10

Earth Stations

10.0 INTRODUCTION

We call the collection of equipment on the surface of the earth for communicating with the satellite an *earth station,* regardless of whether it is a fixed, ground mobile, maritime, or aeronautical terminal. We recognize that, with our broad concept of communications satellites, earth stations can be used in the general case to transmit to and receive from the satellite, but in special applications only to receive or only to transmit. Receive-only stations are of interest for broadcast transmissions from a satellite and transmit-only stations for the still much less developed application of data gathering. Figure 10-1 is a general block diagram of an earth station capable of transmission, reception, and antenna tracking. We identify the following major subsystems:

Transmitter: There may be one or many transmit chains, depending on the number of separate carrier frequencies and satellites with which the station must operate simultaneously.

Receiver: Again, there may be one or many receiver/down-converter chains, depending on the number of separate frequencies and satellites to be received and various operating considerations.

Antenna: Usually one antenna serves for both transmission and reception, but not necessarily. Within the antenna subsystem are the antenna proper, typically a reflector and feed; separate feed systems to permit automatic tracking; and a duplex and multiplex arrangement to permit the simultaneous connection of many transmit and receive chains to the same antenna.

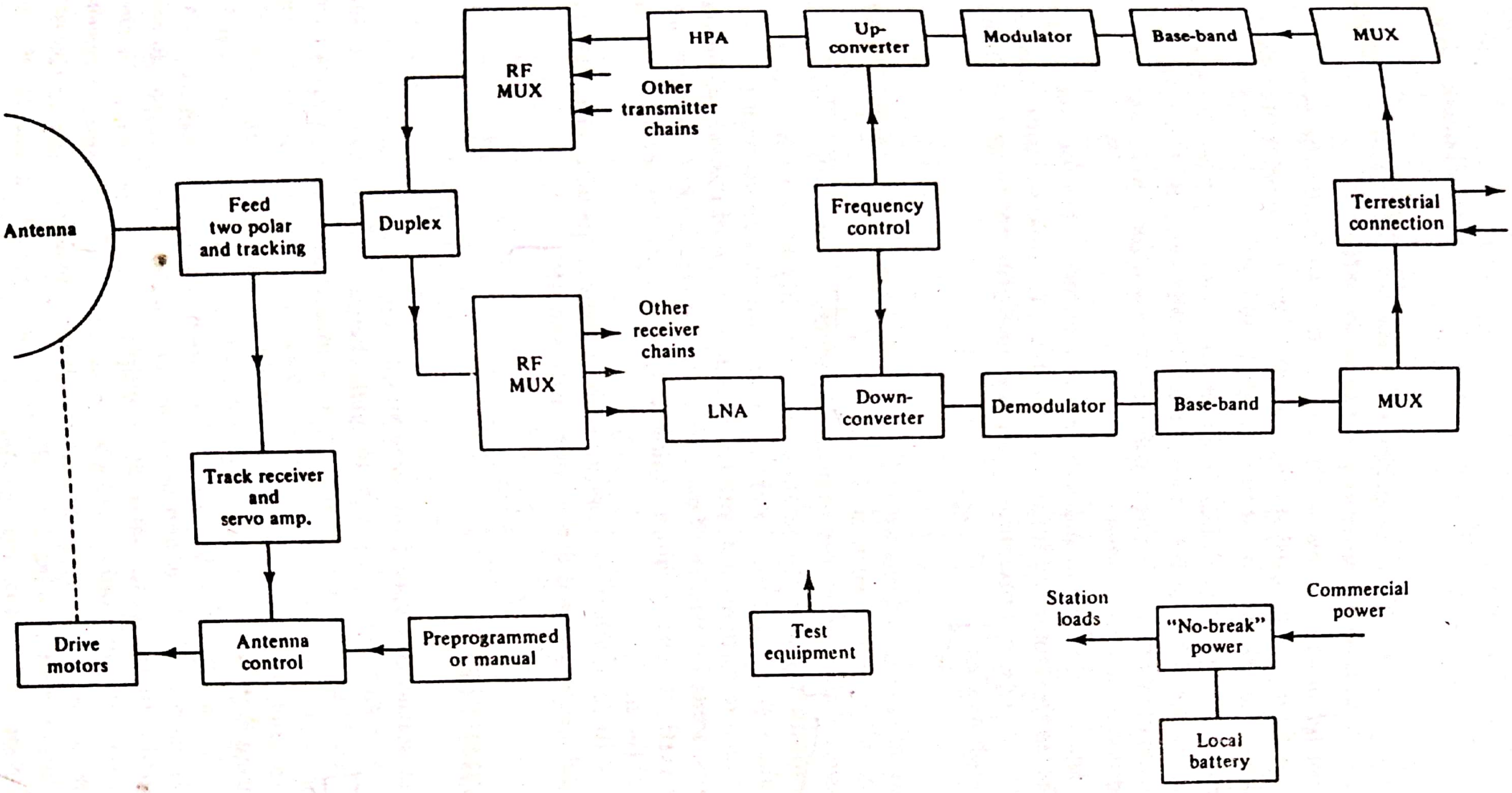


Figure 10-1 General earth station.

Tracking system: This comprises whatever control circuit and drives are necessary to keep the antenna pointed at the satellite.

Terrestrial interface: This is the interconnection with whatever terrestrial system, if any, is involved. In the case of small receive-only or transmit-only stations, the user may be at the earth station itself.

Primary power: This system includes the primary power for running the earth station, whether it be commercial, locally generated, battery supplied, or some combination. It often includes provision for "no break" changeover from one source to another.

Test equipment: This includes the equipment necessary for routine checking of the earth station and terrestrial interface, possible monitoring of satellite characteristics, and occasionally for the measurement of special characteristics such as G/T.

In the following sections we will deal, to varying extents, with each of these subsystems, viewing the earth station from the point of view of the complete system designer. We are interested in those aspects of the earth station that affect its communication link to the satellite and its ability to resist interference from other satellites and terrestrial systems. We are not concerned with the design of earth stations and certainly not with the detailed design of individual subsystems. Nonetheless, some knowledge of each subsystem is required in order intelligently to specify earth stations within a complete system and to ascertain what can be expected of them.

Tables 10-1a and b show ranges for the most conspicuous characteristics of the principal classes of fixed and mobile earth stations.

10.1 TRANSMITTERS

Transmitter subsystems vary from very simple single transmitters of just a few watts for data-gathering purposes to multichannel transmitters using 10-kW amplifiers, such as those found in Intelsat standard A stations. When multiple transmitter chains are required, common wideband traveling-wave tube amplifiers can be used, such as the arrangement shown in Figure 10-2, or each channel can use a separate high-power amplifier, typically a klystron, as shown in Figure 10-3.

Two-for-one redundancy switching is shown, by way of example, with the TWTAs. Numerous methods and levels of redundancy (for example, three-for-two or four-for-three) exist. Similarly, multiplexer and filter arrangements are also multitudinous and only one scheme is shown. The common wideband amplifier is the more usual type, despite its suffering from the familiar problem of intermodulation when nonlinear amplifiers handle more than one carrier simultaneously. Note that a transmitter carrier-to-intermodulation ratio that is not very high must be considered in calculating the overall $(C/N)_T$. It is added to the

Direct Broadcasting satellite.
Very Small Aperture Terminal

TABLE 10-1a TYPICAL EARTH STATIONS FOR FIXED AND BROADCAST SERVICE

	International	Domestic Trunk	Video Distribution	DBS	VSAT
Frequency bands	C, Ku	C, Ku	C, Ku	Ku, Ka	C, Ku
Antenna size (m)	5–20	5–12	5–10	0.5–1.5	1–2
System temperature (K)	35–60	60–200	100–300	200–600	100–300
Transmitter power (W)	1000–10 000	100–5000	N/A	N/A	0.1–10
Multiple access	FDMA, TDMA	FDMA, TDMA	FDMA	FDMA	FDMA, TDMA, CDMA

TABLE 10-1b TYPICAL EARTH STATIONS FOR MOBILE SERVICE

	Marine Mobile	Aeronautical Mobile	Land Mobile	Handheld
Frequency bands	L	L	VHF, L, S	VHF
Antenna size (m)	0.85–2.0	0–1	0–0.5	Omni
System temperature (K)	150	100	300	600
Transmitter power (W)	1-200	5–20	1–10	0–5
Multiple access	FDMA, TDMA	FDMA, TDMA	FDMA, TDMA CDMA	FDMA, CDMA

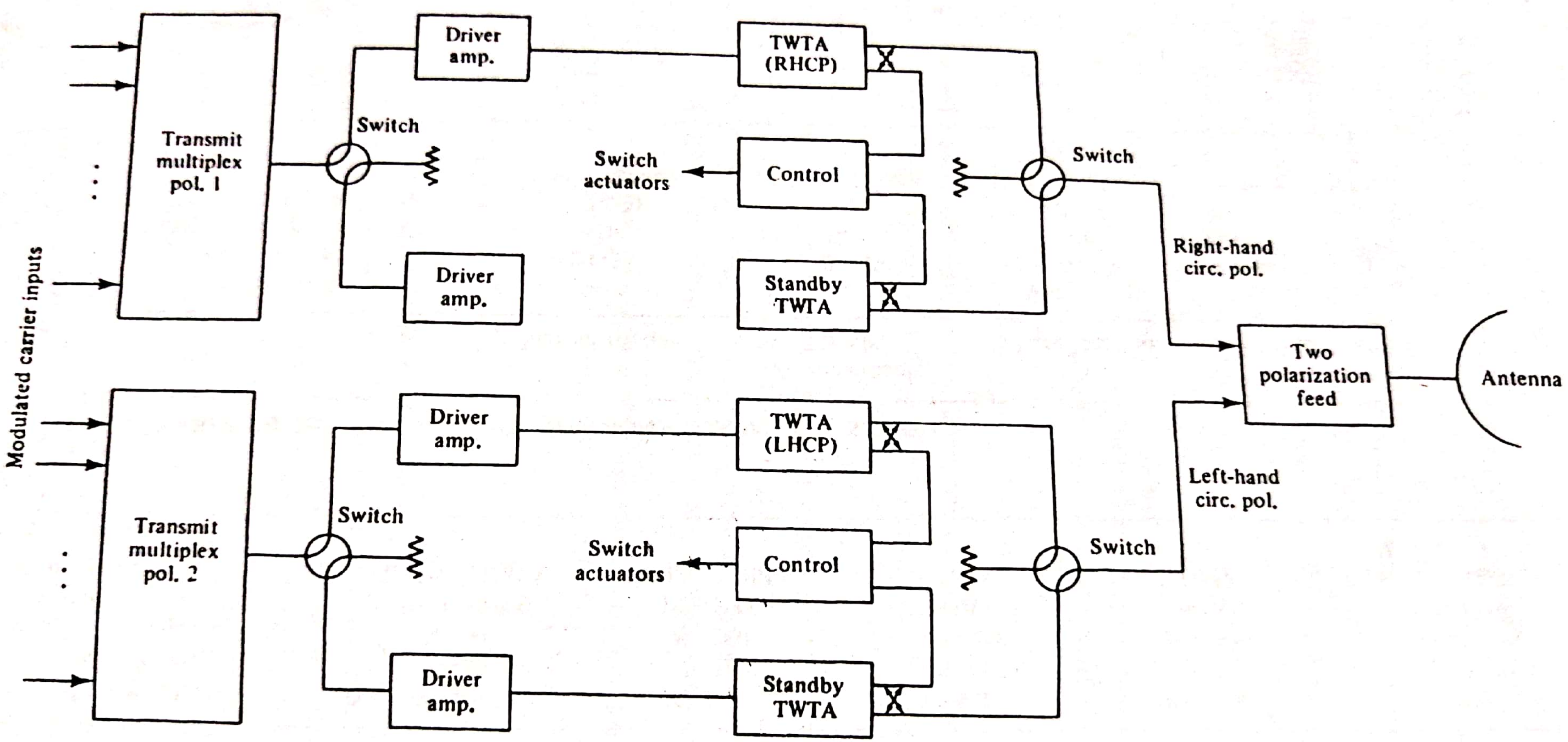


Figure 10-2 Common TWTA transmitted with redundancy (two polarizations).

