Chapter

# 8

# The Earth Segment

## 8.1 Introduction

The earth segment of a satellite communications system consists of the transmit and receive earth stations. The simplest of these are the home *TV receive-only* (TVRO) systems, and the most complex are the terminal stations used for international communications networks. Also included in the earth segment are those stations which are on ships at sea, and commercial and military land and aeronautical mobile stations.

As mentioned in Chap. 7, earth stations that are used for logistic support of satellites, such as providing the *telemetry, tracking, and command* (TT&C) functions, are considered as part of the space segment.

## 8.2 Receive-Only Home TV Systems

Planned broadcasting directly to home TV receivers takes place in the Ku (12-GHz) band. This service is known as *direct broadcast satellite* (DBS) service. There is some variation in the frequency bands assigned to different geographic regions. In the Americas, for example, the downlink band is 12.2 to 12.7 GHz, as described in Sec. 1.4.

The comparatively large satellite receiving dishes [ranging in diameter from about 1.83 m (6 ft) to about 3-m (10 ft) in some locations], which may be seen in some "backyards" are used to receive downlink TV signals at C band (4 GHz). Originally such downlink signals were never intended for home reception but for network relay to commercial TV outlets (VHF and UHF TV broadcast stations and cable TV "head-end" studios). Equipment is now marketed for home reception of C-band signals, and some manufacturers provide dual C-band/Ku-band equipment. A single mesh type reflector may be used which focuses the signals into a dual feedhorn, which has two separate outputs, one for the C-band signals and one for the Ku-band signals. Much of television programming originates as *first generation signals*, also known as *master broadcast quality signals*. These are transmitted via satellite in the C band to the network head-end stations, where they are retransmitted as compressed digital signals to cable and direct broadcast satellite providers. One of the advantages claimed by sellers of C-band equipment for home reception is that there is no loss of quality compared with the compressed digital signals.

To take full advantage of C-band reception the home antenna has to be steerable to receive from different satellites, usually by means of a polar mount as described in Sec. 3.3. Another of the advantages, claimed for home C-band systems, is the larger number of satellites available for reception compared to what is available for direct broadcast satellite systems. Although many of the C-band transmissions are scrambled, there are free channels that can be received, and what are termed "wild feeds." These are also free, but unannounced programs, of which details can be found in advance from various publications and Internet sources. C-band users can also subscribe to pay TV channels, and another advantage claimed is that subscription services are cheaper than DBS or cable because of the multiple-source programming available.

The most widely advertised receiving system for C-band system appears to be 4DTV manufactured by Motorola. This enables reception of:

1. Free, analog signals and "wild feeds"
2. VideoCipher ll plus subscription services
3. Free DigiCipher 2 services
4. Subscription DigiCipher 2 services

VideoCipher is the brand name for the equipment used to scramble analog TV signals. DigiCipher 2 is the name given to the digital compression standard used in digital transmissions. General information about C-band TV reception will be found at http://orbitmagazine.com/ (Orbit, 2005) and http://www.satellitetheater.com/ (Satellite Theater systems, 2005).

The major differences between the Ku-band and the C-band receive-only systems lies in the frequency of operation of the outdoor unit and the fact that satellites intended for DBS have much higher *equivalent isotropic radiated power* (EIRP), as shown in Table 1.4. As already mentioned C-band antennas are considerably larger than DBS antennas. For clarity, only the Ku-band system is described here.

Figure 8.1 shows the main units in a home terminal DBS TV receiving system. Although there will be variations from system to system, the diagram covers the basic concept for analog [*frequency modulated* (FM)] TV. Direct-to-home digital TV, which is well on the way to replacing analog systems, is discussed in Chap. 16. However, the outdoor unit is similar for both systems.

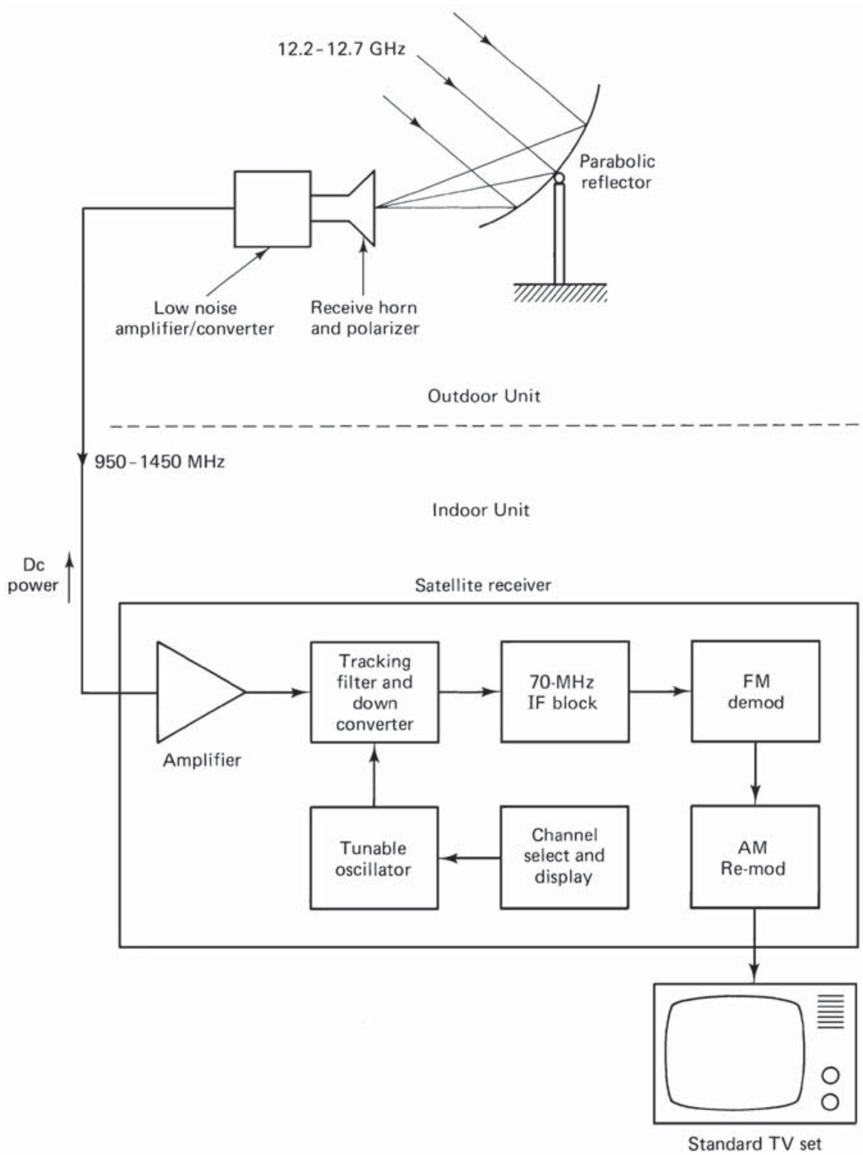


**Figure 8.1** Block diagram showing a home terminal for DBS TV/FM reception.

### 8.2.1 The outdoor unit

This consists of a receiving antenna feeding directly into a low-noise amplifier/converter combination. A parabolic reflector is generally used, with the receiving horn mounted at the focus. A common design is to have the focus directly in front of the reflector, but for better interference rejection, an offset feed may be used as shown.

Huck and Day (1979) have shown that satisfactory reception can be achieved with reflector diameters in the range 0.6 to 1.6 m (1.97–5.25 ft), and the two nominal sizes often quoted are 0.9 m (2.95 ft) and 1.2 m (3.94 ft). By contrast, the reflector diameter for 4-GHz reception can range from 1.83 m (6 ft) to 3 m (10 ft). As noted in Sec. 6.13, the gain of a parabolic dish is proportional to $(D/\lambda)^2$. Comparing the gain of a 3-m dish at 4 GHz with a 1-m dish at 12 GHz, the ratio $D/\lambda$ equals 40 in each case, so the gains will be about equal. Although the free-space losses are much higher at 12 GHz compared with 4 GHz, as described in Chap. 12, a higher-gain receiving antenna is not needed because the DBS operate at a much higher EIRP, as shown in Table 1.4.

The downlink frequency band of 12.2 to 12.7 GHz spans a range of 500 MHz, which accommodates 32 TV/FM channels, each of which is 24-MHz wide. Obviously, some overlap occurs between channels, but these are alternately polarized *left-hand circular* (LHC) and *right-hand circular* (RHC) or vertical/horizontal, to reduce interference to acceptable levels. This is referred to as *polarization interleaving*. A polarizer that may be switched to the desired polarization from the indoor control unit is required at the receiving horn.

The receiving horn feeds into a *low-noise converter* (LNC) or possibly a combination unit consisting of a *low-noise amplifier* (LNA) followed by a converter. The combination is referred to as an LNB, for *low-noise block*. The LNB provides gain for the broadband 12-GHz signal and then converts the signal to a lower frequency range so that a low-cost coaxial cable can be used as feeder to the indoor unit. The standard frequency range of this downconverted signal is 950 to 1450 MHz, as shown in Fig. 8.1. The coaxial cable, or an auxiliary wire pair, is used to carry dc power to the outdoor unit. Polarization-switching control wires are also required.

The low-noise amplification must be provided at the cable input in order to maintain a satisfactory signal-to-noise ratio. An LNA at the indoor end of the cable would be of little use, because it would also amplify the cable thermal noise. Single-to-noise ratio is discussed in more detail in Sec. 12.5. Of course, having to mount the LNB outside means that it must be able to operate over a wide range of climatic conditions, and homeowners may have to contend with the added problems of vandalism and theft.

### 8.2.2 The indoor unit for analog (FM) TV

The signal fed to the indoor unit is normally a wideband signal covering the range 950 to 1450 MHz. This is amplified and passed to a tracking filter which selects the desired channel, as shown in Fig. 8.1.

As previously mentioned, polarization interleaving is used, and only half the 32 channels will be present at the input of the indoor unit for any one setting of the antenna polarizer. This eases the job of the tracking filter, since alternate channels are well separated in frequency.

The selected channel is again downconverted, this time from the 950- to 1450-MHz range to a fixed intermediate frequency, usually 70 MHz although other values in the *very high frequency* (VHF) range are also used. The 70-MHz amplifier amplifies the signal up to the levels required for demodulation. A major difference between DBS TV and conventional TV is that with DBS, frequency modulation is used, whereas with conventional TV, amplitude modulation in the form of *vestigial single sideband* (VSSB) is used. The 70-MHz, FM *intermediate frequency* (IF) carrier therefore must be demodulated, and the baseband information used to generate a VSSB signal which is fed into one of the VHF/UHF channels of a standard TV set.

A DBS receiver provides a number of functions not shown on the simplified block diagram of Fig. 8.1. The demodulated video and audio signals are usually made available at output jacks. Also, as described in Sec. 13.3, an energy-dispersal waveform is applied to the satellite carrier to reduce interference, and this waveform has to be removed in the DBS receiver. Terminals also may be provided for the insertion of IF filters to reduce interference from terrestrial TV networks, and a descrambler also may be necessary for the reception of some programs. The indoor unit for digital TV is described in Chap. 16.

## 8.3 Master Antenna TV System

A *master antenna TV* (MATV) system is used to provide reception of DBS TV/FM channels to a small group of users, for example, to the tenants in an apartment building. It consists of a single outdoor unit (antenna and LNA/C) feeding a number of indoor units, as shown in Fig. 8.2. It is basically similar to the home system already described, but with each user having access to all the channels independently of the other users. The advantage is that only one outdoor unit is required, but as shown, separate LNA/Cs and feeder cables are required for each sense of polarization. Compared with the single-user system, a larger antenna is also required (2- to 3-m diameter) in order to maintain a good signal-to-noise ratio at all the indoor units.

Where more than a few subscribers are involved, the distribution system used is similar to the *community antenna* (CATV) system described in the following section.

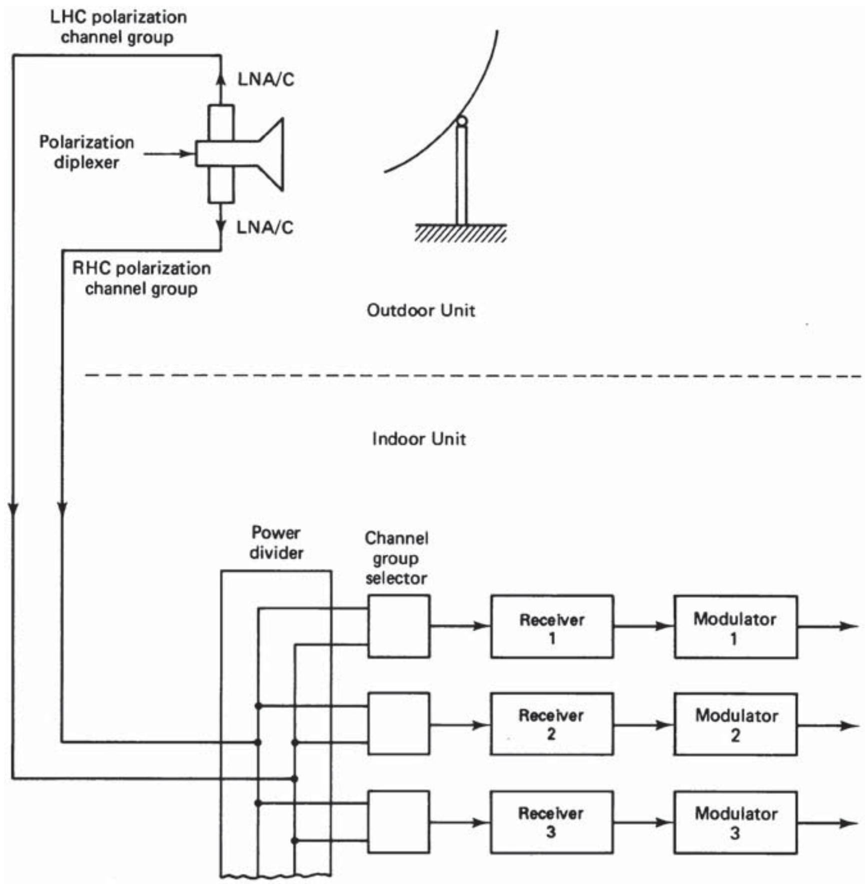


**Figure 8.2** One possible arrangement for a master antenna TV (MATV) system.

## 8.4 Community Antenna TV System

The CATV system employs a single outdoor unit, with separate feeds available for each sense of polarization, like the MATV system, so that all channels are made available simultaneously at the indoor receiver. Instead of having a separate receiver for each user, all the carriers are demodulated in a common receiver-filter system, as shown in Fig. 8.3. The channels are then combined into a standard multiplexed signal for transmission over cable to the subscribers.

In remote areas where a cable distribution system may not be installed, the signal can be rebroadcast from a low-power VHF TV transmitter. Figure 8.4 shows a remote TV station which employs an

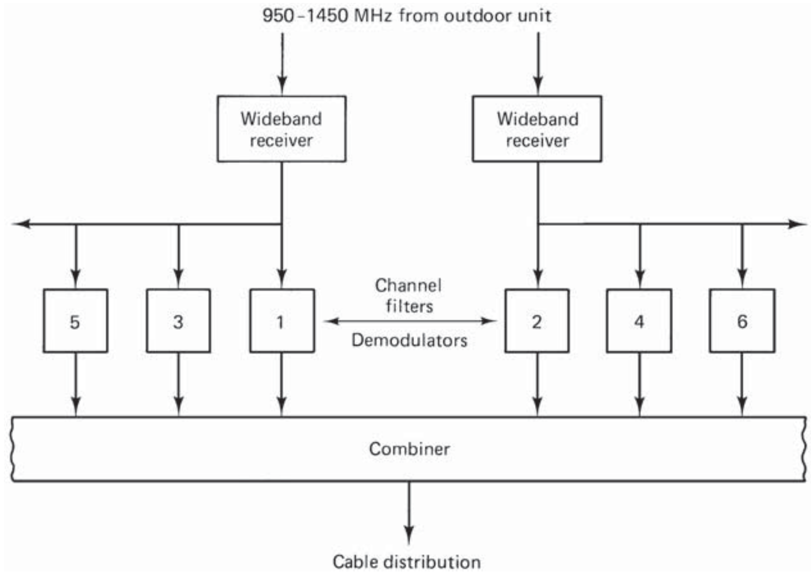


**Figure 8.3** One possible arrangement for the indoor unit of a community antenna TV (CATV) system.

8-m (26.2-ft) antenna for reception of the satellite TV signal in the C band.

