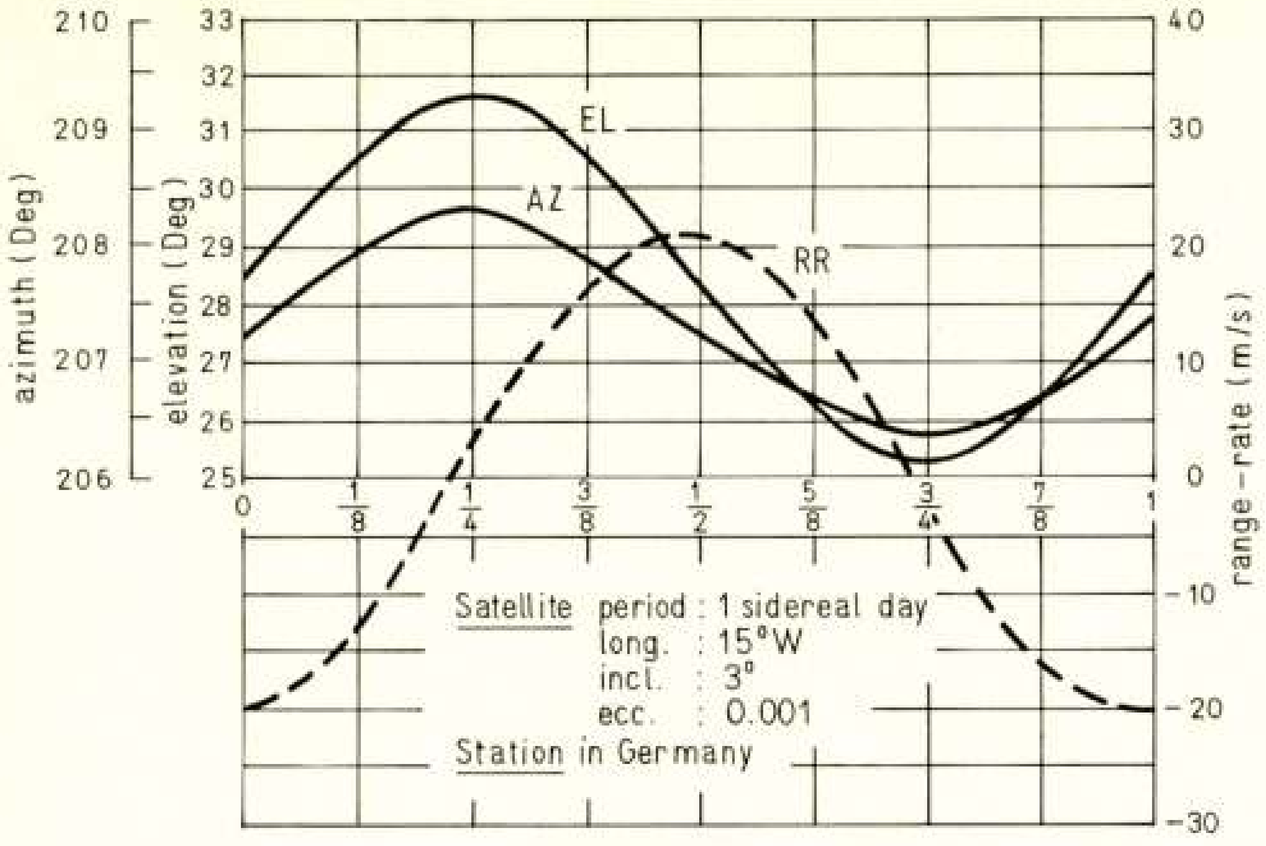

Fig. 3. Look angles and range-rate for synchronous satellite in inclined orbit.

and the input stages of the receiving channels should be at least 50 MHz. For future systems with increased capacity for multiple carrier transmission, also having greater flexibility and anti-interference capability, broadband transmitters and receivers with up to 500 MHz instantaneous RF bandwidths may, however, be required. This requirement affects the manner in which baseband signals are converted up to RF and imposes on the choice of intermediate frequencies conditions for minimizing the effects of unwanted modulation sidebands.

2.3. *Tracking Capability*

The tracking capability in space and frequency can be determined from a consideration of items 1, 10, and 11 of Table 1. Geostationary satellites are most favourable for communications, but although they are intended to remain permanently above the same point on the earth, it is accepted that due to on-board fuel limitations they can move in latitude up to $\pm$ 3°/day at the beginning and end of their life, due to sun and moon gravities, and earth oblateness.

The maximum angular velocity would be about 1.5°/h; the range rate as high as 30 m/s. It is foreseen that in future, satellites may have orbit inclination control for simplifying ground terminal design and for accommodating more satellites in the equatorial belt.

For satellites in near synchronous orbits, which may have advantages for certain applications, the angular velocity would probably be about 1.5°/h with a maximum range rate about 100 m/s.

From these considerations it can be concluded that a terminal tracking speed of at least 1.5°/h with Doppler Shift $\pm$3 kHz would be a minimum tracking requirement. Ranges and rates for a typical location in Europe are shown in Fig. 3.

In the context of tracking capability, it is necessary to consider the outage time due to conjunctions. These will be conjunctions between two satellites (geostationary and moving), and conjunctions with the sun.

As a typical example we mention that two satellites with a relative drift of 1°/h could cause an outage of about 15 minutes assuming a ground antenna beamwidth of 0.2°. Furthermore, sun conjunctions which occur during a few days (maximum 7 days) around the spring and autumn equinoxes could have an effective duration of about 6 minutes each day, with a similar antenna. Consequently, in order to maintain terminal availability at a high level it may be a requirement that two geostationary satellites be used and orbited with an angular separation of at least 4°. Thus, high antenna slewing speeds would become a necessity and speeds of 1°/s in azimuth and elevation are often specified.

Ground terminals which are designed to have elevation coverage of at least 90° and azimuth coverage of more than 360° are capable of tracking satellites in any orbit. For following a geostationary satellite a much smaller coverage is required and it is then possible to design a simple and much cheaper type of antenna mount.

The coverage required for the following of moving equatorial satellites would be somewhere between the two types mentioned above. However, this would still necessitate a mount similar in design to that for wide coverage. For flexibility reasons and for operation with geostationary and moving satellites, it would therefore be desirable to have a terminal which could track up to 90° elevation and $\pm$ 200° azimuth.

2.4. *Polarization*

In order to ease duplexing of transmitter and receiver, satellites use, and will continue to use, circular polarization. 3 dB and lower axial ratios of the polarization ellipse have been achieved in the present satellites. A wave of this kind would produce a maximum ellipticity loss of 0.5 dB when received by an antenna which has an axial ratio of 1.5 dB. This loss can be reduced in the case of geostationary satellites by providing for ellipticity compensation in the ground terminal feed system.

2.5. *Beacon Signal*

As shown in Fig. 1 satellites are always equipped to radiate beacon frequencies for tracking and identification purposes. The beacon may also carry timing information which could be used for network timing, thereby enhancing the performance of communications in three ways:
- by assisting in the reduction of mutual interference among multiple-access signals and facilitating link synchronism;
- by permitting the use of schemes for combating intentional interference;
- by providing a means of synchronization for data streams.

When the characteristics regarding stability, pulse width and pulse rate of the timing pulses are known, the tracking receiver may suitably be designed to extract these pulses.

2.6. *Frequency Supplies*

The heterodyning process in the satellite requires a local frequency supply and this is normally obtained from a stable master clock. It is important to know the short and long term stability characteristics of the local oscillator since this affects the intermodulation noise (IM) in the FM channels and the tracking receiver. We shall consider this matter in more detail in section 10.

Since the virtual local supply frequency of the satellite is about 700 MHz, the local oscillators in the ground terminal must have a stability at least ten times better than that of the satellite. Moreover, due to intermodulation between several carriers passing through the satellite, a careful choice of carrier frequencies is required. This will usually demand a highly stable and flexible source of local frequency supplies in the ground terminal.

2.7. *Up-Link Power Control*

In order to ensure proper sharing of the satellite power it

Table 2. Relationships Between Satellite and Ground Terminal Parameters.

Satellite Characteristics	Ground Terminal Parameters Affected
1. Synchronous and near-synchronous	Servo-drive, tracking receiver
2. Hard-limiting transponders	Modulation and access method
3. Multiple, random-access transponders	Transmitter, receiver, power control, gain stability, local frequency sources
4. Large bandwidth	Up and down conversions, transmitter, local frequency sources
5. Power/bandwidth ratio	Antenna gain/system noise temperature
6. Circular polarization	Feeds, duplexer
7. Sun, moon and satellite conjunctions	Servo-drive, system margin
8. Beacon with timing and identification	Tracking receiver, message and multi-access modulation

is essential that the levels of the accessing carriers do not vary significantly. These levels will be affected by varying ground terminal transmitter power and environmental and propagation effects. This leads to the requirement for the ground terminals that they should be designed so that the transmit channels are inherently stable and the carrier levels are capable of being set to required values accurately and easily. Furthermore, an overall control generally known as System Communications Control (SCC), would normally have to be established when the number of ground terminals in a network is increased beyond about four.

3. Modulation and Access Methods

Up to now we have considered the effect of satellite parameters on the technical design of the ground terminal. The relationships established are shown in Table 2.

The satellite bandwidth, the power output and the modulation methods used in the system (item 5 in Table 2) interact in other ways with the ground terminal and affect the antenna size and the type of receiver front-end to be used. We shall now review the various modulation and access methods and we shall attempt to establish relationships between the access methods and the satellite power and bandwidth in order to indicate the way in which future satellite systems are likely to be developed.

There are numerous modulation and multiple-access methods which can be used, each having different characteristics regarding communications efficiency, suitability for analogue/digital signals, system control, vulnerability to EM interference, etc. The basic modulation stages and access methods are categorized below:

- baseband composition;
- message modulation;
- access method:
  a. Frequency Division Multiple-Access (FDMA);
  b. Time Division Multiple-Access (TDMA);
  c. Code Division Multiple-Access (CDMA).

It should be noted that CDMA covers a variety of techniques which are often considered as separately defined methods. A well-known form is Spread-Spectrum-Multiple-Access (SSMA), wherein the carrier amplitude remains constant but the phase is modulated by the message and the bandspreading code. The transmitted signal occupies a frequency band which is much larger than the bandwidth required for the transmission of the message.

4. Comparison of Multiple Access Methods

The efficiencies of the access methods may be compared by consideration of:

- the percentage of the satellite ERP which can be utilized for the transmission of information;
- the percentage of satellite bandwidth which could be utilized for the transmission of information if the satellite ERP were unlimited;
- operational features;
- vulnerability to external interference.

We shall now consider these factors in turn.

4.1. *Satellite ERP*

For FDMA and SSMA, the percentage of satellite power which can be usefully employed for information transmission is mainly dependent upon the suppression effects in the hard limiter. Assuming that the input level of any one carrier is not greater than the sum of all other carriers, then about 20% of the total satellite ERP would be used by the intermodulation products generated in the hard limiter transponder.

For TDMA, a guard time between transmissions must be allowed for imperfect timing. Moreover, additional time may be required for link synchronization at the beginning of each transmission. Thus, the total percentage time lost is equivalent to power wastage, and has been estimated to be 5 ... 20% [2].

Consequently, there is little to choose between the access methods as regards ERP utilization.

4.2. *Satellite Bandwidth*

Expressions for optimum satellite bandwidth utilization for the various access methods have been developed in Appendix A and can be summarized as follows:

$$\text{FDMA: } E_B = 3.33/N^{1.3}$$

$$\text{TDMA: } E_B = 0.8 - 0.95$$

$$\text{SSMA: } E_B = \frac{3\pi}{16\,R_{min}} \times \frac{1}{1-1/N}$$

where: E_B = fractional satellite bandwidth

N = number of accesses

R_{min} = minimum acceptable carrier – to – intermodulation noise power ratio (C/I).

For FDMA it can be shown that C/I is 9 dB for several equally spaced carriers spread over the bandwidth [3]. This figure would

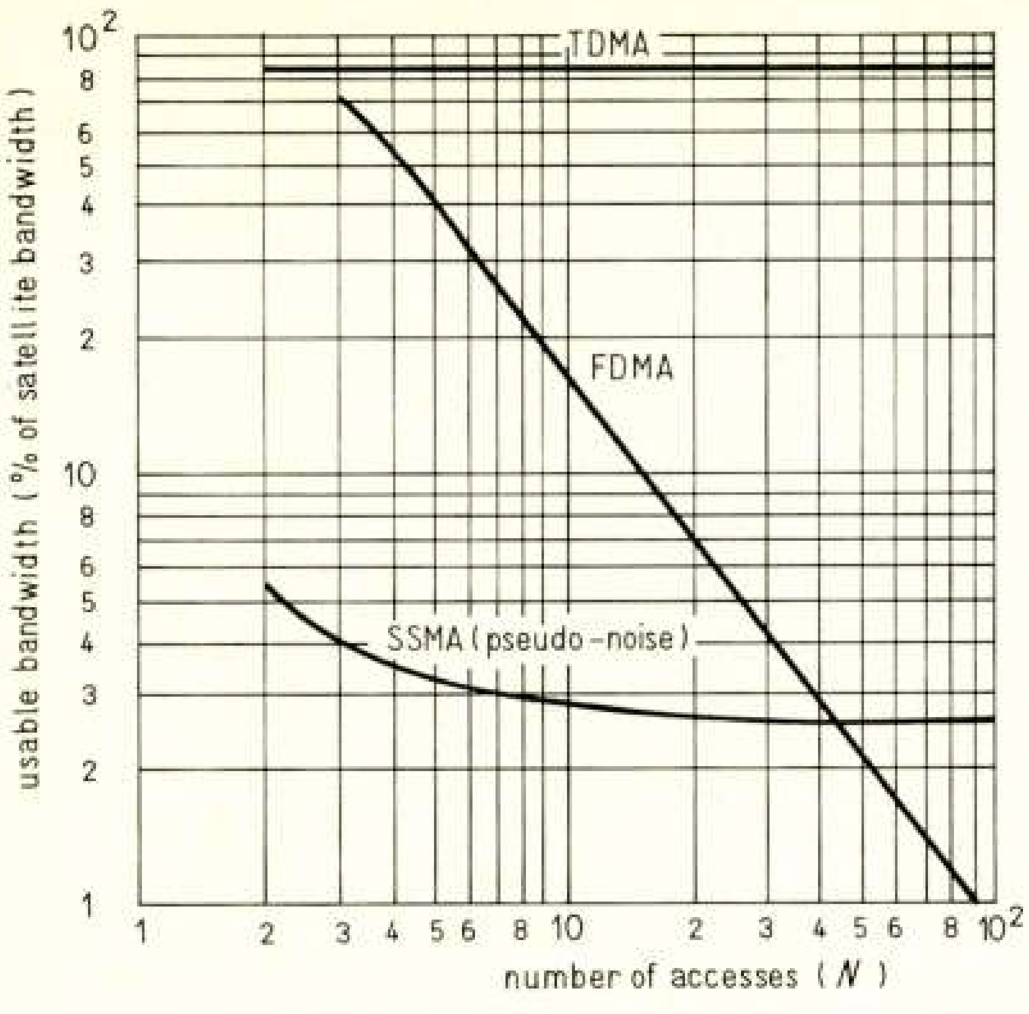

Fig. 4. Usable satellite bandwidth.