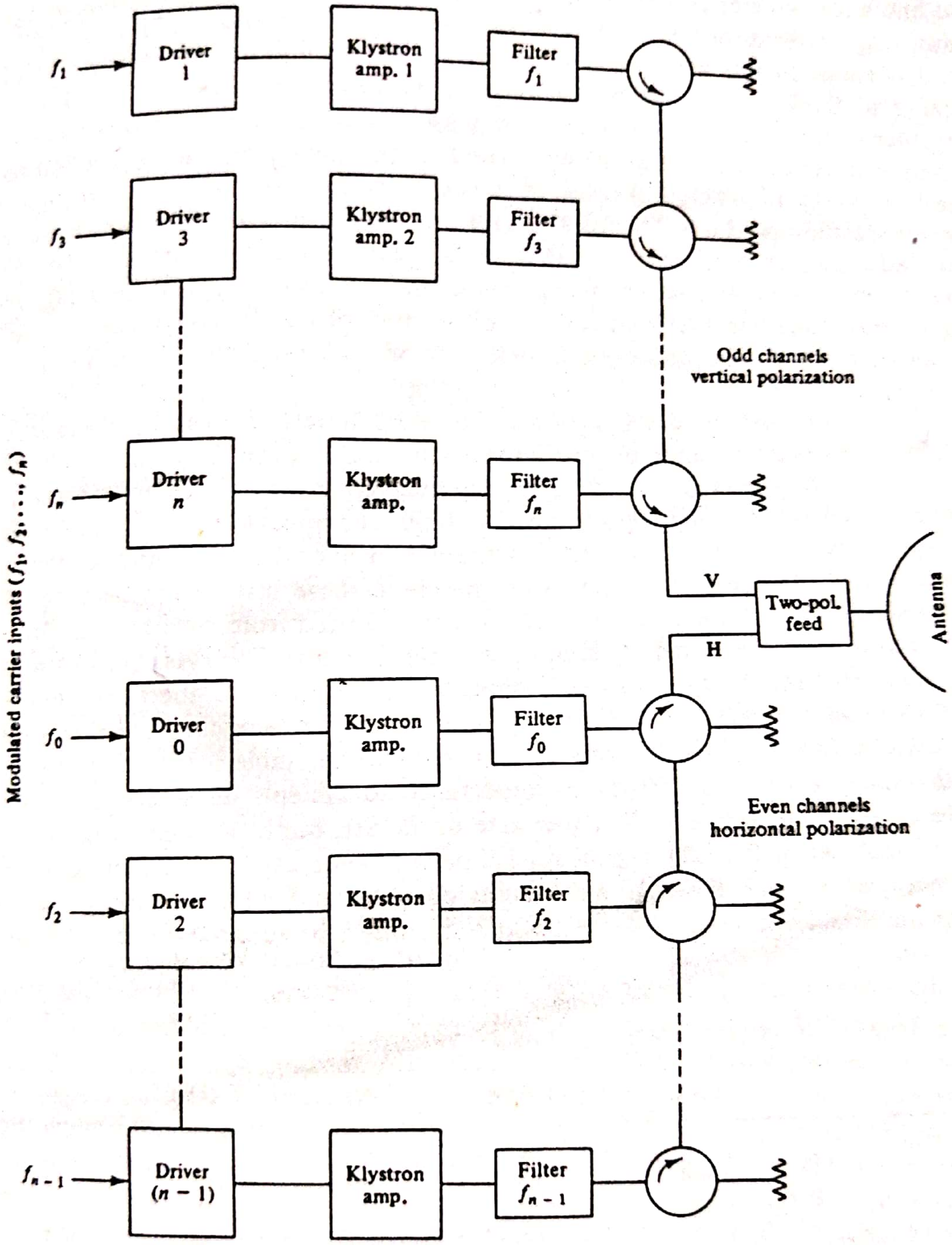


Figure 10-3 Multiple klystron transmitter.

result using the formula derived in Chapter 6; that is, the intermodulation noise is assumed to be additive to the channel thermal noise. This problem is commonly dealt with by the use of back-off, but with considerable reduction in output power. Systems using feedback to reduce the nonlinearity effect are coming into use and allow greater power output. Another increasingly popular technique for improving carrier-to-intermodulation performance in high-power amplifiers uses predistortion. In this method, a low-level nonlinear amplifier of characteristics similar to those of the high-power amplifier is introduced ahead of the HPA, together with appropriate amplitude and phase equalizing, and a reference linear signal in order to generate an inverse, predistorted signal that, when applied to the high-power amplifier, will result in substantially decreased overall third-order intermodulation products. Miya (1985) gives results for three carriers, which show that reductions of as much as 10 dB in third-order intermodulation products for three carriers can be achieved using this method. Because typical back-offs in high-performance earth stations can be on the order of 7 to 10 dB, this technique, by allowing substantial reductions in that back-off, can yield good improvements in efficiency.

The alternative of using separate amplifiers is less flexible in operation. Usually, the separate amplifiers are narrowband and require retuning to change frequencies. The problems of multiplexing many chains on an antenna without interaction among the amplifiers become still more complicated. The simple ways of effecting such combinations, such as the use of hybrids, also produce power losses, typically of the same order of magnitude as those involved in backing off wideband amplifiers. Nonetheless, such systems are used from time to time since klystron amplifiers are currently simpler and cheaper than TWTAs. We can also argue that the reliability of such systems is higher because there are fewer single-point modes of failure. A few typical high-power amplifier specifications are shown in Table 10-2. It is important to note that the table is intended to give typical values of the parameters of importance to systems designers. It is not complete, nor even a summary of the state of the art, but it does give ranges for those critical values that determine system performance and it does highlight the differences among the types. Final system planning and hardware design must rely on manufacturer's rating sheets, which are much more detailed and contain

TABLE 10-2 HIGH POWER AMPLIFIER CHARACTERISTICS

	TWTA	TWTA	TWTA	SSPA	SSPA
Freq. Band	c	K_u	K_a	C	K_u
Power (W)	600	300	100	25	16
Efficiency (%)	25	22	18	15	5
Bandwidth (MHz)	500+	500+	2500	500	500
Gain (dB)	50	70	50	50	50
Noise Fig. (dB)	25	28	35	6	12
Third Order Intercept (dBm)	10	10	10	20	20
AM–PM	2°/dB	2°/dB	2°dB	0.5°/dB	0.5°/dB
Mean Time to Fail. (MTTF) hrs	15–30 000	15–30 000	15–30 000	150 000	150 000

information on voltages, impedances, weights and power consumption, and other characteristics necessary to design the interfaces.

10.2 RECEIVERS

To receive a signal from a satellite, several distinct operations must be performed. The signal must first be amplified, then reduced to a frequency low enough for convenient further amplification and demodulation, then demodulated and delivered to whatever baseband processing equipment is needed. The signal may be used either at the earth terminal itself, say in the case of a home TV receive-only (TVRO) terminal, or converted into a form suitable for transmission elsewhere. When we speak of the receiver chain, we refer ~~here~~ have specifically to the low-noise amplifiers, down-converters, and demodulators. Down conversion can be accomplished either in one step, going directly from the satellite downlink carrier frequency to the intermediate demodulator frequency (characteristically 70 MHz), or it can be accomplished in several steps. Two-stage down conversion is often done when the same receiver is to be tuned to a multiplicity of channels. Figure 10-4 is a prototypical receiver chain for a general case. LNA redundancy is shown to illustrate the switching, but other redundancies, such as for the down-converter, are not indicated, although they are common. Again, as in the case of transmitters, the variety of possibilites for switching and multiplexing is considerable. Figure 10-5 is a general block diagram for a video and audio receive only station. Such stations are widely used in satellite communications. Such receivers are used in cable heads to receive TV programs from satellites. They are then redistributed, usually by cable but sometimes by microwave or wireless cable systems. The same kind of receiver, perhaps simplified, is used where the satellite transmission is intended directly for home or other end-user use. The first down conversion is usually done for the entire band in question and is referred to as block down conversion. There is increasing interest, and indeed several planned systems, for the direct transmission of audio signals from satellites to end users. One application is in use in which a wideband transponder carries 24 digital carriers in FDMA. Each carrier carries something like 1.0 Mb/s in QPSK, enough for high-fidelity stereo transmission. The receivers are still of the type shown. Direct audio broadcast has been proposed for several regions in the world, using digital and FM transmissions of quality ranging from barely recognizable to almost that of compact disks.

The low-noise amplifier is one of the critical elements in determining the earth station performance as a system element. This performance is characterized by the familiar figure of merit, G/T, as shown in Section 6.2. It is determined by the antenna gain, discussed in Section 10.3, and the system temperature, the expression for which was developed in Chapter 6 and is repeated here for convenience.

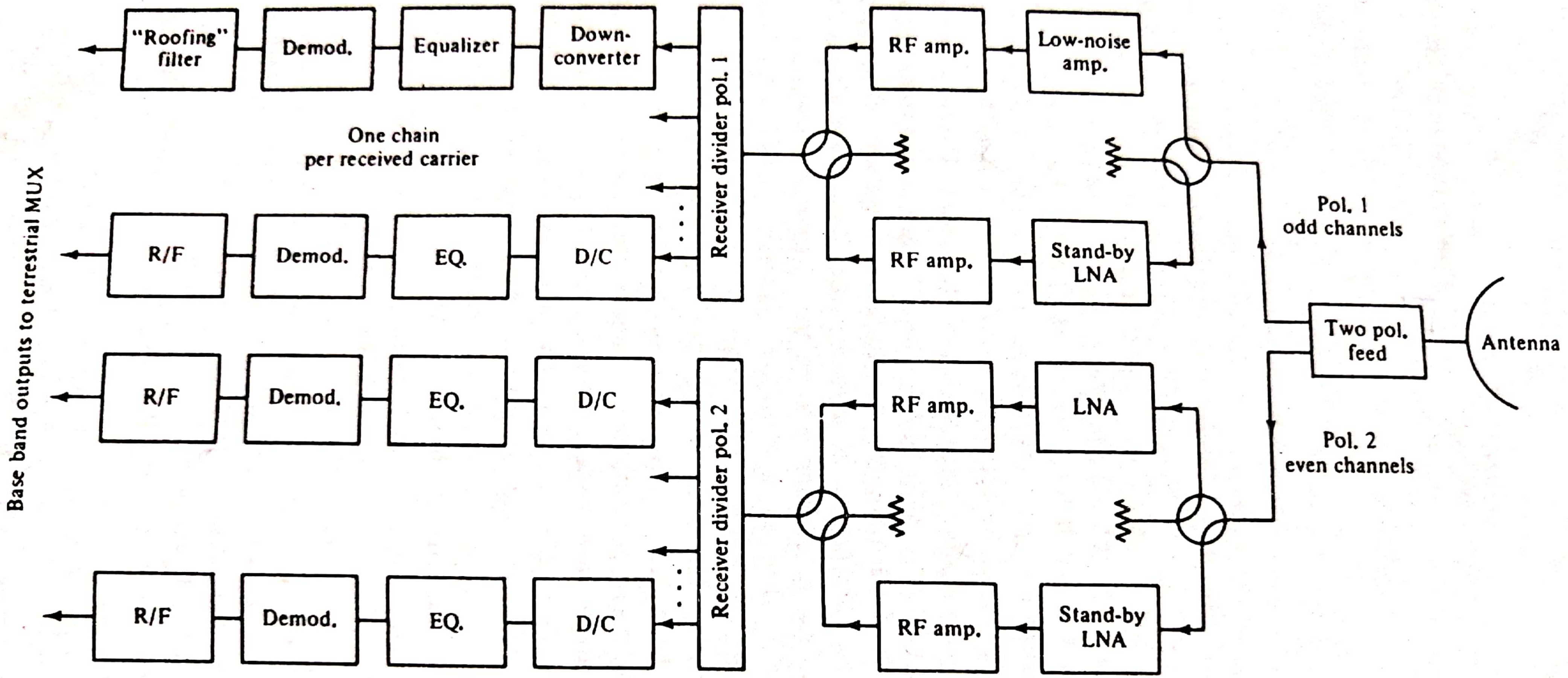


Figure 10-4 Receive subsystem for multicarrier earth station.

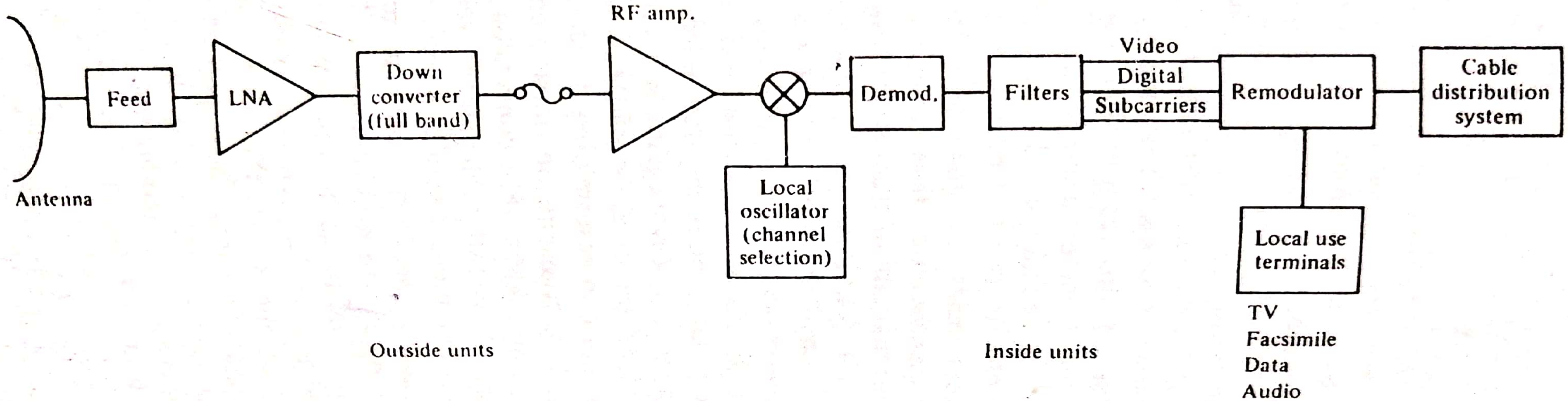


Figure 10-5 General TVRO station-direct reception or cable distribution.

$$T_s = T_a + (L-1)T_0 + LT_R + \frac{L(F-1)}{G_R} T_0 \qquad (6\text{-}54)$$

where all terms are expressed as ratios or in Kelvins.

The second term in this equation is the clue to the reason for many decisions in earth station design. Even if the antenna temperature is very low, as is normally the case in clear weather, and the excess temperature of the reciever T_R is also low, perhaps 50 K, the system temperature can be surprisingly high if there is even a small loss in the transmission line between the antenna and low-noise amplifier. The term $(L-1)T_0$ for a loss of 0.5 dB is 35 K, as high as the receiver itself in low-noise cryogenic systems. This makes it a necessity in high-performance earth stations for this transmission line to be made as short as practical. It explains the almost universal use of Cassegrainian antennas in large, high-performance systems, especially at C-band and below, where antenna temperatures can be expected to run low even in the presence of rain. On the other hand, at higher frequencies, designed of necessity to have a high rain margin, the antenna temperature under those rainy circumstances will be several hundred degrees; a certain amount of waveguide loss is now tolerable since it produces less proportionate deterioration. Additionally, at these higher frequencies, the antennas are smaller since higher levels of transmitter power are normally used in the satellites. Physically smaller antennas are also desirable at higher frequencies to keep the beamwidths from becoming too small. These small antennas make it practical to use prime-focus-fed reflectors (see Section 10.3) and still have adequately short lengths of waveguide to receiver locations.