With the CATV system, local programming material also may be distributed to subscribers, an option which is not permitted in the MATV system.

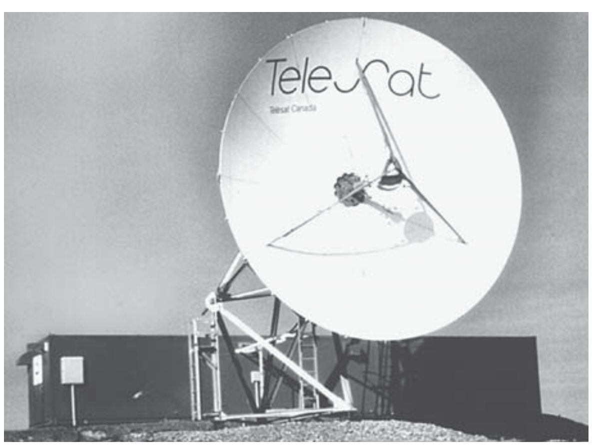


**Figure 8.4** Remote television station. (*Courtesy of Telesat Canada, 1983.*)

## 8.5 Transmit-Receive Earth Stations

In the previous sections, receive-only TV stations are described. Obviously, somewhere a transmit station must complete the uplink to the satellite. In some situations, a transmit-only station is required, for example, in relaying TV signals to the remote TVRO stations already described. Transmit-receive stations provide both functions and are required for telecommunications traffic generally, including network TV. The uplink facilities for digital TV are highly specialized and are covered in Chap. 16.

The basic elements for a redundant earth station are shown in Fig. 8.5. As mentioned in connection with transponders in Sec. 7.7.1, redundancy means that certain units are duplicated. A duplicate, or redundant, unit is automatically switched into a circuit to replace a corresponding unit that has failed. Redundant units are shown by dashed lines in Fig. 8.5.

The block diagram is shown in more detail in Fig. 8.6, where, for clarity, redundant units are not shown. Starting at the bottom of the diagram, the first block shows the interconnection equipment required between satellite station and the terrestrial network. For the purpose of explanation, telephone traffic will be assumed. This may consist of a number of telephone channels in a multiplexed format. Multiplexing is a method of grouping telephone channels together, usually in basic groups of 12, without mutual interference. It is described in detail in Chaps. 9 and 10.

It may be that groupings different from those used in the terrestrial network are required for satellite transmission, and the next block shows the multiplexing equipment in which the reformatting is carried out. Following along the transmit chain, the multiplexed signal is modulated onto a carrier wave at an intermediate frequency, usually 70 MHz. Parallel IF stages are required, one for each microwave carrier to be transmitted. After amplification at the 70-MHz IF, the modulated signal is then upconverted to the required microwave carrier frequency. A number of carriers may be transmitted simultaneously, and although these are at different frequencies they are generally specified by their nominal frequency, for example, as 6-GHz or 14-GHz carriers.

It should be noted that the individual carriers may be multidestination carriers. This means that they carry traffic destined for different stations. For example, as part of its load, a microwave carrier may have telephone traffic for Boston and New York. The same carrier is received at both places, and the designated traffic sorted out by filters at the receiving earth station.

Referring again to the block diagram of Fig. 8.6, after passing through the upconverters, the carriers are combined, and the resulting wideband signal is amplified. The wideband power signal is fed to the antenna

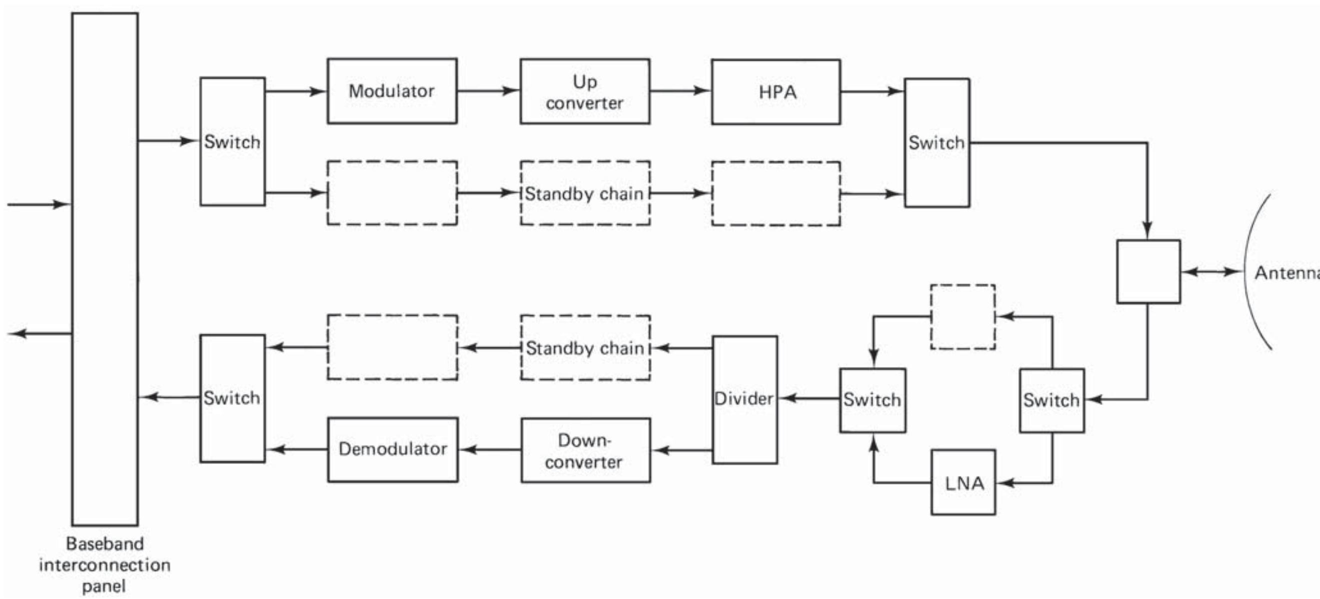


**Figure 8.5** Basic elements of a redundant earth station. (*Courtesy of Telesat Canada, 1983.*)

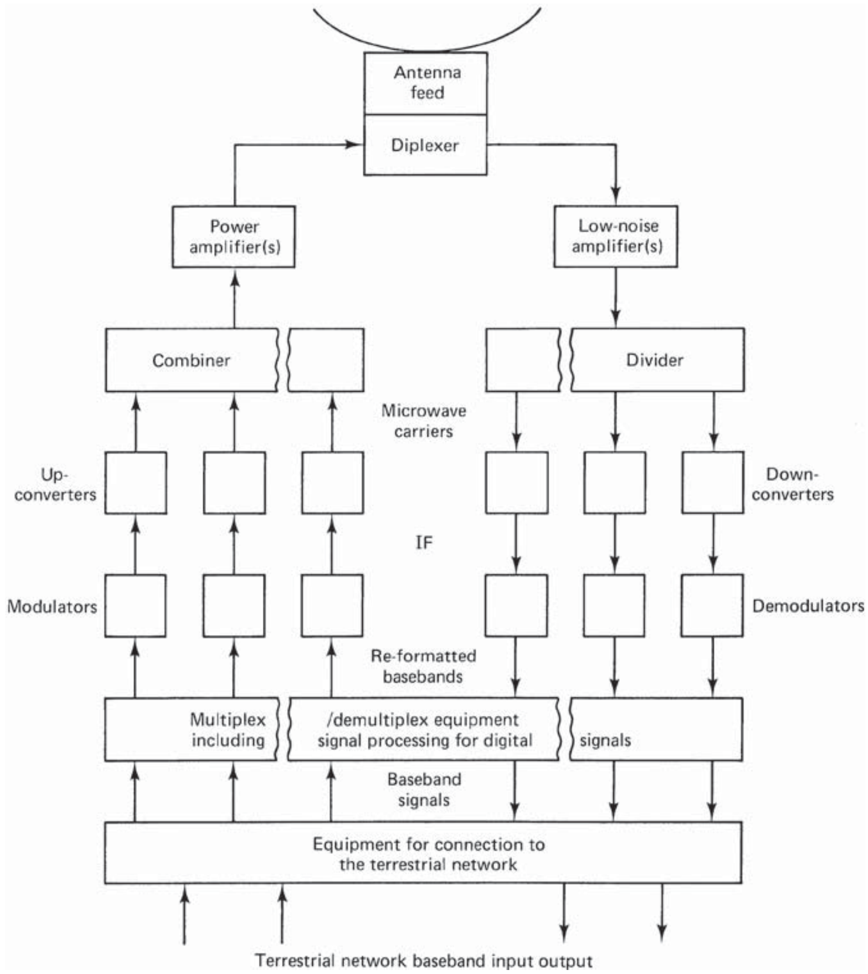


**Figure 8.6** More detailed block diagram of a transmit-receive earth station.

through a diplexer, which allows the antenna to handle transmit and receive signals simultaneously.

The station's antenna functions in both, the transmit and receive modes, but at different frequencies. In the C band, the nominal uplink, or transmit, frequency is 6 GHz and the downlink, or receive, frequency is nominally 4 GHz. In the Ku band, the uplink frequency is nominally 14 GHz, and the downlink, 12 GHz. High-gain antennas are employed in both bands, which also means narrow antenna beams. A narrow beam is necessary to prevent interference between neighboring satellite links. In the case of C band, interference to and from terrestrial microwave links also must be avoided. Terrestrial microwave links do not operate at Ku-band frequencies.

In the receive branch (the right-hand side of Fig. 8.6), the incoming wideband signal is amplified in an LNA and passed to a divider network, which separates out the individual microwave carriers. These are each downconverted to an IF band and passed on to the multiplex block, where the multiplexed signals are reformatted as required by the terrestrial network.

It should be noted that, in general, the signal traffic flow on the receive side will differ from that on the transmit side. The incoming microwave carriers will be different in number and in the amount of traffic carried, and the multiplexed output will carry telephone circuits not necessarily carried on the transmit side.

A number of different classes of earth stations are available, depending on the service requirements. Traffic can be broadly classified as heavy route, medium route, and thin route. In a thin-route circuit, a transponder channel (36 MHz) may be occupied by a number of single carriers, each associated with its own voice circuit. This mode of operation is known as *single carrier per channel* (SCPC), a multiple-access mode which is discussed further in Chap. 14. Antenna sizes range from 3.6 m (11.8 ft) for transportable stations up to 30 m (98.4 ft) for a main terminal.

A medium-route circuit also provides multiple access, either on the basis of *frequency-division multiple access* (FDMA) or *time-division multiple access* (TDMA), multiplexed baseband signals being carried in either case. These access modes are also described in detail in Chap. 14. Antenna sizes range from 30 m (89.4 ft) for a main station to 10 m (32.8 ft) for a remote station.

In a 6/4-GHz heavy-route system, each satellite channel (bandwidth 36 MHz) is capable of carrying over 960 one-way voice circuits simultaneously or a single-color analog TV signal with associated audio (in some systems two analog TV signals can be accommodated). Thus the transponder channel for a heavy-route circuit carries one large-bandwidth signal, which may be TV or multiplexed telephony. The antenna diameter for a heavy-route circuit is at least 30 m (98.4 ft). For international operation such antennas are designed to the INTELSAT specifications for a Standard A earth station (Intelsat, 1982). Figure 8.7 shows a photograph of a 32-m (105-ft) Standard A earth station antenna.

It will be appreciated that for these large antennas, which may weigh in the order of 250 tons, the foundations must be very strong and stable. Such large diameters automatically mean very narrow beams, and therefore, any movement which would deflect the beam unduly must be avoided. Where snow and ice conditions are likely to be encountered, built-in heaters are required. For the antenna shown in Fig. 8.7, deicing

**Figure 8.7** Standard-A (C-band 6/4 GHz) 32-m antenna. (*Courtesy of TIW Systems, Inc., Sunnydale, CA.*)

heaters provide reflector surface heat of 40W/ft$^2$ for the main reflectors and subreflectors, and 3000 W for the azimuth wheels.

Although these antennas are used with geostationary satellites, some drift in the satellite position does occur, as shown in Chap. 3. This, combined with the very narrow beams of the larger earth station antennas, means that some provision must be made for a limited degree of tracking. Step adjustments in azimuth and elevation may be made, under computer control, to maximize the received signal.

The continuity of the primary power supply is another important consideration in the design of transmit-receive earth stations. Apart from the smallest stations, power backup in the form of multiple feeds from the commercial power source and/or batteries and generators is provided. If the commercial power fails, batteries immediately take over with no interruption. At the same time, the standby generators start up, and once they are up to speed they automatically take over from the batteries.

## 8.6 Problems and Exercises

**8.1.** Explain what is meant by DBS service. How does this differ from the home reception of satellite TV signals in the C band?

**8.2.** Explain what is meant by *polarization interleaving*. On a frequency axis, draw to scale the channel allocations for the 32 TV channels in the Ku band, showing how polarization interleaving is used in this.

**8.3.** Why is it desirable to downconvert the satellite TV signal received at the antenna?

**8.4.** Explain why the LNA in a satellite receiving system is placed at the antenna end of the feeder cable.

**8.5.** With the aid of a block schematic, briefly describe the functioning of the indoor receiving unit of a satellite TV/FM receiving system intended for home reception.

**8.6.** In most satellite TV receivers the first IF band is converted to a second, fixed IF. Why is this second frequency conversion required?

**8.7.** For the standard home television set to function in a satellite TV/FM receiving system, a demodulator/remodulator unit is needed. Explain why.

**8.8.** Describe and compare the MATV and the CATV systems.

**8.9.** Explain what is meant by the term *redundant earth station*.

**8.10.** With the aid of a block schematic, describe the functioning of a transmit-receive earth station used for telephone traffic. Describe a multidestination carrier.

## References

Huck, R. W., and J. W. B. Day. 1979. "Experience in Satellite Broadcasting Applications with CTS/HERMES." *XIth International TV Symposium*, Montreux, 27 May–1 June.
INTELSAT. 1982. "Standard A Performance Characteristics of Earth Stations in the INTELSAT IV, IVA, and V Systems." BG-28-72E M/6/77.
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Chapter

# 12

# The Space Link

## 12.1 Introduction

This chapter describes how the link-power budget calculations are made. These calculations basically relate two quantities, the transmit power and the receive power, and show in detail how the difference between these two powers is accounted for.

Link-budget calculations are usually made using decibel or decilog quantities. These are explained in App. G. In this text [square] brackets are used to denote decibel quantities using the basic power definition. Where no ambiguity arises regarding the units, the abbreviation dB is used. For example, Boltzmann's constant is given as −228.6 dB, although, strictly speaking, this should be given as −228.6 decilogs relative to 1 J/K. Where it is desirable to show the reference unit, this is indicated in the abbreviation, for example, dBHz means decibels relative to 1 Hz.

## 12.2 Equivalent Isotropic Radiated Power

A key parameter in link-budget calculations is the *equivalent isotropic radiated power*, conventionally denoted as EIRP. From Eqs. (6.4) and (6.5), the maximum power flux density at some distance $r$ from a transmitting antenna of gain $G$ is

$$\Psi_M = \frac{GP_S}{4\pi r^2} \qquad (12.1)$$

An isotropic radiator with an input power equal to $GP_S$ would produce the same flux density. Hence, this product is referred to as the EIRP, or

$$\text{EIRP} = GP_S \qquad (12.2)$$

EIRP is often expressed in decibels relative to 1 W, or dBW. Let $P_S$ be in watts; then

$$[\text{EIRP}] = [P_S] + [G] \text{ dBW} \tag{12.3}$$

where $[P_S]$ is also in dBW and $[G]$ is in dB.

**Example 12.1** A satellite downlink at 12 GHz operates with a transmit power of 6 W and an antenna gain of 48.2 dB. Calculate the EIRP in dBW.

**Solution**

$$[\text{EIRP}] = 10 \log\left(\frac{6\text{W}}{1\text{W}}\right) + 48.2$$

$$= \underline{\underline{56 \text{ dBW}}}$$

For a paraboloidal antenna, the isotropic power gain is given by Eq. (6.32). This equation may be rewritten in terms of frequency, since this is the quantity which is usually known.

$$G = \eta(10.472fD)^2 \tag{12.4}$$

where $f$ is the carrier frequency in GHz, $D$ is the reflector diameter in m, and $\eta$ is the aperture efficiency. A typical value for aperture efficiency is 0.55, although values as high as 0.73 have been specified (Andrew Antenna, 1985).

