5 UPLINK, DOWNLINK AND OVERALL LINK PERFORMANCE; INTERSATELLITE LINKS

The types of links considered in this chapter are those shown in Figures 1.1 and 4.1:

— *uplinks* from the earth stations to the satellites;
— *downlinks* from the satellites to the earth stations;
— *intersatellite links* between the satellites.

Uplinks and downlinks consist of radio-frequency modulated carriers, while intersatellite links can be either radio frequency or optical. Carriers are modulated by baseband signals conveying information for communications purposes. Connections between end users entail an uplink and a downlink, and possibly one or several intersatellite links.

The performance of the individual links that participate in the connection between the end terminals conditions the quality of service (QoS) for the connection between end users, specified in terms of bit error rate (BER) for digital communications. Chapter 4 has shown how the BER conditions the required value of the ratio of the energy per information bit to the power spectral density of noise (E_b/N_0) and how this impacts the value of the link performance evaluated as the ratio of the received carrier power, C, to the noise power spectral density, N_0, quoted as the ratio C/N_0, expressed in hertz (Hz). This chapter discusses the parameters that impact link performance C/N_0 and provides the means to evaluate the performance of an individual link given the transmit and receive equipment in the link or to dimension the transmit and receive equipment in order to achieve a given link performance.

We first consider the individual link performance, providing the tools to evaluate the carrier-power budget and the noise-contribution budget. Then we introduce the concept of link performance for the overall link from origin to destination station; for connections supported by transparent satellites and regenerative satellites.

5.1 CONFIGURATION OF A LINK

Figure 5.1 represents the elements participating in a link. The transmit equipment consists of transmitter Tx connected by a feeder to the transmit antenna with gain G_T in the direction of the receiver. The power P_T radiated by the transmit equipment in the direction of the receiving

Satellite Communications Systems, Fifth Edition Gérard Maral, Michel Bousquet and Zhili Sun
© 2009 John Wiley & Sons, Ltd.

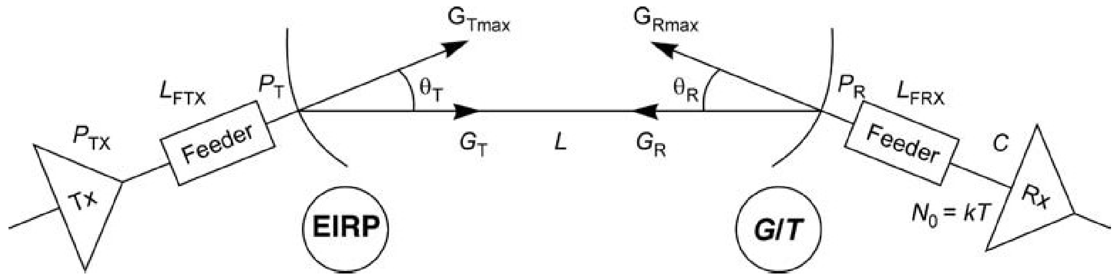

Figure 5.1 Configuration of a link.

equipment and the performance of the transmit equipment is measured by its *effective isotropic radiated power* (EIRP), which is defined as:

$$\mathrm{EIRP} = P_T G_T \quad (\mathrm{W}) \tag{5.1}$$

On its way, the radiated power suffers from path loss L.

The receiving equipment consists of the receive antenna with gain G_R in the direction of the transmit equipment, connected by a feeder to the receiver Rx. At the receiver input, the power of the modulated carrier is C and all sources of noise in the link contribute to the system noise temperature T. This system noise temperature conditions the noise power spectral density N_0, and therefore the link performance C/N_0 can be calculated at the receiver input. The receiving equipment performance is measured by its figure of merit, G/T, where G represents the overall receiving equipment gain.

The following sections present definitions of the relevant parameters that condition the link performance and provide useful equations that permit the calculation of C/N_0.

5.2 ANTENNA PARAMETERS

5.2.1 Gain

The gain of an antenna is the ratio of the power radiated (or received) per unit solid angle by the antenna in a given direction to the power radiated (or received) per unit solid angle by an isotropic antenna fed with the same power. The gain is maximum in the direction of maximum radiation (the electromagnetic axis of the antenna, also called the boresight) and has a value given by:

$$G_{\max} = (4\pi/\lambda^2)A_{\mathrm{eff}} \tag{5.2}$$

where $\lambda = c/f$ and c is the velocity of light $= 3 \times 10^8$ m/s and f is the frequency of the electromagnetic wave. A_{eff} is the effective aperture area of the antenna. For an antenna with a circular aperture or reflector of diameter D and geometric surface $A = \pi D^2/4$, $A_{\mathrm{eff}} = \eta A$, where η is the efficiency of the antenna. Hence:

$$G_{\max} = \eta(\pi D/\lambda)^2 = \eta(\pi Df/c)^2 \tag{5.3}$$

Expressed in dBi (the gain relative to an isotropic antenna), the actual maximum antenna gain is:

$$G_{\max\,\mathrm{dBi}} = 10\log \eta(\pi D/\lambda)^2 = 10\log \eta(\pi Df/c)^2 \quad (\mathrm{dBi})$$

The efficiency η of the antenna is the product of several factors which take account of the illumination law, spill-over loss, surface impairments, ohmic and impedance mismatch losses, and so on:

$$\eta = \eta_i \times \eta_s \times \eta_f \times \eta_z \cdots \tag{5.4a}$$

The *illumination efficiency* η_i specifies the illumination law of the reflector with respect to uniform illumination. Uniform illumination ($\eta_i = 1$) leads to a high level of secondary lobes. A compromise is achieved by attenuating the illumination at the reflector boundaries (aperture edge taper). In the case of a Cassegrain antenna (see Section 8.3.4.3), the best compromise is obtained for an illumination attenuation at the boundaries of 10 to 12 dB which leads to an illumination efficiency η_i of the order of 91%.

The *spill-over efficiency* η_s is defined as the ratio of the energy radiated by the primary source which is intercepted by the reflector to the total energy radiated by the primary source. The difference constitutes the spill-over energy. The larger the angle under which the reflector is viewed from the source, the greater the spill-over efficiency. However, for a given source radiation pattern, the illumination level at the boundaries becomes less with large values of view angle and the illumination efficiency collapses. A compromise leads to a spill-over efficiency of the order of 80%.