In addition to the excess temperature of the receiver, a number of other characteristics are of importance in determining the station performance, notably those that affect the degree of transmission impairments, such as group delay, gain stability, gain flatness, and AM-to-PM conversion factors. Their effects are handled by the assessment of overall performance very much as discussed in Chapter 9. A set of typical low-noise amplifier characteristics is shown in Table 10-3. By way of clarification, the term *intercept point* in connection with intermodulation is worth a note. Low-noise amplifiers, like all amplifiers, saturate and thus have the standard problems of intermodulation. It has become common to specify this intermodulation by the intercept point, the point at which the extended linear portion of the curve third-order intermodulation products extended would intercept the extended linear portion of the amplifier's power transfer characteristic, as seen in Figure 10-6. It is not difficult to derive an expression for carrier-to-third-order modulation products, the more useful measure in computation, from this diagram. The result is

$$\left(\frac{C}{I}\right)_3 = 2(P_x - P_0) \qquad (10\text{-}1)$$

P_0 is the saturated output power and P_x is the intercept point, both usually taken

TABLE 10-3 LOW-NOISE AMPLIFIER CHARACTERISTICS

	L-band	C-band	X-band	K_u-band	K_a-band
Cooling	Uncooled	Uncooled/ cooled	Uncooled/ cooled	Uncooled/ Cooled	Uncooled/ cooled
Frequency range (GHz)	1.5–2.5	3.0–5.0	7.0–10.0	10–14	11–20
Bandwidth (MHz)	50–100	500	500–1000	1000	1000
Noise temperature (K)	40–60	35–60	55–75	65–130	200–300
Gain (dB)	45–60	50–60	50–55	50–60	20–25
Output at 1.0-dB compression (dBm)	13	13	13	13	10
Intercept dB above output	10–13	10–13	10–13	10	10
AM-PM (°/dB)	0.03–0.5	0.03–0.50	0.03–0.50	0.03–0.50	0.03–0.50

in dBm. This result depends on the fair assumption that the total of the third-order modulation products varies with the cube of the input power.

Table 10-3 shows typical values and ranges for the LNA parameters of principal interest to system designers. Several generalizations are useful. Most low-noise amplifiers today (1992) use gallium arsenide field-effect transistors, GaAsFETS or HEMTs. They are usually uncooled or, if very high performance is being sought, they are thermoelectrically cooled. Such cooling typically reduces the receiver temperature about 10 K. Cryogenic receivers using liquid helium and nitrogen, so common in the early days of satellite communication, are largely

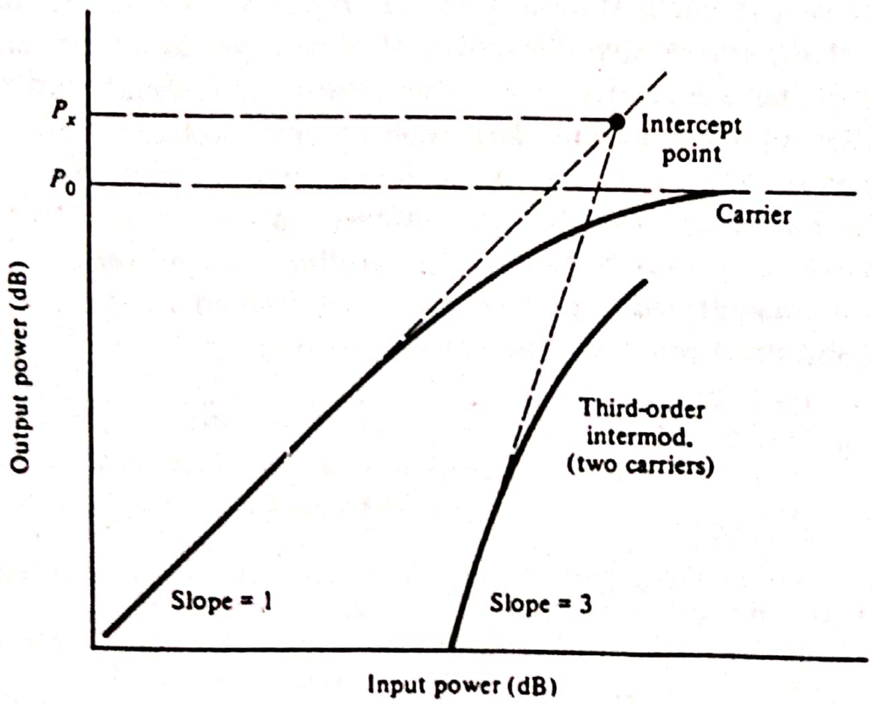


Figure 10-6 Intercept point as a measure of third-order intermodulation level.

disappearing from new designs except in unusual applications like deep-space reception. One reason is their high initial and maintenance costs compared to the marginal increases in system performance. Another is the reality that rain losses, one of the main reasons for needing better receiver performance, also increase the antenna temperatures to a level where the receiver contribution is not so important. The output levels of these amplifiers are given conventionally at the 1.0-dB compression points, the points on the input–output characteristic where the output is 1.0 dB less than it would be if the output continued to increase linearly with the input. This is an easier point to measure precisely than the vaguer saturation level. Intercept point and AM–PM conversion are two simple measures of nonlinearity that help in comparing different amplifiers. For detailed receiver design, it is necessary to have other measures, as discussed in Chapter 9, such as the variations in group delay and amplification. For these and other detailed data, it is necessary to consult the manufacturers' specifications.

10.3 ANTENNAS

10.3.1 General

The parabolic reflector antenna has become the symbol for a satellite communication earth terminal. The synecdoche "dish" for an entire earth terminal has become universal in popular usage. The symbolism is reasonable, not only because the antenna so distinguishes the terminal physically, but also because its characteristics are by far the most important of all in determining the overall earth station performance, both on the uplink and downlink. As we have seen in Chapter 6, the carrier-to-noise ratios achievable on these links, given fixed transmitter powers and geographical coverages, are directly determined by the physical size of the earth station antenna. By now, no one should need further convincing that, for a given terrestrial coverage and desired performance, antennas at K-band must be larger than those at C-band, not smaller. This is because of the rain attenuation. The antenna sizes would otherwise be independent of frequency as long as the noise levels are not determined principally by interference. The reason that K-band antennas are generally found to be smaller is simply that more power is used in the satellite transmitters.

The antenna electrical performance is involved in the system planning in many ways; the most important are the following:

Characteristic	Affects
Overall gain, G	System G/T_s
Antenna temperature, T_a	G/T_s
Sidelobe level (including spillover)	Interference (C/I), antenna temperature
Cross-polarized response	C/I and C/N for entire system
Beam width	Geographical coverage (satellite antennas), tracking requirement

The next section will consider some of the important ideas in antenna design. The relation of these characteristics to antenna geometry is the province of the antenna designer, a major engineering speciality in itself.

The electrical characteristics are by no means independent of each other, and their optimization for a particular system is a joint effort of the antenna designer and systems engineer. For system planning, a generalized antenna pattern is often useful. A good pair of equations for such use is

$$\text{On main lobe:} \quad \frac{G}{G_m} = \left[\frac{\sin 1.39(\theta/\theta_0)}{1.39(\theta_0/\theta_0)}\right]^2 \tag{10-2}$$

$$\text{Far from main lobe:} \quad \frac{G}{G_m} = \frac{1}{1+(\theta/\theta_0)^{2.5}} \tag{10-3}$$

Note that θ_0 is half the half-power beamwidth. The beamwidths of the antenna are also related to its gain by virtue of the beam geometry, quite independently of the particular antenna realization. Gain is defined as the ratio of radiation intensity in a given direction to that it would have were the total radiated power to be radiated isotropically. It comprises two elements: the *directivity*, the component of the ratio determined by the geometry of the antenna system, and the effect of *losses* due to such factors as dissipation and spillover. The directivity part is the more important and thus, as a surprisingly good working relation, we can use

$$G \simeq \frac{4\pi}{\theta_1\theta_2} \simeq K\frac{41\,253}{\theta_1^0\theta_2^0} \tag{10-4}$$

where K is a factor to allow for energy not in the main beam (it is about 0.65). θ_1 and θ_2 are the antenna beamwidths in radians or degrees, as appropriate. The Equation follows from the assumption that the radiated power is confined principally to the main lobe, instead of being radiated over 4π steradians isotropically, and is modified approximately by a factor to allow for total energy in all the side lobes.

Although the parabolic reflector is by far the most important kind of antenna that we find both in earth stations and on the satellite, nonetheless, other types are important, particularly *horns* and *arrays*. Horns are used widely as primary feeds for reflectors and occasionally as principal radiators themselves. Two other kinds are occasionally seen in spacecraft. They are *lenses* (either the dielectric or waveguide types) and *phased arrays*. The latter is not a different type of radiator per se, but simply a controlled combination of any kind of individual element. For instance, horn feeds, dipoles, and even parabolic reflectors can be used in arrays with the composite pattern determined by conventional antenna array theory. The array is controlled by varying the phase and amplitude of the excitation to the individual elements. We expect that this kind of antenna will become increasingly important in spacecraft design as the carrier frequencies get lower. This will be the case for satellite service to small mobile terminals because of the requirement for frequency reuse to provide many channels and the concomitant narrow beam-widths. Arrays will be the easiest way to achieve the large apertures. With small terminals for the direct reception of TV, there is much interest in phased arrays so as to have a flat, easily manufacturable, and cosmetically attractive antenna for home use. Such antennas have been developed from arrays of printed-circuit dipoles and slots protected by plastic dielectric sheets. It is important to understand that their possible advantages are in cost, appearance, and packing convenience. They have no particular advantage over the parabolic reflector in the relationships among beamwidth, sidelobe level, and aperture. With small terminals for the direct reception of TV, there is much interest in phased arrays so as to have a flat, easily manufacturable, and cosmetically attractive antenna for home use. Such antennas have been developed from arrays of printed-circuit dipoles and slots protected by plastic dielectric sheets. It is important to understand that their possible advantages are in cost, appearance, and packing convenience. They have no particular advantage over the parabolic reflector in the relationships among beamwidth, sidelobe level, and aperture. The theory of arrays of arbitrary elements is classical and well developed in many texts (Jasik, 1961; Silver, 1949; Kraus, 1950).

10.3.2 Horn Antennas

Horn antennas are commonly used as primary radiators in reflector systems, elements in arrays, and sometimes as complete radiators when wide beamwidths are required. Frequently, we find horn antennas on board the satellite to provide earth coverage beams. That angle is about 18° from geostationary orbit and simp'y achieved with horns.

We find two kinds of horns in common use: the *pyramidal horn* as an extension of rectangular waveguide and the *conical horn* as an extension of circular guide. Pyramidal horns are easily designed, and the following equations are applicable for those horns that are long compared to a wavelength:

$$G = 10\frac{AB}{\lambda^2}$$

$$\theta_E = 51\frac{\lambda}{B} \tag{10-5}$$

$$\theta_H = 70\frac{\lambda}{A}$$

where A is the longer dimension of the horn aperture. If it is desired to have the shortest length possible, that length, L_1, is given by

$$L_1 = L\left(1 - \frac{a}{2A} - \frac{b}{2B}\right) \tag{10-6}$$

Figure 10-7 permits the design of an optimum-length horn.

Conical horns, which are the natural extension of circular waveguides, are often used and typically exploit higher-mode propagation. If the TM_{11} and TE_{11}

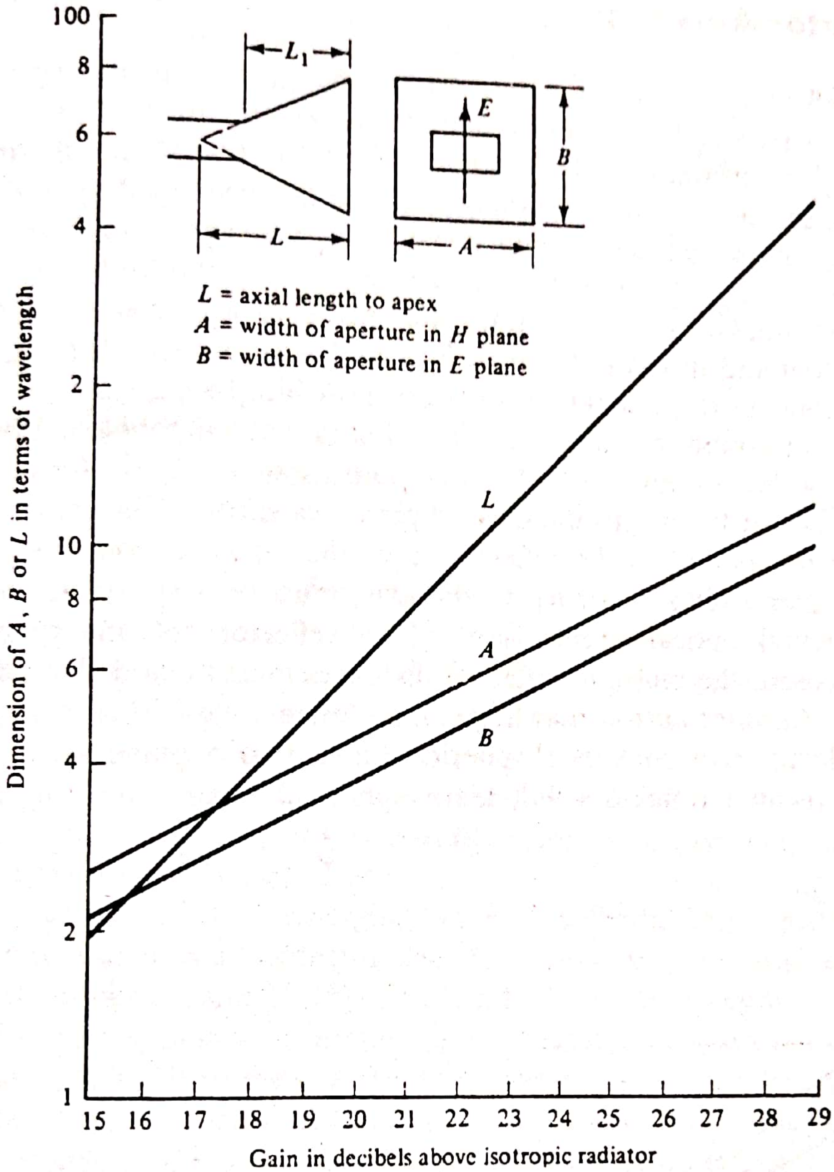
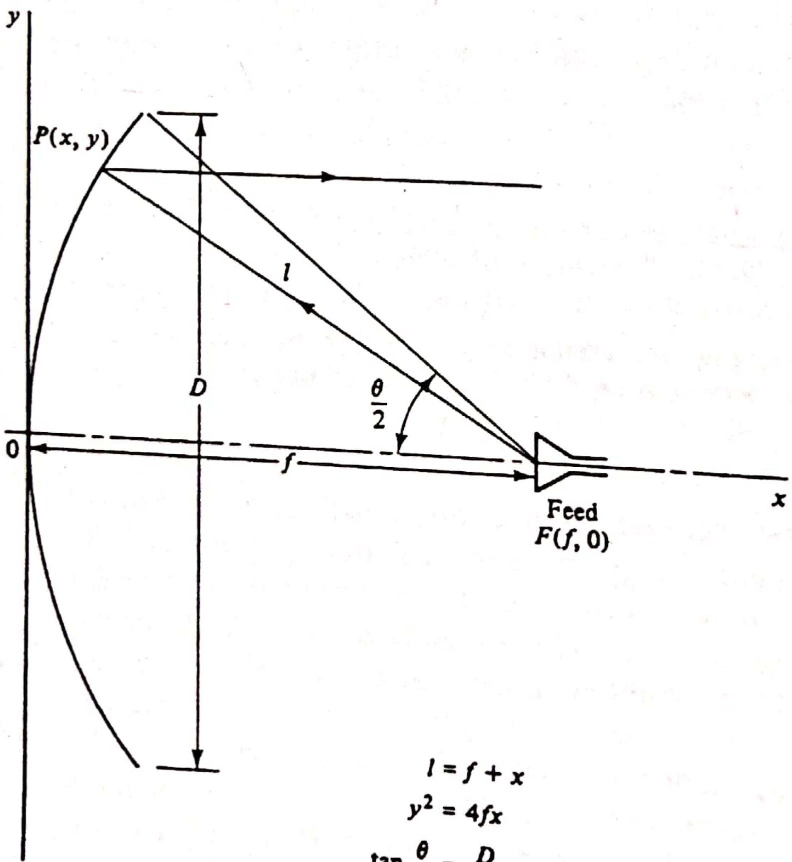