With the diameter $D$ in feet and all other quantities as before, the equation for power gain becomes

$$G = \eta(3.192fD)^2 \tag{12.5}$$

**Example 12.2** Calculate the gain in decibels of a 3-m paraboloidal antenna operating at a frequency of 12 GHz. Assume an aperture efficiency of 0.55.

**Solution**

$$G = 0.55 \times (10.472 \times 12 \times 3)^2 \cong 78168$$

Hence,

$$[G] = 10 \log 78168 = 48.9 \text{ dB}$$

## 12.3 Transmission Losses

The [EIRP] may be thought of as the power input to one end of the transmission link, and the problem is to find the power received at the other end. Losses will occur along the way, some of which are constant.

Other losses can only be estimated from statistical data, and some of these are dependent on weather conditions, especially on rainfall.

The first step in the calculations is to determine the losses for *clear-weather* or *clear-sky conditions*. These calculations take into account the losses, including those calculated on a statistical basis, which do not vary significantly with time. Losses which are weather-related, and other losses which fluctuate with time, are then allowed for by introducing appropriate *fade margins* into the transmission equation.

### 12.3.1 Free-space transmission

As a first step in the loss calculations, the power loss resulting from the spreading of the signal in space must be determined. This calculation is similar for the uplink and the downlink of a satellite circuit. Using Eqs. (12.1) and (12.2) gives the power-flux density at the receiving antenna as

$$\Psi_M = \frac{\text{EIRP}}{4\pi r^2} \tag{12.6}$$

The power delivered to a matched receiver is this power-flux density multiplied by the effective aperture of the receiving antenna, given by Eq. (6.15). The received power is therefore

$$\begin{aligned} P_R &= \Psi_M A_{\text{eff}} \\ &= \frac{\text{EIRP}}{4\pi r^2}\frac{\lambda^2 G_R}{4\pi} \\ &= (\text{EIRP})(G_R)\left(\frac{\lambda}{4\pi r}\right)^2 \end{aligned} \tag{12.7}$$

Recall that $r$ is the distance, or range, between the transmit and receive antennas and $G_R$ is the isotropic power gain of the receiving antenna. The subscript $R$ is used to identify the receiving antenna.

The right-hand side of Eq. (12.7) is separated into three terms associated with the transmitter, receiver, and free space, respectively. In decibel notation, the equation becomes

$$[P_R] = [\text{EIRP}] + [G_R] - 10\log\left(\frac{4\pi r}{\lambda}\right)^2 \tag{12.8}$$

The received power in dBW is therefore given as the sum of the transmitted EIRP in dBW plus the receiver antenna gain in dB minus a third term, which represents the free-space loss in decibels. The free-space loss component in decibels is given by

$$[\text{FSL}] = 10\log\left(\frac{4\pi r}{\lambda}\right)^2 \tag{12.9}$$

Normally, the frequency rather than wavelength will be known, and the substitution $\lambda = c/f$ can be made, where $c = 10^8$ m/s. With frequency in megahertz and distance in kilometers, it is left as an exercise for the student to show that the free-space loss is given by

$$[\text{FSL}] = 32.4 + 20 \log r + 20 \log f \qquad (12.10)$$

Equation (12.8) can then be written as

$$[P_R] = [\text{EIRP}] + [G_R] - [\text{FSL}] \qquad (12.11)$$

The received power $[P_R]$ will be in dBW when the [EIRP] is in dBW, and [FSL] in dB. Equation (12.9) is applicable to both the uplink and the downlink of a satellite circuit, as will be shown in more detail shortly.

**Example 12.3** The range between a ground station and a satellite is 42,000 km. Calculate the free-space loss at a frequency of 6 GHz.

**Solution**

$$[\text{FSL}] = 32.4 + 20 \log 42{,}000 + 20 \log 6000 = \underline{\underline{200.4 \text{ dB}}}$$

This is a very large loss. Suppose that the [EIRP] is 56 dBW (as calculated in Example 12.1 for a radiated power of 6 W) and the receive antenna gain is 50 dB. The receive power would be $56 + 50 - 200.4 = -94.4$ dBW. This is 355 pW. It also may be expressed as $-64.4$ dBm, which is 64.4 dB below the 1-mW reference level.

Equation (12.11) shows that the received power is increased by increasing antenna gain as expected, and Eq. (6.32) shows that antenna gain is inversely proportional to the square of the wavelength. Hence, it might be thought that increasing the frequency of operation (and therefore decreasing wavelength) would increase the received power. However, Eq. (12.9) shows that the free-space loss is also inversely proportional to the square of the wavelength, so these two effects cancel. It follows, therefore, that for a constant EIRP, the received power is independent of frequency of operation.

If the transmit power is a specified constant, rather than the EIRP, then the received power will increase with increasing frequency for given antenna dish sizes at the transmitter and receiver. It is left as an exercise for the student to show that under these conditions the received power is directly proportional to the square of the frequency.

### 12.3.2 Feeder losses

Losses will occur in the connection between the receive antenna and the receiver proper. Such losses will occur in the connecting waveguides, filters, and couplers. These will be denoted by RFL, or [RFL] dB, for *receiver* *feeder losses*. The [RFL] values are added to [FSL] in Eq. (12.11). Similar losses will occur in the filters, couplers, and waveguides connecting the transmit antenna to the *high-power amplifier* (HPA) output. However, provided that the EIRP is stated, Eq. (12.11) can be used without knowing the transmitter feeder losses. These are needed only when it is desired to relate EIRP to the HPA output, as described in Secs. 12.7.4 and 12.8.2.

### 12.3.3 Antenna misalignment losses

When a satellite link is established, the ideal situation is to have the earth station and satellite antennas aligned for maximum gain, as shown in Fig. 12.1*a*. There are two possible sources of off-axis loss, one at the satellite and one at the earth station, as shown in Fig. 12.1*b*.

The off-axis loss at the satellite is taken into account by designing the link for operation on the actual satellite antenna contour; this is described in more detail in later sections. The off-axis loss at the earth station is referred to as the *antenna pointing loss*. Antenna pointing losses are usually only a few tenths of a decibel; typical values are given in Table 12.1.

In addition to pointing losses, losses may result at the antenna from misalignment of the polarization direction (these are in addition to the polarization losses described in Chap. 5). The polarization misalignment losses are usually small, and it will be assumed that the antenna misalignment losses, denoted by [AML], include both pointing and polarization losses resulting from antenna misalignment. It should be noted

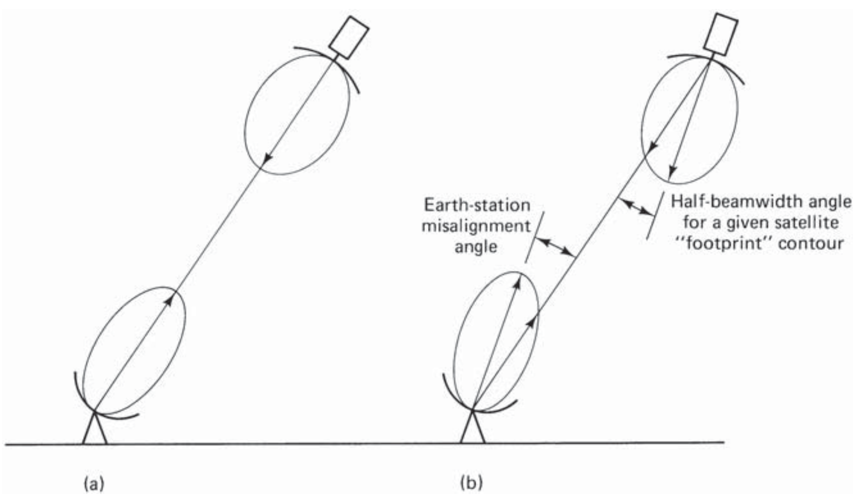



**Figure 12.1** (*a*) Satellite and earth-station antennas aligned for maximum gain; (*b*) earth station situated on a given satellite "footprint," and earth-station antenna misaligned.

**TABLE 12.1 Atmospheric Absorption Loss and Satellite Pointing Loss for Cities and Communities in the Province of Ontario**

| Location | Atmospheric absorption dB, summer | Satellite antenna pointing loss, dB | |
|---|---|---|---|
| | | ¼ Canada coverage | ½ Canada coverage |
| Cat Lake | 0.2 | 0.5 | 0.5 |
| Fort Severn | 0.2 | 0.9 | 0.9 |
| Geraldton | 0.2 | 0.2 | 0.1 |
| Kingston | 0.2 | 0.5 | 0.4 |
| London | 0.2 | 0.3 | 0.6 |
| North Bay | 0.2 | 0.3 | 0.2 |
| Ogoki | 0.2 | 0.4 | 0.3 |
| Ottawa | 0.2 | 0.6 | 0.2 |
| Sault Ste. Marie | 0.2 | 0.1 | 0.3 |
| Sioux Lookout | 0.2 | 0.4 | 0.3 |
| Sudbury | 0.2 | 0.3 | 0.2 |
| Thunder Bay | 0.2 | 0.3 | 0.2 |
| Timmins | 0.2 | 0.5 | 0.2 |
| Toronto | 0.2 | 0.3 | 0.4 |
| Windsor | 0.2 | 0.5 | 0.8 |

SOURCE: Telesat Canada Design Workbook.

that the antenna misalignment losses have to be estimated from statistical data, based on the errors actually observed for a large number of earth stations, and of course, the separate antenna misalignment losses for the uplink and the downlink must be taken into account.

### 12.3.4 Fixed atmospheric and ionospheric losses

Atmospheric gases result in losses by absorption, as described in Sec. 4.2 and by Eq. (4.1). These losses usually amount to a fraction of a decibel, and in subsequent calculations, the decibel value will be denoted by [AA]. Values obtained for some locations in the Province of Ontario, Canada, are shown in Table 12.1. Also, as discussed in Sec. 5.5, the ionosphere introduces a depolarization loss given by Eq. (5.19), and in subsequent calculations, the decibel value for this will be denoted by [PL].

## 12.4 The Link-Power Budget Equation

As mentioned at the beginning of Sec. 12.3, the [EIRP] can be considered as the input power to a transmission link. Now that the losses for the link have been identified, the power at the receiver, which is the power output of the link, may be calculated simply as [EIRP] − [LOSSES] + [$G_R$], where the last quantity is the receiver antenna gain. Note carefully that decibel addition must be used.

The major source of loss in any ground-satellite link is the free-space spreading loss [FSL], as shown in Sec. 12.3.1, where Eq. (12.13) is the basic link-power budget equation taking into account this loss only. However, the other losses also must be taken into account, and these are simply added to [FSL]. The losses for clear-sky conditions are

$$[\text{LOSSES}] = [\text{FSL}] + [\text{RFL}] + [\text{AML}] + [\text{AA}] + [\text{PL}] \quad (12.12)$$

The decibel equation for the received power is then

$$[P_R] = [\text{EIRP}] + [G_R] - [\text{LOSSES}] \quad (12.13)$$

where [PR] = received power, dBW
[EIRP] = equivalent isotropic radiated power, dBW
[FSL] = free-space spreading loss, dB
[RFL] = receiver feeder loss, dB
[AML] = antenna misalignment loss, dB
[AA] = atmospheric absorption loss, dB
[PL] = polarization mismatch loss, dB

**Example 12.4** A satellite link operating at 14 GHz has receiver feeder losses of 1.5 dB and a free-space loss of 207 dB. The atmospheric absorption loss is 0.5 dB, and the antenna pointing loss is 0.5 dB. Depolarization losses may be neglected. Calculate the total link loss for clear-sky conditions.

**Solution** The total link loss is the sum of all the losses:

$$[\text{LOSSES}] = [\text{FSL}] + [\text{RFL}] + [\text{AA}] + [\text{AML}]$$

$$= 207 + 1.5 + 0.5 + 0.5$$

$$= \underline{\underline{209.5 \text{ dB}}}$$

## 12.5 System Noise

It is shown in Sec. 12.3 that the receiver power in a satellite link is very small, on the order of picowatts. This by itself would be no problem because amplification could be used to bring the signal strength up to an acceptable level. However, electrical noise is always present at the input, and unless the signal is significantly greater than the noise, amplification will be of no help because it will amplify signal and noise to the same extent. In fact, the situation will be worsened by the noise added by the amplifier.

The major source of electrical noise in equipment is that which arises from the random thermal motion of electrons in various resistive and active devices in the receiver. Thermal noise is also generated in the lossy components of antennas, and thermal-like noise is picked up by the antennas as radiation.

The available noise power from a thermal noise source is given by

$$P_N = kT_N B_N \tag{12.14}$$

Here, $T_N$ is known as the equivalent noise temperature, $B_N$ is the equivalent noise bandwidth, and $k = 1.38 \times 10^{-23}$ J/K is Boltzmann's constant. With the temperature in kelvins and bandwidth in hertz, the noise power will be in watts. The noise power bandwidth is always wider than the $-3$-dB bandwidth determined from the amplitude-frequency response curve, and a useful rule of thumb is that the noise bandwidth is equal to 1.12 times the $-3$-dB bandwidth, or $B_N \approx 1.12 \times B_{-3\mathrm{dB}}$. The bandwidths here are in hertz (or a multiple such as MHz).

The main characteristic of thermal noise is that it has a *flat frequency spectrum*; that is, the noise power per unit bandwidth is a constant. The noise power per unit bandwidth is termed the *noise power spectral density*. Denoting this by $N_0$, then from Eq. (12.14),

$$N_0 = \frac{P_N}{B_N} = kT_N \text{ J} \tag{12.15}$$

The noise temperature is directly related to the physical temperature of the noise source but is not always equal to it. This is discussed more fully in the following sections. The noise temperatures of various sources which are connected together can be added directly to give the total noise.

**Example 12.5** An antenna has a noise temperature of 35 K and is matched into a receiver which has a noise temperature of 100 K. Calculate (*a*) the noise power density and (*b*) the noise power for a bandwidth of 36 MHz.

**Solution**

$$(a)\ N_0 = (35 + 100) \times 1.38 \times 10^{-23} = \underline{\underline{1.86}} \times \underline{\underline{10^{-21}\text{ J}}}$$

$$(b)\ P_N = 1.86 \times 10^{-21} \times 36 \times 10^6 = \underline{\underline{0.067\text{ pW}}}$$

In addition to these thermal noise sources, intermodulation distortion in high-power amplifiers (see Sec. 12.7.3) can result in signal products which appear as noise and in fact is referred to as *intermodulation noise*. This is discussed in Sec. 12.10.

### 12.5.1 Antenna noise

Antennas operating in the receiving mode introduce noise into the satellite circuit. Noise therefore will be introduced by the satellite receive antenna and the ground station receive antenna. Although the physical origins of the noise in either case are similar, the magnitudes of the effects differ significantly.

The antenna noise can be broadly classified into two groups: noise originating from antenna losses and *sky noise*. Sky noise is a term used to describe the microwave radiation which is present throughout the universe and which appears to originate from matter in any form at finite temperatures. Such radiation in fact covers a wider spectrum than just the microwave spectrum. The equivalent noise temperature of the sky, as seen by an earth-station antenna, is shown in Fig. 12.2. The lower graph is for the antenna pointing directly overhead, while the upper graph is for the antenna pointing just above the horizon. The increased noise in the latter case results from the thermal radiation of the earth, and this in fact sets a lower limit of about 5° at C band and 10° at Ku band on the elevation angle which may be used with ground-based antennas.