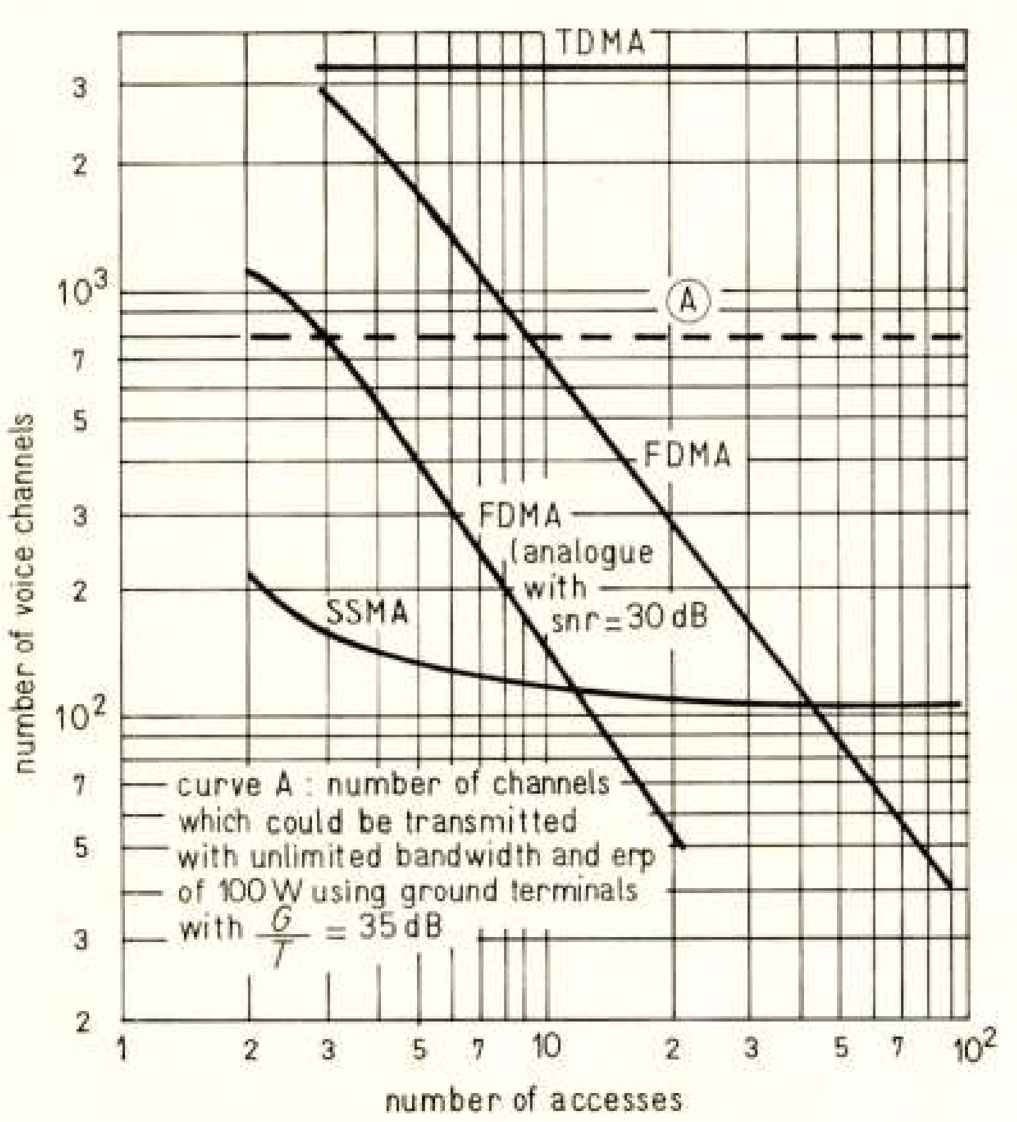

Fig. 5. Maximum number of voice channels through a satellite of 20 MHz bandwidth.

be too low for adequate margin, but it has been shown that a C/I of 18 ... 24 dB is realizable if careful selection of the carrier is made so that intermodulation products fall within unused frequency bands [4]. As shown in Appendix A the bandwidth used would be approximated by the expression given for FDMA.

For TDMA, although full use of the bandwidth could be made by each terminal, guard and synchronization time would reduce utilization from 100% to 80 ... 95% (as for ERP wastage).

For SSMA, the bandwidth utilized for the transmission of information is defined as the hypothetical bandwidth which would be occupied by all accessing carriers with message modulation only. Assuming the system can be considered as being essentially bandwidth limited rather than power limited, then this definition permits a comparison of the transmission capacity using SSMA with that of other access methods.

The value for E_B depends on the type of band-spreading modulation employed. The expression developed has been derived for pseudo-noise modulation.

Fig. 4 shows curves of the percentage of usable bandwidth for all three access methods, in relation to the number (N) of accessing carriers. A value of 14 dB for C/I has been considered to be acceptable.

From the curves of usable bandwidth it is possible to derive the total number of voice channels which can be transmitted through the satellite. This has been done for a satellite of 20 MHz bandwidth assuming digital voice channels of 2400 bits/s. The results are shown in Fig. 5. Also shown (dotted curve A) is the number of voice channels which would be obtained if the satellite ERP were 100 W with unlimited bandwidth and if the ground terminals had a gain-to-system noise temperature ratio (G/T) of 35 dB. This example is quite typical of systems being implemented.

Although the curves shown in Figs. 4 and 5 are based on certain restrictive assumptions, the wide variation in usable bandwidth for the three multiple-access methods warrants certain conclusions to be made:

- TDMA is the most efficient method;
- FDMA is preferred to SSMA, if the number of access carriers does not exceed about 10 ... 20;
- SSMA is preferable for a large number of access carriers, since the utilization is relatively independent of the number of accesses;
- If a ground terminal is to meet a fixed voice channel performance then, for a given value of ERP-to-bandwidth ratio which we shall call (k), the antenna gain-to-noise temperature ratio G/T has an optimum value beyond which no improvement in voice channel signal/noise ratio can be gained. Optimized values for G/T for each access method are tabulated below in Table 3 for $k = 5$ W/MHz (typical for present-day designs).

Table 3. Optimized Values for G/T for each Access Method.

Type of access	Optimum G/T (dB)	Number of accesses
SSMA	32	independent
FDMA	31 ... 38	dependent
TDMA	39	independent

- the optimum value for G/T does not change as long as the ratio 'k' remains constant. Proportional increase of both ERP and bandwidth would, of course, increase the terminal channel capacity:
- increasing 'k' allows G/T to be reduced for optimum performance;
- with an overall antenna efficiency of about 70% which can be shown to be feasible, $G/T = 31$ ... 35 dB would be obtained in either of the following cases:

a. 40 ft antenna having $T = 500$ ... 200 K
b. 20 ft antenna having $T = 125$ ... 50 K

System temperatures only slightly in excess of 200 K can be obtained using an uncooled parametric amplifier. The 20 ft antenna would, of course, require a cooled parametric amplifier. If, however, 'k' were increased to about 20 W/MHz, then an uncooled configuration could be used. It has been shown that for G/T less thann 35 dB, it is cheaper to use an uncooled system

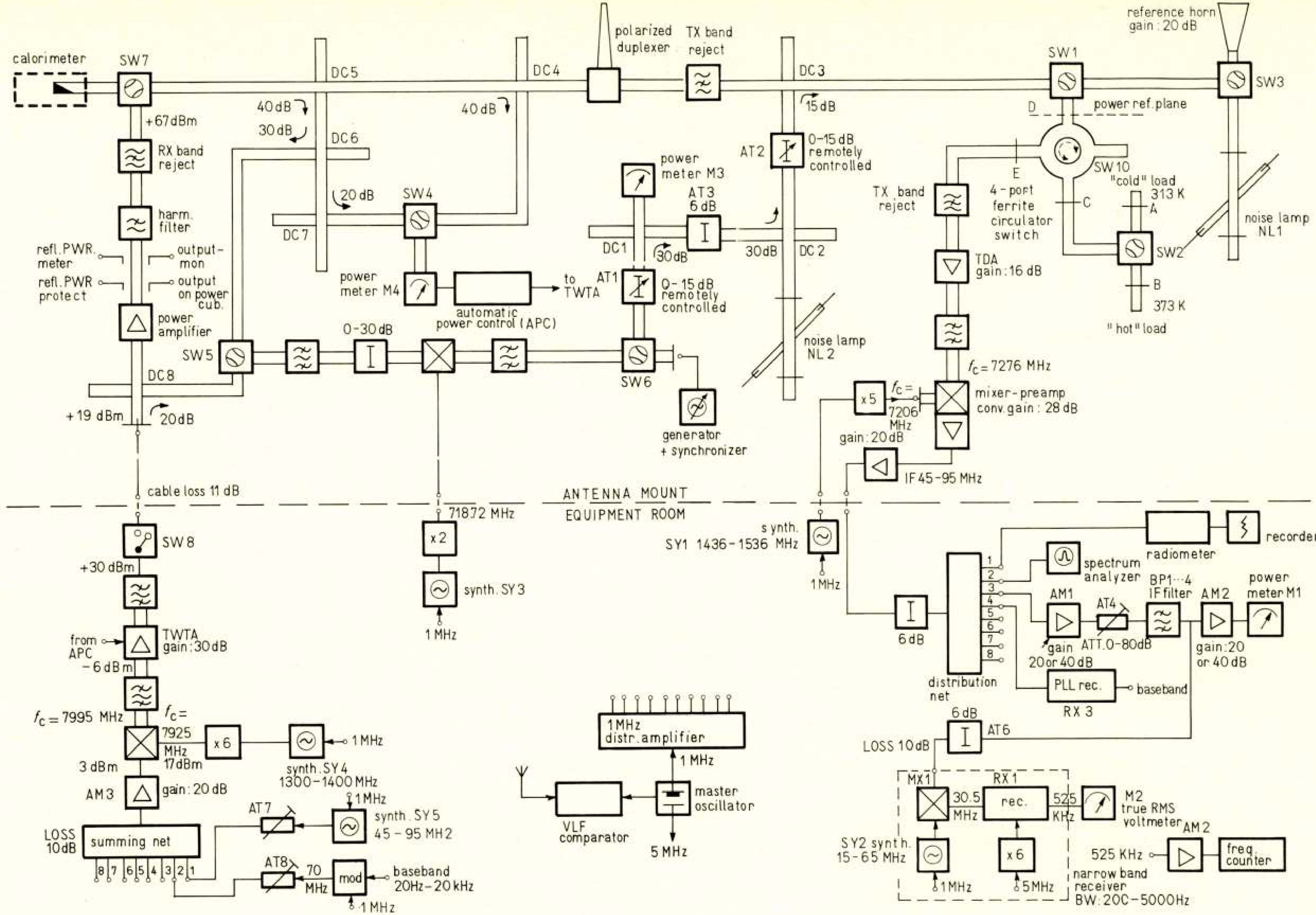

Fig. 6. SHAPE Technical Centre Experimental Satellite Ground Terminal (SET-2).

[5]. The satellite design is therefore very likely to be directed towards increasing 'k'.

4.3. *Operational Features*

Apart from the quantitative aspects of optimizing the efficiency of the three multiple-access methods, it is necessary to take into consideration the limitations and implications of each method as regards equipment availability, system timing and growth, system control, and vulnerability to interference.

From a consideration of the above factors it is found that the characteristics of the three access methods are such that it seems likely they will all be used in future SATCOM systems either singly or in hybrid. Since the design lifetime of a ground terminal may be three to four times longer than that of satellites, appropriate provisions in the design of the link terminal must be made to enable it to work with any modulation method.

5. Minimum Requirements

By considering satellite parameters and modulation/access methods, the minimum requirements of a ground terminal have now been established.

The design and construction of the principal sub-systems will next be reviewed, referring to an experimental ground terminal, called SET-2, designed and constructed at SHAPE Technical Centre.

Fig. 6 shows a detailed block diagram of the communications equipment and the test facilities provided for SET-2. The main terminal characteristics may be summarized as follows:

- 30 ft Cassegrain antenna on a fully steerable elevation over azimuth mount with manual and autotrack facilities. Enclosed in a space-frame radome;
- Parametric amplifier providing a receive-system noise temperature of about 200 K. Alternatively, a tunnel diode amplifier (TDA) giving system noise temperature of 800 K;
- ERP up to 120 dBm calibrated by calorimeter and by calibrated power meter. Instantaneous bandwidth 50 MHz remotely tuned across 7900 ... 8400 MHz band;
- Instantaneous receiver bandwidth about 100 MHz, tunable across 7250 ... 7750 MHz band;
- Dicke radiometer incorporated;
- Equipment using a 10 Megabit spread-spectrum signal for measuring satellite range with accuracies better than 25 meters;
- Second independent access provided either by a second satellite terminal available at SHAPE Technical Centre or by a transmitter connected in parallel with the main high power amplifier.

6. Antenna Sub-System

The main essentials for the antenna sub-system are high efficiency with low noise temperature, both of which imply maximum possible energy concentration in the main beam of the radiation pattern. Also, the feed system loss must be minimized.

6.1. *Antenna Types*

Various types of antenna which can meet the required characteristics to varying degrees are possible. These include the following:

a. Parabolic antenna with focus feed;
b. Parabolic antenna with Cassegrain feed;
c. Parabolic Cassegrain with dielectric feed;
d. Offset Cassegrain;
e. Horn antenna;
f. Casshorn antenna;
g. Phased array.