Figure 10-7 Design of electromagnetic-horn radiator. (From ITT *Reference Data for Radio Engineers*, reprinted by permission of Howard W. Sams and Company, Inc. Publishers, Indianapolis, IN, 1975).

modes circular waveguide are superimposed on each other with suitable control of the relative amplitude and phase, the composite radiation pattern can be improved over that of the single-mode horn in either rectangular or circular guide. Another variation of the horn feed very much used in primary feeds for big earth stations is a *hybrid-mode* horn. Annular corrugations are placed on the inner wall of a circular waveguide in such a way that neither TE or TM modes can be propagated, but instead a hybrid mode is generated. These antennas can be used to improve cross-polarization and sidelobe performance and also to achieve axially symmetric beamwidths. Miya (1985) gives considerable detail on the theory of this particular kind of horn. Conventional single- and multiple-mode horn feeds are discussed in both Silver (1949) and Jasik (1961).

10.3.3 Reflector Antennas

We divide the reflector antennas broadly into two categories: those using a single reflector and horn feed and those using multiple reflectors. In the first category, we have the familiar prime focus feed (Figure 10-8) and the offset-fed parabolic reflectors; in the second, we have a family of antennas developed by analogy to astronomical telescopes and thus called Newtonian, Cassegrainian, and Gregorian. The latter categories depend on whether the subreflector is plane, hyperbolic, or ellipsoidal. These antennas are shown in a convenient summary chart in Figure 10-9 and in more detail in Figures 10-10 through 10-12. Literally dozens of variations on these arrangements are possible. Several themes can be kept in mind to understand the variations. The first is that a paraboloidal reflector will take spherical waves emerging from a point source at the focus of the paraboloid and convert them into the desired plane wavefront. The distance from the focus on the paraboloid to the reflector, plus the distance from the point of reflection to any plane surface normal to the axis, must be a constant. This is in fact the fundamental optical requirement of all reflector antenna systems. If subreflectors are used, the multiple reflected distances must be added in satisfying the requirement. Another notion that helps in understanding is that, whereas the paraboloidal reflector will convert a spherical wave into a plane wave, hyperboloidal and ellipsoidal reflectors will leave spherical waves emerging from a



$$l = f + x$$

$$y^2 = 4fx$$

$$\tan\frac{\theta}{4} = \frac{D}{4f}$$

Figure 10-8 Basic geometry; prime-focus-feed parabolic reflector.

focus unchanged. They simply seem to have originated at a different focus. This explains, for instance, the modified Cassegrainian antenna (Figure 10-12) seen occasionally in which the horn feed located at the vertex of the main reflector is paraboloidal. As a result, it is necessary to use a paraboloidal subreflector to convert the plane waves from the prime source back into spherical waves, which in turn get converted back into plane waves by the main paraboloid.

An important effect of the secondary reflector on a Cassegrainian or Gregorian antenna is to increase the apparent focal length of the antenna. This increase is called *magnification* by analogy to the optical case, and proceeding from the geometric definitions that the ellipse and hyperbola are, respectively, the loci of points for which the sum of or difference between the distances to two fixed points is a constant, the two fixed points being called the foci, we can demonstrate that the equivalent focal length of the Cassegrainian reflector system is given by

$$f_e = mf = \frac{e+1}{e-1}f \tag{10-7}$$

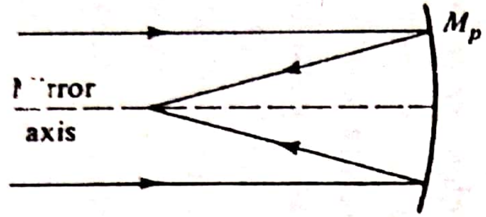
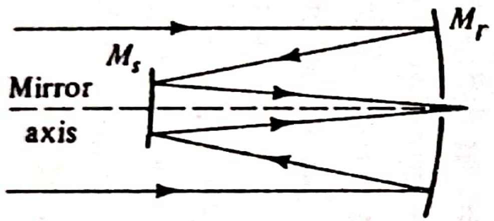
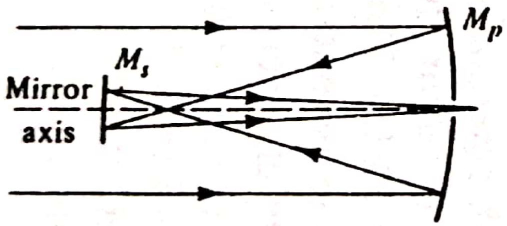
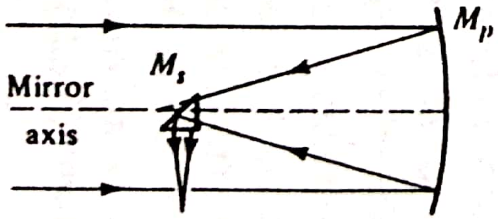
The geometry is seen in Figure 10-10. The angle subtended at the focal point F_2 is very much less than it would be if the feed were located at the virtual focus, F_1. The feed located at F_1 can be designed as if the focal length and thus the ratio f/D were longer by the factor m. This makes the realization of high-aperture efficiencies and low cross-polarization components much easier. Values of m typically range from 2 to 6.

If several horn feeds with emerging beams at different angles are to be used, it is possible to use a main reflector that is circular in one cross section and parabolic in the other (Figure 10-13). This kind of toroidal antenna was first used in large early-warning radars to permit rapid beam scanning over perhaps a 70° sector. As such, it is also useful when one antenna is to be used with several satellites. The circular cross section produces spherical aberration in one axis, which is correctable in the horn feed design. It is important to note that there is no saving in total aperture, but there can be a saving in cost and complexity as the number of beams (feeds) is equal to three or more.

10.3.4 Antenna Performance

The easiest way to compare antennas for system performance is by considering them as illuminated apertures. The secondary radiation pattern from an illuminated aperture can be shown to be a Fourier transform of the primary pattern, general relations can be found among such critical parameters as size, illumination taper, sidelobe level, directivity, and beamwidth. The universal antenna formula relating the effective area (or capture cross section) of the antenna A_{eff} and its gain and wavelength is the familiar

$$A_e = \frac{G\lambda^2}{4\pi} = \eta A \tag{10-8}$$

Type	Ray Diagram	Optical Elements	Pertinent Design Characteristics
Paraboloid		Reflective M_p = paraboloidal mirror	1. Free from spherical aberration 2. Suffers from off-axis coma 3. Available in small and large diameters and f/numbers 4. Low IR loss (reflective) 5. Detector must be located in front of optics
Cassegrain		Reflective M_p = paraboloidal mirror M_s = hyperboloidel mirror	1. Free from spherical aberration 2. Shorter than Gregorian 3. Permits location of detector behind optical system 4. Quite extensively used
Gregorian		Reflective M_p = paraboloidal mirror M_s = ellipsoidal mirror	1. Free from spherical aberration 2. Longer than Cassegrain 3. Permits location of detector behind optical system 4. Gregorian less common than Cassegrain
Newtonian		Reflective M_p = paraboloidal mirror M_s = reflecting prism or plane mirror	1. Suffers from off-axis coma 2. Central obstruction by prism or mirror

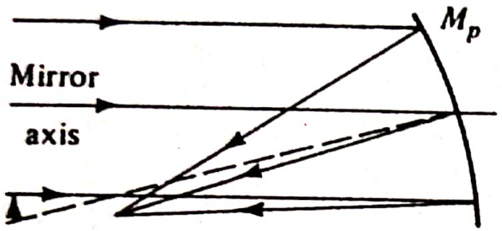
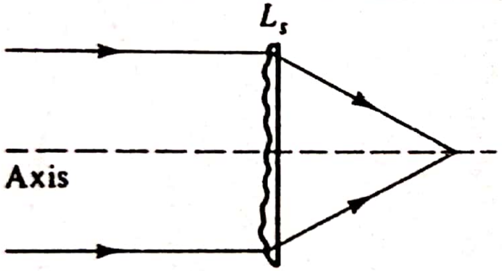
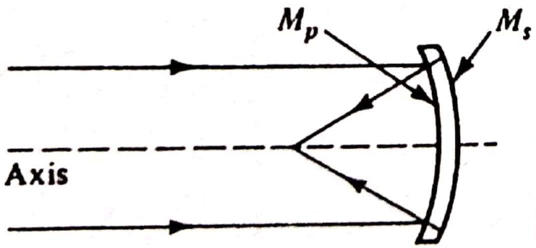
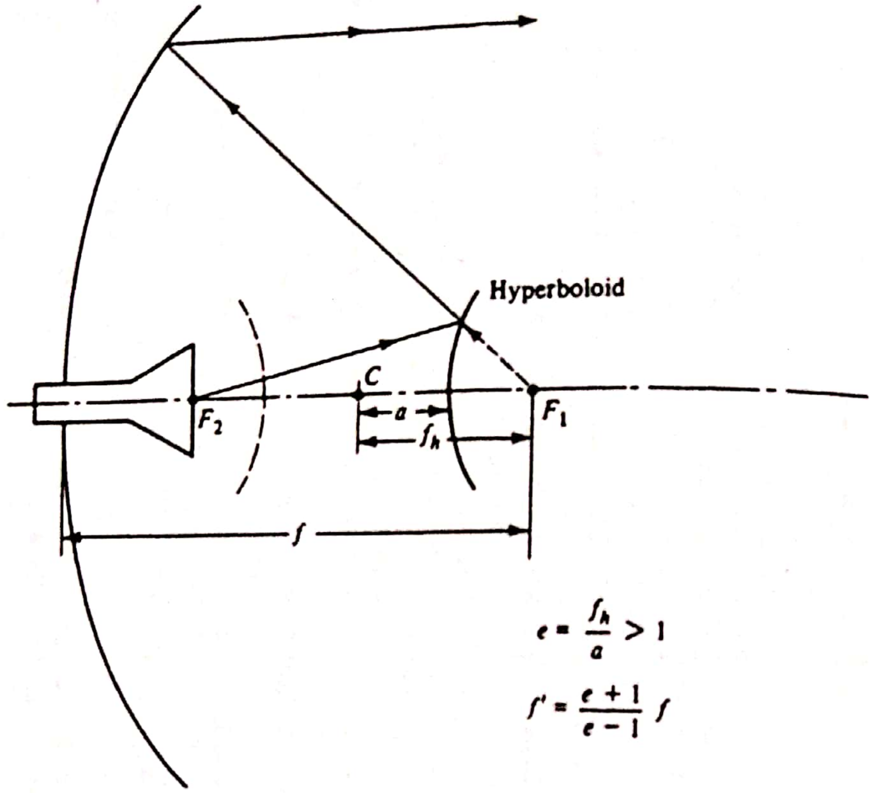
Type	Ray Diagram	Optical Elements	Pertinent Design Characteristics
Herschelian		Reflective M_p = paraboloidal mirror inclined axis	1. Not widely used now 2. No central obstruction by auxiliary lens 3. Simple construction 4. Suffers from some coma
Fresnel lens		Refractive L_p = special Fresnel lens	1. Free of spherical aberration 2. Inherently lighter weight 3. Small axial space 4. Small thickness reduced infrared absorption 5. Difficult to produce with present infrared transmitting materials
Mangin mirror		Refractive–reflective M_p = spherical refractor M_s = spherical reflector	1. Suitable for IR source systems 2. Free of spherical aberration 3. Most suitable for small apertures 4. Covers small angular field 5. Uses spherical surfaces