The *surface finish efficiency* η_f takes account of the effect of surface roughness on the gain of the antenna. The actual parabolic profile differs from the theoretical one. In practice, a compromise must be found between the effect on the antenna characteristics and the cost of fabrication. The effect on the on-axis gain is of the form:

$$\eta_f = \Delta G = \exp[-B(4\pi\varepsilon/\lambda)^2]$$

where ε is the root mean square (rms) surface error, i.e. the deviation between the actual and theoretical profiles measured perpendicularly to the concave face, and B is a factor, less than or equal to 1, whose value depends on the radius of curvature of the reflector. This factor increases as the radius of curvature decreases. For parabolic antennas of focal distance f, it varies as a function of the ratio f/D, where D is the diameter of the antenna. With $f/D = 0.7$, B is of the order of 0.9 considering ε of the order of $\lambda/30$; the surface finish efficiency η_f is of the order of 85%.

The other losses, including ohmic and impedance mismatch losses, are of less importance. In total, the overall efficiency η, the product of the individual efficiencies, is typically between 55% and 75%.

Figure 5.2 gives values of G_{max} in dBi as a function of diameter for different frequencies. It shows a reference case corresponding to a 1 m antenna at frequency 12 GHz. The corresponding gain is $G_{max} = 40$ dBi. It is easy to derive other cases from this reference case: for instance, dividing the frequency by 2 ($f = 6$ GHz) reduces the gain by 6 dB, so $G_{max} = 34$ dBi. Keeping frequency constant ($f = 12$ GHz) and increasing the size of the antenna by a factor of 2 ($D = 2$ m) increases the gain by 6 dB ($G_{max} = 46$ dBi).

5.2.2 Radiation pattern and angular beamwidth

The radiation pattern indicates the variations of gain with direction. For an antenna with a circular aperture or reflector this pattern has rotational symmetry and is completely represented within a plane in polar coordinate form (Figure 5.3a) or Cartesian coordinate form (Figure 5.3b). The main lobe contains the direction of maximum radiation. Side lobes should be kept to a minimum.

The angular beamwidth is the angle defined by the directions corresponding to a given gain fallout with respect to the maximum value. The 3 dB beamwidth, indicated on Figure 5.3a by $\theta_{3\,dB}$, is often used. The 3 dB beamwidth corresponds to the angle between the directions in which the gain falls to half its maximum value. The 3 dB beamwidth is related to the ratio λ/D by a coefficient

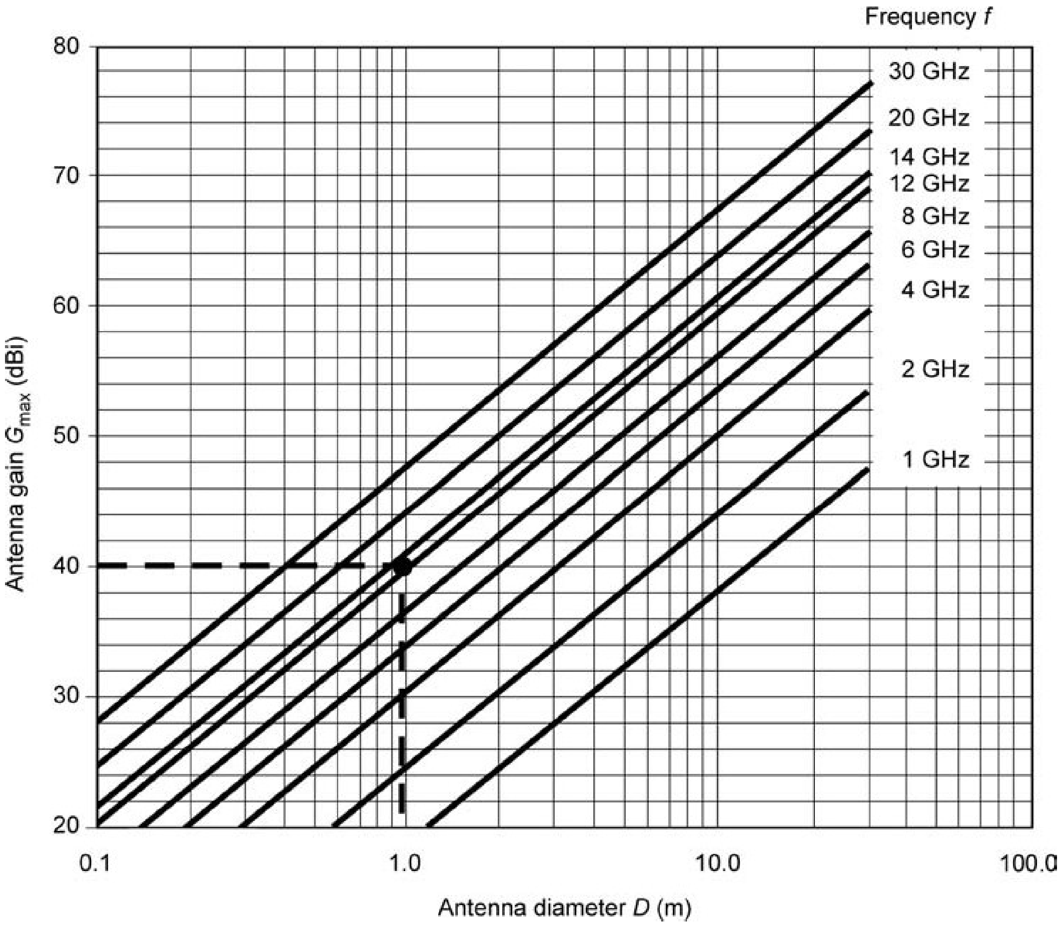

Figure 5.2 Maximum antenna gain as a function of diameter for different frequencies at $\eta = 0.6$. A 1 m antenna at 12 GHz has a gain of 40 dBi.

whose value depends on the chosen illumination law. For uniform illumination, the coefficient has a value of 58.5°. With non-uniform illumination laws, which lead to attenuation at the reflector boundaries, the 3 dB beamwidth increases and the value of the coefficient depends on the particular characteristics of the law. The value commonly used is 70° which leads to the following

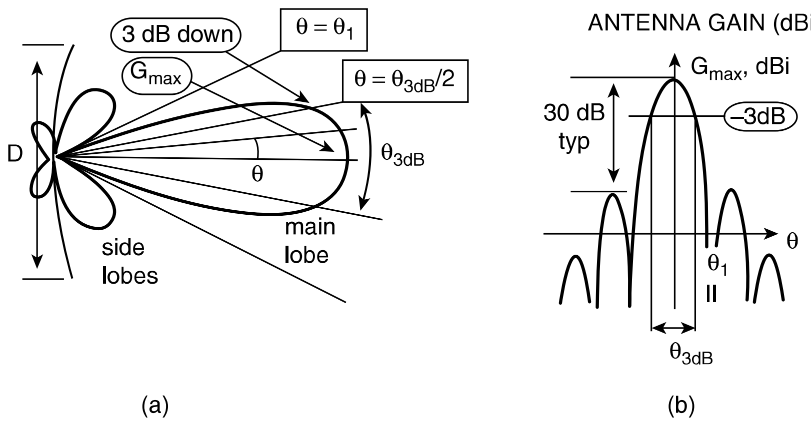

Figure 5.3 Antenna radiation pattern: (a) polar representation and (b) Cartesian representation.