Certain known data are available regarding the above antenna types and an appraisal of the merits and limitations of the antennas is given below:

- type (a) is well-known and a great amount of data is available;
- type (d) is an attempt to reduce the noise temperature of a standard Cassegrain due to sub-reflector spillover;
- type (e) has high efficiency and low noise qualities but is cumbersome and expensive;
- type (f) attempts to combine the high qualities of a horn with reasonable configuration;
- type (g) may include limited electronic steering but could only be used with synchronous satellites.

Despite the main disadvantage of a Cassegrain system – i.e. beam blockage – this system is generally preferred due to:

- favourable location of feeds and RF pre-amp equipment;
- freedom in primary feed design;
- pattern control using sub-reflector characteristics;
- increased focal length due to hyperboloid magnification.

6.2. *Antenna Efficiency*

The various loss factors with their approximate range values associated with a Cassegrain antenna system are given as:

η_1	= Feed and line	0.98 – 0.80
η_s	= Spillover	0.76 – 0.60
η_a	= Aperture	
η_t	= Reflector surface irregularities	0.95 – 0.80
η_b	= Blockage	0.95 – 0.85
	Total:	0.68 – 0.33

Each loss factor will now be considered separately, and the methods of estimation of their values will be shown.

η_1 can be limited to a few tenths of 1 dB for a Cassegrain system by using simple feeds and short waveguide runs.

η_b is given for a 10 dB illumination taper by the expression:

$$\eta_b \approx 1 - 4\,(d/D)^2$$

where d/D is the ratio of sub-to-main reflector diameters. d/D would normally be chosen for maximum blockage of the feed and sub-reflector combination using the expression

$$d = \sqrt{2\lambda F_m}$$

where F_m is the focal length of the main reflector.

η_t depends on the RMS design tolerance of the antenna surface. The gain reduction factor is given by the expression [6]:

$$\eta_t = \exp\left[-\left(\frac{4\pi}{\lambda}\,\varepsilon\right)^2\right]$$

where ε is the RMS deviation from true surface, measured in wavelengths. It will be seen that the surface tolerance sets the ultimate limit to the upper utilisable frequency for the structure. η_t is usually between $\lambda/60$ and $\lambda/25$.

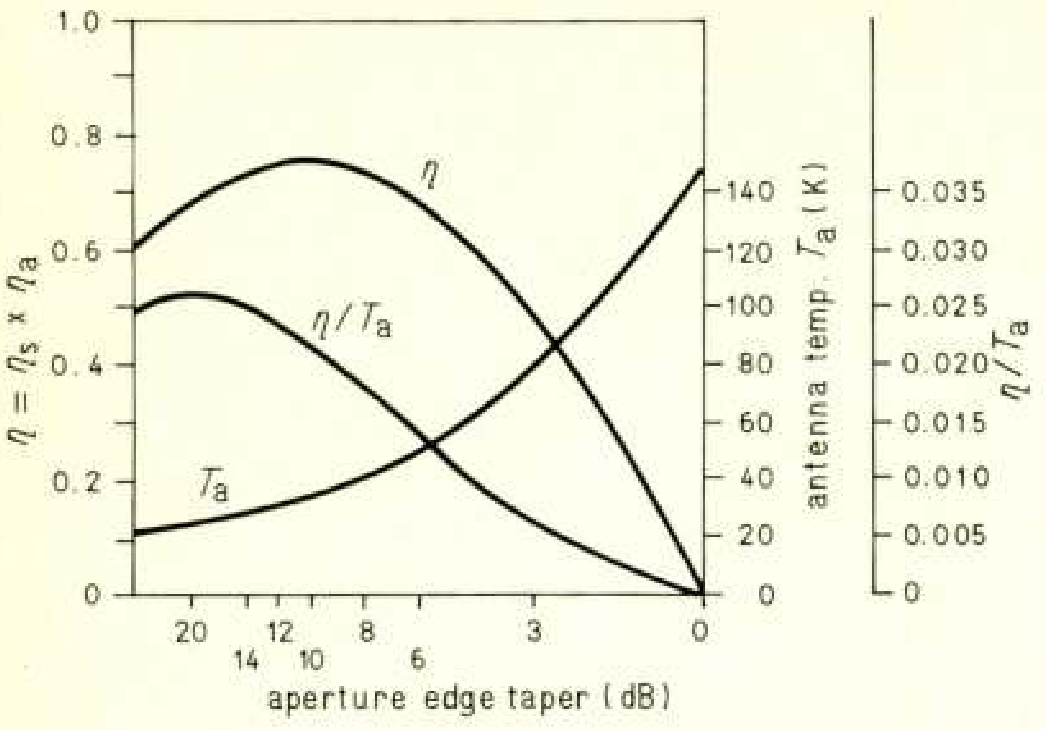

Fig. 7. Antenna efficiency and noise temperature.

η_s & η_a are of course related to the feed pattern in that attempts to achieve high gain by employing a nearly uniformly illuminated aperture will result in large spillover. A plot of η (the product of η_s and η_a) against illumination taper is shown in Fig. 7. The antenna noise temperature (T_a) and the ratio η/T_a are also shown in the same figure. It will be seen that:

- The taper giving maximum gain is about 10 ... 12 dB;
- η/T_a is a maximum at about 18 dB taper, but although the noise temperature would be decreased by about 10% at this point the gain would be about 0.5 dB below maximum. Consequently, for terminals where cooled preamplifiers are not required, and where the system noise temperature would be about 200 K, the antenna should be designed for maximum gain rather than maximum η/T_a.

For conventional antennas net efficiencies of about 55% are often stated. However, to obtain the desirable features of high aperture efficiency and low spillover, a 'shaped' Cassegrain reflector system has recently been developed. Here a normal high efficiency feed is used with a high taper so that the subreflector spillover is low. The sub-reflector is then shaped to provide essentially uniform aperture illumination. The phase condition thereby violated is restored by slightly changing the main reflector from the parabolic contour. It is claimed that net antenna efficiencies over 70% can be obtained by this procedure. This surface shaping technique is also beneficial in that it reduces the noise which normally results from energy spilling over the sub-reflector and from diffraction losses.

6.3. *Antenna Feeds*

Feed assemblies normally fall into three categories:

- four-horn feed;
- five-horn feed;
- multi-mode feed.

In a four-horn monopulse feed assembly, the sum channel aperture has a double cosine distribution in the H-plane which reduces the overall gain factors ($\eta_s \times \eta_a$) to less than 60% for 10 dB taper and increases the noise temperature to about 60 K [7]. The errorchannel sensitivity however is good.

In the five-horn assembly the central horn is designed to give optimum gain or gain/temperature ratio and is used to derive the reference (sum) signal for tracking and, in conjunction with a polarized duplexer, for combination of transmit and receive.

Efficiencies of 70 ... 80% with broadband characteristics are possible. The four outer feeds are connected in pairs to provide azimuth and elevation error channels for automatic tracking.

Feeds with still higher efficiency have been developed based on utilizing higher modes in the conical or square waveguide. Multi-mode feeds have an enclosed power of about 95% with typical illumination tapers, but are somewhat bandwidth limited [8].

6.4. *Feeds for Autotrack*

6.4.1. *Static Split*

The amplitude comparison monopulse technique (static split) is often used for autotracking, mainly for its simplicity of implementation compared to conical-scanning which requires either a nutating feed or rotating reflector with attendant mechanical complications. Moreover, conical scan causes a loss of the communication signal.

The amplitude comparison monopulse employs two overlapping antenna patterns to obtain the angular error in one coordinate. It is possible to produce the required static split characteristics by using either a single or double multi-mode horn or a multi-horn array. The multi-mode horn is capable of high efficiencies, but requires tedious adjustments of the phases and amplitudes of the various modes to produce satisfactory performance over a 10% working bandwidth. Most designs use either four of five element feed arrangements.

Four-horn arrangements give a sum channel gain which is lower and an antenna temperature which is higher than that achievable with other designs. The error-channel sensitivities, however, are good.

A practical feed system which does not result in a serious compromise in the gain and noise temperature of the aperture used for a communication channel is the five-feed arrangement. Moreover, as the outer feeds are used only for reception, their hybrid circuitry can be of low-power design.

6.4.2. *Five-Horn Monopulse*

The arrangement of the five horns is shown in Fig. 8. The central horn which can be designed to give optimum gain or gain/temperature ratio is used to derive the reference (sum) signal for tracking as well as for communication transmit and receive. The central horn dimensions for the two cases in terms of the wavelength λ, and the antenna focal length F and diameter D are given

for maximum gain G		$a = 1.5\,\lambda F/D,\ b = 2.1\,\lambda F/D.$
for maximum	G/T	$a = 1.7\,\lambda F/D,\ b = 2.5\,\lambda F/D.$

For best difference-channel sensitivity the radiation patterns of the outer feeds should cross over as close as possible to their first zeros. If wider feeds are employed, the slope of the difference pattern reduces since the beam slope decreases rapidly

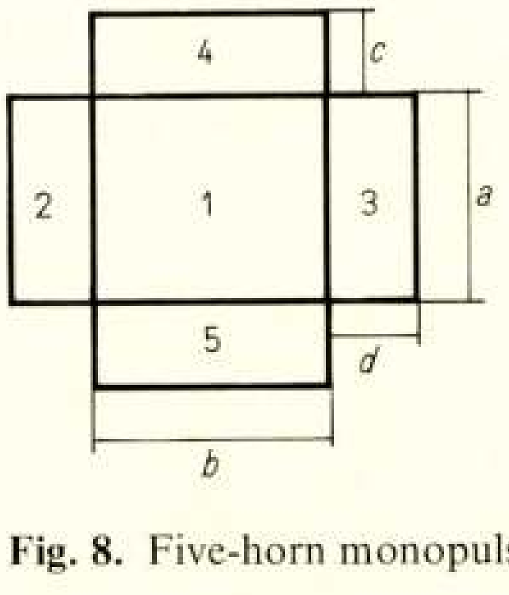

Fig. 8. Five-horn monopulse feed arrangement.

as the cross-over point approaches the peak of the first sidelobe; conversely, narrow feeds, i.e. smaller c and d, have a higher basic slope, but the absolute difference-channel sensitivity reduces due to the lower feed gain. The optimum values for c and d are given below;

for maximum G : $c = 0.55\ \lambda F/D$, $d = 0.42\ \lambda F/D$.
for maximum G/T: $c = 0.5\ \lambda F/D$, $d = 0.3\ \lambda F/D$.

It can be shown that the difference-channel sensitivity defined as the slope of the difference pattern in volts per sum-channel beamwidth (for unit sum-channel voltage) is:

1.14 V for the E-plane;
0.64 V for the H.plane.

If the central horn were designed for optimum G/T, which requires a greater aperture, the difference-channel sensitivity would be reduced due to the smaller aperture of the outer horns.

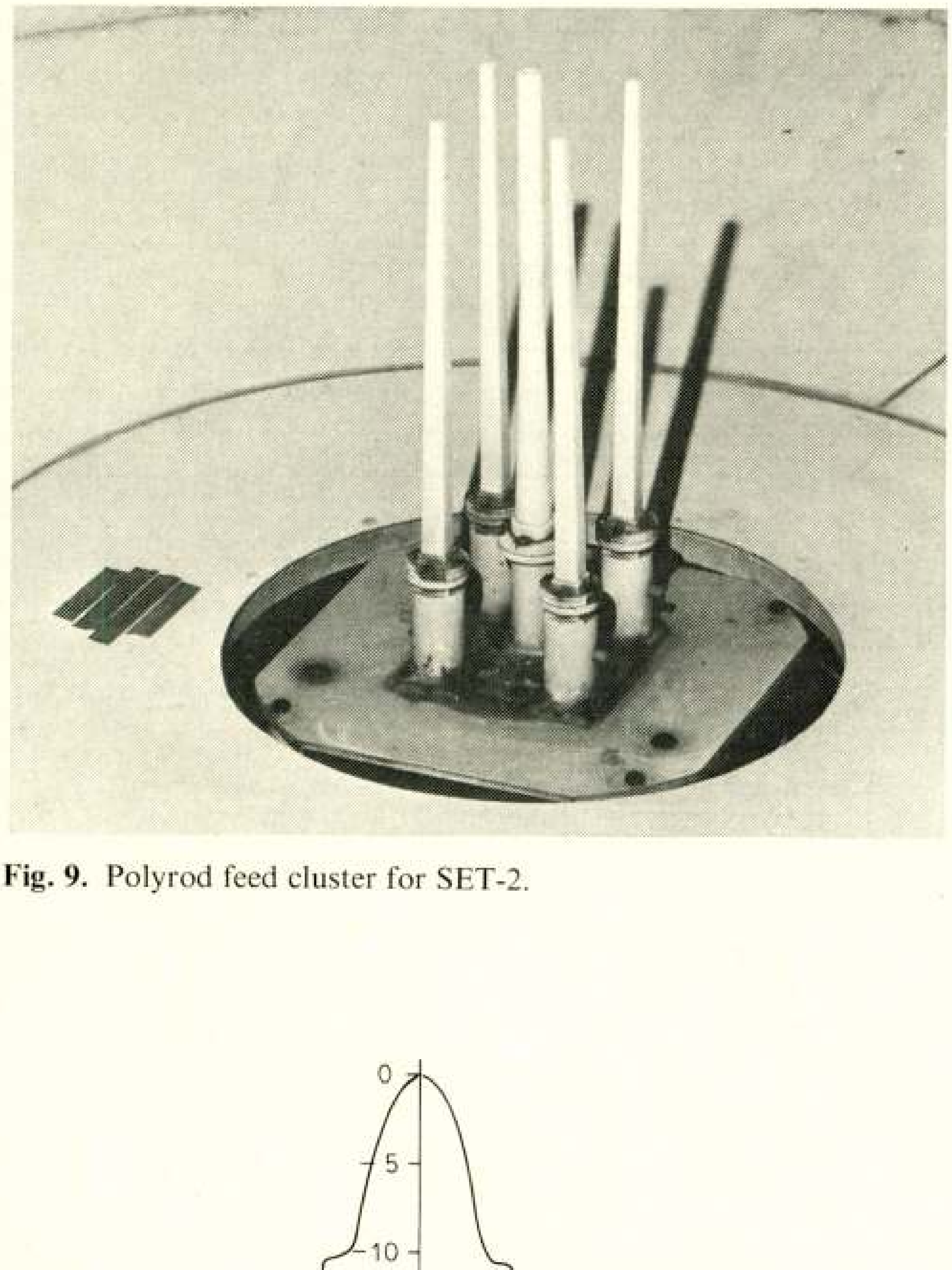

Fig. 9. Polyrod feed cluster for SET-2.