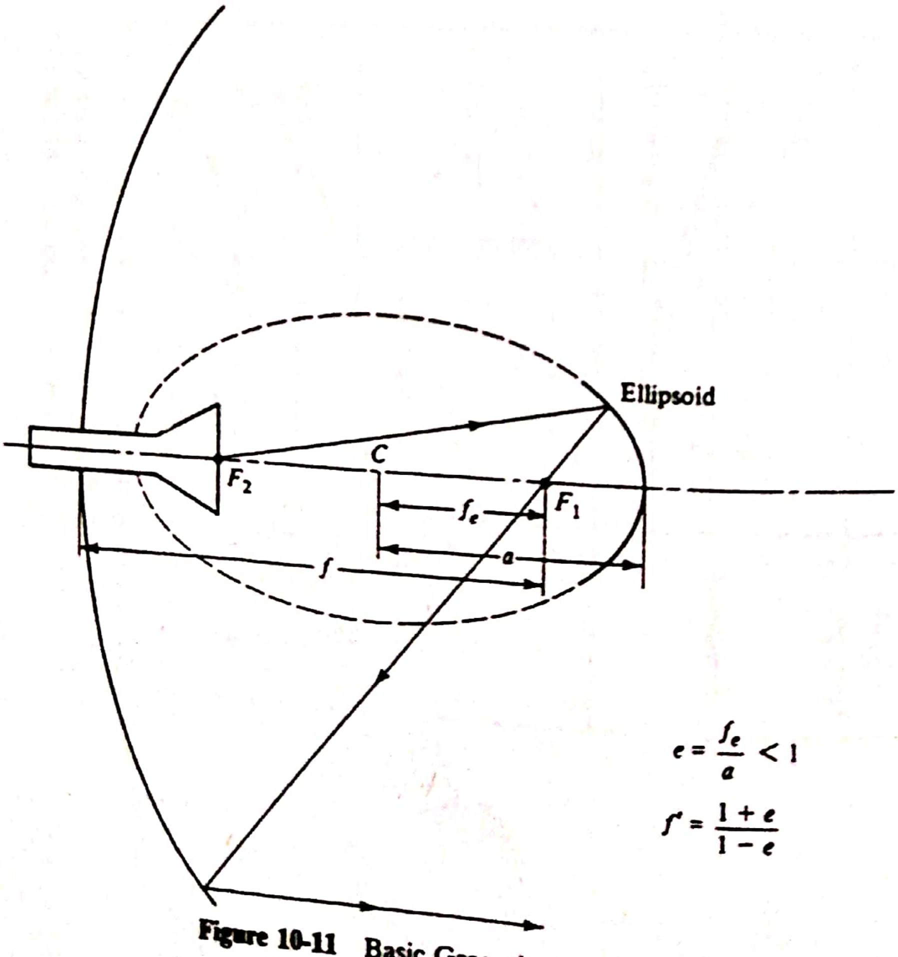
Figure 10-9 Quasi-optical apertures. (Reprinted with permission from G. F. Levy, "Infrared System Design", *Electrical Design Views*, May, 1958, Table 1).



$$e = \frac{f_h}{a} > 1$$

$$f' = \frac{e+1}{e-1}\, f$$

Figure 10-10 Basic Cassegrainian antenna.



$$e = \frac{f_e}{a} < 1$$

$$f' = \frac{1+e}{1-e}$$

Figure 10-11 Basic Gregorian antenna.

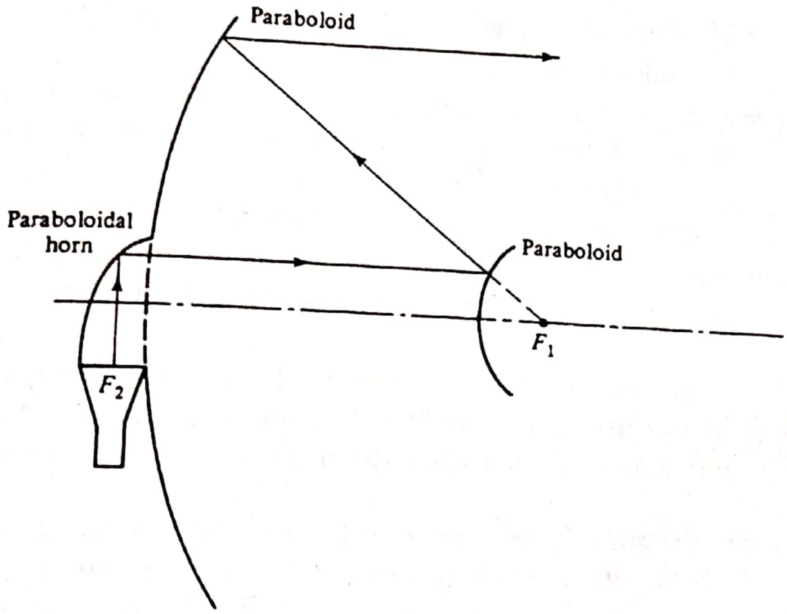


Figure 10-12 Near-field or modified Cassegrainian antenna.

The effective or capture area is related to the physical area A by the overall efficiency η. This overall efficiency η, which must be used in calculating received carrier level, is itself the product of several constituent efficiencies. thus

$$\eta = \eta_a \eta_b \eta_s \eta_p \eta_e \eta_L \tag{10-9}$$

where

η_a = *aperture efficiency*, the result of nonuniform illumination, phase errors, and so on; it *increases* as the sidelobe level increases

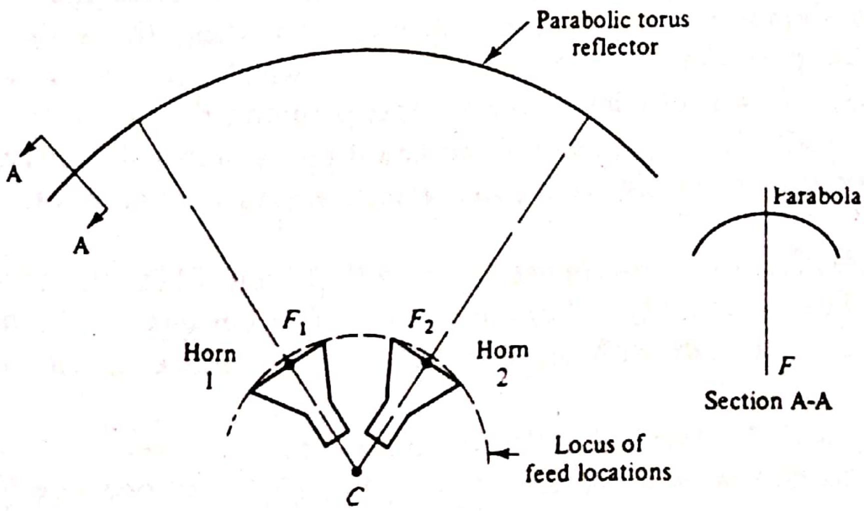


Figure 10-13 Multifeed toroidal antenna.

η_b = *blockage efficiency*, resulting from blockage of main reflector by the subreflector or feeds

η_s = *spillover efficiency*, the loss of energy because the subreflectors and main reflector do not intercept all the energy directed toward them

η_p = *cross-polarization efficiency*, the loss of energy due to energy coupled into the polarization orthogonal to that desired

η_e = *surface efficiency*, the loss in gain resulting from surface irregularities, the statistical departure from a theoretically correct surface

η_L = *ohmic and mismatch efficiency*, the loss from energy reflected at the input terminals (VSWR > 1.0) and that dissipated in ohmic loss in the conducting surfaces, dielectric lenses, and so on

Aperture efficiency is an important but subtle concept. It describes the degree to which an illuminated aperture achieves the *directivity*, and thus the *gain*, that its area would imply (Kraus, 1950; Silver, 1949). The aperture efficiency, η_a, is equal to unity for an aperture that is illuminated uniformly in amplitude and phase, in which case the directivity is the maximum for the given area. In this case, however, the gain of the first sidelobe is only 13 dB below that of the main lobe.

By tapering the intensity of illumination toward the edge of the aperture, the relative sidelobe gain decreases, but so also does η_a. Thus the on-axis gain decreases and, as a corollary, the width of the main beam increases [see Eq. (10-4)]. At the same time, the taper of illumination likely reduces the *spillover* (a cause of loss of overall efficiency). The taper, however, must be produced by narrowing the beam of the feed system. This in turn implies an increase in the size of the feedhorn (or a subreflector) and thus may increase *blockage*, giving decrease in overall efficiency. The reduced spillover will not only increase the value of η_s but also may improve the antenna temperature, T_a (see Chapter 6), since less energy is interchanged with the ground. Thus the compromise to achieve optimum G/T is indeed complex. To assist the system designer in knowing the possibilities, we have Table 10-4, which gives beamwidths, aperture efficiencies, and sidelobe levels for several common primary patterns achievable with horn feeds. The chart assumes constant phase across the aperture. Since this is not easy to achieve perfectly, the actual aperture efficiencies will be slightly lower.

In practice, the compromises are made by controlling the edge illumination or taper. For a cosine-type horn feed, practical designs can be arrived at in a simpler way. We start with the desired efficiency and a certain edge taper, say from Figure 10-14.

Note that this taper in reflector illumination has two components: one due to the horn feed pattern, as shown in Table 10-4, and one due to the inherent reflector geometry that would be present even with a uniform primary pattern.

TABLE 10-4 APERTURE CHARACTERISTICS VERSUS ILLUMINATION PATTERN

Illumination across Aperture		Aperture Efficiency, ηa	Half-power Beam Width[a]	First Null[b]	First Sidelobe, dB[c]
Uniform	$n = 0$	1.00	50	57	−13.2
	$n = 1$	0.810	69	86	−23
$\cos^n \frac{\pi x}{2a}$	$n = 2$	0.667	83	115	−32
	$n = 3$	0.575	95	143	−40
	$n = 4$	—	111	11	−48
$1 - \varepsilon x^2$	$\varepsilon = 1$	0.833	66	82	−20.6
	$\varepsilon = 0.5$	0.970	56	65	−17

[a] As a multiple of λ/D, where D is the physical diameter of the aperture.
[b] From axis as a multiple of λ/D.
[c] With respect to on-axis gain.

The second term is sometimes called *space attenuation* and is simply the difference in inverse square-law loss between the edge and center of the aperture. From the geometry of the parabola, it can be shown that this loss is given by

$$\text{space attenuation} = \left(\frac{R}{f}\right)^2 = \sec^4 \frac{\theta}{4} \tag{10-10}$$

where θ is the full angle subtended by the reflector at the horn.

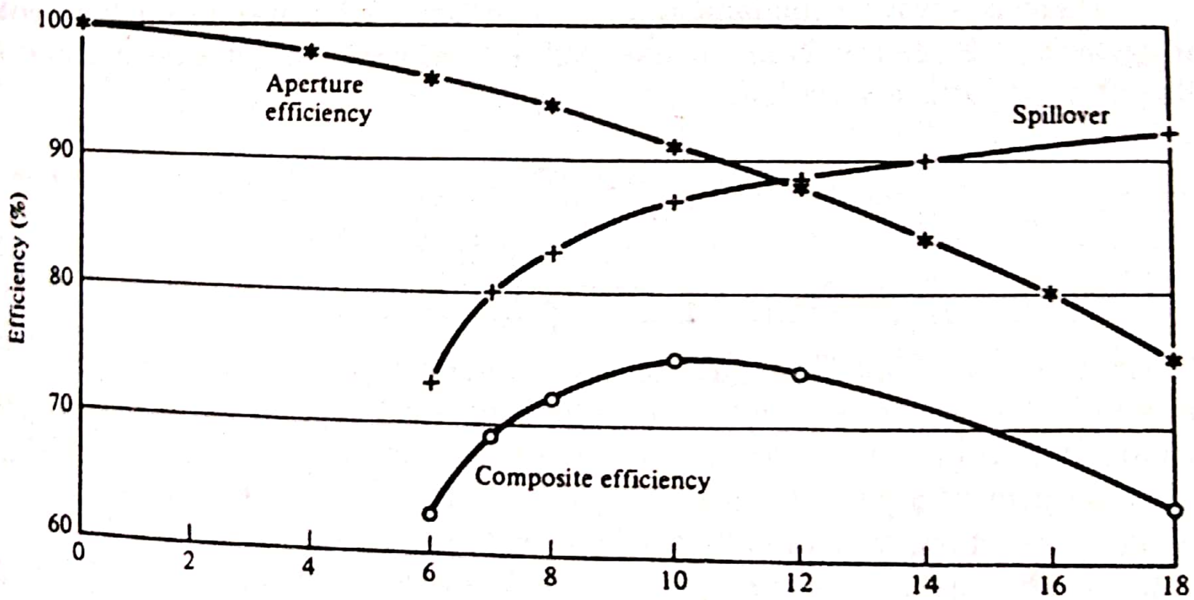


Figure 10-14 Typical spillover and aperture efficiencies as a function of illumination taper.

A good approximation to a cosine horn pattern (in decibels) is simply $10(\theta/\beta_{10})^2$, where β_{10} is the horn beamwidth at the tenth power point. For preliminary planning, it can be used as a universal feed pattern. Thus the net edge taper T is

$$T = 10 \log \sec^4 \frac{\theta}{4} + 10\left(\frac{\theta}{\beta_{10}}\right)^2 \tag{10-11}$$

and the horn feed beamwidth β_{10} can be chosen so as to achieve the desired taper. θ is determined from the f/D ratio and the reflector system geometry. The main reflector beamwidth is calculated using the beamwidth factor that goes with the chosen aperture efficiency in Table 10-4.

In addition to the aperture efficiency just discussed, we have a variety of other factors that contribute to the overall capture efficiency. Some of them, notably spillover, tend to vary contrary to the aperture characteristic as seen in Figure 10-14. The optimum system performance may be achieved using a slightly higher taper than shown, since the spillover has the doubly bad effect of both reducing that constituent of the gain and deteriorating the antenna temperature.

Aperture blockage is a significant problem, especially in Cassegrainian and Gregorian antennas. As a good approximation, the related efficiency η_b is given by $[1 - \eta_a(A_B/A)^2]$, where A_B is the blocked area and A is the total aperture area.

Cross-polarization efficiency η_p is another important problem in satellite antennas. For a symmetrically illuminated antenna, it depends on the curvature of the reflector and thus on the f/D ratio. Cassegrainian antennas with magnifications of 2 or greater are extremely good in this respect, but focal point feeds can be somewhat poorer. Off-center feeds are still worse. The loss will depend on the fourth power of the subtended angle at the primary feed.