expression:

$$\theta_{3\,\text{dB}} = 70(\lambda/D) = 70(c/f\,D) \quad \text{(degrees)} \tag{5.4b}$$

In a direction θ with respect to the boresight, the value of gain is given by:

$$G(\theta)_{\text{dBi}} = G_{\text{max,dBi}} - 12(\theta/\theta_{3\,\text{dB}})^2 \quad \text{(dBi)} \tag{5.5}$$

This expression is valid only for sufficiently small angles (θ between 0 and $\theta_{3\,\text{dB}}/2$).

Combining equations (5.3) and (5.4b), it can be seen that the maximum gain of an antenna is a function of the 3 dB beamwidth and this relation is independent of frequency:

$$G_{\text{max}} = \eta(\pi Df/c)^2 = \eta(\pi 70/\theta_{3\,\text{db}})^2 \tag{5.6}$$

If a value $\eta = 0.6$ is considered, this gives:

$$G_{\text{max}} = 29\,000/(\theta_{3\,\text{dB}})^2 \tag{5.7}$$

in which $\theta_{3\,\text{dB}}$ is expressed in degrees.

Figure 5.4 shows the relationship between 3 dB beamwidth and maximum gain for three values of antenna efficiency. The gain is expressed in dBi and the 3 dB beamwidth in degrees.

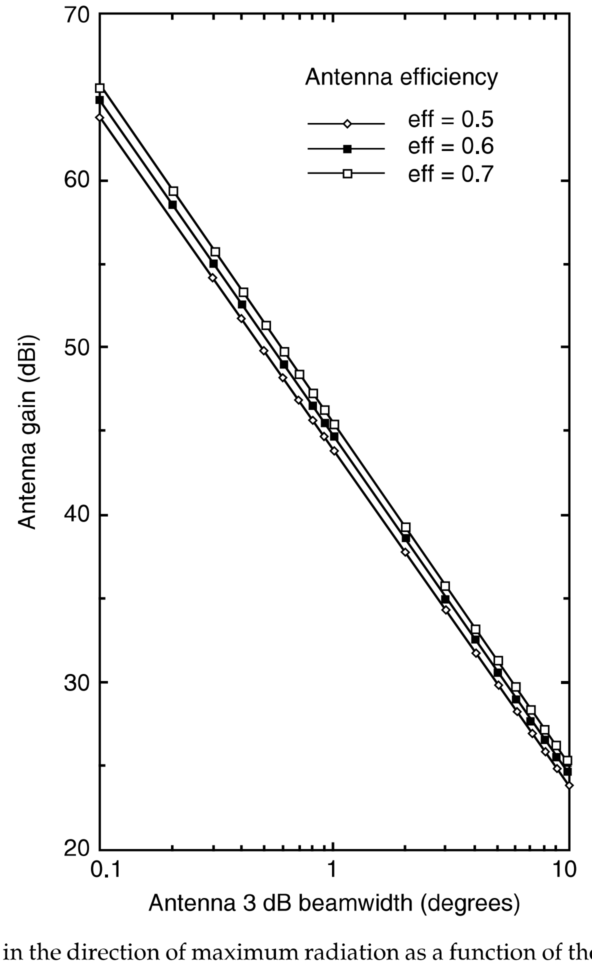

Figure 5.4 Antenna gain in the direction of maximum radiation as a function of the angular beamwidth $\theta_{3\,\text{dB}}$ for three values of efficiency ($\eta = 0.5$, $\eta = 0.6$ and $\eta = 0.7$).

$$G_{\max \text{ d Bi}} = 44.6 - 20 \log \theta_{3 \text{ dB}} \quad \text{(dBi)}$$
$$\theta_{3 \text{ dB}} = 170/10^{\frac{G_{\max \text{ dBi}}}{20}} \quad \text{(degrees)}$$

By differentiating equation (5.5) with respect to θ, we obtain:

$$\frac{dG(\theta)}{d\theta} = -\frac{24\theta}{\theta_{3 \text{ dB}}^2} \quad \text{(dB/deg)}$$

This allows us to calculate the gain fallout ΔG in dB at angle θ degrees from the boresight, for a depointing angle $\Delta\theta$ degrees about the θ direction:

$$\Delta G = -\frac{24\theta}{\theta_{3 \text{ dB}}^2} \Delta\theta \quad \text{(dB)} \tag{5.8}$$

The gain fall-out is maximum at the edge of 3 dB beamwidth ($\theta = \frac{1}{2}\theta_{3 \text{ dB}}$ in the above formula), and is equal to:

$$\Delta G = -\frac{12\Delta\theta}{\theta_{3 \text{ dB}}^2} \quad \text{(dB)} \tag{5.9}$$

5.2.3 Polarisation

The wave radiated by an antenna consists of an electric field component and a magnetic field component. These two components are orthogonal and perpendicular to the direction of propagation of the wave; they vary at the frequency of the wave. By convention, the polarisation of the wave is defined by the direction of the electric field. In general, the direction of the electric field is not fixed; i.e., during one period, the projection of the extremity of the vector representing the electric field onto a plane perpendicular to the direction of propagation of the wave describes an ellipse; the polarisation is said to be elliptical (Figure 5.5).

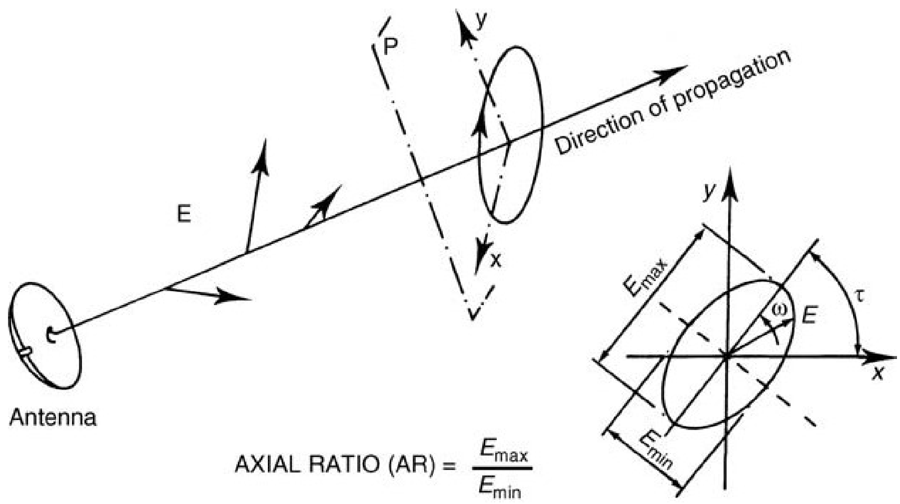

Figure 5.5 Characterisation of the polarisation of an electromagnetic wave.