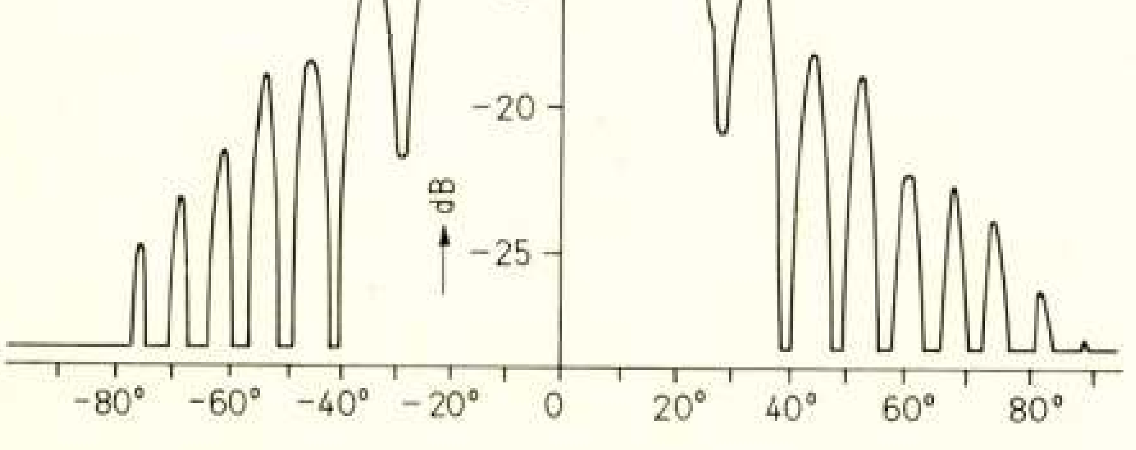

Fig. 10. Radiation pattern of central dielectric rod (SET-2).

6.4.3. *Five-Polyrod Feed System*

Polyrod feed

The phase centres of the outer feeds can be brought closer together by loading the horns with a suitable low-loss dielectric material or by using dielectric rods (polyrods) as feeds. We adopted the latter arrangement in our ground terminal [9] using polyrods in a five-feed configuration as shown in Fig. 9. The spacing of the feeds is such that, apart from an optimized radiation pattern for the communication channel, we derive not only the difference signal but also the sum signal from the outer feeds so that the tracking and communication functions are completely separated. This provides great flexibility for independent experimentation in the two channels.

In the dielectric rod radiator the launching section consists of a circular or rectangular waveguide which couples part of the input power into a surface wave, which travels along the rod to the termination where it radiates into space. The ratio of power in the surface wave to the total input power is usually between 65 and 75%. Power not coupled into the surface wave is directly radiated by the feed in a pattern resembling that radiated by the feed itself. The polyrod may be tapered at the feed end and along the body. The feed taper increases the efficiency of excitation and also affects the shape of the feed pattern while the body taper suppresses sidelobes and increases bandwidth. The length of a polyrod determines its gain and beamwidth; for 10 dB beamwidth of 30° the polyrod will be about 10 λ long. Fig. 10 shows the radiation pattern of a polyrod. The sidelobe level is about -10 dB and the 3-dB bandwidth is about $\pm 10\%$. If the polyrod is to be used for the transmitting feed it must also have the required power handling capacity. The material used for our feeds is polytetrafluoroethylene which has the following properties at X-band frequencies:

dielectric constant (ε)	$= 2.1$
power factor ($\tan \delta$)	$= 1.5 \times 10^{-4}$
thermal conductivity	$= 2.1$ cal/cm^2/h

Calculations based on these figures, assuming no conduction losses, indicate that the polyrod temperature would be less than about 100 °C when handling 5 kW of CW power, and in fact measurements confirmed this.

Difference and Sum-Channel Patterns

The arrangement of the five polyrod feeds, designated A, B, C and D, together with the comparator unit to derive sum and difference signals for tracking is shown in Fig. 11. Let us assume that the central feed is designed to give 10 dB taper with a gain factor of 0.75 and that the outer feeds have 4 dB taper and a gain factor of 0.6.

The radiation patterns of the polyrods can be approximated by the expression:

$$g(u) = \mathrm{A} + \mathrm{B}\left[1 - \left(\frac{2r}{D}\right)^2\right] \tag{1}$$

where D is the antenna reflector diameter and r is the radial distance.

The secondary pattern due to the central feed is then given by:

$$g_c(u) = 0.6\,\frac{\mathrm{J}_1(u)}{u} + 5.6\,\frac{\mathrm{J}_2(u)}{u} \tag{2}$$

and the secondary pattern due to the outer feeds by:

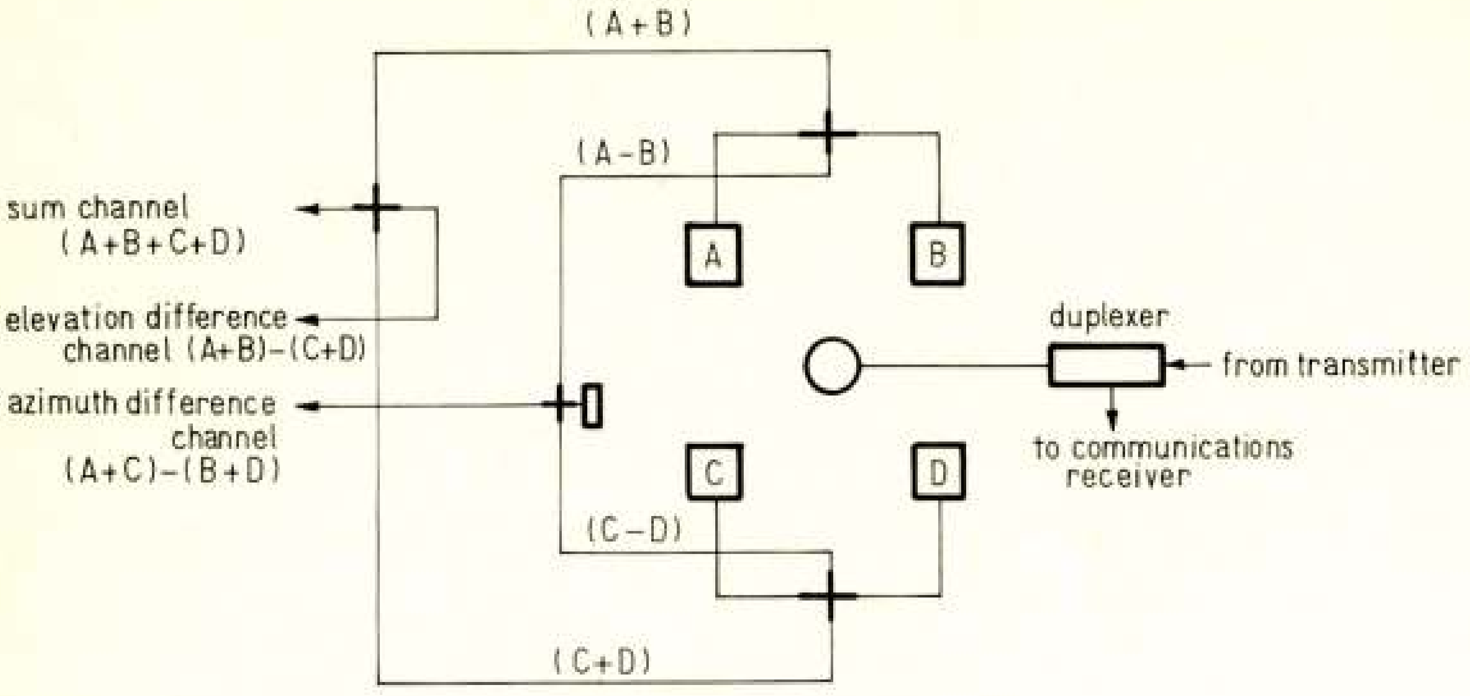

Fig. 11. Five polyrod feed cluster and comparator unit.

$$g_o(u) = 1.2\ \frac{J_1\ (u)}{u} + 3.2\ \frac{J_2\ (u)}{u^2} \qquad (3)$$

where J_1 and J_2 denote first and second order Bessel functions respectively and $u = \frac{\pi D}{\lambda}\ \theta$ with θ as angle off boresight.

From equations (2) and (3) we see that $g(u) = 1$ for $u = 0$.

If we take into account the difference (1 dB) in the gain factors, equation (3) becomes:

$$g_o(u) = 1.07\ \frac{J_1\ (u)}{u} + 2.85\ \frac{J_2\ (u)}{u^2} \qquad (4)$$

If we assume that the cross-over level between (A + B) and (C + D) is 10 dB below the peak of the central feed and that there is no coupling between the feeds, we can show from equation (4) that for $\lambda = 3.85$ cm and $\frac{F}{D} = 1.9$ the separation between the outer polyrods would be 10.2 cm, i.e. just over 2.5 λ.

The sum and difference channel pattern of the SET-2 antenna designed on the basis of the foregoing considerations are shown in Fig. 12. Theoretical results agree very well with measured radiation patterns. As will be seen, the gains of the sum and the difference channels are respectively 7 and 3 dB below that of the central polyrod. The difference-channel sensitivity can be shown to be 1.4 V which is better than that of the five-horn assembly.

7. Antenna Mount and Control Sub-System

This sub-system consists of an antenna mount, servo-drive and control and a beacon receiver. Most satellite communication terminals use an elevation-over-azimuth mount.

7.1. *Servo-Drive and Control*

Since the major disturbances in tracking satellites (gusting wind and noise) are essentially Gaussian in distribution, the tracking accuracy of the antenna drive control is normally specified in statistical form. The 'one sigma' angular error required is often given as one-tenth of the 3 dB bandwidth.

The primary operational modes usually required are:
- manual positioning and track;
- automatic track;
- slew.

There are also several possible sub-modes of interest including:
- pre-set position;
- spatial acquisition scan;
- automatic frequency acquisition;
- standby.

Regarding speeds and angular ranges which we have already considered, it is worthy to note that a feature whereby the servo-system would be in the standby mode, but will automatically revert into autotrack for short periods whenever the error signals exceed certain limits, could significantly increase the lifetime of the servo-drive unit. The pre-set position sub-mode may be very useful in switching rapidly and accurately from satellite to satellite and to other positions.

7.2. *Beacon Receiver*

The main function is to receive a sum signal, an elevation error signal and an azimuth error signal from the feed/comparator unit, and to derive d.c. error signals for the servo-drive in automatic tracking mode, beacon identification, and timing pulses. The following provisions should be made:
- drive the antenna in automatic tracking mode;
- permit manual track on both beacon and communication signals;
- provide flexibility to re-tune in the event of beacon frequency change;
- operate satisfactorily with beacon carrier-to-noise density ratios down to, say, 25 dB/Hz;
- automatically search a frequency spectrum and lock on the beacon frequency.

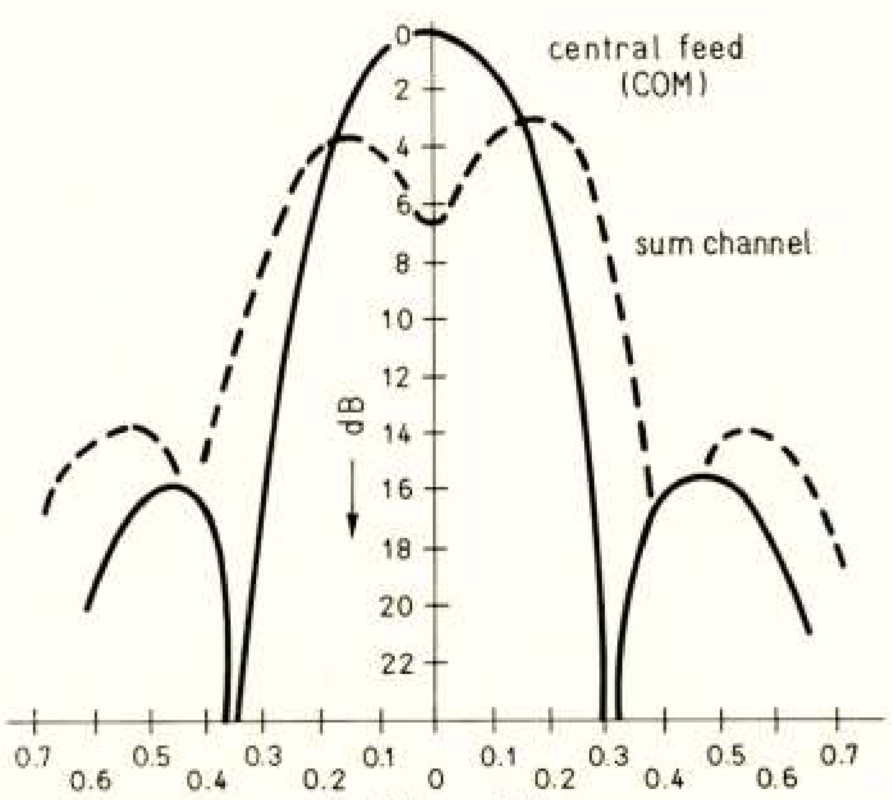

Fig. 12. SET-2 antenna patterns.

A number of filters should be included to change the noise bandwidth of the tracking loop to give double-sided noise bands in the ranges 50 Hz ... 1,000 Hz.

8. Transmitter

Reference has already been made to the advisability of designing a ground terminal to incorporate provisions for operation using any modulation or access method. Such provisions relate to:
- first IF interface;
- amplitude/frequency and phase linearity characteristics;
- number of up-conversions.

Consideration of possible traffic patterns in operational systems indicates that the maximum instantaneous IF bandwidth required is very likely to be determined by the SSMA equipment which may be used with the terminal. Depending on the spread used for SSMA the IF can be 70, 105 MHz or higher. In the near future the spread is likely not to exceed about 50 MHz.

Klystrons (for the HPA) with a power output of several kilowatts and 50 ... 100 MHz instantaneous bandwidth are readily available. They should be capable of being tuned across the 500 MHz band (Fig. 2), and would thus be usable for different modulation and access methods.