There is always a fundamental loss in efficiency because of random *surface* irregularities. Ruze (1952) in a classic paper developed the following equation for the effect of surface variation:

$$\eta_e = \frac{G}{G_0} = e^{-k(4\pi\delta/\lambda)^2} \tag{10-12}$$

$$k = \frac{1}{1 + (D/4f)^2} \approx 1 \tag{10-13}$$

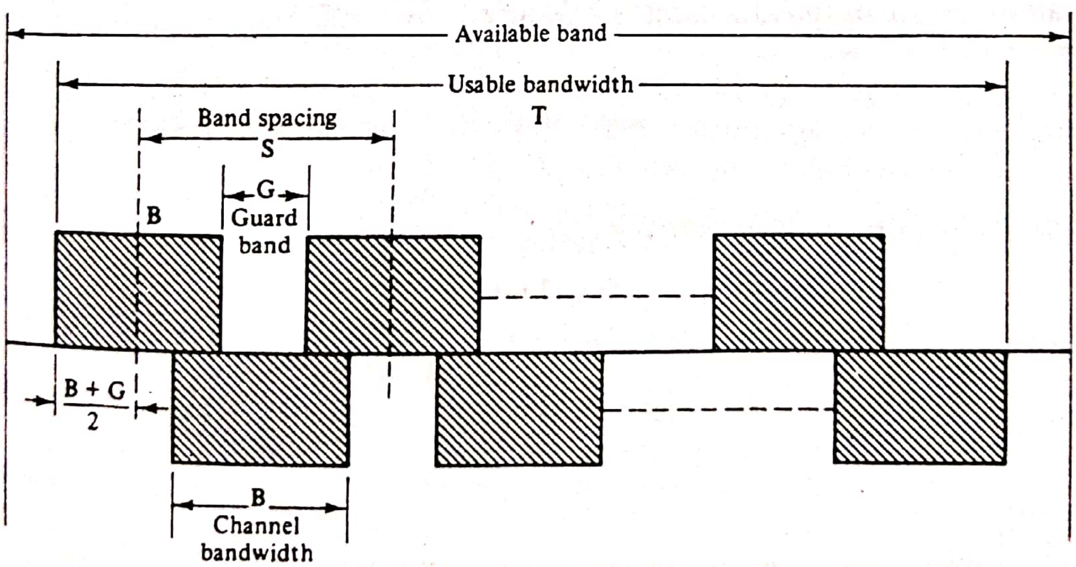
These equations hold for Gaussian distribution of phase errors due to surface imperfections. Correlation intervals should be small compared to aperture and comparable to or greater than a wavelength. δ is the mass surface deviation and G_0 is the gain of a perfect surface reflector. It is a good, practical equation. For multiple reflectors, δ^2 should be the sum of squares of the values for the various surfaces. The effect of Ruze's equation is to put a practical upper limit on the gain of a constant-size antenna as the wavelength decreases. This limit seems to be somewhere around 70 dB today, and any system plan calling for an intenna gain in that region should be reviewed carefully with antenna experts.

An often overlooked loss in gain is that due to resistive losses and *antenna mismatch*. This loss is not part of the radiation characteristic of the antenna, but is nonetheless effective in reducing the system performance. Not only do resistive losses reduce the gain, but they also increase the noise temperature.

The overall gain, taking into account all these effects, is the one to be used in system planning. The composite of all the efficiency factors for large and expensive earth stations today seem to run between 0.6 and 0.7; for smaller cheaper stations, between 0.55 and 0.6. Home terminals could well be assumed conservatively at 0.5 for system planning.

10.3.5 Cross-polarized systems

Often, the satellite channel capacity is limited by the available bandwidth and, if sufficient power is available, it is desirable that the assigned frequency band be used as many times as possible. This can be achieved in the satellite by spot beam antennas, as discussed in Chapter 9, but in addition, and now almost universally, two polarizations are used in satellite communications systems to achieve at least a two-for-one frequency reuse. Polarizations used can be either crossed linear, that is, vertical versus horizontal, or counterrotating (left versus right-hand circular polarization). Keep in mind that circular polarization may be considered simply the combination of two linearly polarized waves in both axis and time quadrature. Typical dual-polarization channelization plans are shown in Figure 10-15. As a first approximation, both systems are equally effective at isolating two beams at the same frequency from each other, and the choice between the



$$S = B(1+g)$$
$$N = \frac{2T - B(1-g)}{2B(1+g)} \qquad (10\text{-}14)$$

Figure 10-15 Interlaced, cross-polarized channelization.

systems becomes a matter of practical technique. Circular polarization systems have the advantage of not requiring polarization orientation, which can be important in simple systems, but they suffer more from depolarization during heavy rain. Conversely, crossed linear systems require polarization orientation and sometimes readjustment as satellite pointing is changed, but tend to perform slightly better in the presence of rain. The crossed-polarization isolation of an antenna is thus an important parameter. In this respect, the Cassegrainian family of antennas is extremely good because of both the long effective focal length and the axial symmetry. Offset antennas, sometimes used for small terminals and multiple beams, suffer from a lack of axial symmetry and therefore have poor crossed-polarization characteristics. On the other hand, such antennas are free of the aperture blockage problems that plague the Cassegrainian family. As a result, their sidelobe levels can be extremely low, not much above the theoretical level attributable to the illumination taper and aperture size.

To produce circularly polarized waves, it is necessary to shift one linear component (either vertical or horizontal) by 90° in time relative to the other and to maintain this phase shift over a wide band of frequencies. This is not easy, and the antenna curvature itself disturbs the phase relationship. If the required axis and phase quadratures are not maintained, or if the two components are not of equal magnitude, an *elliptical* rather than circular polarization is created. This can be shown to be resolvable into two circularly polarized signals of opposite hand, with field strength magnitudes E_L and E_R, respectively. Assume that for this channel left-hand polarization was intended. The E_R component will now represent a type of crosstalk into the receiver channel that is intended for our other signal (with right-hand polarization intended). Thus, the systems' discrimination between polarizations is reduced by "ellipticity" to the *circular polarization ratio*.

$$p = \frac{E_R}{E_L} \tag{10-15}$$

which in dB form is called *the cross-polarization* discrimination

$$XPD = 20 \log p \tag{10-16}$$

The ratio between the amplitudes of the electric fields along the major and minor axes o. the polarization ellipse, the *axial ratio*, is

$$r = \frac{E_L + E_R}{E_L - E_R} \tag{10-17}$$

The two ratios, r and p, are related by

$$r = \frac{p+1}{p-1} \qquad p = \frac{r+1}{r-1} \tag{10-18}$$

Imperfect cross-polarization discrimination, whether for linear or circular polarization systems, effectively yields another carrier-to-interference ratio that must be considered in assessing overall performance.

10.3.6 Antenna Mounts

The antenna must be pointed at the satellite. This pointing is occasionally rudimentary. A dipole-like antenna pattern produces hemispherical coverage and an antenna gain just in excess of 3.0 dB. This is convenient for small, mobile terminals, either two-way or transmit-only. At the other extreme, we have narrow-beam antennas that must be pointed continuously at the satellite, and with great care. In between, we have several levels of pointing and tracking precision depending on the service and performance parameters. Sometimes, but rarely, this pointing is fixed permanently, sometimes it is occasionally adjusted, and in some installations it is continually driven by a tracking system. Such tracking systems will be discussed in Section 10.4, but in the meantime we note that every earth station antenna must be capable of some adjustment in pointing, even if only for initial setup. Such adjustments come in three categories geometrically. The simplest, and indeed the most flexible, is the *elevation over azimuth system*, usually called *Az-El* (Figure 10-16a). In this system the azimuth is determined by rotation about a vertical axis normal to the local horizontal plane, and the elevation is adjusted about a horizontal elevation axis, which is in turn normal to the vertical azimuth axis. This system is simple, effective, and capable of use almost anywhere. There is some difficulty with automatic tracking systems if the earth station is to be located near the equator. Tracking through the zenith is awkward with Az-El systems, and if that is required, it is common to use a system in which one axis is parallel to the ground and another axis is normal to it, as shown in Figure 10-16b. This kind of mount, commonly called *XY*, can also be used to point anywhere, but is awkward for tracking close to the horizon. An interesting third possibility, borrowed from astronomy, is the equatorial mount (Figure 10-16c). One axis, the *hour-angle axis*, is parallel to the axis of the earth, and the other axis, normal to it, produces the desired *declination*. This mount is used universally for astronomical telescopes since it permits the automatic tracking of a celestial object despite the rotation of the earth. It is easily applicable to satellite systems. An interesting complication appears if the mount is to be sufficiently flexible, as most mounts are, to point anywhere on the visible geostationary arc. If the earth station were located on the equator, it would be necessary only to make an adjustment in the hour-angle axis. The declination would be everywhere 0°. Even for locations at other latitudes, the change in declination as the hour angle is varied to point along the geostationary arc is rather small. If the beamwidth of the antennas is wide compared to the associated change in declination, as is often the case with small terminals, it is possible to point to satellites at different longitudes by varying only one axis. The error is small enough to be acceptable in many systems. We expect that this approach will be used increasingly in small- and medium- sized antenna terminals.

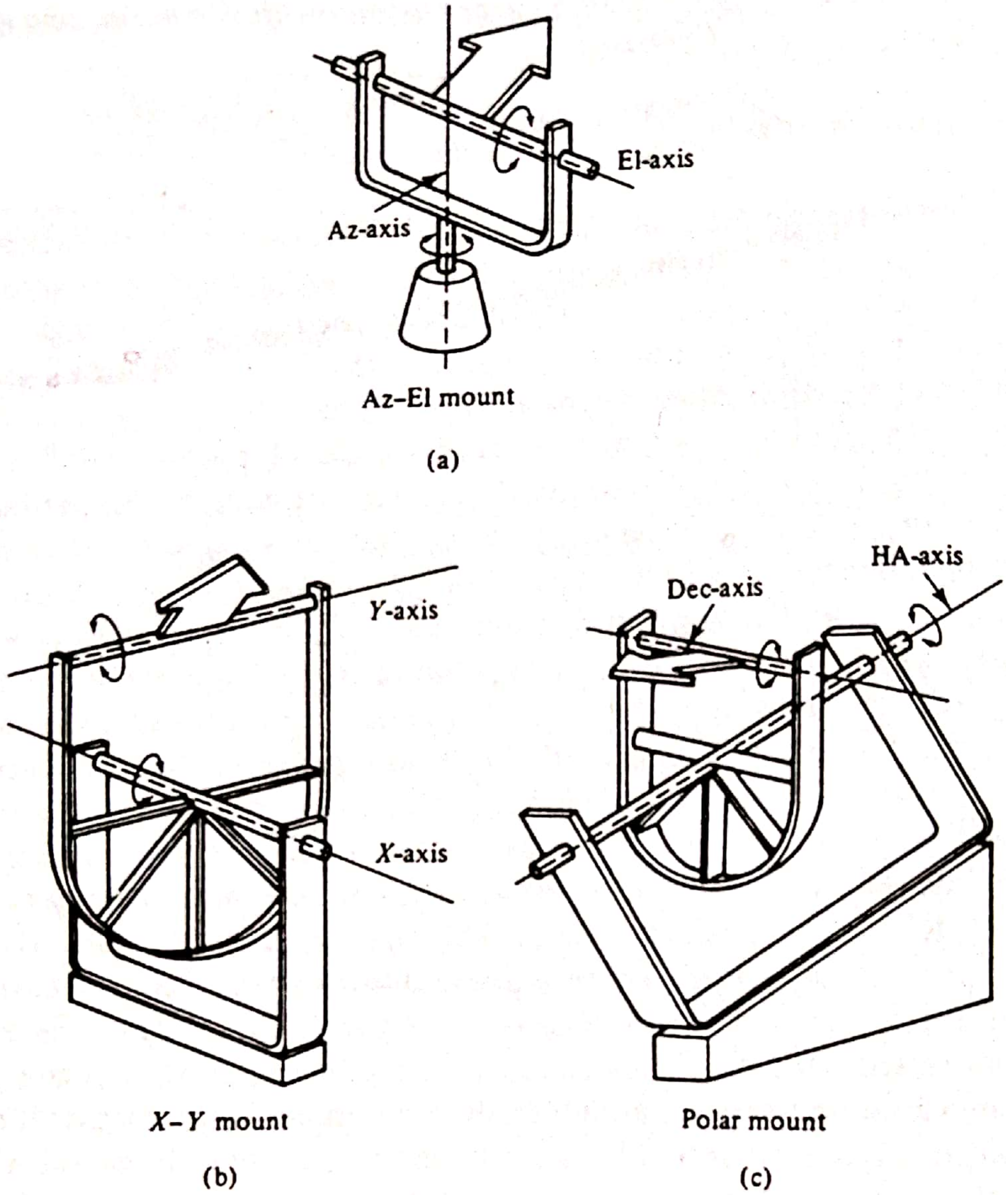


Figure 10-16 Antenna mounts: (a) elevation over azimuth; (b) Y over X; (c) equatorial or polar. (Reprinted with permission from K. Miya, *Satellite Communications Engineering*, 2nd ed., Lattice Company, Tokyo, p. 232.)

The error in single axis pointing can be calculated several ways. Equations (3-61) and (3-62) for the correction of parallax in declination can be used to calculate the change in δ at the limits of coverage, or a lesser range of longitudes if desired, assuming that it is set correctly for a satellite on the local meridian. It can also be calculated in a straightforward way using the geometry of Figure 10-17. P_0 is the satellite location on the observer's meridian at which the declination is set equal to δ_0. At some other longitude λ, the declination is δ and difference is the error. The result is given by Eq. (10-19) and is plotted for λ at the edge of coverage as a function of latitude in Figure 10-18.

$$\delta - \delta_0 = \cos^{-1}\frac{p}{d} - \cos^{-1}\frac{p}{d_0} \tag{10-19}$$

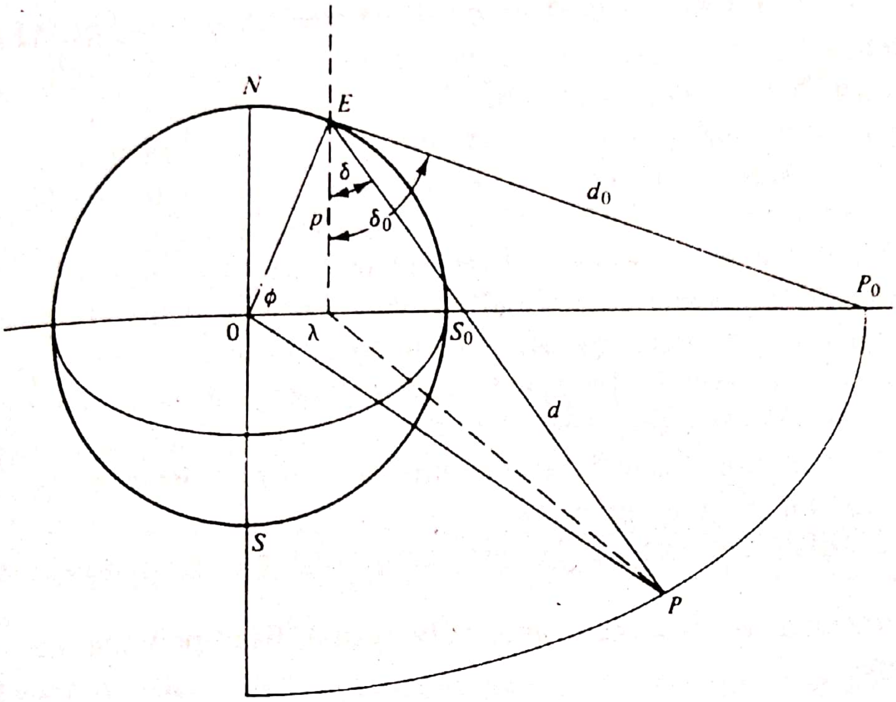


Figure 10-17 Calculation of the error in single axis pointing.