Polarisation is characterised by the following parameters:

— *direction of rotation* (with respect to the direction of propagation): right-hand (clockwise) or left-hand (counter-clockwise);
— *axial ratio* (AR): AR $= E_{max}/E_{min}$, that is the ratio of the major and minor axes of the ellipse. When the ellipse is a circle (axial ratio $= 1 = 0$ dB), the polarisation is said to be circular. When the ellipse reduces to one axis (infinite axial ratio: the electric field maintains a fixed direction), the polarisation is said to be linear;
— *inclination* τ of the ellipse.

Two waves are in orthogonal polarisation if their electric fields describe identical ellipses in opposite directions. In particular, the following can be obtained:

— two orthogonal circular polarisations described as right-hand circular and left-hand circular (the direction of rotation is for an observer looking in the direction of propagation);
— two orthogonal linear polarisations described as horizontal and vertical (relative to a local reference).

An antenna designed to transmit or receive a wave of given polarisation can neither transmit nor receive in the orthogonal polarisation. This property enables two simultaneous links to be established at the same frequency between the same two locations; this is described as frequency re-use by orthogonal polarisation. To achieve this either, two polarised antennas must be provided at each end or, preferably, one antenna which operates with the two specified polarisations may be used. This practice must, however, take account of imperfections of the antennas and the possible depolarisation of the waves by the transmission medium (Section 5.7.1.2). These effects lead to mutual interference of the two links.

This situation is illustrated in Figure 5.6 which relates to the case of two orthogonal linear polarisations (but the illustration is equally valid for any two orthogonal polarisations). Let a and b be the amplitudes, assumed to be equal, of the electric field of the two waves transmitted simultaneously with linear polarisation, a_C and b_C the amplitudes received with the same polarisation and a_X and b_X the amplitudes received with orthogonal polarisations. The following

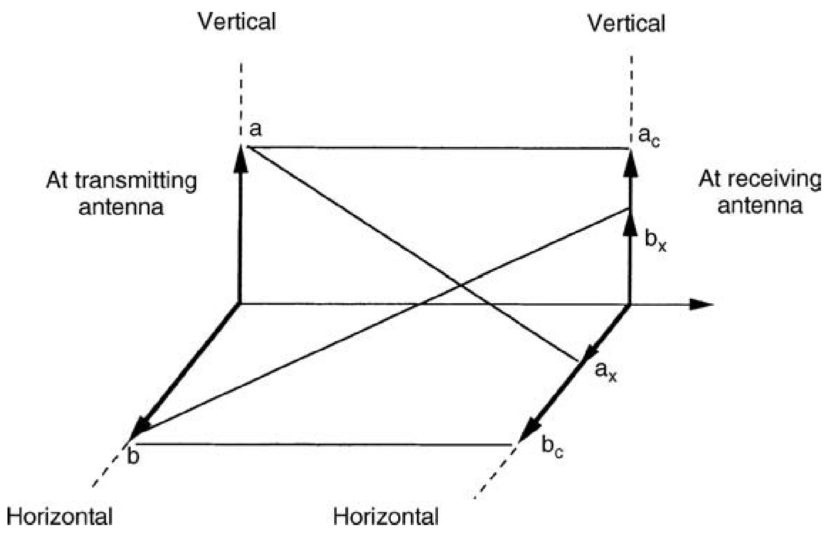

Figure 5.6 Amplitude of the transmitted and received electric field for the case of two orthogonal linear polarisations.

are defined:

— The *cross-polarisation isolation: XPI* $= a_C/b_X$ or b_C/a_X, hence:

$$\text{XPI (dB)} = 20\log(a_C/b_X) \text{ or } 20\log(b_C/a_X) \quad \text{(dB)}$$

— The *cross-polarisation discrimination* (when a single polarisation is transmitted): XPD $= a_C/a_X$, hence:

$$\text{XPD (dB)} = 20\log(a_C/a_X) \quad \text{(dB)}$$

In practice, XPI and XPD are comparable and are often included in the term 'isolation'.

For a quasi-circular polarisation characterised by its value of axial ratio AR, the cross-polarisation discrimination is given by:

$$\text{XPD} = 20\log[(\text{AR}+1)/(\text{AR}-1)] \quad \text{(dB)}$$

Conversely, the axial ratio AR can be expressed as a function of XPD by:

$$\text{AR} = (10^{\text{XPD}/20}+1)/(10^{\text{XPD}/20}-1)$$

The values and relative values of the components vary as a function of direction with respect to the antenna boresight. The antenna is thus characterised for a given polarisation by a radiation pattern for nominal polarisation (copolar) and a radiation pattern for orthogonal polarisation (cross-polar). Cross-polarisation discrimination is generally maximum on the antenna axis and degrades for directions other than that of maximum gain.

5.3 RADIATED POWER

5.3.1 Effective isotropic radiated power (EIRP)

The power radiated per unit solid angle by an isotropic antenna fed from a radio-frequency source of power P_T is given by:

$$P_T/4\pi \quad \text{(W/steradian)}$$

In a direction where the value of transmission gain is G_T, any antenna radiates a power per unit solid angle equal to:

$$G_T P_T/4\pi \quad \text{(W/steradian)}$$

The product $P_T G_T$ is called the 'effective isotropic radiated power' (EIRP). It is expressed in W.

5.3.2 Power flux density

A surface of area A situated at a distance R from the transmitting antenna subtends a solid angle A/R^2 at the transmitting antenna (see Figure 5.7). It receives a power equal to:

$$P_R = (P_T G_T/4\pi)(A/R^2) = \Phi A \quad \text{(W)} \tag{5.10}$$

The magnitude $\Phi = P_T G_T/4\pi R^2$ is called the *power flux density*. It is expressed in W/m^2.