The graphs show that at the low-frequency end of the spectrum, the noise decreases with increasing frequency. Where the antenna is zenith-pointing, the noise temperature falls to about 3 K at frequencies between

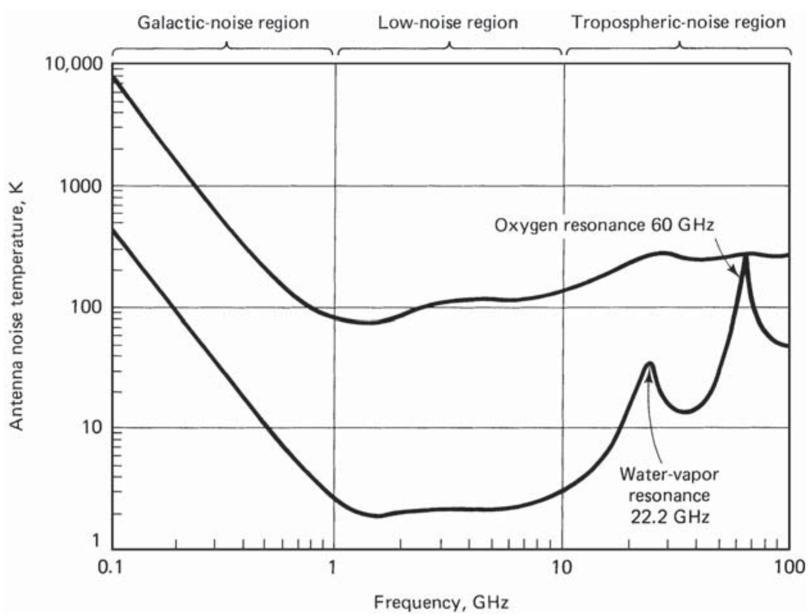



**Figure 12.2** Irreducible noise temperature of an ideal, ground-based antenna. The antenna is assumed to have a very narrow beam without sidelobes or electrical losses. Below 1 GHz, the maximum values are for the beam pointed at the galactic poles. At higher frequencies, the maximum values are for the beam just above the horizon and the minimum values for zenith pointing. The low-noise region between 1 and 10 GHz is most amenable to application of special, low-noise antennas. (*From Philip F. Panter, "Communications Systems Design," McGraw-Hill Book Company, New York, 1972. With permission.*)

about 1 and 10 GHz. This represents the residual background radiation in the universe. Above about 10 GHz, two peaks in temperature are observed, resulting from resonant losses in the earth's atmosphere. These are seen to coincide with the peaks in atmospheric absorption loss shown in Fig. 4.2.

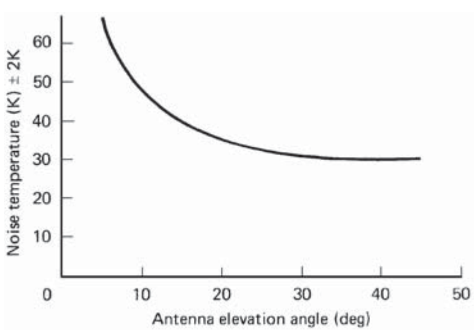
Any absorptive loss mechanism generates thermal noise, there being a direct connection between the loss and the effective noise temperature, as shown in Sec. 12.5.5. Rainfall introduces attenuation, and therefore, it degrades transmissions in two ways: It attenuates the signal, and it introduces noise. The detrimental effects of rain are much worse at Ku-band frequencies than at C band, and the downlink rain-fade margin, discussed in Sec. 12.9.2, must also allow for the increased noise generated.

Figure 12.2 applies to ground-based antennas. Satellite antennas are generally pointed toward the earth, and therefore, they receive the full thermal radiation from it. In this case the equivalent noise temperature of the antenna, excluding antenna losses, is approximately 290 K.

Antenna losses add to the noise received as radiation, and the total antenna noise temperature is the sum of the equivalent noise temperatures of all these sources. For large ground-based C-band antennas, the total antenna noise temperature is typically about 60 K, and for the Ku band, about 80 K under clear-sky conditions. These values do not apply to any specific situation and are quoted merely to give some idea of the magnitudes involved. Figure 12.3 shows the noise temperature as a function of angle of elevation for a 1.8-m antenna operating in the Ku band.

### 12.5.2 Amplifier noise temperature

Consider first the noise representation of the antenna and the *low noise amplifier* (LNA) shown in Fig. 12.4*a*. The available power gain of the amplifier is denoted as $G$, and the noise power output, as $P_{no}$. For the




| | |
|---|---|
| Operating frequency (GHz) | 11.7–12.2 |
| Polarization | Linear |
| Gain $dB_i$ ± 0.2 dB at 11.95 GHz | 45.0 |
| Noise temperature (K) | |
| At 10° elevation | 47 |
| At 20° elevation | 35 |
| At 30° elevation | 32 |
| VSWR maximum | 1.3 |
| Isolation (dB) | 35 |
| Half-power beamwidth (deg) | 0.91 |
| Pattern envelope (PE) number | 4374 |
| Output flanges | WR-75 |

**Figure 12.3** Antenna noise temperature as a function of elevation for 1.8-m antenna characteristics. (*Andrew Bulletin 1206; courtesy of Andrew Antenna Company, Limited.*)

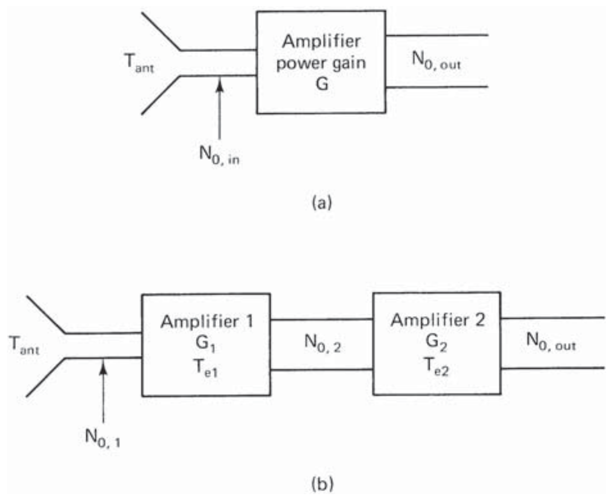



**Figure 12.4** Circuit used in finding equivalent noise temperature of (*a*) an amplifier and (*b*) two amplifiers in cascade.

moment we will work with the noise power per unit bandwidth, which is simply noise energy in joules as shown by Eq. (12.15). The input noise energy coming from the antenna is

$$N_{0,\text{ant}} = kT_{\text{ant}} \tag{12.16}$$

The output noise energy $N_{0,\text{out}}$ will be $GN_{0,\text{ant}}$ plus the contribution made by the amplifier. Now all the amplifier noise, wherever it occurs in the amplifier, may be *referred to the input* in terms of an equivalent input noise temperature for the amplifier $T_e$. This allows the output noise to be written as

$$N_{0,\text{out}} = Gk(T_{\text{ant}} + T_e) \tag{12.17}$$

The total noise referred to the input is simply $N_{0,\text{out}}/G$, or

$$N_{0,\text{in}} = k(T_{\text{ant}} + T_e) \tag{12.18}$$

$T_e$ can be obtained by measurement, a typical value being in the range 35 to 100 K. Typical values for $T_{\text{ant}}$ are given in Sec. 12.5.1.

### 12.5.3 Amplifiers in cascade

The cascade connection is shown in Fig. 12.4*b*. For this arrangement, the overall gain is

$$G = G_1G_2 \tag{12.19}$$

The noise energy of amplifier 2 referred to its own input is simply $kT_{e2}$. The noise input to amplifier 2 from the preceding stages is $G_1 k(T_{\text{ant}} + T_{e1})$, and thus the total noise energy *referred to amplifier 2 input* is

$$N_{0,2} = G_1 k(T_{\text{ant}} + T_{e1}) + kT_{e2} \tag{12.20}$$

This noise energy may be referred to amplifier 1 input by dividing by the available power gain of amplifier 1:

$$\begin{aligned} N_{0,1} &= \frac{N_{0,2}}{G_1} \\ &= k\left(T_{\text{ant}} + T_{e1} + \frac{T_{e2}}{G_1}\right) \end{aligned} \tag{12.21}$$

A system noise temperature may now be defined as $T_S$ by

$$N_{0,1} = kT_S \tag{12.22}$$

and hence it will be seen that $T_S$ is given by

$$T_S = T_{\text{ant}} + T_{e1} + \frac{T_{e2}}{G_1} \tag{12.23}$$

This is a very important result. It shows that the noise temperature of the second stage is divided by the power gain of the first stage when referred to the input. Therefore, in order to keep the overall system noise as low as possible, the first stage (usually an LNA) should have high power gain as well as low noise temperature.

This result may be generalized to any number of stages in cascade, giving

$$T_S = T_{\text{ant}} + T_{e1} + \frac{T_{e2}}{G_1} + \frac{T_{e3}}{G_1 G_2} + \cdots \tag{12.24}$$

### 12.5.4 Noise factor

An alternative way of representing amplifier noise is by means of its *noise factor, F*. In defining the noise factor of an amplifier, the source is taken to be at *room temperature*, denoted by $T_0$, usually taken as 290 K. The input noise from such a source is $kT_0$, and the output noise from the amplifier is

$$N_{0,\text{out}} = FGkT_0 \tag{12.25}$$

Here, $G$ is the available power gain of the amplifier as before, and $F$ is its noise factor.

A simple relationship between noise temperature and noise factor can be derived. Let $T_e$ be the noise temperature of the amplifier, and let the source be at room temperature as required by the definition of $F$. This means that $T_{\text{ant}} = T_0$. Since the same noise output must be available whatever the representation, it follows that

$$Gk(T_0 + T_e) = FGkT_0$$

or

$$T_e = (F - 1)\, T_0 \tag{12.26}$$

This shows the direct equivalence between noise factor and noise temperature. As a matter of convenience, in a practical satellite receiving system, noise temperature is specified for low-noise amplifiers and converters, while noise factor is specified for the main receiver unit.

The *noise figure* is simply $F$ expressed in decibels:

$$\text{Noise figure} = [F] = 10 \log F \tag{12.27}$$

**Example 12.6** An LNA is connected to a receiver which has a noise figure of 12 dB. The gain of the LNA is 40 dB, and its noise temperature is 120 K. Calculate the overall noise temperature referred to the LNA input.

**Solution** 12 dB is a power ratio of 15.85:1, and therefore,

$$T_{e2} = (15.85 - 1) \times 290 = 4306 \text{ K}$$

A gain of 40 dB is a power ratio of $10^4$:1, and therefore,

$$T_{\text{in}} = 120 + \frac{4306}{10^4}$$

$$= \underline{\underline{120.43 \text{ K}}}$$

In Example 12.6 it will be seen that the decibel quantities must be converted to power ratios. Also, even though the main receiver has a very high noise temperature, its effect is made negligible by the high gain of the LNA.

### 12.5.5 Noise temperature of absorptive networks

An *absorptive network* is one which contains resistive elements. These introduce losses by absorbing energy from the signal and converting it to heat. Resistive attenuators, transmission lines, and waveguides are all examples of absorptive networks, and even rainfall, which absorbs energy from radio signals passing through it, can be considered a form of absorptive network. Because an absorptive network contains resistance, it generates thermal noise.

Consider an absorptive network, which has a power loss $L$. The power loss is simply the ratio of input power to output power and will always be greater than unity. Let the network be matched at both ends, to a terminating resistor, $R_T$, at one end and an antenna at the other, as shown in Fig. 12.5, and let the system be at some ambient temperature $T_x$. The noise energy transferred from $R_T$ into the network is $kT_x$. Let the network noise be represented at the output terminals (the terminals connected to the antenna in this instance) by an equivalent noise temperature $T_{\mathrm{NW,0}}$. Then the noise energy radiated by the antenna is

$$N_{\mathrm{rad}} = \frac{kT_x}{L} + kT_{\mathrm{NW,0}} \tag{12.28}$$

Because the antenna is matched to a resistive source at temperature $T_x$, the available noise energy which is fed into the antenna and radiated is $N_{\mathrm{rad}} = kT_x$. Keep in mind that the antenna resistance to which the network is matched is fictitious, in the sense that it represents radiated power, but it does not generate noise power. This expression for $N_{\mathrm{rad}}$ can be substituted into Eq. (12.28) to give

$$T_{\mathrm{NW,0}} = T_x\left(1 - \frac{1}{L}\right) \tag{12.29}$$

This is the equivalent noise temperature of the network referred to the output terminals of the network. The equivalent noise at the output can be transferred to the input on dividing by the network power gain, which by definition is $1/L$. Thus, the equivalent noise temperature of the network referred to the network input is

$$T_{\mathrm{NW},i} = T_x(L - 1) \tag{12.30}$$

Since the network is bilateral, Eqs. (12.29) and (12.30) apply for signal flow in either direction. Thus, Eq. (12.30) gives the equivalent noise

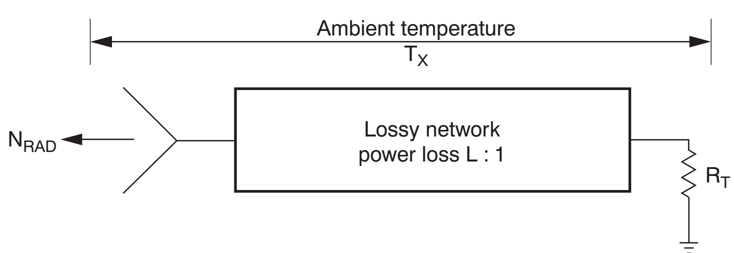



**Figure 12.5** Network matched at both ends, to a terminating resistor $R_T$ at one end and an antenna at the other.

temperature of a lossy network referred to the input at the antenna when the antenna is used in receiving mode.

If the lossy network should happen to be at room temperature, that is, $T_x = T_0$, then a comparison of Eqs. (12.26) and (12.30) shows that

$$F = L \tag{12.31}$$

This shows that at room temperature the noise factor of a lossy network is equal to its power loss.

### 12.5.6 Overall system noise temperature

Figure 12.6*a* shows a typical receiving system. Applying the results of the previous sections yields, for the system noise temperature referred to the input,

$$T_S = T_{\text{ant}} + T_{e1} + \frac{(L - 1)T_0}{G_1} + \frac{L(F - 1)T_0}{G_1} \tag{12.32}$$

The significance of the individual terms is illustrated in the following examples.

**Example 12.7** For the system shown in Fig. 12.6*a*, the receiver noise figure is 12 dB, the cable loss is 5 dB, the LNA gain is 50 dB, and its noise temperature 150 K. The antenna noise temperature is 35 K. Calculate the noise temperature referred to the input.

**Solution** For the main receiver, $F = 10^{1.2} = 15.85$. For the cable, $L = 10^{0.5} = 3.16$. For the LNA, $G = 10^5$. Hence,

$$T_S = 35 + 150 + \frac{(3.16 - 1) \times 290}{10^5} + \frac{3.16 \times (15.85 - 1) \times 290}{10^5}$$

$$\cong \underline{\underline{185 \text{ K}}}$$

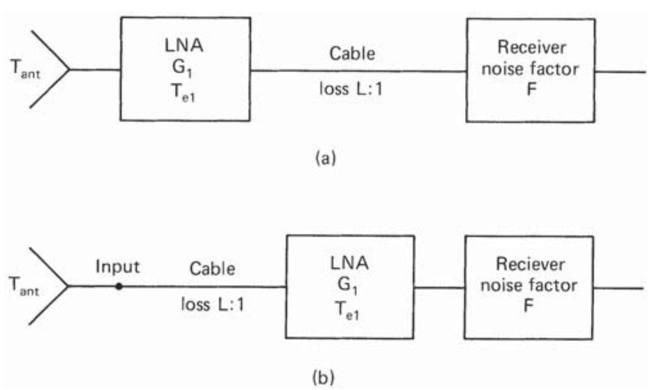



**Figure 12.6** Connections used in examples illustrating overall noise temperature of system, Sec. 12.5.6.

**Example 12.8** Repeat the calculation when the system of Fig. 12.6*a* is arranged as shown in Fig. 12.6*b*.

**Solution** In this case the cable precedes the LNA, and therefore, the equivalent noise temperature referred to the cable input is

$$T_S = 35 + (3.16 - 1) \times 290 + 3.16 \times 150 + \frac{3.16 \times (15.85 - 1) \times 290}{10^5}$$

$$= \underline{\underline{1136 \text{ K}}}$$

Examples 12.7 and 12.8 illustrate the important point that the LNA must be placed ahead of the cable, which is why one sees amplifiers mounted right at the dish in satellite receive systems.