In a network comprising several ground terminals, intercommunication could be provided in two different ways:
- by making use of the inherent multi-destination feature of satellite communications whereby each terminal transmits one carrier;
- by transmitting several carriers from each terminal, each carrier being destined for reception at a particular terminal.

Obviously, FDM would be restricted to the use of 1 or 2 carriers per terminal, due to the fact that satellite bandwidth utilization decreases rapidly with increasing number of accessing carriers. Further growth would make it necessary to design the terminal so to permit easy and accurate change of carrier frequency.

Translation of the IF signals to the appropriate radio frequency band may be accomplished in one, two, or three stages of up-conversion. Fig. 13 shows a very flexible arrangement of three stages of up-conversion which is capable of transmitting multiple carriers within the frequency range 7.9-8.4 GHz and in which carrier frequencies or blocks of carrier frequencies may be changed without any tuning of the various filters involved in the chain.

When travelling wave-tubes of 500 MHz bandwidth become available, which can deliver several kilowatts of power, an arrangement as depicted in the lower chain may be adopted. This has advantages with regard to flexibility and simple provision of standby facilities, always required in terminals with high reliability.

A typical specification for the transmitter sub-system would also cover the following characteristics:

a. *Phase-Linearity and Amplitude-Response*

This would depend on the characteristics of the bandspreading equipment. Assuming that the phase linearity requirement for this would be about as stringent as for multi-channel FM signals, the maximum allowed phase distortion may be calculated as shown in Appendix B. It can be shown that satisfactory performance would be obtained if the phase non-linearity is not more than about 0.05 radians over the centre of the band, reaching to about 0.3 radians at the band edges. The amplitude response would also be determined by the spread-spectrum equipment characteristics, as well as by the power control for multi-carrier operation. The response is often quoted as 1 dB over 50 MHz or more.

b. *Matching between HPA and Feeder*

Klystron tubes as used in power amplifiers, are known to be capable of operation with a VSWR of up to 1.5 without suffering damage. The echo effect which this mismatch could cause to an SSMA signal should be kept small by applying a line-feed not longer than about a meter [10].

c. *Level Stability*

Level stability as determined by the gain variation due to all causes should be within 1 dB over several hours. The level stability required dictates the use of automatic gain control which is applied to the units liable to gain change. This is shown in Fig. 13.

d. *Terminal Availability*

Terminal availability which is often desired to be in excess of 99%, necessitates the provision of standby facilities. When the phasing of high-power amplifiers is successfully achieved it would be desirable to arrange the 'worker' and 'standby' power amplifiers so that they may be combined to double the power output thus limiting interference.

9. Communications Receiver

For a satellite with $k = 5$ W/MHz, being accessed by more than

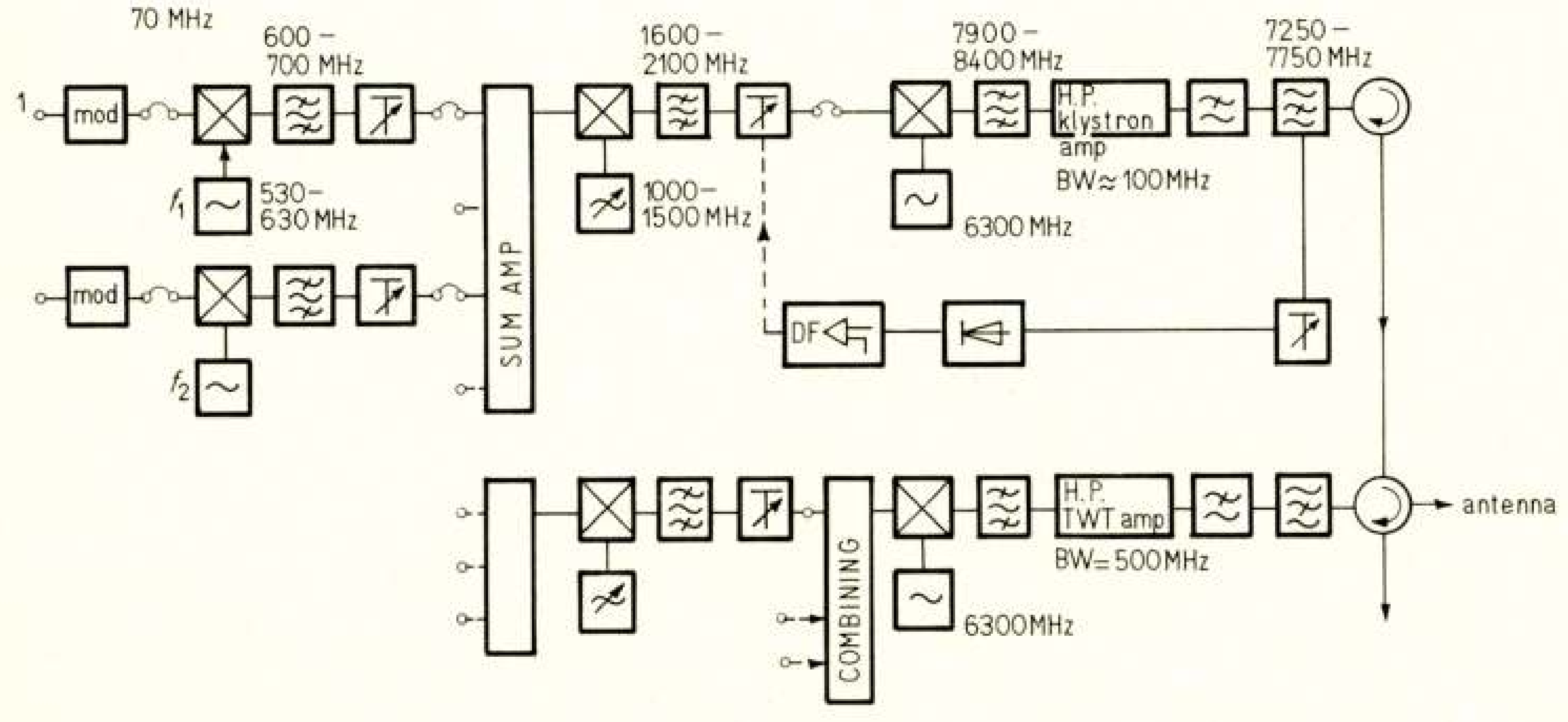

Fig. 13. Transmitter with triple up-conversion.

about 8 carriers, a terminal with $\frac{G}{T} = 35$ dB appears to be optimum when operating in FDMA and/or SSMA. This performance may be achieved by a 40-ft terminal employing a room temperature parametric amplifier, whereas optimum performance for TDMA would require $\frac{G}{T} = 39$ dB. This would need either a 40 ft antenna using a cooled parametric amplifier or a 60 ft terminal employing a non-cooled amplifier. Successful TDMA operation, however, will still take some time to be developed and for some years to come it is very likely that satellites with increased ERP-to-bandwidth ratios will be used in FDMA and SSMA modes. It appears reasonable, therefore, to assume that a terminal with an antenna of 40 ft diameter, employing an uncooled parametric amplifier would be, as it were, the universal terminal required to operate in the frequency range 7.25-7.75 GHz.

Figure 14 shows a flexible communications receiver for such a terminal. To enable any carrier within the 500 MHz band (CCIR) to be received would necessitate the pre-amplifier having an instantaneous bandwidth of 500 MHz, and a noise temperature not exceeding about 200 K. We have already shown that parametric amplifiers with 120 K can meet this system noise requirement. Room temperature parametric amplifiers with this noise performance are commercially available but only with an instantaneous bandwidth of about 50 MHz, which, however, can be tuned within the required 500 MHz. It may be expected that in the next few years the successful development will be shown of a room temperature parametric amplifier with the required wide bandwidth and other desired characteristics regarding gain, stability and reliability.

The gain of the amplifier should be sufficiently high to keep the noise contribution of the mixer stage at a reasonably low level. It is also necessary that the gain variations be kept within about 1 dB. The gain stability of the amplifier is governed by the amplitude and frequency stability of the microwave pump source used to drive it. It can be shown that for 1 dB stability at 20 dB gain, the pump level must be held stable to 0.1 dB. A pump source which is temperature-controlled for frequency stability and amplitude-controlled for gain stability, and using all-solid state construction, would probably be needed to obtain the performance, long life and reliability required.

Demodulation of the various carriers, which may be located anywhere within the 500 MHz bandwidth, is best accomplished by allocating contiguous carriers into sub-bands of, say, 50 ... 100 MHz wide which are then down-translated by using separate converters. Consequently, a branching network is required to feed the required number of down-converters.

The demodulators should be provided with automatic gain and frequency control facilities. The dynamic range should be consistent with expected nominal carrier levels and changes due to precipitation as well as long-term gain changes in the receive chain. Provision of a frequency search mode with a range consistent with expected frequency changes would make carrier acquisition easy and thus reduce communication outages. Other desirable features to facilitate corrective action when the link performance is degraded are:

- monitoring of carrier levels down to threshold;
- monitoring loop error voltages in the PLLs;
- monitoring of noise power.

The amplitude and phase response, gain stability, and phase jitter requirements for the receiver sub-system should be as good as for the transmitter sub-system.

10. Local Frequency Supplies

It is essential that the long- and short-term stabilities of all local frequency supplies be very high, and, moreover, it must be possible to set up any desired frequency within a certain given accuracy. In a multi-station communication complex where each terminal may be in communication with many others, keeping frequency uncertainties to a minimum results in simpler, cheaper and more reliable terminal equipment, and easier terminal maintenance and alignment. Derivation of all local frequencies from one station master oscillator with good long-term stability using frequency synthesis techniques is therefore very desirable. This would keep the search ranges and times of VCOs and pre-detection bandwidths in communications demodulators and tracking receivers to a minimum. Oscillators with long-term stabilities of 1 in 10^9 or better per month are readily available and should be used. Since the required frequencies are derived by multiplication (N) of the master oscillator frequency, any required spectrum purity at X-band implies at least an N times better performance of the master oscillator.

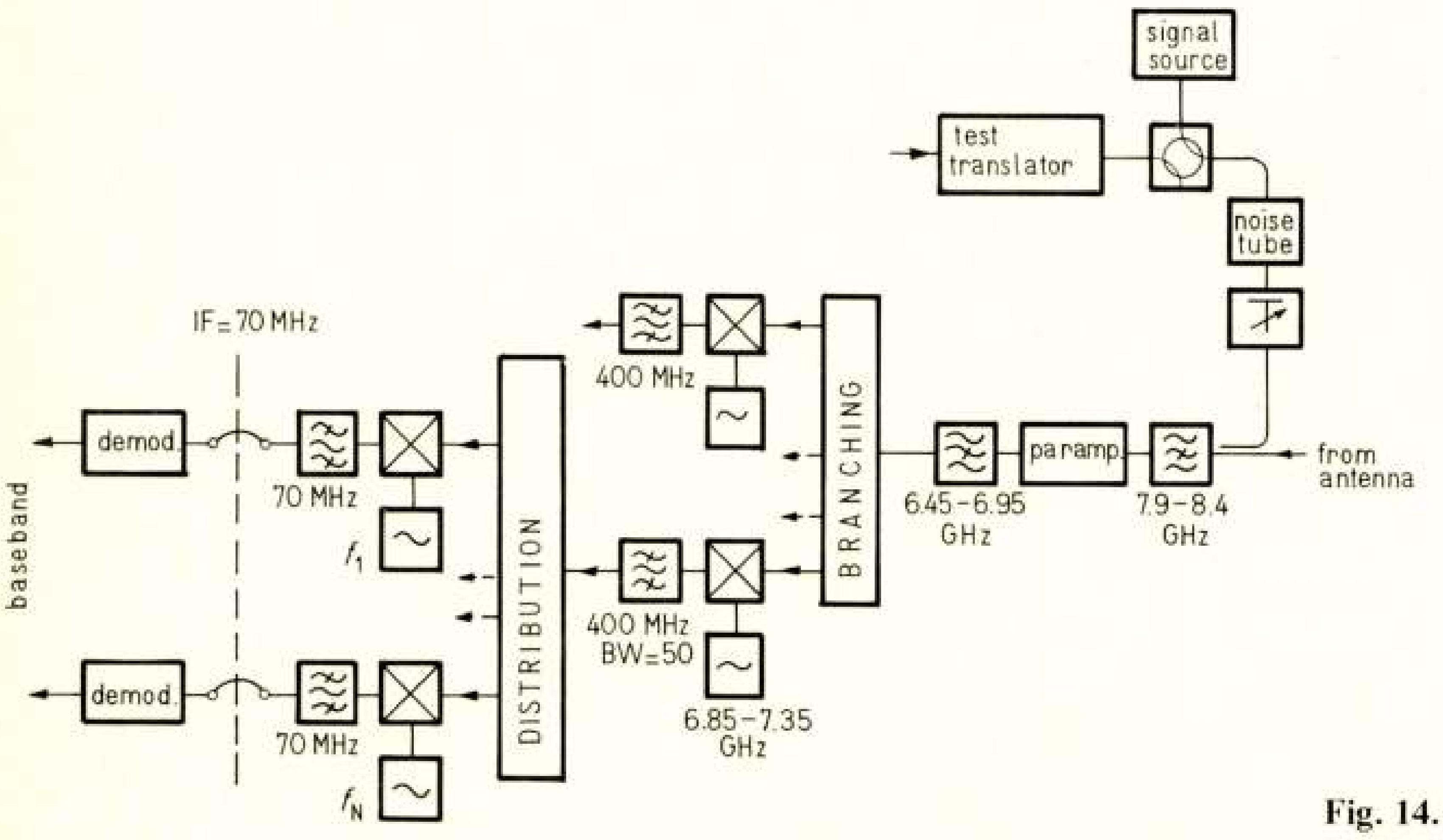

Fig. 14. Communications receiver with double conversion.

The short-term stability of a crystal oscillator is usually defined as the RMS fractional frequency deviation for averaging times ranging from 100 µs ... 10 s. At the longer averaging times it is predominantly determined by oscillator defects but for very short averaging times it approaches the limits set by the thermal noise of the crystal. By way of example, the stability characteristic of the SET-2 Master Oscillator is shown in Fig. 15.