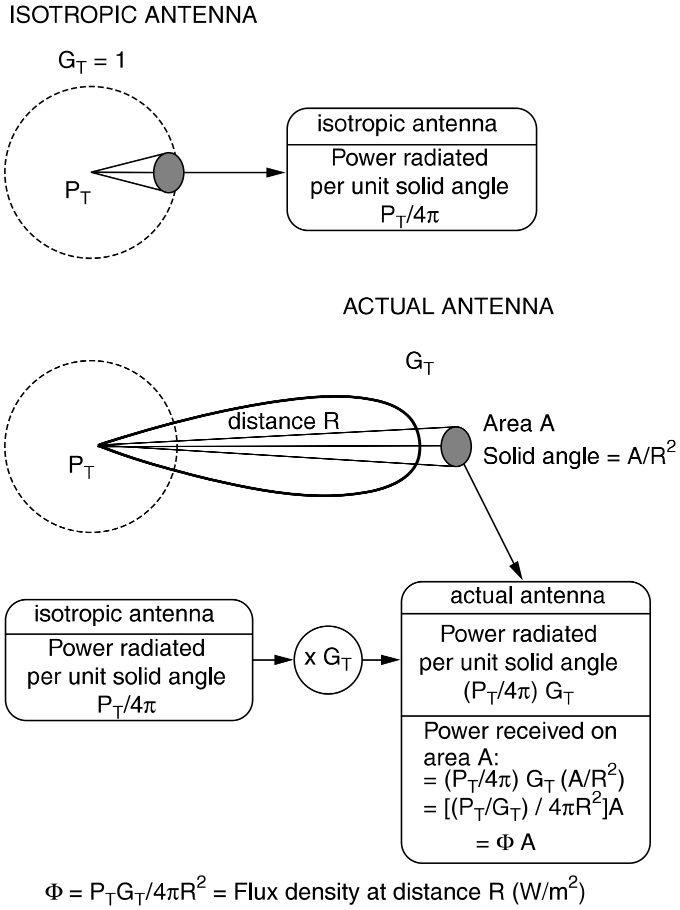

Figure 5.7 Power flux density.

5.4 RECEIVED SIGNAL POWER

5.4.1 Power captured by the receiving antenna and free space loss

As shown in Figure 5.8, a receiving antenna of effective aperture area A_{Reff} located at a distance R from the transmitting antenna receives power equal to:

$$P_{\text{R}} = \Phi A_{R\ \text{eff}} = (P_{\text{T}} G_{\text{T}}/4\pi R^2) A_{\text{R eff}} \quad (\text{W}) \tag{5.11}$$

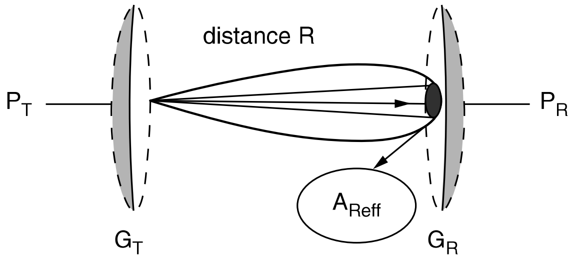

Figure 5.8 The power received by a receiving antenna.

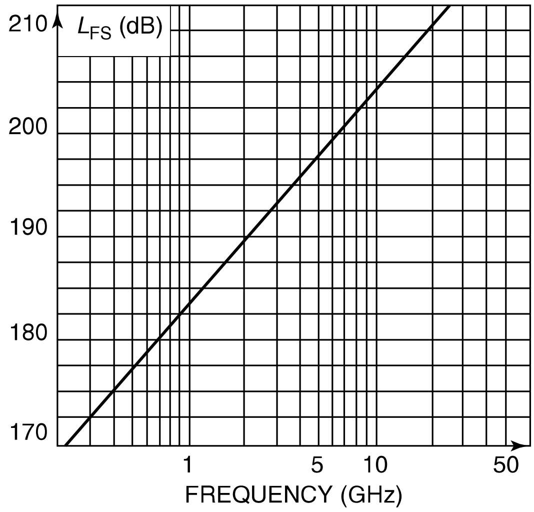

Figure 5.9 Free space loss attenuation at geostationary sub-satellite point: $L_{FS}(R_0)$.

The effective area of an antenna is expressed as a function of its receiving gain G_R according to equation (5.2):

$$A_{R\,eff} = G_R/(4\pi/\lambda^2) \quad (m^2) \tag{5.12}$$

Hence an expression for the received power:

$$\begin{aligned} P_R &= (P_T G_T/4\pi R^2)(\lambda^2/4\pi)G_R \\ &= (P_T G_T)(\lambda/4\pi R)^2 G_R \\ &= (P_T G_T)(1/L_{FS})G_R \quad (W) \end{aligned} \tag{5.13}$$

where $L_{FS} = (4\pi R/\lambda)^2$ is called the *free space loss* and represents the ratio of the received and transmitted powers in a link between two isotropic antennas. Figure 5.9 gives the value of $L_{FS}(R_0)$ as a function of frequency for a geostationary satellite and a station situated at the sub-satellite point at a distance $R = R_0 = 35\,786$ km equal to the altitude of the satellite. Notice that L_{FS} is of the order of 200 dB. For any station whose position is represented by its relative latitude and longitude l and L with respect to the geostationary satellite (since the satellite is situated in the equatorial plane, l is the geographical latitude of the station), the value of $L_{FS}(R_0)$ provided by Figure 5.9 must be corrected by the term $(R/R_0)^2$, hence:

$$L_{FS} = (4\pi R/\lambda)^2 = (4\pi R_0/\lambda)^2 (R/R_0)^2 = L_{FS}(R_0)(R/R_0)^2$$

where $(R/R_0)^2 = 1 + 0.42(1 - \cos l \cos L)$ (see Chapter 2, equation (2.62)). The value of $(R/R_0)^2$ is between 1 and 1.356 (0 to 1.3 dB).

5.4.2 Example 1: Uplink received power

Consider the transmitting antenna of an earth station equipped with an antenna of diameter $D = 4$ m. This antenna is fed with a power P_T of 100 W, that is 20 dBW, at a frequency $f_U = 14$ GHz. It radiates this power towards a geostationary satellite situated at a distance of 40 000 km from the station on the axis of the antenna. The beam of the satellite receiving antenna has a width $\theta_{3\,dB} = 2°$.

It is assumed that the earth station is at the centre of the region covered by the satellite antenna and consequently benefits from the maximum gain of this antenna. The efficiency of the satellite antenna is assumed to be $\eta = 0.55$ and that of the earth station to be $\eta = 0.6$.