## 12.6 Carrier-to-Noise Ratio

A measure of the performance of a satellite link is the ratio of carrier power to noise power at the receiver input, and link-budget calculations are often concerned with determining this ratio. Conventionally, the ratio is denoted by $C/N$ (or CNR), which is equivalent to $P_R/P_N$. In terms of decibels,

$$\left[\frac{C}{N}\right] = [P_R] - [P_N] \tag{12.33}$$

Equations (12.17) and (12.18) may be used for $[P_R]$ and $[P_N]$, resulting in

$$\left[\frac{C}{N}\right] = [\text{EIRP}] + [G_R] - [\text{LOSSES}] - [k] - [T_S] - [B_N] \tag{12.34}$$

The $G/T$ ratio is a key parameter in specifying the receiving system performance. The antenna gain $G_R$ and the system noise temperature $T_S$ can be combined in Eq. (12.34) as

$$[G/T] = [G_R] - [T_S] \text{ dBK}^{-1} \tag{12.35}$$

Therefore, the link equation [Eq. (12.34)] becomes

$$\left[\frac{C}{N}\right] = [\text{EIRP}] + \left[\frac{G}{T}\right] - [\text{LOSSES}] - [k] - [B_N] \tag{12.36}$$

The ratio of carrier power to noise power density $P_R/N_0$ may be the quantity actually required. Since $P_N = kT_NB_N = N_0\text{B}_\text{N}$, then

$$\left[\frac{C}{N}\right] = \left[\frac{C}{N_0B_N}\right]$$

$$= \left[\frac{C}{N_0}\right] - [B_N]$$

and therefore

$$\left[\frac{C}{N_0}\right] = \left[\frac{C}{N}\right] + [B_N] \tag{12.37}$$

$[C/N]$ is a true power ratio in units of decibels, and $[B_N]$ is in decibels relative to 1 Hz, or dBHz. Thus, the units for $[C/N_0]$ are dBHz.

Substituting Eq. (12.37) for $[C/N]$ gives

$$\left[\frac{C}{N_0}\right] = [\text{EIRP}] + \left[\frac{G}{T}\right] - [\text{LOSSES}] - [k] \tag{12.38}$$

**Example 12.9** In a link-budget calculation at 12 GHz, the free-space loss is 206 dB, the antenna pointing loss is 1 dB, and the atmospheric absorption is 2 dB. The receiver $[G/T]$ is 19.5 dB/K, and receiver feeder losses are 1 dB. The EIRP is 48 dBW. Calculate the carrier-to-noise spectral density ratio.

**Solution** The data are best presented in tabular form and in fact lend themselves readily to spreadsheet-type computations. For brevity, the units are shown as decilogs, (see App. G) and losses are entered as negative numbers to take account of the minus sign in Eq. (12.38). Recall that Boltzmann's constant equates to −228.6 decilogs, so $-[k]$ = 228.6 decilogs, as shown in the following table.

Entering data in this way allows the final result to be entered in a table cell as the sum of the terms in the rows above the cell, a feature usually incorporated in spreadsheets and word processors. This is illustrated in the following table.

| Quantity | Decilogs |
|---|---|
| Free-space loss | −206 |
| Atmospheric absorption loss | −2 |
| Antenna pointing loss | −1 |
| Receiver feeder losses | −1 |
| Polarization mismatch loss | 0 |
| Receiver $G/T$ ratio | 19.5 |
| EIRP | 48 |
| $-[k]$ | 228.6 |
| $[C/N_0]$, Eq. (12.38) | 86.1 |

The final result, 86.10 dBHz, is the algebraic sum of the quantities as given in Eq. (12.38).

## 12.7 The Uplink

The uplink of a satellite circuit is the one in which the earth station is transmitting the signal and the satellite is receiving it. Equation (12.38) can be applied to the uplink, but subscript $U$ will be used to denote specifically that the uplink is being considered. Thus Eq. (12.38) becomes

$$\left[\frac{C}{N_0}\right]_U = [\text{EIRP}]_U + \left[\frac{G}{T}\right]_U - [\text{LOSSES}]_U - [k] \qquad (12.39)$$

In Eq. (12.39) the values to be used are the earth station EIRP, the satellite receiver feeder losses, and satellite receiver $G/T$. The free-space loss and other losses which are frequency-dependent are calculated for the uplink frequency. The resulting carrier-to-noise density ratio given by Eq. (12.39) is that which appears at the satellite receiver.

In some situations, the flux density appearing at the satellite receive antenna is specified rather than the earth-station EIRP, and Eq. (12.39) is modified as explained next.

### 12.7.1 Saturation flux density

As explained in Sec. 7.7.3, the *traveling-wave tube amplifier* (TWTA) in a satellite transponder exhibits power output saturation, as shown in Fig. 7.21. The flux density required at the receiving antenna to produce saturation of the TWTA is termed the *saturation flux density*. The saturation flux density is a specified quantity in link budget calculations, and knowing it, one can calculate the required EIRP at the earth station. To show this, consider again Eq. (12.6) which gives the flux density in terms of EIRP, repeated here for convenience:

$$\Psi_M = \frac{\text{EIRP}}{4\pi r^2}$$

In decibel notation this is

$$[\Psi_M] = [\text{EIRP}] + 10\log\frac{1}{4\pi r^2} \qquad (12.40)$$

But from Eq. (12.9) for free-space loss we have

$$-[\text{FSL}] = 10\log\frac{\lambda^2}{4\pi} + 10\log\frac{1}{4\pi r^2} \qquad (12.41)$$

Substituting this in Eq. (12.40) gives

$$[\Psi_M] = [\text{EIRP}] - [\text{FSL}] - 10\log\frac{\lambda^2}{4\pi} \qquad (12.42)$$

The $\lambda^2/4\pi$ term has dimensions of area, and in fact, from Eq. (6.15) it is the effective area of an isotropic antenna. Denoting this by $A_0$ gives

$$[A_0] = 10\log\frac{\lambda^2}{4\pi} \qquad (12.43)$$

Since frequency rather than wavelength is normally known, it is left as an exercise for the student to show that with frequency $f$ in gigahertz, Eq. (12.43) can be rewritten as

$$[A_0] = -(21.45 + 20 \log f) \qquad (12.44)$$

Combining this with Eq. (12.42) and rearranging slightly gives the EIRP as

$$[\text{EIRP}] = [\Psi_M] + [A_0] + [\text{FSL}] \qquad (12.45)$$

Equation (12.45) was derived on the basis that the only loss present was the spreading loss, denoted by [FSL]. But, as shown in the previous sections, the other propagation losses are the atmospheric absorption loss, the polarization mismatch loss, and the antenna misalignment loss. When allowance is made for these, Eq. (12.45) becomes

$$[\text{EIRP}] = [\Psi_M] + [A_0] + [\text{FSL}] + [\text{AA}] + [\text{PL}] + [\text{AML}] \qquad (12.46)$$

In terms of the total losses given by Eq. (12.12), Eq. (12.46) becomes

$$[\text{EIRP}] = [\Psi_M] + [A_0] + [\text{LOSSES}] - [\text{RFL}] \qquad (12.47)$$

This is for clear-sky conditions and gives the *minimum* value of [EIRP] which the earth station must provide to produce a given flux density at the satellite. Normally, the saturation flux density will be specified. With saturation values denoted by the subscript $S$, Eq. (12.47) is rewritten as

$$[\text{EIRP}_S]_U = [\Psi_S] + [A_0] + [\text{LOSSES}]_U - [\text{RFL}] \qquad (12.48)$$

**Example 12.10** An uplink operates at 14 GHz, and the flux density required to saturate the transponder is $-120$ dB(W/m$^2$). The free-space loss is 207 dB, and the other propagation losses amount to 2 dB. Calculate the earth-station [EIRP] required for saturation, assuming clear-sky conditions. Assume [RFL] is negligible.

**Solution** At 14 GHz,

$$[A_0] = -(21.45 + 20 \log 14) = -44.37 \text{ dB}$$

The losses in the propagation path amount to 207 + 2 = 209 dB. Hence, from Eq. (12.48),

$$[\text{EIRP}_S]_U = -120 - 44.37 + 209$$
$$= \underline{\underline{44.63 \text{ dBW}}}$$

### 12.7.2 Input backoff

As described in Sec. 12.7.3, where a number of carriers are present simultaneously in a TWTA, the operating point must be backed off to a linear portion of the transfer characteristic to reduce the effects of intermodulation distortion. Such multiple carrier operation occurs with *frequency-division multiple access* (FDMA), which is described in Chap. 14. The point to be made here is that *backoff* (BO) must be allowed for in the link-budget calculations.

Suppose that the saturation flux density for single-carrier operation is known. Input BO will be specified for multiple-carrier operation, referred to the single-carrier saturation level. The earth-station EIRP will have to be reduced by the specified BO, resulting in an uplink value of

$$[\text{EIRP}]_U = [\text{EIRP}_S]_U - [\text{BO}]_i \tag{12.49}$$

Although some control of the input to the transponder power amplifier is possible through the ground TT&C station, as described in Sec. 12.7.3, input BO is normally achieved through reduction of the [EIRP] of the earth stations actually accessing the transponder.

Equations (12.48) and (12.49) may now be substituted in Eq. (12.39) to give

$$\left[\frac{C}{N_0}\right]_U = [\Psi_S] + [A_0] - [\text{BO}]_i + \left[\frac{G}{T}\right]_U - [k] - [\text{RFL}] \tag{12.50}$$

**Example 12.11** An uplink at 14 GHz requires a saturation flux density of $-91.4$ dBW/m$^2$ and an input BO of 11 dB. The satellite [$G/T$] is $-6.7$ dBK$^{-1}$, and receiver feeder losses amount to 0.6 dB. Calculate the carrier-to-noise density ratio.

**Solution** As in Example 12.9, the calculations are best carried out in tabular form.

[$A_0$] = $-44.37$ dBm$^2$ for a frequency of 14 GHz is calculated by using Eq. (12.44) as in Example 12.10.

| Quantity | Decilogs |
|---|---|
| Saturation flux density | −91.4 |
| [$A_0$] at 14 GHz | −44.4 |
| Input BO | −11.0 |
| Satellite saturation [$G/T$] | −6.7 |
| −[$k$] | 228.6 |
| Receiver feeder loss | −0.6 |
| Total | 74.5 |

Note that $[k] = -228.6$ dB, so $-[k]$ in Eq. (12.50) becomes 228.6 dB. Also, [RFL] and $[\text{BO}]_i$ are entered as negative numbers to take account of the minus signs attached to them in Eq. (12.50). The total gives the carrier-to-noise density ratio at the satellite receiver as 74.5 dBHz.

Since fade margins have not been included at this stage, Eq. (12.50) applies for *clear-sky* conditions. Usually, the most serious fading is caused by rainfall, as described in Sec. 12.9.

### 12.7.3 The earth station HPA

The earth station HPA has to supply the radiated power plus the transmit feeder losses, denoted here by TFL, or [TFL] dB. These include waveguide, filter, and coupler losses between the HPA output and the transmit antenna. Referring back to Eq. (12.3), the power output of the HPA is given by

$$[P_{\text{HPA}}] = [\text{EIRP}] - [G_T] + [\text{TFL}] \tag{12.51}$$

The [EIRP] is that given by Eq. (12.49) and thus includes any input BO that is required at the satellite.

The earth station itself may have to transmit multiple carriers, and its output also will require back off, denoted by $[\text{BO}]_{\text{HPA}}$. The earth station HPA must be rated for a saturation power output given by

$$[P_{\text{HPA,sat}}] = [P_{\text{HPA}}] + [\text{BO}]_{\text{HPA}} \tag{12.52}$$

Of course, the HPA will be operated at the backed-off power level so that it provides the required power output $[P_{\text{HPA}}]$. To ensure operation well into the linear region, an HPA with a comparatively high saturation level can be used and a high degree of BO introduced. The large physical size and high power consumption associated with larger tubes do not carry the same penalties they would if used aboard the satellite. Again, it is emphasized that BO at the earth station may be required quite independently of any BO requirements at the satellite transponder. The power rating of the earth-station HPA should also be sufficient to provide a fade margin, as discussed in Sec. 12.9.1.

## 12.8 Downlink

The downlink of a satellite circuit is the one in which the satellite is transmitting the signal and the earth station is receiving it. Equation (12.38) can be applied to the downlink, but subscript $D$ will be used to denote specifically that the downlink is being considered. Thus Eq. (12.38) becomes

$$\left[\frac{C}{N_0}\right]_D = [\text{EIRP}]_D + \left[\frac{G}{T}\right]_D - [\text{LOSSES}]_D - [k] \tag{12.53}$$

In Eq. (12.53) the values to be used are the satellite EIRP, the earth-station receiver feeder losses, and the earth-station receiver $G/T$. The free space and other losses are calculated for the downlink frequency. The resulting carrier-to-noise density ratio given by Eq. (12.53) is that which appears at the detector of the earth station receiver.

Where the carrier-to-noise ratio is the specified quantity rather than carrier-to-noise density ratio, Eq. (12.38) is used. This becomes, on assuming that the signal bandwidth $B$ is equal to the noise bandwidth $B_N$:

$$\left[\frac{C}{N}\right]_D = [\mathrm{EIRP}]_D + \left[\frac{G}{T}\right]_D - [\mathrm{LOSSES}]_D - [k] - [B] \qquad (12.54)$$

**Example 12.12** A satellite TV signal occupies the full transponder bandwidth of 36 MHz, and it must provide a $C/N$ ratio at the destination earth station of 22 dB. Given that the total transmission losses are 200 dB and the destination earth-station $G/T$ ratio is 31 dB/K, calculate the satellite EIRP required.

**Solution** Equation (12.54) can be rearranged as

$$[\mathrm{EIRP}]_D = \left[\frac{C}{N}\right]_D - \left[\frac{G}{T}\right]_D + [\mathrm{LOSSES}]_D + [k] + [B]$$

Setting this up in tabular form, and keeping in mind that $+[k] = -228.6$ dB and that losses are numerically equal to $+200$ dB, we obtain

| Quantity | Decilogs |
|---|---|
| $[C/N]$ | 22 |
| $-[G/T]$ | −31 |
| [LOSSES] | 200 |
| $[k]$ | −228.6 |
| $[B]$ | 75.6 |
| [EIRP] | 38 |

The required EIRP is 38 dBW or, equivalently, 6.3 kW.

Example 12.12 illustrates the use of Eq. (12.54). Example 12.13 shows the use of Eq. (12.53) applied to a digital link.

**Example 12.13** A QPSK signal is transmitted by satellite. Raised-cosine filtering is used, for which the rolloff factor is 0.2 and a *bit error rate* (BER) of $10^{-5}$ is required. For the satellite downlink, the losses amount to 200 dB, the receiving earth-station G/T ratio is 32 $\mathrm{dBK}^{-1}$, and the transponder bandwidth is 36 MHz. Calculate (*a*) the bit rate which can be accommodated, and (*b*) the EIRP required.

**Solution** Equation (10.16) gives

$$R_b = \frac{2B}{1 + \rho}$$
$$= \frac{2 \times 36 \times 10^6}{1.2}$$
$$= 60 \text{ Mbps}$$

Hence,

$$[R_b] = 10 \log\left(\frac{60 \times 10^6}{1\text{s}^{-1}}\right)$$
$$= 77.78 \text{ dBbps}$$

For BER $= 10^{-5}$, Fig. 10.17 gives an $[E_b/N_0] = 9.6$ dB.

From Eq. (10.24) the required $C/N_0$ ratio is

$$\left[\frac{C}{N_0}\right] = \left[\frac{E_b}{N_0}\right] + [R_b]$$
$$= 77.78 + 9.6$$
$$= 87.38 \text{ dBHz}$$

From Eq. (12.53),

$$[\text{EIRP}]_D = \left[\frac{C}{N_0}\right]_D - \left[\frac{G}{T}\right]_D + [\text{LOSSES}]_D + [k]$$
$$= 87.38 - 32 + 200 - 228.6$$
$$\cong \underline{\underline{26.8 \text{ dBW}}}$$

### 12.8.1 Output back-off

Where input BO is employed as described in Sec. 12.7.2, a corresponding output BO must be allowed for in the satellite EIRP. As the curve of Fig. 7.21 shows, output BO is not linearly related to input BO. A rule of thumb, frequently used, is to take the output BO as the point on the curve which is 5 dB below the extrapolated linear portion, as shown in Fig. 12.7. Since the linear portion gives a 1:1 change in decibels, the relationship between input and output BO is $[\text{BO}]_0 = [\text{BO}]_i - 5$ dB. For example, with an input BO of $[\text{BO}]_i = 11$ dB, the corresponding output BO is $[\text{BO}]_0 = 11 - 5 = 6$ dB.