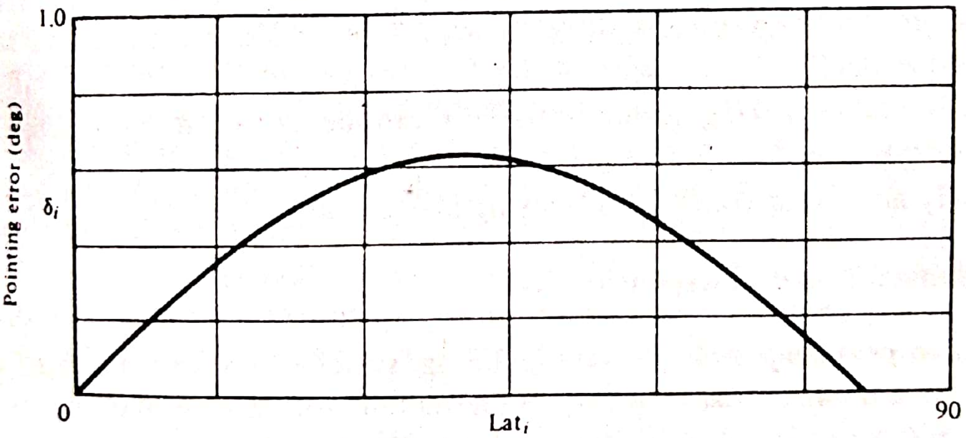


Figure 10-18 Plot of edge of coverage pointing error vs. latitude.

Note that for small antennas with beamwidths of several degrees or more this pointing error is negligible.

TRACKING SYSTEMS

Tracking satellites, as distinguished from simply pointing at them in an initial orientation or switching from one satellite to another, is required whenever the antenna beam widths are narrower than the expected geocentric satellite orbital motions. We encounter such motions with nongeostationary satellites and in the residual motions of geostationary satellites. The problem is relative and depends on the orbit and antenna beamwidth. In general, nongeostationary orbits require more tracking than geostationary, but there are many exceptions. For instance, messaging systems for ground mobile service from low earth orbit often use hemispherical coverage antennas, as do aeronautical and many marine terminals, and require no tracking, which is operationally and economically desirable. On the other hand, there are successful mobile services to vehicles using Ku band, with the inevitable narrow beams and tracking because of the vehicular motion. Some fixed-service stations operating with GEO satellites have large antennas relative to the wavelength, beamwidths well under a degree, and full tracking systems to accommodate the residual motions. Normally, GEO satellites are held within $\pm 0.1°$, but antennas with beamwidths even several times that value will use tracking to minimize pointing losses.

We identify a hierarchy of pointing and tracking categories as follows:

1. No tracking is necessary and only initial fixed-pointing adjustment is required.
2. Repointing of the antenna is needed to switch from one satellite to another and possibly to correct for satellite motion. This repointing can be needed rarely or frequently.
3. Tracking is required, but it is satisfactory to drive the antenna in two axes and to preprogram this drive in accordance with the calculated satellite motion.
4. Automatic tracking is necessary but can be achieved by a simple *step-tracking* system.
5. Fully automatic continuous tracking is necessary.

Some comment on each type is useful.

Fixed-pointing only. Fixed-pointing systems are usually restricted to wide beam antennas. The geometry of the mounts is as discussed in the preceding section; screw drives are available for initial adjustments.

Occasional repointing. The adjustments are flexible enough so that they can be changed manually without difficulty. Simple motor drives may be added to do it remotely. One-axis mounts are common.

Preprogrammed. Once motor drives are available for one- or two-axis control, a variety of methods, both automatic and preprogrammed, can be used. The orbital position of any satellite is calculatable to a high precision even allowing for gravitational anomalies using the methods outlined in Chapter 3. This applies equally well to GEO satellites and their drift orbits. If the antenna beamwidth is wide relative to the prediction error, it can be preprogrammed to track open loop. Often the principal apparent GEO satellite motion is that due to imperfect inclination control. This motion, for small inclinations and otherwise perfect orbits, is a figure eight with a period of one sidereal day. Its vertical height is twice the orbital inclination, and its width is only a small fraction of that value. The methods of Section 3.5 are easily used to calculate the figure eight in detail, if necessary. If the orbit has zero inclination but has a small eccentricity e, the amplitude of the maximum longitudinal departure is $2e$ radians. We can calculate the satellite orbital position as a function of time, considering both effects, and then correct to azimuth and elevation. This tracking method has been used frequently in large stations.

Step tracking. Step tracking uses a primitive servomechanism in which the antenna is moved a discrete amount in a step, and if the signal level increases, it is moved again in this direction. As soon as the signal level does not increase, it returns to the previous position. The fineness of this method obviously depends on the size of step. Nonetheless, it is satisfactory for all but rather demanding applications.

Fully automatic. Fully automatic tracking can be provided using methods originally developed for the pointing of radar antennas. The most common is the *monopulse* or *simultaneous lobing system,* in which four beams are generated in an auxiliary feed, and combinations of the signals from these four beams provide left–right and up–down error signals. These error signals are detected, amplified, and used to generate control signals for driving the antenna. A block diagram of a general automatic tracking system is seen in Figure 10-19. Such systems are complicated and expensive and are required only for narrow beamwidths. Note that it requires four extra antenna feeds in addition to the electronics and precise two-axis drives. Such precision usually precludes the use of single-axis mounts. It is possible to derive the error signals either with multihorn systems or by the use of higher modes in the main antenna feed. The multiple-horn feeds use four horns grouped together, or sometimes four horns grouped around a single larger horn, whereas the higher-mode error-determining signals use circular waveguide modes such as TM_{01} or TE_{01}, which have no field component on the axis. The secondary radiation patterns produced by such modes thus have a null on the main axis of the antenna rather than a peak, and departures from this null can be used to generate an error signal. Many variations using different modes and combinations are possible, and the choice is the antenna designer's and depends on the polarization method. Horn feed design for tracking and cross-polarized systems is complicated. One possible block diagram is shown in Figure 10-20.

10.5 TERRESTRIAL INTERFACE

The terrestrial interface comprises a wide variety of equipment. At one extreme, when the terminal is a mobile or receive-only station, there may be no terrestrial interface equipment at all. The operating devices, such as TV receivers,

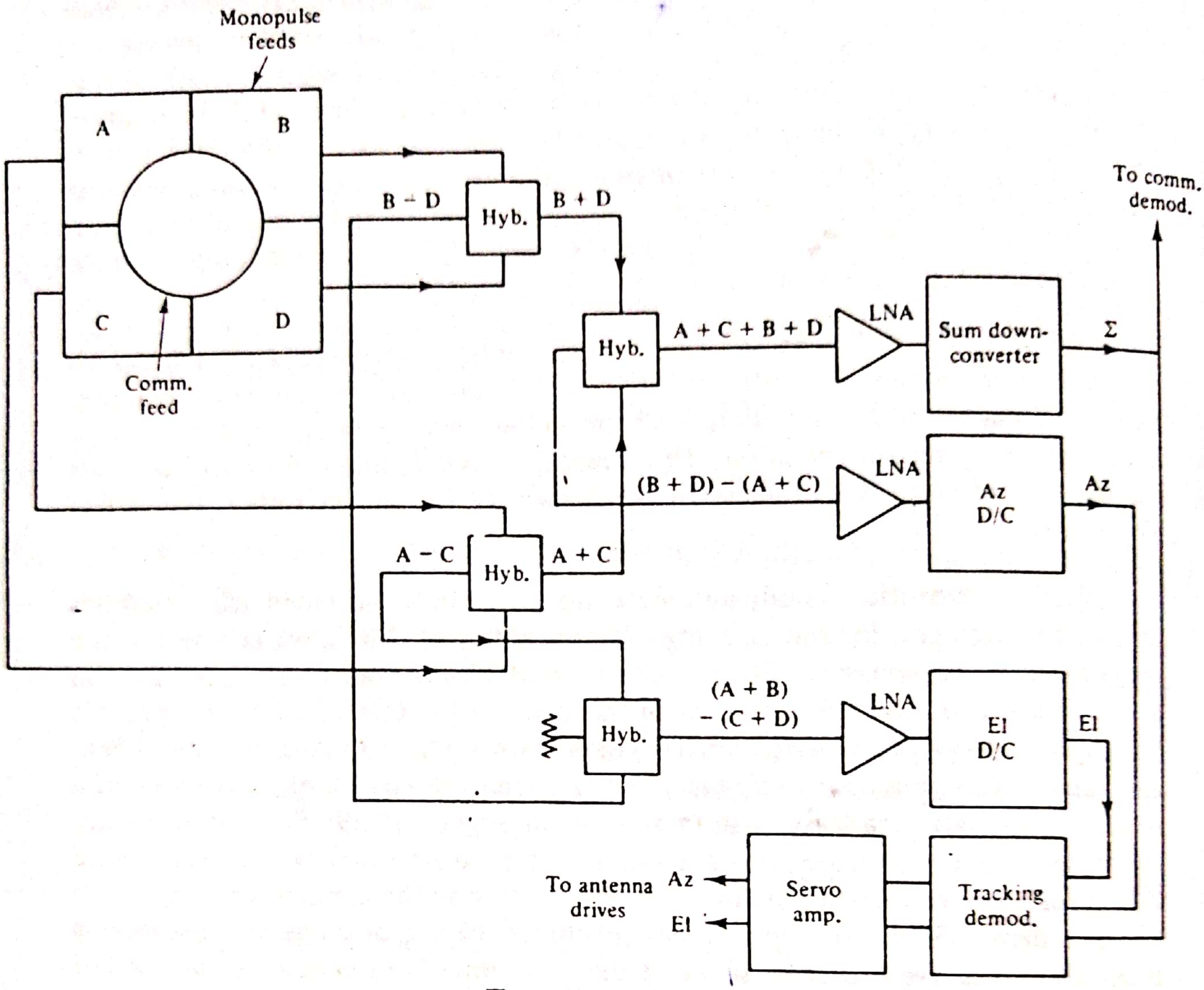


Figure 10-19 Tracking system.

telephones, data sets, and so on, are used right at the earth station. At the other extreme, we find the interface equipment necessary in a large commercial satellite system for fixed service. In such cases, hundreds of telephone channels, together with data and video, are brought to the station by microwave and cable systems using either frequency- or time-division terrestrial multiplex methods. The signals must be changed from those formats into formats suitable for satellite transmission. In an easy case, frequency-division multiplex groups and supergroups, as brought in from terrestrial transmission facilities, can be transmitted directly or with simple translation in basebound frequency from the satellite after modulation and up-conversion, but in many cases it is necessary to reformat extensively for terrestrial circuits. Individual telephone channels, for instance, may all be transmitted on the same carrier, which is received by many earth stations in the network. The return channels for particular conversation circuits will be coming

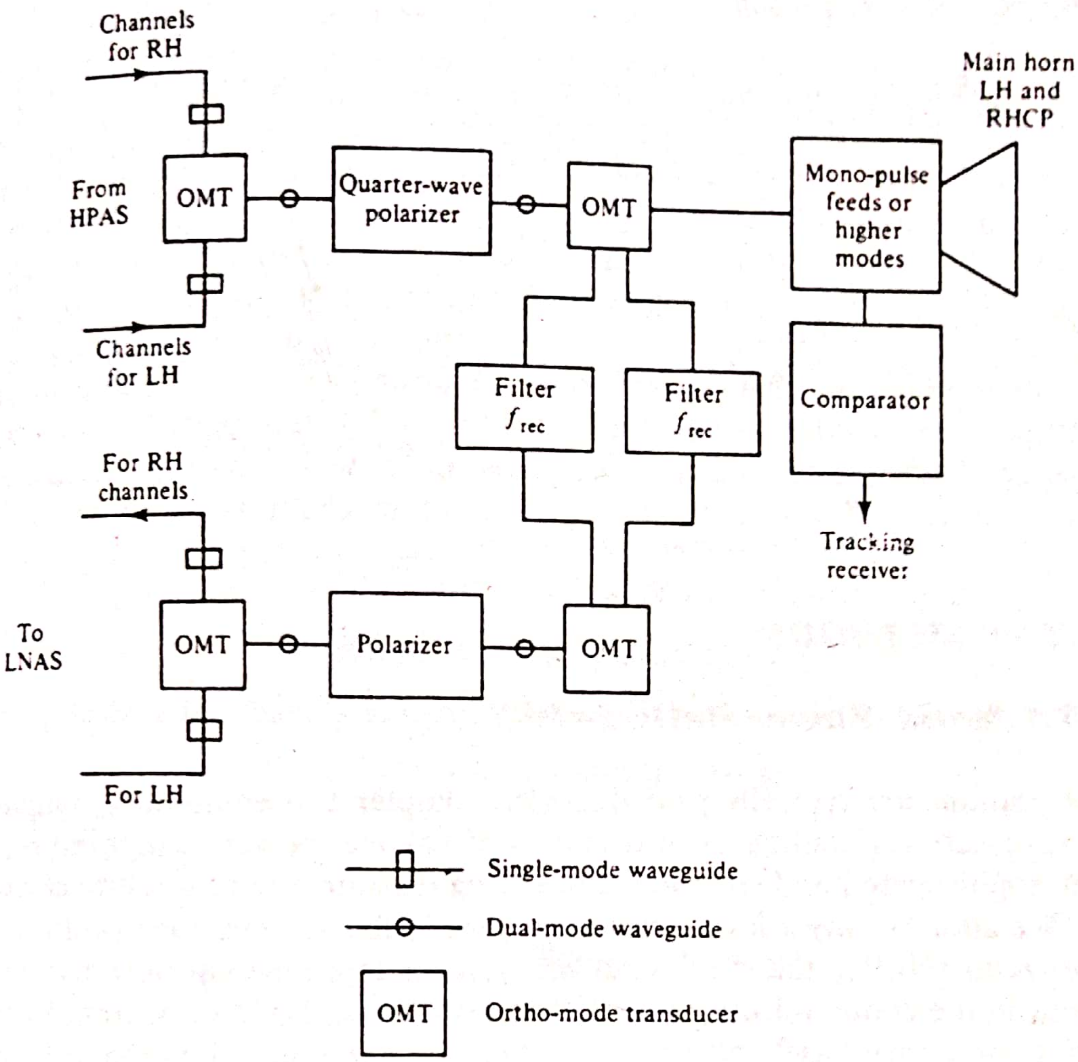


Figure 10-20 Two-polarization horn-feed system.

in on various carrier frequencies, depending on their source, and they must be tagged and put together with the corresponding outgoing circuit to make up a terrestrial circuit. This can be a complex process. The presence of video and data complicates matters further.