— The power flux density at the satellite situated at earth station antenna boresight is calculated as:

$$\Phi_{\text{max}} = P_{\text{T}}G_{\text{Tmax}}/4\pi R^2 \quad (\text{W/m}^2)$$

The gain of the earth station antenna, from equation (5.3), is:

$$\begin{aligned} G_{\text{Tmax}} &= \eta(\pi D/\lambda_{\text{U}})^2 = \eta(\pi D f_{\text{U}}/c)^2 \\ &= 0.6(\pi \times 4 \times 14 \times 10^9/3 \times 10^8)^2 = 206\,340 = 53.1\ \text{dBi} \end{aligned}$$

The effective isotropic radiated power of the earth station (on the axis) is given by:

$$(\text{EIRP}_{\text{max}})_{\text{ES}} = P_{\text{T}}G_{\text{Tmax}} = 53.1\ \text{dBi} + 20\ \text{dBW} = 73.1\ \text{dBW}$$

The power flux density is given by:

$$\begin{aligned} \Phi_{\text{max}} &= P_{\text{T}}G_{\text{Tmax}}/4\pi R^2 = 73.1\ \text{dBW} - 10\log(4\pi(4 \times 10^7)^2) \\ &= 73.1 - 163 = -89.9\ \text{dBW/m}^2 \end{aligned}$$

— The power received (in dBW) by the satellite antenna is obtained using equation (5.13):

$$P_R = \text{EIRP} - \text{attenuation of free space} + \text{gain of receiving antenna}$$

The attenuation of free space $L_{\text{FS}} = (4\pi R/\lambda_{\text{U}})^2 = (4\pi R f_{\text{U}}/c)^2 = 207.4$ dB.
The gain of the satellite receiving antenna $G_{\text{R}} = G_{\text{Rmax}}$ is obtained using equation (5.3):

$$G_{\text{Rmax}} = \eta(\pi D/\lambda_{\text{U}})^2$$

The value of D/λ_{U} is obtained using equation (5.7), hence $\theta_{3\,\text{dB}} = 70(\lambda_{\text{U}}/D)$, from which

$$D/\lambda_{\text{U}} = 70/\theta_{3\,\text{dB}} \text{ and } G_{\text{Rmax}} = \eta(70\pi/\theta_{3\,\text{dB}})^2 = 6650 = 38.2\ \text{dBi}.$$

Notice that the antenna gain does not depend on frequency when the beamwidth, and hence the area covered by the satellite antenna, is imposed. In total:

$$P_{\text{R}} = 73.1 - 207.4 + 38.2 = -96.1\ \text{dBW}, \quad \text{that is } 0.25\ \text{nW or } 250\ \text{pW}$$

5.4.3 Example 2: Downlink received power

Consider the transmitting antenna of a geostationary satellite fed with a power P_{T} of 10 W, that is, 10 dBW at a frequency $f_{\text{D}} = 12$ GHz, and radiating this power in a beam of width $\theta_{3\,\text{dB}}$ equal to 2°. An earth station equipped with a 4 m diameter antenna is located on the axis of the antenna at a distance of 40 000 km from the satellite. The efficiency of the satellite antenna is assumed to be $\eta = 0.55$ and that of the earth station to be $\eta = 0.6$.

— The power flux density at the earth station situated at the satellite antenna boresight is calculated as:

$$\Phi_{\text{max}} = P_{\text{T}}G_{\text{Tmax}}/4\pi R^2 \quad (\text{W/m}^2)$$

The gain of the satellite antenna is the same in transmission as in reception since the beamwidths are made the same (notice that this requires two separate antennas on the satellite since the diameters cannot be the same and are in the ratio $f_U/f_D = 14/12 = 1.17$). Hence:

$$(\text{EIRP}_{\max})_{\text{SL}} = P_T G_{T\max} = 38.2\,\text{dBi} + 10\,\text{dBW} = 48.2\,\text{dBW}$$

The power flux density is:

$$\begin{aligned}\Phi_{\max} &= P_T G_{T\max}/4\pi R^2 = 48.2\,\text{dBW} - 10\log(4\pi(4\times 10^7)^2) = 48.2 - 163\\ &= -114.8\,\text{dBW/m}^2\end{aligned}$$

— The power (in dBW) received by the antenna of the earth station is obtained using equation (5.13):

$$P_R = \text{EIRP} - \text{attenuation of free space} + \text{gain of the receiving antenna}$$

The attenuation of free space is $L_{FS} = (4\pi R/\lambda_D)^2 = 206.1$ dB.

The gain $G_R = G_{R\max}$ of the ground station receiving antenna is obtained using equation (5.3), hence:

$$G_{R\max} = \eta(\pi D/\lambda_D)^2 = 0.6(\pi \times 4/0.025)^2 = 151\,597 = 51.8\,\text{dB}$$

In total:

$$P_R = 48.2 - 206.1 + 51.8 = -106.1\,\text{dBW}, \quad \text{that is } 25\,\text{pW}$$

5.4.4 Additional losses

In practice, it is necessary to take account of additional losses due to various causes:

— attenuation of waves as they propagate through the atmosphere;
— losses in the transmitting and receiving equipment;
— depointing losses;
— polarisation mismatch losses.

5.4.4.1 Attenuation in the atmosphere

The attenuation of waves in the atmosphere, denoted by L_A, is due to the presence of gaseous components in the troposphere, water (rain, clouds, snow and ice) and the ionosphere. A quantitative presentation of these effects is given in Section 5.7. The overall effect on the power of the received carrier can be taken into account by replacing L_{FS} in equation (5.13) by the path loss, L, where:

$$L = L_{FS} L_A \tag{5.14}$$

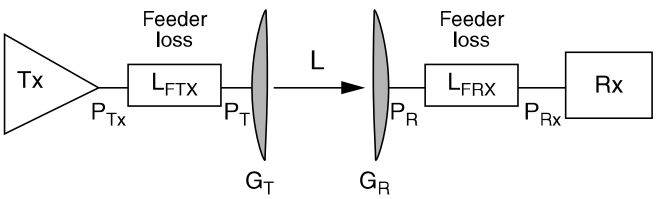

Figure 5.10 Losses in the terminal equipment.