If the satellite EIRP for saturation conditions is specified as $[\text{EIRP}_S]_D$, then $[\text{EIRP}]_D = [\text{EIRP}_S]_D - [\text{BO}]_0$ and Eq. (12.53) becomes

$$\left[\frac{C}{N_0}\right]_D = [\text{EIRP}_S]_D - [\text{BO}]_0 + \left[\frac{G}{T}\right]_D - [\text{LOSSES}]_D - [k] \qquad (12.55)$$

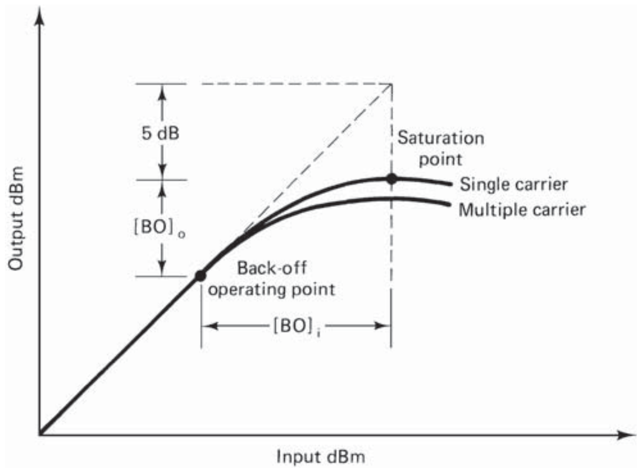



**Figure 12.7** Input and output back-off relationship for the satellite traveling-wave-tube amplifier; $[\text{BO}]_i = [\text{BO}]_0 + 5$ dB.

**Example 12.14** The specified parameters for a downlink are satellite saturation value of EIRP, 25 dBW; output BO, 6 dB; free-space loss, 196 dB; allowance for other downlink losses, 1.5 dB; and earth-station $G/T$, 41 $\text{dBK}^{-1}$. Calculate the carrier-to-noise density ratio at the earth station.

**Solution** As with the uplink budget calculations, the work is best set out in tabular form with the minus signs in Eq. (12.55) attached to the tabulated values.

| Quantity | Decilogs |
|---|---|
| Satellite saturation [EIRP] | 25.0 |
| Free-space loss | −196.0 |
| Other losses | −1.5 |
| Output BO | −6.0 |
| Earth station [$G/T$] | 41.0 |
| −[$k$] | 228.6 |
| Total | 91.1 |

The total gives the carrier-to-noise density ratio at the earth station in dBHz, as calculated from Eq. (12.55).

For the uplink, the saturation flux density at the satellite receiver is a specified quantity. For the downlink, there is no need to know the saturation flux density at the earth-station receiver, since this is a terminal point, and the signal is not used to saturate a power amplifier.

### 12.8.2 Satellite TWTA output

The satellite power amplifier, which usually is a TWTA, has to supply the radiated power plus the transmit feeder losses. These losses include the waveguide, filter, and coupler losses between the TWTA output and the satellite's transmit antenna. Referring back to Eq. (12.3), the power output of the TWTA is given by

$$[P_{\text{TWTA}}] = [\text{EIRP}]_D - [G_T]_D + [\text{TFL}]_D \tag{12.56}$$

Once $[P_{\text{TWTA}}]$ is found, the saturated power output rating of the TWTA is given by

$$[P_{\text{TWTA}}]_S = [P_{\text{TWTA}}] + [\text{BO}]_0 \tag{12.57}$$

**Example 12.15** A satellite is operated at an EIRP of 56 dBW with an output BO of 6 dB. The transmitter feeder losses amount to 2 dB, and the antenna gain is 50 dB. Calculate the power output of the TWTA required for full saturated EIRP.

**Solution** Equation (12.56):

$$\begin{aligned}[P_{\text{TWTA}}] &= [\text{EIRP}]_D - [G_T]_D + [\text{TFL}]_D \\ &= 56 - 50 + 2 \\ &= 8 \text{ dBW}\end{aligned}$$

Equation (12.57):

$$\begin{aligned}[P_{\text{TWTA}}]_S &= 8 + 6 \\ &= \underline{\underline{14 \text{ dBW (or 25 W)}}}\end{aligned}$$

## 12.9 Effects of Rain

Up to this point, calculations have been made for clear-sky conditions, meaning the absence of weather-related phenomena which might affect the signal strength. In the C band and, more especially, the Ku band, rainfall is the most significant cause of signal fading. Rainfall results in attenuation of radio waves by scattering and by absorption of energy from the wave, as described in Sec. 4.4. Rain attenuation increases with increasing frequency and is worse in the Ku band compared with the C band. Studies have shown (CCIR Report 338-3, 1978) that the rain attenuation for horizontal polarization is considerably greater than for vertical polarization.

Rain attenuation data are usually available in the form of curves or tables showing the fraction of time that a given attenuation is exceeded or, equivalently, the probability that a given attenuation will be exceeded (see Hogg et al., 1975; Lin et al., 1980; Webber et al., 1986). Some yearly average Ku-band values are shown in Table 12.2.

The percentage figures at the head of the first three columns give the percentage of time, averaged over any year, that the attenuation exceeds

**TABLE 12.2 Rain Attenuation for Cities and Communities in the Province of Ontario**

| | Rain attenuation, dB | | |
|---|---|---|---|
| Location | 1% | 0.5% | 0.1% |
| Cat Lake | 0.2 | 0.4 | 1.4 |
| Fort Severn | 0.0 | 0.1 | 0.4 |
| Geraldton | 0.1 | 0.2 | 0.9 |
| Kingston | 0.4 | 0.7 | 1.9 |
| London | 0.3 | 0.5 | 1.9 |
| North Bay | 0.3 | 0.4 | 1.9 |
| Ogoki | 0.1 | 0.2 | 0.9 |
| Ottawa | 0.3 | 0.5 | 1.9 |
| Sault Ste. Marie | 0.3 | 0.5 | 1.8 |
| Sioux Lookout | 0.2 | 0.4 | 1.3 |
| Sudbury | 0.3 | 0.6 | 2.0 |
| Thunder Bay | 0.2 | 0.3 | 1.3 |
| Timmins | 0.2 | 0.3 | 1.4 |
| Toronto | 0.2 | 0.6 | 1.8 |
| Windsor | 0.3 | 0.6 | 2.1 |

SOURCE: Telesat Canada Design Workbook.

the dB values given in each column. For example, at Thunder Bay, the rain attenuation exceeds, on average throughout the year, 0.2 dB for 1 percent of the time, 0.3 dB for 0.5 percent of the time, and 1.3 dB for 0.1 percent of the time. Alternatively, one could say that for 99 percent of the time, the attenuation will be equal to or less than 0.2 dB; for 99.5 percent of the time, it will be equal to or less than 0.3 dB; and for 99.9 percent of the time, it will be equal to or less than 1.3 dB.

Rain attenuation is accompanied by noise generation, and both the attenuation and the noise adversely affect satellite circuit performance, as described in Secs. 12.9.1 and 12.9.2.

As a result of falling through the atmosphere, raindrops are somewhat flattened in shape, becoming elliptical rather than spherical. When a radio wave with some arbitrary polarization passes through raindrops, the component of electric field in the direction of the major axes of the raindrops will be affected differently from the component along the minor axes. This produces a depolarization of the wave; in effect, the wave becomes elliptically polarized (see Sec. 5.6). This is true for both linear and circular polarizations, and the effect seems to be much worse for circular polarization (Freeman, 1981). Where only a single polarization is involved, the effect is not serious, but where frequency reuse is achieved through the use of orthogonal polarization (as described in Chap. 5), depolarizing devices, which compensate for the rain depolarization, may have to be installed.

Where the earth-station antenna is operated under cover of a radome, the effect of the rain on the radome must be taken into account. Rain falling on a hemispherical radome forms a water layer of constant thickness. Such a layer introduces losses, both by absorption and by reflection. Results presented by Hogg and Chu (1975) show an attenuation of about 14 dB for a 1-mm-thick water layer. It is desirable, therefore, that earth station antennas be operated without radomes where possible. Without a radome, water will gather on the antenna reflector, but the attenuation produced by this is much less serious than that produced by the wet radome (Hogg and Chu, 1975).

### 12.9.1 Uplink rain-fade margin

Rainfall results in attenuation of the signal and an increase in noise temperature, degrading the $[C/N_0]$ at the satellite in two ways. The increase in noise, however, is not usually a major factor for the uplink. This is so because the satellite antenna is pointed toward a "hot" earth, and this added to the satellite receiver noise temperature tends to mask any additional noise induced by rain attenuation. What is important is that the uplink carrier power at the satellite must be held within close limits for certain modes of operation, and some form of *uplink power control* is necessary to compensate for rain fades. The power output from the satellite may be monitored by a central control station or in some cases by each earth station, and the power output from any given earth station may be increased if required to compensate for fading. Thus the earth-station HPA must have sufficient reserve power to meet the fade margin requirement.

Some typical rain-fade margins are shown in Table 12.2. As an example, for Ottawa, the rain attenuation exceeds 1.9 dB for 0.1 percent of the time. This means that to meet the specified power requirements at the input to the satellite for 99.9 percent of the time, the earth station must be capable of providing a 1.9-dB margin over the clear-sky conditions.

### 12.9.2 Downlink rain-fade margin

The results given by Eqs. (12.53) and (12.54) are for clear-sky conditions. Rainfall introduces attenuation by absorption and scattering of signal energy, and the absorptive attenuation introduces noise as discussed in Sec. 12.5.5. Let $[A]$ dB represent the rain attenuation caused by absorption. The corresponding power loss ratio is $A = 10^{[A]/10}$, and substituting this for $L$ in Eq. (12.29) gives the effective noise temperature of the rain as

$$T_{\text{rain}} = T_a\left(1 - \frac{1}{A}\right) \tag{12.58}$$

Here, $T_a$, which takes the place of $T_x$ in Eq. (12.29), is known as the *apparent absorber temperature*. It is a measured parameter which is a function of many factors including the physical temperature of the rain and the scattering effect of the rain cell on the thermal noise incident upon it (Hogg and Chu, 1975). The value of the apparent absorber temperature lies between 270 and 290 K, with measured values for North America lying close to or just below freezing (273 K). For example, the measured value given by Webber et al. (1986) is 272 K.

The total sky-noise temperature is the clear-sky temperature $T_{CS}$ plus the rain temperature:

$$T_{sky} = T_{CS} + T_{rain} \tag{12.59}$$

Rainfall therefore degrades the received $[C/N_0]$ in two ways: by attenuating the carrier wave and by increasing the sky-noise temperature.

**Example 12.16** Under clear-sky conditions, the downlink [C/N] is 20 dB, the effective noise temperature of the receiving system being 400 K. If rain attenuation exceeds 1.9 dB for 0.1 percent of the time, calculate the value below which [C/N] falls for 0.1 percent of the time. Assume $T_a = 280$ K.

**Solution** 1.9 dB attenuation is equivalent to a 1.55:1 power loss. The equivalent noise temperature of the rain is therefore

$$T_{rain} = 280\,(1 - 1/1.55) = 99.2 \text{ K}$$

The new system noise temperature is 400 + 99.2 = 499.2 K. The decibel increase in noise power is therefore [499.2] − [400] = 0.96 dB. At the same time, the carrier is reduced by 1.9 dB, and therefore, the [C/N] with 1.9-dB rain attenuation drops to 20 − 1.9 − 0.96 = 17.14 dB. This is the value below which [C/N] drops for 0.1 percent of the time.

It is left as an exercise for the student to show that where the rain power attenuation $A$ (not dB) is entirely absorptive, the downlink $C/N$ power ratios (not dBs) are related to the clear-sky value by

$$\left(\frac{N}{C}\right)_{rain} = \left(\frac{N}{C}\right)_{CS}\left(A + (A - 1)\frac{T_a}{T_{S,CS}}\right) \tag{12.60}$$

where the subscript CS is used to indicate clear-sky conditions and $T_{S,CS}$ is the system noise temperature under clear-sky conditions. Note that noise-to-carrier ratios, rather than carrier-to-noise ratios are required by Eq. (12.60).

For low frequencies (6/4 GHz) and low rainfall rates (below about 1 mm/h), the rain attenuation is almost entirely absorptive. At higher rainfall rates, scattering becomes significant, especially at the higher frequencies. When scattering and absorption are both significant, the total attenuation must be used to calculate the reduction in carrier power and the absorptive attenuation to calculate the increase in noise temperature.

As discussed in Chap. 9, a minimum value of [$C/N$] is required for satisfactory reception. In the case of frequency modulation, the minimum value is set by the threshold level of the FM detector, and a *threshold margin* is normally allowed, as shown in Fig. 9.12. Sufficient margin must be allowed so that rain-induced fades do not take the [$C/N$] below threshold more than a specified percentage of the time, as shown in Example 12.17.

**Example 12.17** In an FM satellite system, the clear-sky downlink [$C/N$] ratio is 17.4 dB and the FM detector threshold is 10 dB, as shown in Fig. 9.12. (*a*) Calculate the threshold margin at the FM detector, assuming the threshold [$C/N$] is determined solely by the downlink value. (*b*) Given that $T_a$ = 272 K and that $T_{S,\mathrm{CS}}$ = 544 K, calculate the percentage of time the system stays above threshold. The curve of Fig. 12.8 may be used for the downlink, and it may be assumed that the rain attenuation is entirely absorptive.

**Solution** (*a*) Since it is assumed that the overall [$C/N$] ratio is equal to the downlink value, the clear-sky input [$C/N$] to the FM detector is 17.4 dB. The threshold level for the detector is 10 dB, and therefore, the rain-fade margin is 17.4 − 10 = 7.4 dB.

(*b*) The rain attenuation can reduce the [$C/N$] to the threshold level of 10 dB (i.e., it reduces the margin to zero), which is a ($C/N$) power ratio of 10:1 or a downlink $N/C$ power ratio of 1/10.

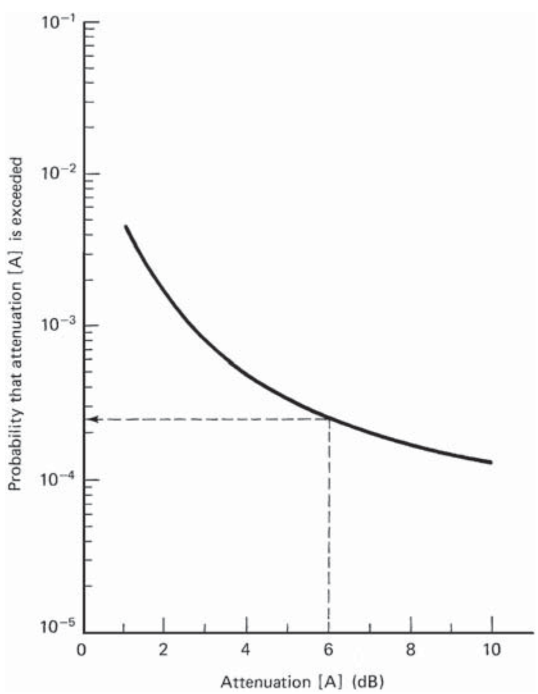



**Figure 12.8** Typical rain attenuation curve used in Example 12.17.

For clear-sky conditions, $[C/N]_{CS} = 17.4$ dB, which gives an $N/C$ ratio of 0.0182. Substituting these values in Eq. (12.60) gives

$$0.1 = 0.0182 \times \left(A + \frac{(A - 1) \times 272}{544}\right)$$

Solving this equation for $A$ gives $A = 4$, or approximately 6 dB. From the curve of Fig. 12.8, the probability of exceeding the 6-dB value is $2.5 \times 10^{-4}$, and therefore, the availability is $1 - 2.5 \times 10^{-4} = 0.99975$, or 99.975 percent.

For digital signals, the required $[C/N_0]$ ratio is determined by the acceptable BER, which must not be exceeded for more than a specified percentage of the time. Figure 10.17 relates the BER to the $[E_b/N_0]$ ratio, and this in turn is related to the $[C/N_0]$ by Eq. (10.24), as discussed in Sec. 10.6.4.

For the downlink, the user does not have control of the satellite [EIRP], and thus the downlink equivalent of uplink power control, described in Sec. 12.9.1, cannot be used. In order to provide the rain-fade margin needed, the gain of the receiving antenna may be increased by using a larger dish and/or a receiver front end having a lower noise temperature. Both measures increase the receiver $[G/T]$ ratio and thus increase $[C/N_0]$ as shown by Eq. (12.53).