It is shown in Appendix C that if the satellite transponder were to use a master oscillator with the characteristics of Fig. 15, the jitter noise would be about 2 $(Hz)^2$. This would correspond to a negligible testtone-to-noise ratio of about 77 dB assuming a channel RMS deviation of 10 kHz. Using the same oscillator in ground terminals where the up and down frequency translations over a link would be 10 times greater than the translation within the transponder, the jitter noise would be about 400 $(Hz)^2$. This would result in a testtone-to-jitternoise ratio of about 54 dB.

There are other sources of phase jitter which will have to be added to the synthesizer noise to calculate the overall noise due to phase perturbations of the carrier signal. The principal sources are the high power amplifier (HPA), the parametric amplifier (PA) and the modems.

For example, if we require that the channel SNR due to jitter will not be worse than 51 dB, then we can assign 400 $(Hz)^2$ to the noise due to sources other than the synthesizers. If all this noise were produced by the HPA we could determine that noise in terms of phase jitter in the following way:

The phase modulation noise spectrum of a high-power klystron amplifier, including the effects of vibration and acoustic noise, may be expressed by:

$$\Delta\Phi = \frac{\alpha}{f}$$

where $\Delta\Phi$ = RMS phase deviation in a bandwidth of 1 Hz,
and f = frequency of modulation.

The power per cycle bandwidth relative to carrier is therefore given by:

$$\eta(f) = \frac{(\Delta\Phi)^2}{4} = \frac{\alpha^2}{4f^2}$$

and the RMS noise deviation in any channel is given by:

$$(\Delta f)^2 = 2\int_{f_1}^{f_2} f^2\eta(f)\,\mathrm{d}f$$

For a standard 3.1 kHz voice channel this gives:

$$(\Delta f)^2 = \frac{3100}{2}\alpha^2$$

Hence for $(\Delta f)^2 = 400\ \mathrm{Hz}^2$ we obtain $\alpha \approx \frac{1}{2}$.

Thus the total RMS phase jitter in a band extending from 100 Hz is given by:

$$\sqrt{\left(\int_{100}^{\infty}\frac{1}{4f^2}\,\mathrm{d}f\right)} = 0.05 \text{ radians}$$

This is achievable with existing klystron power amplifiers.

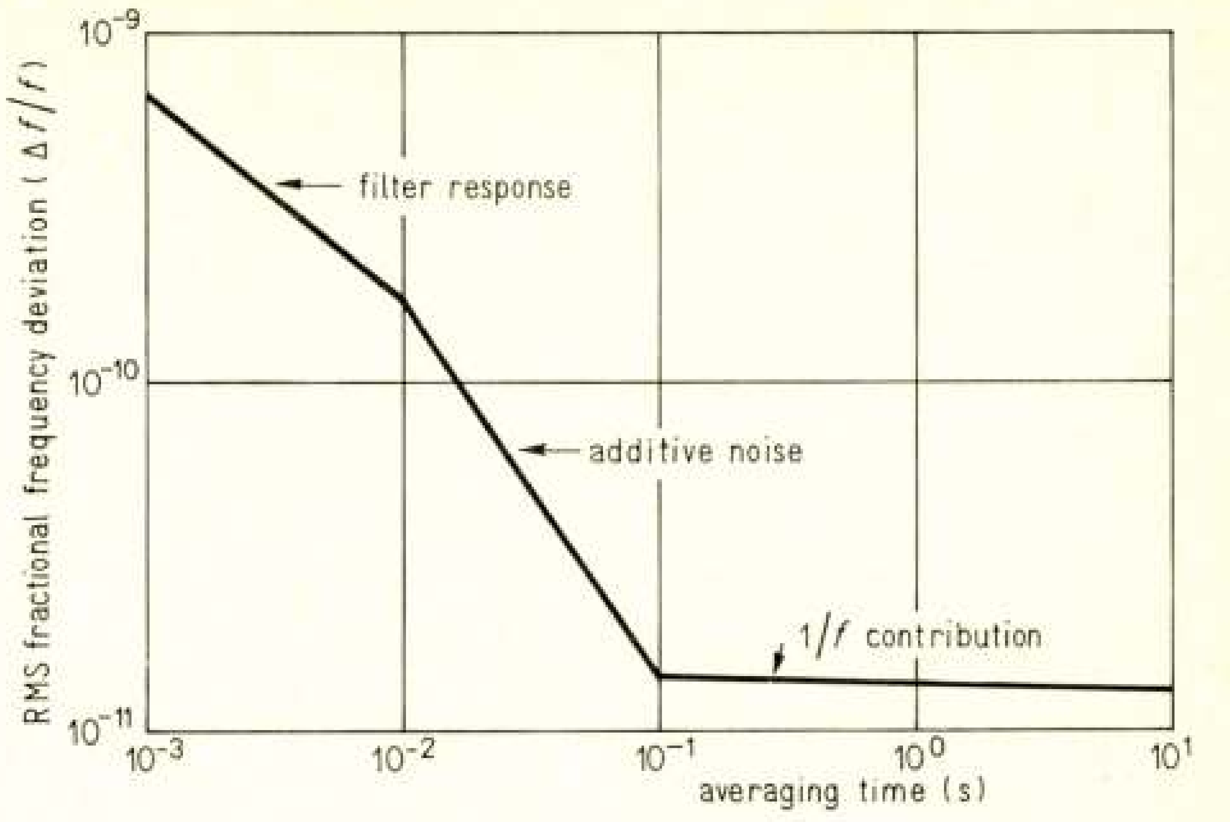

Fig. 15. SET-2 5-MHz precision crystal oscillator stability characteristics.

11. Reliability

11.1 *Methods Applied*

Reliability is generally defined quantitatively as the probability that the unit, equipment, or system concerned, will perform its intended function without failure for the required period of time under stated operating conditions. In our case the terminal reliability is derived from the desired overall communications reliability which may be expressed as user-to-user circuit availability. This is dependent on the satellite and ground terminal reliability and on the overall system configuration.

A reliability figure of 99.9% is often specified for all channels, corresponding to an out-of-service period of less than 9 h/a. Measures usually taken include:
- satellites with redundant internal units and assemblies and capable of being rapidly repositioned;
- configuration of ground terminals and interconnection facilities suitable for rapid re-routing of circuits;
- ground terminals with redundant units, facilities for rapid fault location, and reliable primary power.

11.2 *Communications System Control*

If all the measures discussed above were taken than there would still remain the variations in receive signal level and system noise power due to external factors. These external factors are:
- excess attenuation caused by cloud and rain. For example, this could exceed 2.5 dB for 1% of the time in Central Europe;
- excess noise due to rain and cloud attenuation, manmade interference, and to sun and moon conjunctions;
- interference, intentional or otherwise, received via the satellite;
- tracking degradation due to strong winds – assuming, of course, no radome. For example, wind speeds of 45 m.p.h., gusting to 60 m.p.h., may occur for 0.1% of the time;
- orbital variation due to precession, earth sensor noise producing a loss of about 0.5 dB;
- moon conjunction giving rise to a degradation of about 3 dB;
- conjunction with other satellites.

Other effects, that in addition to the above-mentioned external factors, cause performance degradation, are:
- increase in receiver noise temperature and in demodulator threshold due to changes in ambient temperature and ageing;
- errors of measurement of quantities such as transmitter power and receive carrier level;

- the fact that control action may not be initiated, particularly for a manual control system, until a certain error is built up.

From the factors we have referred to, the total reduction of carrier-to-noise density, which could be exceeded for 1% of the time, has been estimated at about 3 dB.

Quantities which are controlled in real time to counteract the undesirable variations are:
- terminal transmitter power;
- in extreme cases, information capacity.

The complex interrelationships which exist between the different carriers transmitted, and the degrees of interconnection between terminals, together indicate that any control action must consider the system as a whole and the following data must be monitored and measured regularly by the terminals:
- transmitter output power;
- level of each received carrier;
- signal-to-noise ratio for each received baseband;
- terminal alarm status;
- transmit and receive pilot levels;
- communications mode used for each carrier.

11.3. *Use of Radomes*

Link terminal reliability, with the attendant problems of structural design and the maintaining of reflector tolerance, is considerably eased if the antenna is operated in a controlled environment.

Fig. 16. General view of SET-2 antenna system with radome lifted.

The use of a radome may ease the requirements on the complete antenna and control system which would otherwise have to be designed to work, and ultimately survive, under adverse weather conditions. It would, however, introduce some loss and an increase in antenna noise. Figures quoted are about 0.5 dB dry weather loss and about 10 ... 15 K noise temperature increase [11]. In rain conditions, these figures would increase but such increase should be compared with increase of loss due to tracking error under wind conditions. At SHAPE Technical Centre we acquired a space-frame radome three years ago for our terminal, intended for three purposes: to increase terminal availability which was limited by inadequate drive torque, to provide protection for personnel and equipment, and finally, to conduct loss and noise tests and to gain experience in the use and maintenance of radomes.

Figure 16 shows the SET-2 terminal with the radome being lifted.

12. Calibration of Terminals

12.1 *Error Analysis*

We have just seen that for terminals which are used either in an operational communications system or as an instrument there is the requirement for measuring accurately:
- the level of the received signal power;
- the terminal effective radiated power;
- the noise temperature.

It is clear that the assessment of the possible error in any measured quantity is of fundamental importance in science, and in operating a communications system efficiently. The accuracies required in an on-orbit testing of X-band communications satellites are as stringent as those figures that would be obtained under laboratory conditions and Table 4 gives the values achieved in the SET-2 terminal.

Table 4. SET-2 Measurement Accuracies.

Item	Value	Accuracy
Standard gain horn	nom. 16 and 20 dB	0.06 and 0.1 dB
Noise power standard (noise tube)	10530 K *)	2.1%
Hot and cold loads	375.24/313.04 K	1.4-0.4 K
Radiometer, power linearity	—	2%
Antenna axial ratio at transmitter frequencies	1.4 dB	0.2 dB
Antenna axial ratio at receiver frequencies	1.6 dB	0.2 dB
Polarization loss	—	0.2 dB
Antenna gain at 7995 MHz	50.82 dB	0.34 dB
Antenna gain at 7276 MHz	50.74 dB	0.38 dB
Transmitter power	max. +67 dBm	0.26 dB
Antenna noise temperature	45 K	10 K
Receiver noise temperature	836 K	27 K
Receiver noise bandwidths	nom. 1 MHz and 5 kHz	2.5%
Received signal power measurement	from −90 dBm to −120 dBm	0.3 dB
Attenuators/couplers	up to 40 dB	0.03-0.05 dB
Multipath	—	0.04 dB

*) All noise temperatures are with reference to plane D in Fig. 6. All noise sources are calibrated against a National Bureau of Standards (NBS) certified noise standard.

The error assessment involves estimating:
a. the systematic errors;
b. the accidental errors.
Certain systematic errors may be eliminated by making the measurements in special ways or in particular combinations; others may be estimated by comparing the measurements with those obtained when a quite different method or different instrument is employed. In all cases the aim is to correct thus the readings as to ensure that all remaining errors are accidental. Sometimes the experiments are so arranged and the measurements are so randomized that any remaining systematic errors acquire the nature of accidental errors. The accidental error is assessed by applying certain statistical concepts and techniques.

12.2. *Received Signal Power Measurement*

There are several methods which may be used for measuring the absolute power level of a received CW signal but in SET-2 we employed two methods:
- CW signal substitution method;
- noise power substitution method.

The received signal is compared either with a CW test signal of known level or with the noise power of a calibrated noise lamp. All powers are measured at a reference point or referred to it. This point is chosen as the input of the ferrite switch plane D (Fig. 6).

The CW signal is generated either by the test-loop translator or by an external SHF generator locked to a synchronizer. The power level is measured and adjusted by means of a power meter followed by attenuators and couplers which have been calibrated with an overall accuracy better than 0.1 dB. The attenuators are remotely adjusted so that the test signal level is equal to that of the received signal at the receiver output. These signal levels are measured by a power meter at the IF output or by a true RMS voltmeter at the output of a narrow band receiver (BW: 200 Hz ... 5 kHz).

The advantage of this method is that the receiver section after the reference point is used only as an indicator for the equality of two levels. One difficulty with this method is the fairly high shielding necessary to prevent coupling of the test signal into the signal path by routes other than the calibrated path.

The RMS error (EP) in dB of the received signal power P is given by the square root of the sum of the squares of all individual error contributions in dB. With the error values for SET-2 given in section 5 we get EP = 0.21 dB as RMS error for the CW signal substitution method.

Turning now to the noise power substitution method we measure at the IF output the ratio, y_1, of the power received when the antenna is connected to the reference plane D, to that obtained with the noise lamp switched to the reference plane. It can be shown that the received power, P, is given by the expression:

$$P = \frac{kB}{g(f)}[\,y_1\,(T_{NL}+T_R)-(T_A+T_R)]$$

B = Noise bandwidth of receiver, $g(f)$ = IF filter transfer function: at reference Plane D
T_{NL} = Noise tube temperature, T_R = Receiver noise temperature: at reference Plane D
T_A = Antenna noise temperature
y_1 = Y-factor

In order to evaluate this expression, among other quantities, the antenna and receiver noise temperatures have to be measured. For T_A, a Dicke gain modulation radiometer is used for comparing the antenna noise temperature with the temperature of two highly stable (0.1 K) constant temperature loads, T_1 and T_2 which were calibrated against an NBS calibrated noise tube (accuracy 0.1 dB). T_A has been measured to be 45 K.

The RMS error EP of the measured power can be written as:

$$(EP_{dB})^2 = \left(\frac{10}{\ln 10}\right)^2 \left(\frac{kB}{g(f)\,P}\right)^2 \left[y_1^2\,(ET_{NL})^2 + (ET_A)^2 + (y_1-1)^2\,(ET_R)^2 + y_1^{\,2}\,(T_{NL}+T_R)\left(\frac{Ey_1}{y_1}\right)^2\right] + \left(\frac{10}{\ln 10}\right)^2\left(\frac{EB}{B}\right)^2 + (Eg_{dB})^2 + (\alpha_{dB})^2$$

where E denotes error in the quantity to which it is coupled. With the following parameter values – as given in Table 4 – the above expression shows that (EP) decreases with decreasing signal power and increasing bandwidth, as expected, and that for received signal power in the range −90 to −120 dBm, (EP) can be better than 0.3 dB.