5.4.4.2 Losses in the transmitting and receiving equipment

Figure 5.10 clarifies these losses:

— The feeder loss L_{FTX} between the transmitter and the antenna: to feed the antenna with a power P_T it is necessary to provide a power P_{TX} at the output of the transmission amplifier such that:

$$P_{TX} = P_T L_{FTX} \quad (W) \tag{5.15}$$

Expressed as a function of the rated power of the transmission amplifier, the EIRP can be written:

$$\text{EIRP} = P_T G_T = (P_{TX} G_T)/L_{FTX} \quad (W) \tag{5.16}$$

— The feeder loss L_{FRX} between the antenna and the receiver: the signal power P_{RX} at the input of the receiver is equal to:

$$P_{RX} = P_R/L_{FRX} \quad (W) \tag{5.17}$$

5.4.4.3 Depointing losses

Figure 5.11 shows the geometry of the link for the case of imperfect alignment of the transmitting and receiving antennas. The result is a fallout of antenna gain with respect to the maximum gain on transmission and on reception, called *depointing loss*. These depointing losses are a function of the misalignment of angles of transmission (θ_T) and reception (θ_R) and are evaluated using equation (5.5). Their value is given by:

$$\begin{aligned} L_T &= 12(\theta_T/\theta_{3\,dB})^2 \quad (dB) \\ L_R &= 12(\theta_R/\theta_{3\,dB})^2 \quad (dB) \end{aligned} \tag{5.18}$$

5.4.4.4 Losses due to polarisation mismatch

It is also necessary to consider the polarisation mismatch loss L_{POL} observed when the receiving antenna is not oriented with the polarisation of the received wave. In a link with circular polarisation, the transmitted wave is circularly polarised only on the axis of the antenna and becomes elliptical off this axis. Propagation through the atmosphere can also change circular into

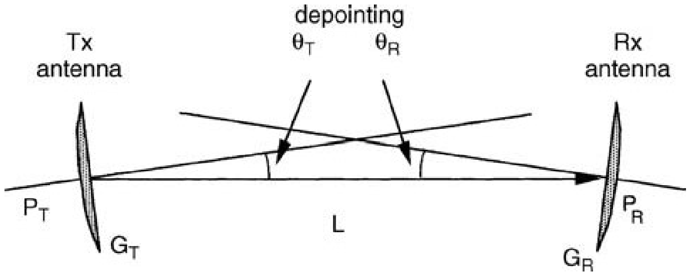

Figure 5.11 Geometry of the link.

elliptical polarisation (see Section 5.7). In a linearly polarised link, the wave can be subjected to a rotation of its plane of polarisation as it propagates through the atmosphere. Finally, with linear polarisation, the receiving antenna may not have its plane of polarisation aligned with that of the incident wave. If ψ is the angle between the two planes, the polarisation mismatch loss L_{POL} (in dB) is equal to $-20\log\cos\psi$. In the case where a circularly polarised antenna receives a linearly polarised wave, or a linearly polarised antenna receives a circularly polarised wave, L_{POL} has a value of 3 dB. Considering all sources of loss, the signal power at the receiver input is given by:

$$P_{RX} = (P_{TX}G_{Tmax}/L_T L_{FTX})(1/L_{FS}L_A)(G_{Rmax}/L_R L_{FRX} L_{POL}) \quad \text{(W)} \tag{5.19}$$

5.4.5 Conclusion

Equations (5.13) and (5.19), which express the received power at the input to the receiver, are of the same form; they are the product of three factors:

— EIRP, which characterises the transmitting equipment:

$$\text{EIRP} = (P_{TX}G_{Tmax}/L_T L_{FTX}) \quad \text{(W)}$$

This expression takes account of the losses L_{FTX} between the transmission amplifier and the antenna and the reduction in antenna gain L_T due to misalignment of the transmitting antenna.

— 1/L, which characterises the transmission medium;

$$1/L = 1/L_{FS}L_A$$

The path loss L takes account of the attenuation of free space L_{FS} and the attenuation in the atmosphere L_A.

— the gain of the receiver, which characterises the receiving equipment:

$$G = G_{Rmax}/L_R L_{FRX} L_{POL}$$

This expression takes account of the losses L_{FRX} between the antenna and the receiver, the loss of antenna gain L_R due to misalignment of the receiving antenna and the polarisation mismatch losses L_{POL}.

5.5 NOISE POWER SPECTRAL DENSITY AT THE RECEIVER INPUT

5.5.1 The origins of noise

Noise consists of all unwanted contributions whose power adds to the wanted carrier power. It reduces the ability of the receiver to reproduce correctly the information content of the received wanted carrier.

The origins of noise are as follows:

— the noise emitted by natural sources of radiation located within the antenna reception area;
— the noise generated by components in the receiving equipment.

Carriers from transmitters other than those which it is wished to receive are also classed as noise. This noise is described as interference.

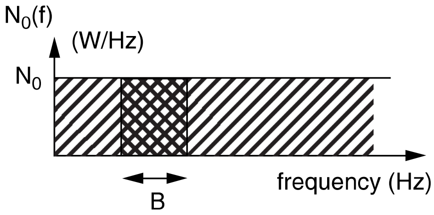

Figure 5.12 Spectral density of white noise.

5.5.2 Noise characterisation

Harmful noise power is that which occurs in the bandwidth B of the wanted modulated carrier. A popular noise model is that of white noise, for which the power spectral density N_0 (W/Hz) is constant in the frequency band involved (Figure 5.12). The equivalent noise power N (W) captured by a receiver with equivalent noise bandwidth B_N, usually matched to B ($B = B_N$), is given by:

$$N = N_0 B_N \quad \text{(W)} \tag{5.20}$$

Real noise sources do not always have a constant power spectral density, but the model is convenient for representation of actual noise observed over a limited bandwidth.

5.5.2.1 *Noise temperature of a noise source*

The noise temperature of a two-port noise source delivering an available noise power spectral density N_0 is given by:

$$T = N_0/k \text{ (K)} \tag{5.21}$$

where k is Boltzmann's constant $= 1.379 \times 10^{-23} = -228.6\,\text{dBW/Hz K}$, T represents the thermodynamic temperature of a resistance which delivers the same available noise power as the source under consideration (Figure 5.13). Available noise power is the power delivered by the source to a device which is impedance matched to the source.

5.5.2.2 *Effective input noise temperature*

The effective input noise temperature T_e of a four-port element is the thermodynamic temperature of a resistance which, placed at the input of the element assumed to be noise-free, establishes the same available noise power at the output of the element as the actual element without the noise

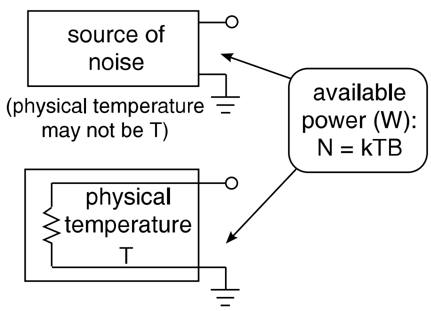

Figure 5.13 Definition of the noise temperature of a noise source.