## 12.10 Combined Uplink and Downlink *C/N* Ratio

The complete satellite circuit includes an uplink and a downlink, as sketched in Fig. 12.9*a*. Noise will be introduced on the uplink at the satellite receiver input. Denoting the noise power per unit bandwidth by $P_{NU}$ and the average carrier at the same point by $P_{RU}$, the carrier-to-noise ratio on the uplink is $(C/N_0)_U = (P_{RU}/P_{NU})$. It is important to note that power levels, and not decibels, are being used here.

The carrier power at the end of the space link is shown as $P_R$, which of course is also the received carrier power for the downlink. This is equal to $\gamma$ times the carrier power input at the satellite, where $\gamma$ is the system power gain from satellite input to earth-station input, as shown in Fig. 12.9*a*. It includes the satellite transponder and transmit antenna gains, the downlink losses, and the earth-station receive antenna gain and feeder losses.

The noise at the satellite input also appears at the earth station input multiplied by $\gamma$, and in addition, the earth station introduces its own noise, denoted by $P_{ND}$. Thus the end-of-link noise is $\gamma P_{NU} + P_{ND}$.

The $C/N_0$ ratio for the downlink alone, not counting the $\gamma P_{NU}$ contribution, is $P_R/P_{ND}$, and the combined $C/N_0$ ratio at the ground receiver is

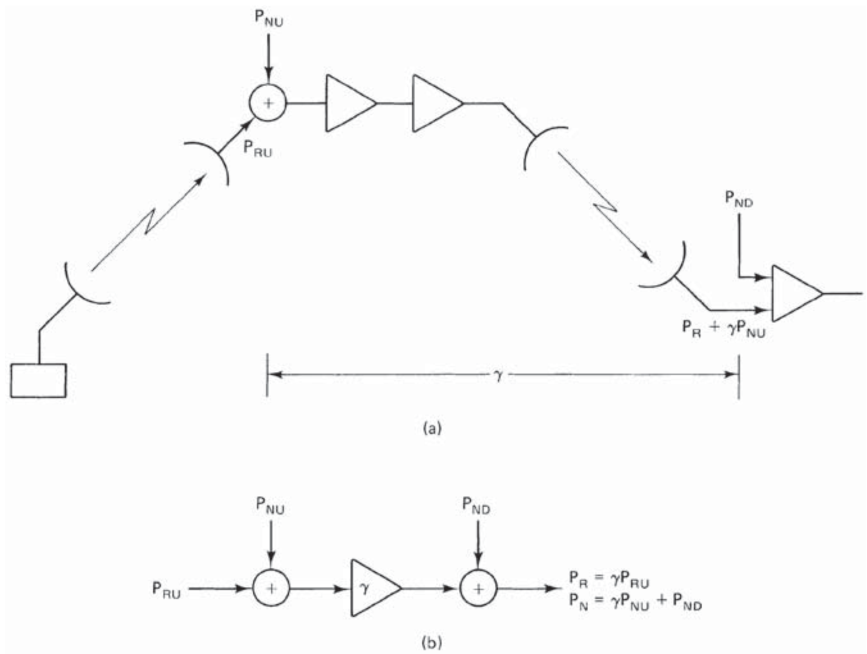



**Figure 12.9** (*a*) Combined uplink and downlink; (*b*) power flow diagram for (*a*).

$P_R/(\gamma P_{\mathrm{NU}} + P_{\mathrm{ND}})$. The power flow diagram is shown in Fig. 12.9*b*. The combined carrier-to-noise ratio can be determined in terms of the individual link values. To show this, it is more convenient to work with the noise-to-carrier ratios rather than the carrier-to-noise ratios, and again, these must be expressed as power ratios, not decibels. Denoting the combined noise-to-carrier ratio value by $N_0/C$, the uplink value by $(N_0/C)_U$, and the downlink value by $(N_0/C)_D$ then,

$$\begin{aligned}\frac{N_0}{C} &= \frac{P_N}{P_R}\\ &= \frac{\gamma P_{\mathrm{NU}} + P_{\mathrm{ND}}}{P_R}\\ &= \frac{\gamma P_{\mathrm{NU}}}{P_R} + \frac{P_{\mathrm{ND}}}{P_R}\\ &= \frac{\gamma P_{\mathrm{NU}}}{\gamma P_{\mathrm{RU}}} + \frac{P_{\mathrm{NU}}}{P_R}\\ &= \left(\frac{N_0}{C}\right)_U + \left(\frac{N_0}{C}\right)_D\end{aligned} \tag{12.61}$$

Equation (12.61) shows that to obtain the combined value of $C/N_0$, the reciprocals of the individual values must be added to obtain the $N_0/C$ ratio and then the reciprocal of this taken to get $C/N_0$. Looked at in another way, the reason for this reciprocal of the sum of the reciprocals method is that a single signal power is being transferred through the system, while the various noise powers, which are present are additive. Similar reasoning applies to the carrier-to-noise ratio, $C/N$.

**Example 12.18** For a satellite circuit the individual link carrier-to-noise spectral density ratios are: uplink 100 dBHz; downlink 87 dBHz. Calculate the combined $C/N_0$ ratio.

**Solution**

$$\frac{N_0}{C} = 10^{-10} + 10^{-8.7} = 2.095 \times 10^{-9}$$

Therefore,

$$\left[\frac{C}{N_0}\right] = -10 \log(2.095 \times 10^{-9})$$

$$= \underline{\underline{86.79 \text{ dBHz}}}$$

Example 12.18 illustrates the point that when one of the link $C/N_0$ ratios is much less than the other, the combined $C/N_0$ ratio is approximately equal to the lower (worst) one. The downlink $C/N$ is usually (but not always) less than the uplink $C/N_0$, and in many cases it is much less. This is true primarily because of the limited EIRP available from the satellite.

Example 12.19 illustrates how BO is taken into account in the link-budget calculations and how it affects the $C/N_0$ ratio.

**Example 12.19** A multiple carrier satellite circuit operates in the 6/4-GHz band with the following characteristics.

*Uplink:*

Saturation flux density $-67.5$ dBW/m$^2$; input BO 11 dB; satellite $G/T$ $-11.6$ dBK$^{-1}$.

*Downlink:*

Satellite saturation EIRP 26.6 dBW; output BO 6 dB; free-space loss 196.7 dB; earth station $G/T$ 40.7 dBK$^{-1}$. For this example, the other losses may be ignored. Calculate the carrier-to-noise density ratios for both links and the combined value.

**Solution** As in the previous examples, the data are best presented in tabular form, and values are shown in decilogs. The minus signs in Eqs. (12.50) and (12.55) are attached to the tabulated numbers:

| | Decilog values |
|---|---|
| Uplink | |
| Saturation flux density | −67.5 |
| $[A_0]$ at 6 GHz | −37 |
| Input BO | −11 |
| Satellite saturation $[G/T]$ | −11.6 |
| $-[k]$ | 228.6 |
| $[C/N_0]$ from Eq. (12.50) | 101.5 |
| Downlink | |
| Satellite [EIRP] | 26.6 |
| Output BO | −6 |
| Free-space loss | −196.7 |
| Earth station $[G/T]$ | 40.7 |
| $-[k]$ | 228.6 |
| $[C/N_0]$ from Eq. (12.55) | 93.2 |

Application of Eq. (12.61) provides the combined $[C/N_0]$:

$$\frac{N_0}{C} = 10^{-10.15} + 10^{-9.32} = 5.49 \times 10^{-10}$$

$$\left[\frac{C}{N_0}\right] = -10\log(5.49 \times 10^{-10})$$

$$= \underline{\underline{92.6 \text{ dBHz}}}$$

Again, it is seen from Example 12.19 that the combined $C/N_0$ value is close to the lowest value, which is the downlink value.

So far, only thermal and antenna noise has been taken into account in calculating the combined value of $C/N_0$ ratio. Another source of noise to be considered is intermodulation noise, which is discussed in the following section.

## 12.11 Intermodulation Noise

Intermodulation occurs where multiple carriers pass through any device with nonlinear characteristics. In satellite communications systems, this most commonly occurs in the traveling-wave tube HPA aboard the satellite, as described in Sec. 7.7.3. Both amplitude and phase nonlinearities give rise to intermodulation products.

As shown in Fig. 7.20, third-order intermodulation products fall on neighboring carrier frequencies, where they result in interference. Where a large number of modulated carriers are present, the intermodulation products are not distinguishable separately but instead appear as a type of noise which is termed *intermodulation noise*.

The carrier-to-intermodulation-noise ratio is usually found experimentally, or in some cases it may be determined by computer methods. Once this ratio is known, it can be combined with the carrier-to-thermal-noise ratio by the addition of the reciprocals in the manner described in Sec. 12.10. Denoting the intermodulation term by $(C/N_0)_{\mathrm{IM}}$ and bearing in mind that the reciprocals of the $C/N_0$ power ratios (and not the corresponding dB values) must be added, Eq. (12.61) is extended to

$$\frac{N_0}{C} = \left(\frac{N_0}{C}\right)_U + \left(\frac{N_0}{C}\right)_D + \left(\frac{N_0}{C}\right)_{IM} \tag{12.62}$$

A similar expression applies for noice-to-carrier ($N/C$) ratios.

**Example 12.20** For a satellite circuit the carrier-to-noise ratios are uplink 23 dB, downlink 20 dB, intermodulation 24 dB. Calculate the overall carrier- to-noise ratio in decibels.

**Solution** From Eq. (12.62),

$$\frac{N}{C} = 10^{-2.4} + 10^{-2.3} + 10^{-2} = 0.0019$$

$$\left[\frac{C}{N}\right] = -10 \log(0.0019)$$

$$= \underline{\underline{17.2 \text{ dBHz}}}$$

In order to reduce intermodulation noise, the TWT must be operated in a BO condition as described previously. Figure 12.10 shows how the $[C/N_0]_{\mathrm{IM}}$ ratio improves as the input BO is increased for a typical TWT. At the same time, increasing the BO decreases both $[C/N_0]_U$ and $[C/N_0]_D$, as shown by Eqs. (12.50) and (12.55). The result is that there is an optimal point where the overall carrier-to-noise ratio is a maximum. The component $[C/N_0]$ ratios as functions of the TWT input are sketched in Fig. 12.11. The TWT input in dB is $[\Psi]_S - [\mathrm{BO}]_i$, and therefore, Eq. (12.50) plots as a straight line. Equation (12.55) reflects the curvature in the TWT characteristic through the output BO, $[\mathrm{BO}]_0$, which is not linearly related to the input BO, as shown in Fig. 12.7. The intermodulation curve is not easily predictable, and only the general trend is shown. The overall $[C/N_0]$, which is calculated from Eq. (12.62), is also sketched. The optimal operating point is defined by the peak of this curve.

## 12.12 Inter-Satellite Links

*Inter-satellite links* (ISLs) are radio frequency or optical links that provide a connection between satellites without the need for intermediate

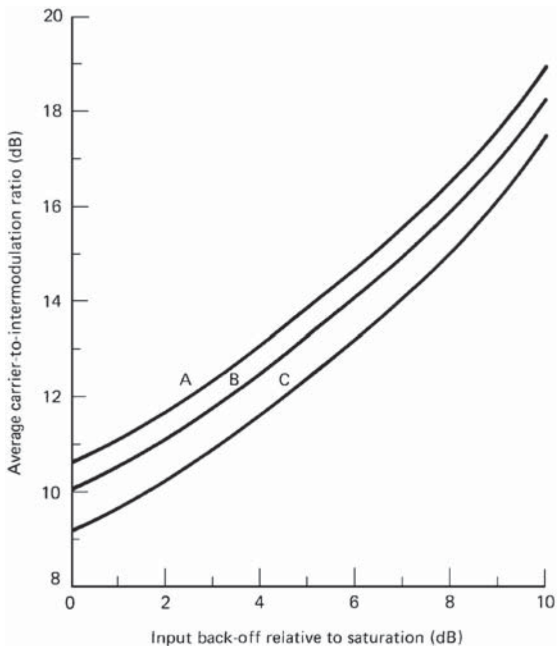



**Figure 12.10** Intermodulation in a typical TWT. Curve A, 6 carriers; curve B, 12 carriers; curve C, 500 carriers. (*From CCIR, 1982b. With permission from the International Telecommunications Union.*)

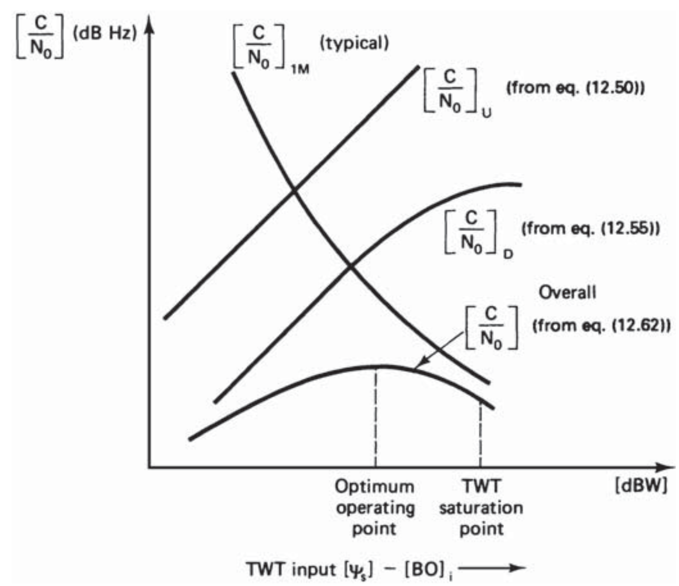



**Figure 12.11** Carrier-to-noise density ratios as a function of input back-off.

ground stations. Although many different links are possible, the most useful ones in operation are:

- *low earth orbiting* (LEO) satellites—LEO ↔ LEO
- *geostationary earth orbiting* (GEO) satellites—GEO ↔ GEO
- LEO ↔ GEO

Consider first some of the applications to GEOs. As shown in Chap. 3, the antenna angle of elevation is limited to a minimum of about 5 degrees because of noise induced from the earth. The limit of visibility as set by the minimum angle of elevation is a function of the satellite longitude and earth-station latitude and longitude as shown in Sec. 3.2. Figure 12.12 shows the situation where earth station *A* is beyond the range of satellite *S*2, a problem that can be overcome by the use of two satellites connected by an ISL. Thus, a long distance link between earth stations *A* and *B* can be achieved by this means. A more extreme example is where an intercontinental service may require a number of "hops." For example a Europe-Asia circuit requires three hops (Morgan, 1999): Europe to eastern U.S.A; eastern U.S.A. to western U.S.A; western U.S.A. to Asia; and of course each hop required an uplink and a downlink. By using an ISL only one uplink and one downlink is required. Also, as will be discussed shortly, the ISL frequencies are well outside the standard uplink and downlink bands so that spectrum use is conserved. The cost of the ISL is more than offset by not having to provide the additional earth stations required by the three-hop system.

The distance *d* for the ISL is easily calculated. From Fig. 12.13:

$$d = 2a_{\text{GSO}} \sin \frac{\Delta\phi}{2} \tag{12.63}$$

where $\Delta\phi$ is the longitudinal separation between satellites *S*1 and *S*2, and $a_{\text{GSO}}$ is the radius of the geostationary orbit [see Eq. (3.2)], equal to 42164 km. For example the western limits for the continental United States (CONUS) arc are at 55° and 136° (see Prob. 3.11). Although there are no satellites positioned exactly at these longitudes they can be used

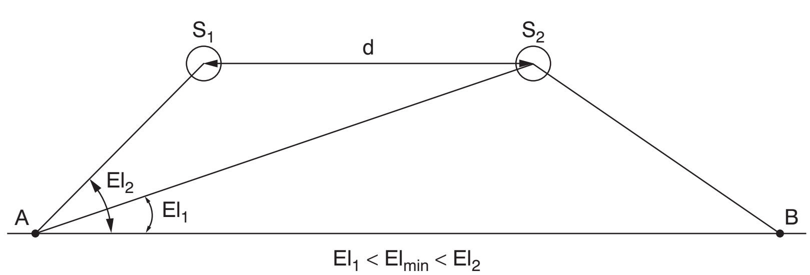



**Figure 12.12** Angle of elevation as determined by an ISL.

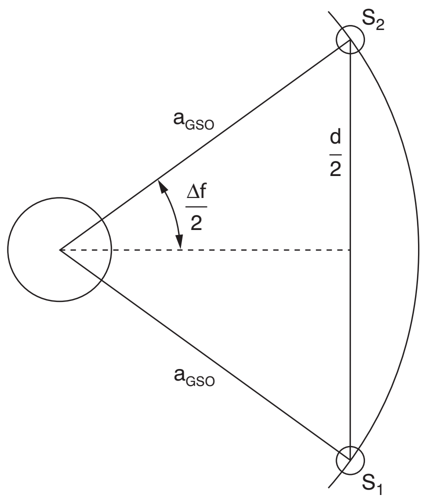



**Figure 12.13** Finding the distance $d$ between two GEO satellites.

to get an estimate of distance $d$ for an ISL spanning CONUS.

$$d = 2 \times 42164 \times \sin\frac{(136^\circ - 55^\circ)}{2} = 54767 \text{ km}$$

Although this may seem large, the range from earth station to satellite is in the order of 41000 km, (Prob. 3.11) so distance involved with three uplinks and three downlinks is 246000 km!