If the satellite transmission is single channel per carrier, it is necessary to bring each terrestrial carrier down to baseband before remodulation. The interfaces between terrestrial time-division and satellite frequency-division systems, and vice versa, are complicated and can be accomplished in a variety of ways. Television video signals must often be separated from order wire channels, program sound channels, cueing channels, and so on, and then matched up again at the proper point.

The underlying theory for understanding what must be done and how it must be done is presented in Chapters 7 and 8. Usually, in the systems engineering and programming planning phase it is only necessary to be alert to the problems and possibilities; the detailed design can be saved until later in the program.

10.5 PRIMARY POWER

Primary power systems vary from plain battery- or solar-cell-operated remote transmitters for data gathering to huge, combined commercial power and diesel generator systems for large stations. Most transmit and receive earth stations require some kind of "no-break" power system, that is emergency power to continue the communications during commercial power outages. Such power outages are frequent, even in highly organized industrial areas, if for no reason other than thunderstorms. The no-break transition derives its name from the necessity to make the change over from one power system to another without any interruption in service. Almost all systems today use batteries to effect this transition. Some systems have been devised in which motor generators store enough energy in flywheels to permit a smooth mechanical transition.

10.7 TEST METHODS

10.7.1 Noise Power Ratio (NPR)

Earth stations are typically provided with complex test equipment, ranging from that necessary for routine measurements of voltage, power, temperature, and so on, to sophisticated and specialized measurements unique to satellite communication. We address only a few of the latter class in this section. One of these is *noise power ratio* (NPR), the traditional measure of intermodulation noise for FDM systems in the communications field. The principle of NPR measurement involves loading the entire baseband spectrum, save for the one voice-frequency channel slot, with noise, simulating in total the loading of the system by actual voice traffic in all but that channel. Noise appearing in the unloaded slot is a manifestation of intermodulation. The ratio of that noise power to the per-channel loading noise power is the NPR. NPR is measured by a setup as shown in Figure 10-21. The system can be between any two points of interest. The noise generator *band* is limited by filters to the baseband, and the noise generator *level* is set to simulate full load according to the CCIR formulas

$$P = -15 + 10 \log N \text{ dBmO}, \qquad N \geq 240 \tag{10-20}$$

$$P = -1 + 4 \log N \text{ dBmO}, \qquad N < 240 \tag{10-21}$$

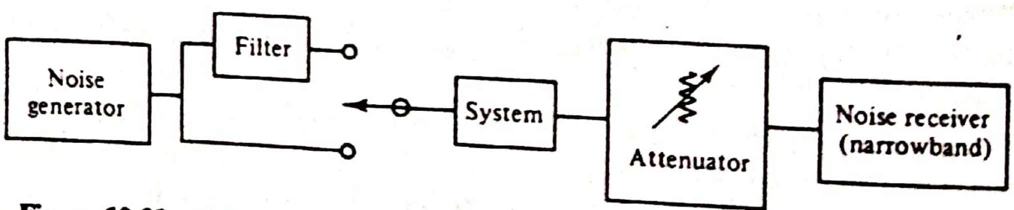


Figure 10-21 Noise power ratio test setup. Noise generator band is limited by filters to baseband; noise generator level is set to simulate full load.

These CCIR expressions give equivalent Gaussian noise to simulate N speech channels during busy hours (see Chapter 7). The filter has a stopband corresponding to the selected voice channel. The receive passband corresponds to that same channel. The difference between the noise receiver readings with the filter in and out is the NPR.

Measurement is typically made at low, center, and highest telephone channels. If they are carried out at different levels (and deviations for FM), they have a shape indicative of the system nonlinearities and frequency response. NPR is usually converted to an equivalent per-channel signal-to-noise ratio

$$\text{BWR} = 10 \log \frac{\text{baseband total bandwidth}}{\text{single channel bandwidth}} \tag{10-22}$$

$$\begin{aligned}\text{NLR} &= 10 \log \frac{\text{baseband noise test power}}{\text{test-tone power per channel}} \\ &= \text{dBmO of loading calculation}\end{aligned} \tag{10-23}$$

The equivalent baseband signal-to-noise ratio due to intermodulation is then

$$S/N = \text{NPR} + \text{BWR} - \text{NLR} \tag{10-24}$$

If

$$960 \text{ channels}$$

$$\text{NPR} = 55 \text{ dB in 3-kHz slot}$$

$$B = 4028 - 60$$

then

$$\text{BWR} = \frac{4028 - 60}{3} = 31.2 \text{ dB}$$

$$\text{NLR} = 10 \log 960 - 15 - 14.8 \text{ dBmO}$$

$$(S/N)_{\text{equiv}} = 71.4 \text{ dB}$$

10.7.2 The Measurement of G/T

System temperature T_S can be determined by conventional laboratory noise generator measurement of receiver noise figure and radiometric measurements of antenna temperature. The basic system parameter G/T_S also requires a knowledge of antenna gain, and as the antennas get larger, this characteristic is not so easy to get. The gain of smaller antennas, say less than 7 or 8 m, can be found from pattern measurements on a range or by comparison to a gain standard, but these methods are cumbersome and may be impractical for larger antennas.

Large earth stations, with antenna sizes up from 10 m, can sometimes use a carefully calibrated satellite signal to measure G/T_S. In effect, G/T_S is calculated from the link equation, knowing the other variables. This method is often used with intermediate-sized antennas (from 5 to 15 m).

An ingenious method has been developed for the measurement of G/T_S for large antennas using the known radio noise characteristics of stellar sources, usually called *radio stars*. These characteristics, particularly S, the flux density of the source in $W/m^2 \cdot Hz$, have been accurately measured by radio astronomers and are shown in Table 10-5.

The accurate implementation of the method requires care and many detailed corrections (Wait et al., 1974; Price, 1982) too lengthy for this section. Nonetheless, understanding the idea is important.

The basic measurement is that for *Y factor*. By definition, Y factor is the ratio of the output noise measured when the receiver is connected to a hot noise source (T_h), to the output noise measured when connected to a cold source (T_c). This is a familiar measurement in receiver work and is discussed at length in Mumford and Scheibe (1968). Using the equations of Chapter 6, it is straightforward to show that the receiver excess noise T_e is related to the Y factor by

$$T_w = \frac{T_h - YT_c}{Y - 1} \tag{10-25}$$

If the cold source is the normal sky and the hot source the radio star, the operating system temperature T_s, the sum of T_c and T_e given above, is easily shown to be

$$T_s = \frac{T_h - T_c}{Y - 1} = \frac{\Delta T_a}{Y - 1} \tag{10-26}$$

where ΔT_a is the increase in *antenna* temperature when changing from a radio source to the cold sky. The apparent increase in antenna temperature is related to the noise density increase by Boltzmann's constant k. If S is the randomly polarized flux density for the given star in $W/m^2 \cdot Hz$, only one polarization is received by an antenna of gain G, and a is a factor to allow for atmospheric loss ($a > 1$), then, from the universal antenna formula,

$$\Delta T_a = \frac{S}{2ak}\frac{G\lambda^2}{4\pi} \tag{10-27}$$

and

$$\frac{G}{T_s} = \frac{G(Y-1)}{\Delta T_a} = \frac{8\pi k}{S\lambda^2 a}(Y-1) \tag{10-28}$$

If a is the atmospheric absorption at the zenith, then at an elevation angle θ,

$$\frac{G}{T_s} = \frac{8\pi k}{S\lambda^2 aA}(Y-1)\sin\theta \tag{10-29}$$

If the stellar source is not randomly polarized, another correction factor is needed. Cassiopeia A, the most commonly used source, does not need this correction.

TABLE 10-5 RADIO STARS

	Location[a] Perturbation					Polarization Position Angle		Flux Density ($W/m^2\,Hz \times 10^{-26}$)	
Radio Source	Right Ascension	Declination	Shape	Size	Spectral Indix[b]	4170 MHz	6390 MHz	4000 MHz	6390 MHz
CasA	23h 21m 11.4s / 2.71s	58°31.9′ / 0.33′	Annular	Diameter, 4′	−0.792	—	—	1067[c]	774
TauA	05h 31m 30s / 3.61s	21°59.3′ / 0.4′	Elliptical (Gaussian)	Major axis, 4.3′ Minor axis, 2.7′	−0.287	5.7% / 143°	7.0% / 147°	679	604
CygA	19h 57m 44.5s / 2.08	40°35.8 / 0.16′	Dual point source	Separation 2′	−1.198	3.0% / 160°	5.7% / 148°	483	297

Source: Reprinted with permission (Miya, 1985).

[a] Perturbation per annum. Location $(1950 + x)$ = location (1950) + perturbation $\times x$.

[b] 1 to 16 GHz.

[c] Value for January 1965.

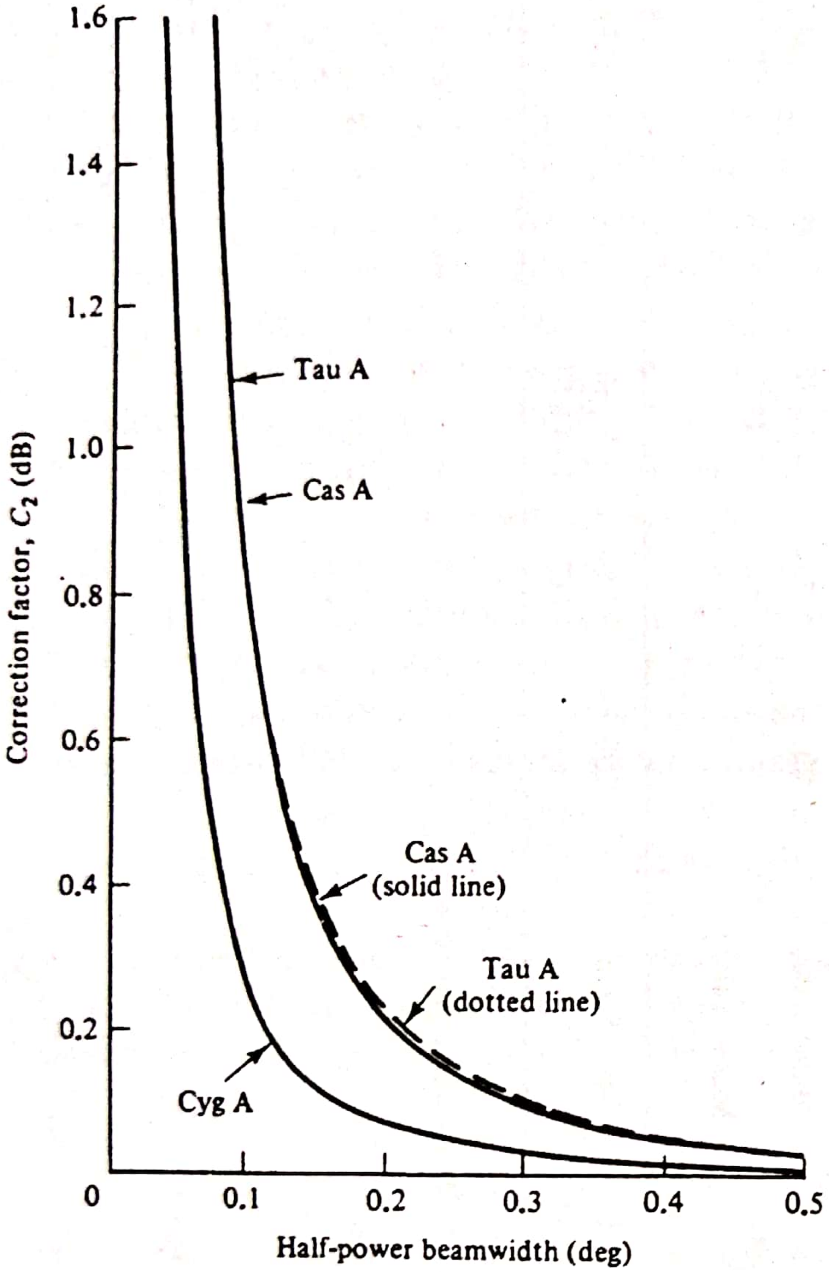


Figure 10-22 Correction factor for G/T measurement using extended sources. (Reprinted with permission from K. Miya, *Satellite Communications Engineering Technology*, 2nd ed. KDD Engineering & Consulting Co. 1985).

Some further correction may be necessary if the beamwidth of the antenna under test is narrow compared to the stellar radio source. An extended source of varying brightness can be considered as equivalent to a Rayleigh–Jeans blackbody radiator and the brightness integrated over the extent of the source. Correction factors can be arrived at (see Figure 10-22). Note that they are significant for narrowbeam antennas.

If the method is to be used at higher frequencies than given in Table 10-5, corrected values of S must be used. Wait et al. (1974) present data to justify assuming that S varies as $f^{-0.875}$.