12.3. *Antenna Gain*

The measurements of terminal ERP and satellite ERP, require the determination of the antenna gain and axial ratio for both transmit and receive frequencies. The calibrations of the SET-2 antenna for these quantities were done by using different methods of measurement and different instruments to eliminate systematic errors and, by repeating a measurement many times, to allow a statistical assessment of the quantity measured.

The error analysis conducted took into account errors such as those due to multipath, power variations, impedance mismatches, attenuator losses, and the measurement of range and frequency, and those due to pointing and polarization losses.

Knowing the axial ratios and the gains of the standard horns together with the values of all possible sources of error involved in the measurements it was possible to measure the gains of the antenna, for transmit and receive frequency bands using a combination of four essentially different methods, namely:
a) transmission method where a known amount of power is transmitted between the antenna and the horn located in the far-field and from the power received the gain is calculated and associated errors estimated;
b) substitution method where the antenna is compared with a standard horn using a signal source such as a satellite in the far-field;
c) comparison method where the gain of the antenna is compared with another antenna of known gain using a satellite;
d) stellar measurements, where the radio frequency emissions from Cassiopeia A and Taurus A are received by the antenna and the gain is calculated from the receive power measured. These stars have almost identical flux at X-band though their polarization and visibilities are different.

The mean values of several results obtained in measurements using these different methods, involving different signal sources, instruments, antenna positions etc. are given in Table 4.

12.4. *Link Availability*

A simple example will be given below to illustrate how measurement accuracy affects the link availability in a system which employs multiple terminals working in FM/FDMA mode with manual control. Let the system margin be 5 dB under nominal conditions. The following parameters and errors are relevant:

Parameter	*Error (dB)*
Path loss	0.3
Antenna gain	0.4
Transmitter power	0.3
Receiver noise	0.2
Receiver threshold	0.5
Satellite EIRP	0.5
Intermodulation noise	0.7

The total measurement error for the example above would therefore be about 1.2 dB. Assuming further that no control action would be initiated for changes less than, say, 1 dB, the net margin under clear-weather conditions would be 2.8 dB. As mentioned in Section 11, this margin would be exceeded by about 2% of the time due to weather and other external factors, thus giving a link availability of 98%. If now the measurement accuracy were to be improved by 0.2 dB the link availability would increase to about 99%, thus reducing the link outage by one half.

Appendix A

Bandwidth Utilization

1. *General*

In this Appendix values of E_B (the ratio of the bandwidth B_u, which can be utilized for the transmission of information, to the satellite bandwidth B_S) are calculated with the assumption that satellite ERP is not a limiting factor. Approximate results are derived for FDMA and SSMA, while the assumptions made are such that the results are applicable to message modulation methods such as PSK and FM with threshold extension.

2. *Bandwidth Utilization for FDMA*

With the assumption that the spectrum of each carrier passing through the satellite occupies the same bandwidth B_1, then

$$B_U = NB_1 \tag{1}$$

and

$$E_B = B_u/B_S = NB_1/B_S = N/M \tag{2}$$

where N is the number of carriers, and $M = B_S/B_1$ is the maximum number of carriers which could be relayed by an ideally linear repeater. If 3rd order intermodulation products generated by the hard limiting process do not interfere with any carrier, the maximum value for N can be obtained from Fig. 3 of Ref. [4]. For $N > 3$ the curve in that figure can be approximated by

$$N = (3.33M)^{0.435}$$
$$\therefore M = 0.3N^{2.3} \tag{3}$$

Inserting this expression into equation (2) the following expression for E_B is obtained:

$$E_B = \frac{3.33}{N^{1.3}} \tag{4}$$

3. *Bandwidth Utilization for SSMA*

In this case it is assumed that there are N carriers with a spectral density function which is of the form $\left(\frac{\sin x}{x}\right)^2$ with the first nulls of the function coinciding with the edges of the available bandwidth. In addition, each carrier is uncorrelated with any other and occupies the same message bandwidth B, then, the level C_{SS} of the wanted received carrier is given by

$$C_{SS} = \frac{\pi}{4} \times \frac{C_T}{N} \tag{5}$$

where C_T is the received level of the total satellite ERP. The remaining power $C_2 = (1-1/N)\, C_T$, which is a mixture of intermodulation products and other uncorrelated carriers, causes interference to the wanted signal. The level, N_u, of this interference is given by:

$$N_u = \left(1 - \frac{1}{N}\right) C_T \frac{4}{3} \frac{B_1}{B_S} \tag{6}$$

The signal to interference ratio R is then given by:

$$R = \frac{C_{SS}}{N_u} = \frac{1}{\left(1 - \frac{1}{N}\right)} \frac{B_S}{NB_1} \frac{3\pi}{16} \tag{7}$$

where $NB_1 = B_u$ is the bandwidth utilized for the transmission of all message modulated carriers. On substituting and solving equation (7) for B_U/B_S

$$\frac{B_U}{B_S} = \frac{3\pi}{16R} \frac{1}{\left(1 - \frac{1}{N}\right)} \tag{8}$$

If R_{min} is the minimum carrier-to-interference ratio required for proper reception:

$$\frac{B_U}{B_S} = E_B = \frac{3\pi}{16\, R_{min}} \cdot \frac{1}{1 - \frac{1}{N}} \quad \text{for } N > 1 \tag{9}$$

A value for R_{min} of 14 dB will permit the transmission of FM and PSK modulated signals with a sufficient link margin, provided that the satellite ERP is sufficient to dominate the thermal noise of the receiver.

Appendix B

Intermodulation in FM Caused by Phase Non-Linearities

Assuming amplitude non-linearities to be negligibly small [12] phase non-linearity in a practical bandpass filter would cause intermodulation to an FM signal as calculated below:

The output phase θ_o from the filter will be the initial phase plus phase shifts produced by the filter:

$$\theta_o = \theta + (\theta_F + k_1\theta + k_2\theta^2 + \ldots\ldots\ldots\ldots)$$

where the coefficients k_1, k_2 .... k_r depend on the type of filter used.

It can then be shown that the noise power ratio (NPR) in the top baseband channel is given by:

$$\text{NPR} = \frac{10}{8\pi^4 (\Delta F)^2 f_2^2 k_2^2 + 19.2\pi^6 (\Delta F)^4 f_2^2 k_3^2} \tag{1}$$

with baseband loading according to CCIR/CCITT recommendations.

The values of the coefficients k_2 and k_3 depend on the particular type of filter used. If the RF signal is centred in the pass-band of a symmetrical filter, $k_2 = 0$. Let us assume that the filter is a 3-pole Butterworth filter. Using the lowpass equivalent, the transfer function is:

$$G(j\omega) = \frac{1}{\left(1-\frac{2\omega^2}{\omega_3^2}\right)+j\left(\frac{2\omega}{\omega_3}-\frac{\omega^3}{\omega_3^3}\right)}$$

where ω_3 = angular frequency (rad/s) at the 3 dB point. The phase is given by:

$$\theta = \tan^{-1}\frac{\omega}{\omega^3}\left(2\omega_3^2-\omega^2\right)\left(1-\frac{2\omega}{\omega_3^2}\right)^{-1}$$

from which by series expansion we obtain

$$\theta(\omega^3) = \frac{\omega^3}{3\omega_3^3} \qquad (2)$$

and hence

$$k = \frac{1}{3\omega_3^3}$$

Since $\omega_3 = \frac{2\pi B}{2}$, where B = 3 dB bandwidth of the equivalent band-pass filter, we have

$$k_3 = \frac{1}{3\pi^3 B^3}$$

From equation (1) above the NPR becomes:

$$\text{NPR} = 4.68\ \frac{B^6}{(\Delta F)^4 f_2^2} \qquad (3)$$

For a 5-pole Butterworth filter, the third-order phase coefficient is given by:

$$k_3 = \frac{1}{2.42\,\pi^3 B^3}$$

i.e. not significantly different from that for the 3-pole filter.

For $B = B_{RF}$ equation (3) can be written as:

$$\text{NPR} = 4.68\ \frac{B_{RF}^{\ 6}}{(B_{RF/2}-f_2)^4 f_2^2}$$

where B_{RF} is Carson's bandwidth given by $B_{RF} = 2(f_2 + \Delta F)$.

For f_2 = 48 kHz and B_{RF} = 1 MHz, the expression above gives NPR = 46 dB which would be considered as of very good quality. A pre-detection bandwidth in an FM system as given by the expression of Carson:

$B_{RF} = 2(f_2 + \Delta F)$

could therefore be regarded as a good compromise between a low threshold which requires small B_{RF} and a low degree of intermodulation which requires large bandwidth.

It should be noted that when the third-order distortion is predominant the phase deviation from linearity is given by:

$$\theta(\omega_1^3) = \frac{\omega^3}{3\omega_3^3}$$

For $\frac{\omega}{\omega_3} = \frac{1}{2}$, departure from linearity is ± 0.04 radians; at the band edges this is ± 0.33 radians.

Appendix C

Local Oscillator Stability

The quantities of interest related to stability of the satellite local oscillator are:

a. Initial frequency offset;
b. Long-term drift;
c. Short-term stability (phase jitter).

A typical value for (a) and (b) is better than one part in 10^6 per year, and (c), stated as the maximum RMS fractional-frequency deviation $\left(\frac{\Delta f}{f}\right)$ may be as good as 1×10^{-10} for an averaging time of 1 second.

The long-term drift and frequency offset affect the search range of local oscillators and/or VCOs in PLL and AFC circuits. The frequency uncertainty due to the satellite local oscillator drift and offset may therefore exceed $\pm$ 8 kHz. Short-term stability, which is related to the purity of the oscillator output, determines the minimum noise bandwidth of PLL. Oscillators having a value of $\left(\frac{\Delta f}{f}\right)$ as good as the figure quoted above possess good spectral purity and can therefore be tracked with very small bandwidths (a few cycles). In a communication receiver which has a bandwidth of many kilocycles the phase jitter (residual FM) would appear at the baseband and could affect the performance of the voice channels.

In order to assess this degradation quantitatively, the power spectrum of the phase jitter (or of the instantaneous frequency) has to be known. Most oscillators are specified by a graph showing RMS-fractional frequency deviation $\left(\frac{\Delta f}{f}\right)$ versus averaging time (τ). Raven [13] gives a method to obtain the power spectrum of the instantaneous frequency S_ϕ from such a graph [13]. He assumes that S_ϕ is given by

$$S_\phi = \frac{\alpha}{|\omega|}+\beta+\gamma\,\frac{\omega^2}{\omega^2+\Omega^2} \qquad (1)$$

where the first term is due to flicker noise, the second term is due to internal noise of the oscillator, and the last term is additive external noise, which is band-limited by a single pole filter.

He then shows that $\left(\frac{\Delta f}{f}\right)$ is given by

$$(\Delta f)^2 = \frac{\alpha \log_e N}{\pi}+\frac{\beta}{\tau}+\frac{\gamma}{\tau\left(1+\frac{\Omega\tau}{2}\right)} \qquad (2)$$

where τ is the averaging time and N is given by the method of measurement; typically $N \approx 100$.

Fig. 15 shows a graph of $\left(\frac{\Delta f}{f}\right)$ versus τ as defined by equation (2).

The values of τ_1, τ_2, and τ_3, are approximately given by:

$$\tau_1 \approx \frac{2}{\Omega}, \ \tau_2 \approx \frac{2\gamma}{\Omega\beta}, \ \tau_3 \approx \frac{\pi\beta}{\alpha \log_e N}$$

The values of τ_1, τ_2, τ_3, and $\frac{\alpha \log_e N}{\pi} = K$ can easily be obtained from a plot of $\left(\frac{\Delta f}{f}\right)$ versus τ and are then given by:

$$\alpha = \frac{K\pi}{\log_e N}, \ \beta = K\tau_3, \ \gamma = K\frac{\tau_2\tau_3}{\tau_1}, \ \Omega = \frac{2}{\tau_1}$$

In many oscillators the internal noise is always smaller than either flicker noise or external noise, so that only two corners can be observed on the graph ($\tau_2 = \tau_3$). In this case only the third term of equation (1) will affect the performance of an FM system as flicker noise is limited to frequencies below the lower cut-off frequency of the baseband.

As an example, let us assume that an oscillator with $\left(\frac{\Delta f}{f}\right)$ given by Fig. 15 is used in the satellite. Then

$$\frac{K}{(2\pi f)^2} = (1.2 \times 10^{-11})^2$$

$$\tau_1 = 10^{-2} \text{ s}$$

$$\tau_2 = \tau_3 = 0.1 \text{ s}$$

For practical down-and-up conversion schemes, f may be taken to be approximately 800 MHz.

$$K = (1.2 \times 10^{-11} \times 2\pi \times 800 \times 10^6)^2 = 3.6 \times 10^{-3}$$

$$\alpha = \frac{10^{-2}}{10^{-2}} \times 3.6 \times 10^{-3} = 3.6 \times 10^{-3} \, (\text{rad/s})^2 \text{ per Hz}$$

$$\Omega = \frac{2}{\tau_1} = 200 \text{ rad/s}$$

The total noise power in a voice channel is given by:

$$(\Delta f)^2 = \frac{1}{4\pi^2} \int_{\omega_1}^{\omega_2} S_\phi \, d\omega$$

or as we are interested in the 3rd term of equation (1) only

$$(\Delta f)^2 = \frac{1}{4\pi^2} \int_{\omega_1}^{\omega_2} \frac{\gamma\omega^2}{\omega^2 + \Omega^2} \, d\omega =$$

$$= \frac{\gamma}{4\pi^2}\left[2\pi b - \Omega\left(\tan^{-1}\frac{\omega_2}{\Omega} - \tan^{-1}\frac{\omega_1}{\Omega}\right)\right] \qquad (3)$$

where $b = 3.1$ kHz is the bandwidth of the voice channel. Equation (3) gives in the lowest channel (0.3-3.4 kHz): $(\Delta f)^2 = 1.7$ $(\text{Hz})^2$ and in higher channels $(\Delta f) = 1.8$ $(\text{Hz})^2$ and assuming a test-tone channel deviation of 10 kHz the testtone-to-noise ratio would be about 77 dB.

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Korte technische berichten

Elektronisch schoolbord

Op 13 maart jl. werd door prof. Bordewijk en zijn medewerkers voor de pers een bijeenkomst belegd in de Afdeling der Elektrotechniek van de T.H. Delft, waarin de voortgang werd besproken bij de ontwikkeling van een elektronisch schoolbord. Aan dit project wordt te Delft sinds 1970 gewerkt.