GEO satellites are often arranged in clusters at some nominal longitude. For example, there are a number of EchoStar satellites at longitude 119°W. The separation between satellites is typically about 100 km, the corresponding longitudinal separation for this distance being, from Eq. (12.63), approximately 0.136°. Because the satellites are relatively close together they are subject to the same perturbing and drift forces which simplifies positional control. Also, all satellites in the cluster are within the main lobe of the earth-station antenna.

Because LEO satellites are not continuously visible from a given earth location, an intricate network of satellites is required to provide continuous coverage of any region. A typical LEO satellite network will utilize a number of orbits, with equispaced satellites in each orbit. For example, the Iridium system uses 6 orbital planes with 11 equispaced satellites in each plane, for a total of 66 satellites. Communication between two earth stations via the network will appear seamless as message handover occurs between satellites via intersatellite links.

Radio frequency ISLs make use of frequencies that are highly attenuated by the atmosphere, so that interference to and from terrestrial systems using the same frequencies is avoided. Figure 4.2 shows the atmospheric absorption peaks at 22.3 GHz and 60 GHz. Table 12.3 shows the frequency bands in use:

**TABLE 12.3 ISL Frequency Bands**

| Frequency band, GHz | Available bandwidth, MHz | Designation |
|---|---|---|
| 22.55–23.55 | 1000 | ISL-23 |
| 24.45–24.75 | 300 | ISL-24 |
| 25.25–27.5 | 2250 | ISL-25 |
| 32–33 | 1000 | ISL-32 |
| 54.25–58.2 | 3950 | ISL-56 |
| 59–64 | 5000 | ISL-60 |
| 65–71 | 6000 | ISL-67 |
| 116–134 | 18000 | ISL-125 |
| 170–182 | 12000 | |
| 185–190 | 5000 | |

SOURCE: Morgan, 1999.

Antennas for the ISL are steerable and the beamwidths are sufficiently broad to enable a tracking signal to be acquired to maintain alignment. Table 12.4 gives some values for an ISL used in the Iridium system.

**Example 12.21** Calculate the free-space loss for the ISL parameters tabulated in Table 12.4. Given that the system margin for transmission loss is 1.8 dB, calculate the received power.

**Solution** From Eq. (12.10):

$$
\begin{aligned}
[\text{FSL}] &= 32.4 + 20 \log r + 20 \log f \\
&= 32.4 + 20 \log 4400 + 20 \log(23.28 \times 10^3) \\
&= \underline{\underline{192.6 \text{ dB}}}
\end{aligned}
$$

**TABLE 12.4 Iridium ISL**

| | East-West ISL (without sun) |
|---|---|
| Frequency, GHz | 23.28 |
| Range, km | 4400.0 |
| **Transmitter** | |
| Power, dBW | 5.3 |
| Antenna gain, dB | 36.7 |
| Circuit loss, dB | 1.8 |
| Pointing loss, dB | 1.8 |
| **Receiver** | |
| Pointing loss, dB | 1.8 |
| Antenna gain, dB | 36.7 |
| Noise temperature, K | 720.3 |
| Noise bandwidth, dBHz | 71 |

SOURCE: Motorola, 1992.

The [EIRP] is

$$[\text{EIRP}] = [P_T] + [G_T] - [\text{AML}]_T - [\text{TFL}]$$
$$= 5.3 + 36.7 - 1.8 - 1.8$$
$$= 38.4 \text{ dBW}$$

The total losses, including the link margin and the receiver misalignment (pointing) loss are:

[LOSSES] = 192.6 + 1.8 + 1.8 = 196.2 dB. The received power is, from Eq. (12.13)

$$[P_R] = [\text{EIRP}] + [G_R] - [\text{LOSSES}]$$
$$= \underline{\underline{-121.1 \text{ dBW}}}$$

Radio ISLs have the advantage that the technology is mature, so the risk of failure is minimized. However, the bandwidth limits the bit rate that can be carried, and optical systems, with their much higher carrier (optical) frequencies, have much greater bandwidth. Optical ISLs have a definite advantage over rf ISLs for data rates in excess of about 1 Gbps Also, telescope apertures are used which are considerably smaller than their rf counterparts, and generally, optical equipment tends to be smaller and more compact (see Optical Communications and IntersatelliteLinks, undated, at www.wtec.org/loyola/satcom2/03_06.htm-22k-). The optical beamwidth is typically 5 $\mu$rad (Maral et al., 2002). Table 12.5 lists properties of some solid state lasers.

The free-space loss given by Eq. (12.9) is repeated here:

$$[\text{FSL}] = 10 \log\left(\frac{4\pi r}{\lambda}\right)^2$$

**TABLE 12.5 Solid State Lasers**

| Type | Wavelength, $\mu$m | Power | Beam diameter, mm | Beam divergence |
|---|---|---|---|---|
| GaAs/GaAlAs | 0.78–0.905 | 1–40 mW Avg | | 10° × 35° |
| InGaAsP | 1.1–1.6 | 1–10 mW | | 10° × 30° − 20° × 40° |
| Nd: YAG Pulsed | 1.064 | Up to 600 W Avg | 1– 10 | 0.3–20 mrad |
| Nd:YAG Diode pumped | 1.064 | 0.5–10 mW | 1–2 | 0.5–2.0 mrad |
| Nd:YAG (cw) | 1.064 | 0.04–600 W | 0.7–8 | 2–25 mrad |

NOTES: Al—aluminum; As—arsenide; Ga—gallium; Nd—neodymium; P— phosphorus; YAG—yttrium-aluminum garnet; cw—continuous wave; $\mu$m—micron = $10^{-6}$ m; mm—millimeter; mrad—milliradian; mW—milliwatt; W—watt.

SOURCE: Extracted from Chen, 1996.

Here, wavelength rather than frequency is used in the equation as this is the quantity usually specified for a laser, and of course $r$ and $\lambda$ must be in the same units.

The intensity distribution of a laser beam generally follows what is termed a *Gaussian law*, for which the intensity falls off in an exponential manner in a direction transverse to the direction of propagation. The *beam radius* is where the transverse electric field component drops to 1/e of its maximum value, where e ≈ 2.718. The diameter of the beam (twice the radius) gives the total beamwidth. The on-axis gain (similar to the antenna gain defined in Sec. 6.6 is given by (Maral et al., 2002)

$$G_T = \frac{32}{\theta_T^2} \tag{12.64}$$

where $\theta_T$ is the total beamwidth.

On the receive side, the telescope aperture gain is given by:

$$G_R = \left(\frac{\pi D}{\lambda}\right)^2 \tag{12.65}$$

where $D$ is the effective diameter of the receiving aperture.

The optical receiver will receive some amount of optical power $P_R$. The energy in a photon is $hc/\lambda$, where $h$ is Plank's constant ($6.6256 \times 10^{-34}$ J-s) and $c$ is the speed of light in vacuum (approximately $3 \times 10^8$ m/s). For a received power $P_R$ the number of photons received per second is therefore $P_R\lambda/hc$. The detection process consists of photons imparting sufficient energy to valence band electrons to raise these to the conduction band. The *quantum efficiency* of a photo-diode is the ratio (average number of conduction electrons generated)/(average number of photons received). Denoting the quantum efficiency by $\eta$, the average number of electrons released is $\eta P_R\lambda/hc$ and the photo current is:

$$I_{ph} = \frac{q\eta P_R\lambda}{hc} \tag{12.66}$$

where $q$ is the electron charge. The *responsivity* of a photodiode is defined as the ratio of photo current to incident power. Denoting responsivity by $R_0$ and evaluating the constants in Eq. (12.66) gives

$$R_0 = \frac{\eta\lambda}{1.24} \quad \text{with } \lambda \text{ in } \mu\text{m} \tag{12.67}$$

The energy band gap of the semiconductor material used for the photodiode determines the wavelengths that it can respond to. The requirement in general is that the bandgap energy must be less than the photon energy, or $E_G < hc/\lambda$. On this basis it turns out that silicon is useful for wavelengths shorter that 1 $\mu$m. Germanium is useful at 1.3 $\mu$m and InSb and InAs at 1.55 $\mu$m.

The total current flowing in a photodiode consists of the actual current generated by the photons, plus what is termed the *dark current*. This is the current that flows even when no signal is present. Ideally it should be zero, but in practice it is of the order of a few nanoamperes, and it can contribute to the noise. Denoting the dark current by $I_d$ the total current is

$$I = I_{ph} + I_d \tag{12.68}$$

This current is accompanied by *shot noise* (a name that is a hangover from vacuum tube days), the mean square spectral density of which is $2qI$ in $A^2$/Hz with $I$ in amperes. This would be for a diode without any internal amplification such as a PIN diode. An *avalanche photo diode* (APD) multiplies the signal current by a factor $M$ and at the same time generates excess noise, represented by an *excess noise factor, F*, so that the mean square spectral density is $(2qI)MF$. $F$ increases with increase in $M$ (see Jones, 1988, p. 240). Typical values are of the order $M = 100$, $F = 12$. For a PIN diode, $M = 1$ and $F = 1$.

The analysis to follow is based on that given in Jones (1988). The equivalent circuit for the input stage of an optical receiver is shown in Fig. 12.14, where $R$ is the parallel combination of diode resistance, input load resistance and preamplifier input resistance, and $C$ is the parallel combination of diode capacitance, preamplifier input capacitance and stray circuit capacitance. Resistance $R$ will generate thermal noise current, the spectral density of which is $4kT/R$ ($k$ is Boltzmann's constant and $T$ is the temperature, which may be taken as room temperature). The preamplifier transistor will also generate shot noise, which when

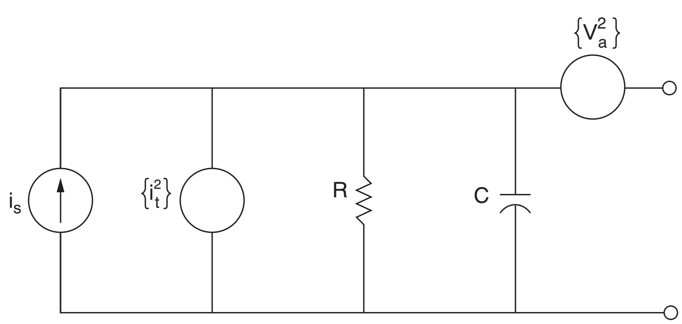



**Figure 12.14** The equivalent input circuit for an optical receiver. $\{i_t^2\}$ is the mean square spectral density for the current noise source, and $\{v_a^2\}$ the mean square spectral density for the voltage noise source. $i_s$ is the signal current.

referred to the input has a spectral density $2q/I_{in}$ in $A^2$/Hz, where $I_{in}$ is the transistor gate (JFET) or base (BJT) current in amperes. Using curly brackets {.} for the spectral density values (see Jones, 1988), the total noise current spectral density at the input is

$$\{i_t^2\} = 2qIM^2F + \frac{4kT}{R} + 2qI_{tr} \ \text{A}^2/\text{Hz} \tag{12.69}$$

The pre-amplifier also has a noise voltage component shown as $v_a$ resulting from the shot noise in the drain or collector current. The mean square noise voltage spectral density is $2qI_{tr}/g_m^2 \text{V}^2$/Hz where $I_{tr}$ is the transistor drain (JFET) or collector (BJT) current in amperes and $g_m$ is the device transconductance in siemens. The total noise voltage spectral density at the input is therefore

$$\{v_n^2\} = \{i_t^2\}|Z_L|^2 + \frac{2qI_{tr}}{g_m^2} \ \text{V}^2/\text{Hz} \tag{12.70}$$

where $Z_L$ is the impedance of $R$ and $C$ in parallel.

The average signal voltage is $MI_{ph}R$. Increasing $R$ should result in an increase in signal to noise ratio since the signal voltage is proportional to $R$ and the $R$ component of noise current density is inversely proportional to $R$. However, the bandwidth of the RC input network is inversely proportional to $R$ and therefore sets a limit to how large $R$ can be. An equalizing filter, which has a transfer function, given by $H_{eq}(f) = 1 + j2\pi f\text{RC}$ can be included in the overall transfer function, which compensates for the input impedance frequency response over the signal bandwidth. The effective spectral density for the mean square noise voltage at the input is then $\{v_n^2\}|H(f)|^2$. The mean square noise voltage $V_n^2$ at the input is obtained by integrating this expression over the signal bandwidth. Only the result will be given here:

$$V_n^2 = \left[\left(2qIM^2F + \frac{4kT}{R} + 2qI_{tr}\right)R^2 + \frac{2qI_{tr}}{g_m^2}\left(1 + \frac{(2\pi\text{RCB})^2}{3}\right)\right]B \tag{12.71}$$

With the average signal voltage given by $V_s = MI_{ph}R$ the signal to noise ratio is

$$\frac{S}{N} = \frac{V_s^2}{V_n^2}$$

$$= \frac{(MI_{ph}R)^2}{\left[\left(2qIM^2F + \frac{4kT}{R} + 2qI_{in}\right)R^2 + \frac{2qI_{tr}}{g_m^2}\left(1 + \frac{(2\pi RCB)^2}{3}\right)\right]B} \tag{12.72}$$

This can be simplified on dividing through by $M^2R^2$ to get:

$$\frac{S}{N} = \frac{I_{ph}^2}{\left[2qIF + \frac{4kT}{M^2R} + \frac{2qI_{in}}{M^2} + \frac{2qI_{tr}}{M^2g_m^2}\left(\frac{1}{R^2} + \frac{(2\pi CB)^2}{3}\right)\right]B} \qquad (12.73)$$

**Example 12.22** An optical receiver utilizes an APD for which $R_0 = 0.65$ A/W, $M = 100$, $F = 4$, $I_{in} = 0$, gm = 3000 mS and $I_{tr} = 0.15$ mA. The dark current may be neglected. The input load consists of a 600 Ω resistor in parallel with a 10pF capacitance. The signal bandwidth is 25 MHz and equalization is employed. Calculate the resultant signal-to-noise ratio for an input signal power of 1 $\mu$W.

**Solution** The photocurrent is $I_{ph} = 0.65 \times 10^{-6} = 0.65$ $\mu$A. The individual terms in the denominator are, with $I_d = 0$ and $I_{in} = 0$:

$$2qIF = 6.408 \times 10^{-25} \text{ A}^2/\text{Hz}$$

$$\frac{4kT}{M^2R} = 0.027 \times 10^{-25} \text{ A}^2/\text{Hz}$$

$$\frac{2qI_{tr}}{M^2g_m^2}\left(\frac{1}{R^2} + \frac{(2\pi \text{CB})^2}{3}\right) = 0.011 \times 10^{-25} \text{ A}^2/\text{Hz}$$

$$\frac{S}{N} = \frac{(0.65 \times 10^{-6})^2 \times 10^{25}}{(6.408 + 0.027 + 0.011) \times 25 \times 10^6}$$

$$= 2.622 \times 10^4$$

In decibels this is 44.2 dB

## 12.13 Problems and Exercises

*Note:* In problems where room temperature is required, assume a value of 290 K. In calculations involving antenna gain, an efficiency factor of 0.55 may be assumed.

**12.1.** Give the decibel equivalents for the following quantities: (*a*) a power ratio of 30:1; (*b*) a power of 230 W; (*c*) a bandwidth of 36 MHz; (*d*) a frequency ratio of 2 MHz/3 kHz; (*e*) a temperature of 200 K.

**12.2.** (*a*) Explain what is meant by EIRP. (*b*) A transmitter feeds a power of 10 W into an antenna which has a gain of 46 dB. Calculate the EIRP in (*i*) watts; (*ii*) dBW.

**12.3.** Calculate the gain of a 3-m parabolic reflector antenna at a frequency of (*a*) 6 GHz; (*b*) 14 GHz.