Het elektronisch schoolbord is een systeem voor het overdragen van op papier geschreven teksten en tekeningen, welke aan de ontvangzijde op het scherm van een televisie-apparaat nagenoeg gelijktijdig weer zichtbaar kunnen worden gemaakt. De ontvangzijde kan via een kabel of een radioverbinding met de zendkant verbonden zijn. Het is echter ook mogelijk, de door de zendzijde afgegeven signalen çerst op registratieband vast te leggen om deze aan de ontvangzijde later op een gelegen moment aan de band te ontlenen.

Het project is een studie in het kader van ontwikkelingswerk, speciaal voor Indonesië. Op een centraal punt geeft een ervaren leerkracht zijn les, welke desnoods duizenden kilometers verderop kan worden gevolgd. De leerlingen horen de stem van de onderwijzer en zien op het elektronisch schoolbord tekeningen, grafieken en woorden verschijnen, die een onzichtbare hand daar voor hen optekent.

Helemaal nieuw is het elektronisch schoolbord niet. Het systeem, dat de werkgroep van de T.H. Delft introduceert, is echter wel aanmerkelijk goedkoper dan alle andere systemen die op dit gebied bekend zijn. De essentie van de vinding is, dat zij financieel mogelijk maakt, wat andere systemen die op de markt zijn niet vermogen. Het is een tussenoplossing voor het onderwijs per radio en dat per televisie. Radio heeft het nadeel dat het visuele aspect aan de les ontbreekt. Televisie is duur en levert zonder steunzenders of satellieten, die het systeem nog duurder zouden maken, een geografisch slechts gering verzorgingsgebied op.

De transmissiesignalen voor het beeld ('tekeningensignaal') nemen in dit systeem een geringe bandbreedte in beslag en kunnen samen met het begeleidende gesproken woord in een spraakkanaal worden ondergebracht.

Aan de ontvangzijde zijn alleen nodig een heel eenvoudige radio-ontvanger, een apparaat dat het tekeningensignaal omzet in videosignalen, en een televisie-ontvanger voor het opwekken van geluid en beeld. Het geheel behoeft niet meer te kosten dan een kleurentelevisie-ontvanger.

Aan de zendzijde is niet veel meer nodig dan een omroepzender, een schrijftableau en een schakeling voor het samenvoegen van de signalen voor het geluid en de beeldopbouw.

De leerkracht bereidt normaal zijn les voor en geeft die les met behulp van het elektronisch schrijftableau. Op dit tableau bevindt zich het vel papier, dat hij gebruikt voor het optekenen van de les. De tekenstift levert de beeldsignalen.

Om technische en didactische ervaring te verwerven, werden vanuit Delft twee rekenlessen gegeven door het hoofd van een basisschool te Zevenhuizen voor de leerlingen van de zesde klas van deze school. (Zevenhuizen is ongeveer 15 km van Delft gelegen.) Het experiment werd gevolgd door deskundigen van het Pedagogisch-didactisch Instituut voor de leraarsopleiding aan de Rijksuniversiteit te Utrecht en van de inspectie voor het voortgezet onderwijs.

Voor het onderwijs in Nederland lijkt het onderwijs per elektronisch schoolbord minder interessant dan voor het werk in ontwikkelingslanden, hoewel bijv. de Fryske Akademy belangstelling heeft getoond in verband met het onderwijs in de Friese taal.

Het systeem houdt verder mogelijkheden in voor de ontwikkeling van een 'tekeningentelefoon'. Een groep Delftse onderzoekers gaat deze mogelijkheid bestuderen.

*

T.H. Delft.

Raisting krijgt derde antenne

Medio 1972 zal in het grondstation Raisting (Opper-Beieren) de derde antenne-installatie voor het Intelsat communicatiesysteem operationeel worden.[1])

De eerste twee antennes zijn in gebruik voor respectievelijk de Intelsat-satellieten boven de Atlantische Oceaan en de Indische Oceaan. Daar voor transatlantische satelliet-communicatie nu twee satellieten in gebruik zijn, is in Raisting een derde antenne nodig, om verbinding te onderhouden met de tegenstations in het Atlantische gebied, die via de tweede Atlantische satelliet hun verkeer uitwisselen.

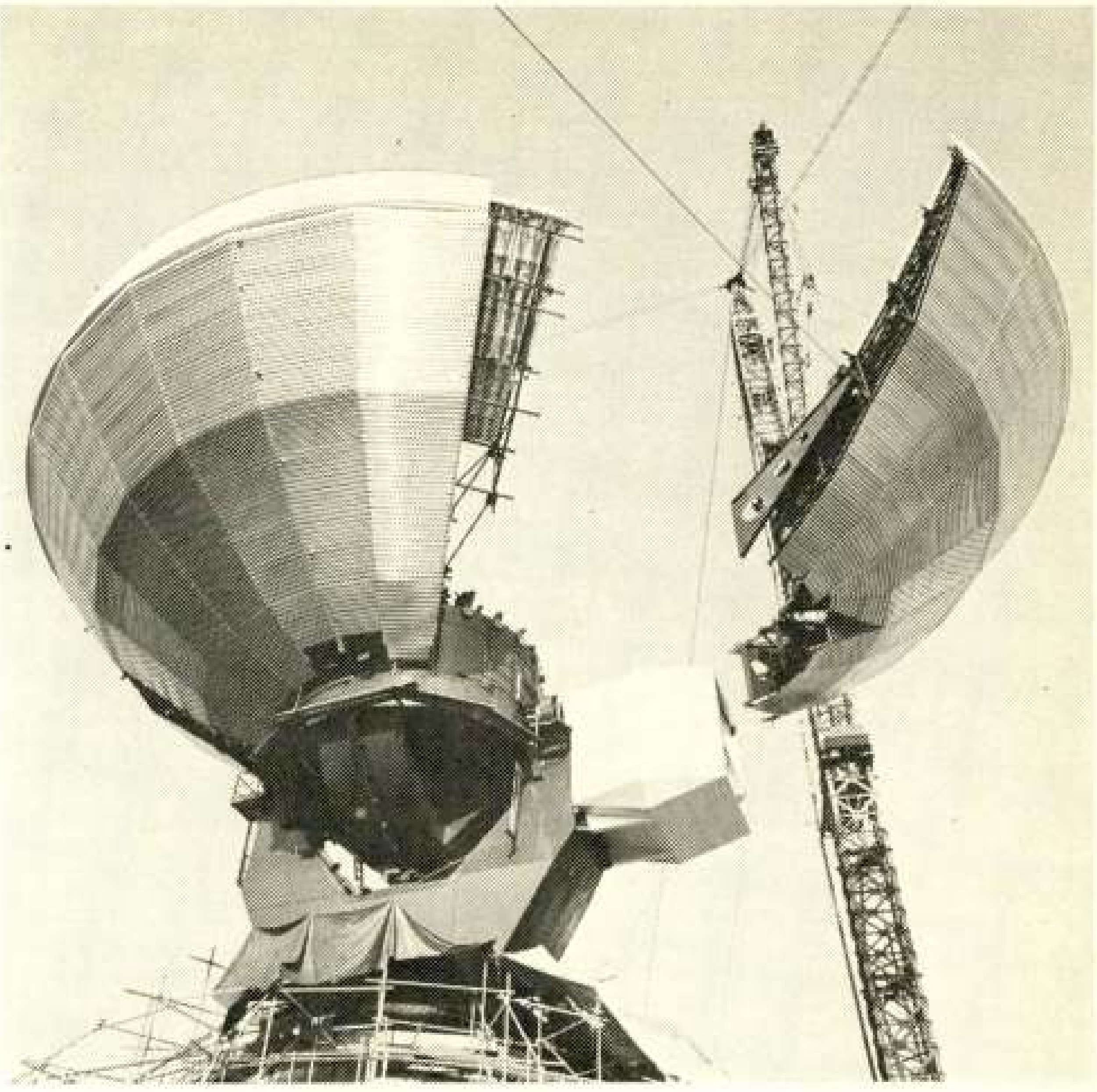

Fig. 1. Montage van de parabolische spiegel van Raisting III.

Evenals destijds bij de tweede antenne, heeft de Duitse Bundespost aan Siemens de opdracht gegeven, de installatie te bouwen. De antenne Raisting III krijgt evenals de Raisting II een parabolische spiegel met een diameter van 28,5 m. Ook deze antenne zal zonder radome worden opgesteld; achter de antennespiegel worden ongeveer 5000 infraroodstralers gemonteerd, die het afzetten van sneeuw en ijs op de antenne moeten voorkomen. De antenne is naar alle richtingen wendbaar, weegt 225 t en kan met een volgnauwkeurigheid van 0,01 graad op een der synchrone satellieten worden gericht.

De elektronische uitrusting van de nieuwe antenne is volgens de nieuwste inzichten ontworpen, waardoor deze ook dienst zal kunnen blijven verrichten wanneer omstreeks 1980 de Intelsat V satellieten in een baan om de aarde komen. Deze zullen een capaciteit van 50 000 ... 100 000 telefoonkanalen bezitten.

De zender van Raisting III kan een aantal draaggolven in het 6 GHz-gebied gelijktijdig uitzenden. De eindversterker is een lopende-golfbuis van groot vermogen. De ontvanger heeft een breedbandige ruisarme ingang voor 4 GHz, uitgerust met een parametrische versterker, die met vloeibare helium wordt gekoeld. De versterkers die aan ontvang- en zendzijde met de antenne zijn verbonden, draaien in azimuth met de antenne mee; in elevatie loopt de signaalverbinding met de antenne via een draaikoppeling.

De achtereenvolgende series van Intelsat-satellieten schonken bij elke nieuwe generatie aanzienlijk meer mogelijkheden. Intelsat I, in 1965 gelanceerd, werkte met een bandbreedte van 25 MHz bij een transponderuitgangsvermogen van 12 W. Per richting konden 240 telefoongesprekken of één televisieprogramma worden uitgezonden. Hierna volgde in 1967 Intelsat II met een bandbreedte van 126 MHz en een uitgangsvermogen van 35 W. De transponder liet gebruik volgens het principe van meervoudige toegankelijkheid toe. In 1968 volgde de eerste satelliet

[1]) Zie ook 'De Ingenieur' 1968, 22 maart, blz. ET 46-47.

van de serie Intelsat III, waarmee een begin werd gemaakt met een wereldomvattend satelliet-communicatienetwerk. Deze satellieten zijn uitgerust met twee transponders, met een bandbreedte van 225 MHz. Het effectief zendvermogen bedraagt 150 W. Gelijktijdig kunnen 1200 telefoongesprekken worden overgedragen; ook kan vanuit twee richtingen over de satelliet een televisieprogramma worden doorgegeven. Van de generatie Intelsat IV staat de eerste satelliet vanaf januari 1971 boven de Atlantische Oceaan. De capaciteit van deze bedraagt 6000 telefoonkanalen en verscheidene televisiekanalen. De satelliet is voorzien van 12 transponders, elk met een bandbreedte van 36 MHz, waardoor intermodulatie-effecten bij het meervoudig gelijktijdig gebruik van de transponders tot een minimum beperkt blijven.

Bij de Olympische Zomerspelen 1972 in München zal Raisting III al een rol spelen. Men voorziet, dat antenne I via een Intelsat III satelliet de verbinding met Azië en Australië zowel voor telefoon als voor televisie zal verzorgen. Antenne Raisting II zal via een Intelsat IV satelliet met het Amerikaanse continent en met Iran in contact staan, terwijl ook televisieprogramma's voor Amerika en Afrika hierover doorgezonden zullen worden. Raisting III zal via een satelliet van het type Intelsat III minstens nog een televisieprogramma over de Atlantische verbinding doorgeven.

*

Persbericht Siemens.

Centraal Laboratorium van PTT bestaat 25 jaar

Naar aanleiding van het 25-jarig bestaan van het Centraal Laboratorium van de PTT werd eind april 1972 in één der vleugels van het Dr. Neher-Laboratorium te Leidschendam een tentoonstelling geopend, waarin historie en toekomst van de techniek op het gebied van de telefonie, telegrafie, postverwerking en automatisering bij de PTT is uitgebeeld.

Het Centraal Laboratorium van de PTT werd in het leven geroepen bij de beschikking van 19 juli 1946 van de toenmalige directeur-generaal L. Neher, als samenbundeling van een aantal bedrijfslaboratoria van het Staatsbedrijf.

Op 17 mei 1955 werd de nieuwe behuizing te Leidschendam in gebruik genomen, welke ter ere van de stichter de naam van Dr. Neher-Laboratorium ontving. Het complex te Leidschendam heeft een nuttige vloeroppervlakte van bijna 9000 m^2. De investering in terrein en opstallen bedroeg destijds bijna 9 miljoen gulden.

Jaarlijks wordt ca. $\frac{3}{4}\%$ van de omzet van PTT aan speur- en ontwikkelingswerk besteed; de investeringen van het Dr. Neher-Laboratorium bewegen zich rond 0,5% van de bedrijfsinvesteringen.

*

Het PTT-Bedrijf.

Varia

'Satellite Systems for Mobile Communications and Surveillance'

An international conference on Satellite Systems for Mobile Communications and Surveillance is to be held at the IEE *from 13 ... 15 March 1973* and organized by the Electronics Division of the Institution of Electrical Engineers, in association with the Institute of Navigation, the Institution of Electronic & Radio Engineers and the Royal Aeronautical Society. It is intended that the conference will cover the following topics in both the civil and military fields:

1. General communications and surveillance system environments:
- operational requirements and constraints;
- traffic predictions;
- other systems applications, e.g. search and rescue;
- emergency communications.

2. Technical factors and constraints:
- choice of frequency bands;
- performance;
- interference and noise;
- modulation methods;
- signal processing;

3. Space segment and satellite control:
- basic system configurations;
- system redundancy and reliability;
- system calibration and control;
- satellite design.

4. Terminals for mobile services:
- for ships;
- for aircraft;
- for land vehicles;
- for associated base stations.

5. Overall system management and economics.

The organizing committee invites offers of contributions not exceeding 2000 words, for consideration for inclusion in the conference programme. Those intending to offer a contribution should *submit a 250 word synopsis* to the Conference Department by *17 April 1972.*

Further details may be obtained from the Manager, IEE Conference Department, Savoy Place, London WC2R OBL.

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