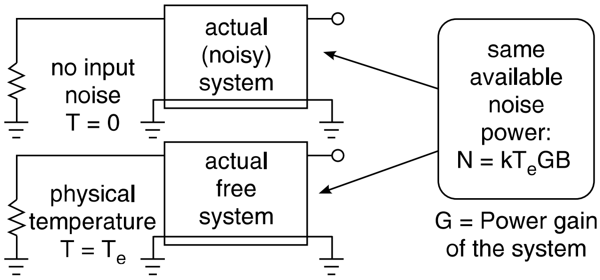

Figure 5.14 Effective input noise temperature of a four-port element.

source at the input (Figure 5.14). T_e is thus a measure of the noise generated by the internal components of the four-port element.

The noise figure of this four-port element is the ratio of the total available noise power at the output of the element to the component of this power engendered by a source at the input of the element with a noise temperature equal to the reference temperature $T_0 = 290\,\mathrm{K}$.

Assume that the element has a power gain G, a bandwidth B and is driven by a source of noise temperature T_0; the total power at the output is $Gk(T_e + T_0)B$. The component of this power originating from the source is GkT_0B. The noise figure is thus:

$$F = [Gk(T_e + T_0)B]/[GkT_0B] = (T_e + T_0)/T_0 = 1 + T_e/T_0 \tag{5.22}$$

The noise figure is usually quoted in decibels (dB), according to:

$$F(\mathrm{dB}) = 10 \log F$$

Figure 5.15 displays the relationship between noise temperature and noise figure (dB).

5.5.2.3 Effective input noise temperature of an attenuator

An attenuator is a four-port element containing only passive components (which can be classed as resistances) all at temperature T_{ATT} which is generally the ambient temperature. If L_{ATT} is the attenuation caused by the attenuator, the effective input noise temperature of the attenuator is:

$$T_{eATT} = (L_{ATT} - 1)T_{ATT} \quad (\mathrm{K}) \tag{5.23}$$

If $T_{ATT} = T_0$, the noise figure of the attenuator from a comparison of equations (5.22) and (5.23), is:

$$F_{ATT} = L_{ATT}$$

5.5.2.4 Effective input noise temperature of cascaded elements

Consider a chain of N four-port elements in cascade, each element j having a power gain $G_j (j = 1,2,\ldots,N)$ and an effective input noise temperature T_{ej}.

The overall effective input noise temperature is:

$$T_e = T_{e1} + T_{e2}/G_1 + T_{e3}/G_1G_2 + \cdots + T_{iN}/G_1G_2,\ldots,G_{N-1} \quad (\mathrm{K}) \tag{5.24}$$

The noise figure is obtained from equation (5.22):

$$F = F_1 + (F_2 - 1)/G_1 + (F_3 - 1)/G_1G_2 + \cdots + (F_N - 1)/G_1G_2,\ldots,G_{N-1} \tag{5.25}$$

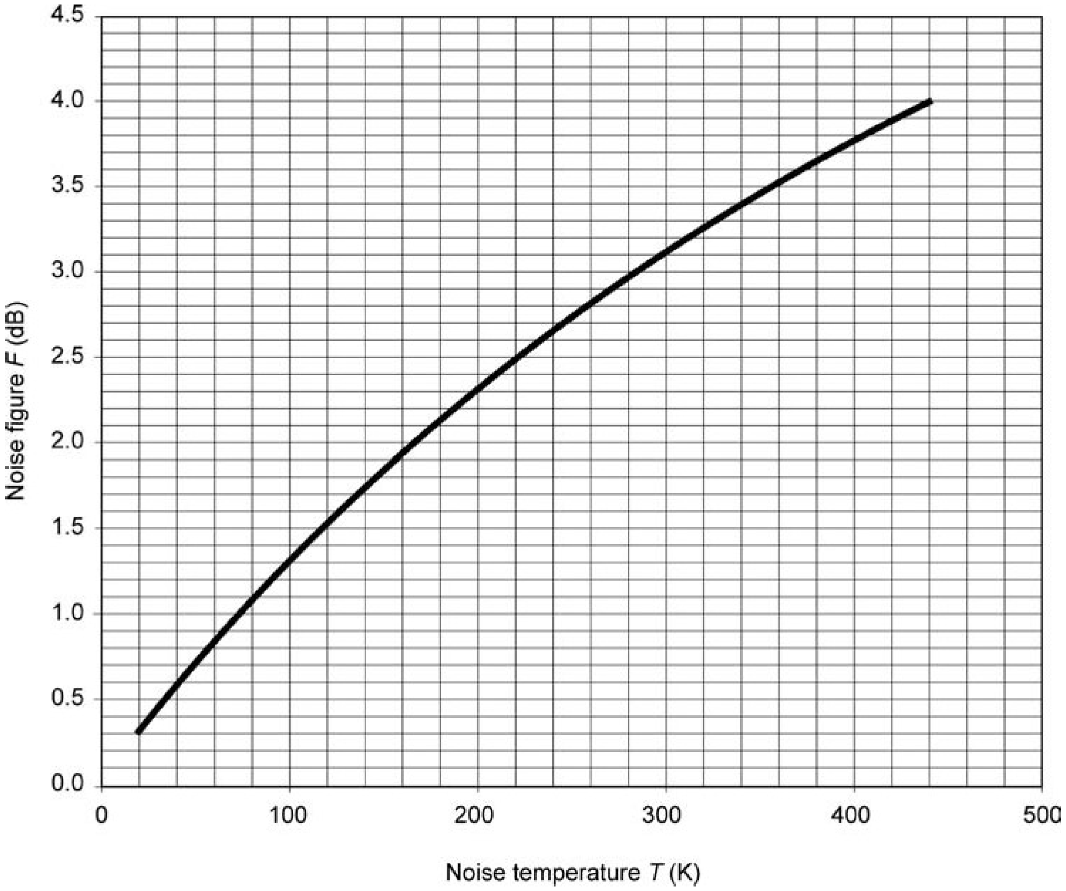

Figure 5.15 Noise figure versus noise temperature: $F(\text{dB}) = 10\log(1 + T/T_0)$ with $T_0 = 290\,\text{K}$.

5.5.2.5 *Effective input noise temperature of a receiver*

Figure 5.16 shows the arrangement of the receiver. By using equation (5.24), the effective input noise temperature T_{eRX} of the receiver can be expressed as:

$$T_{eRX} = T_{LNA} + T_{MX}/G_{LNA} + T_{IF}/G_{LNA}G_{MX} \quad (\text{K}) \tag{5.26}$$

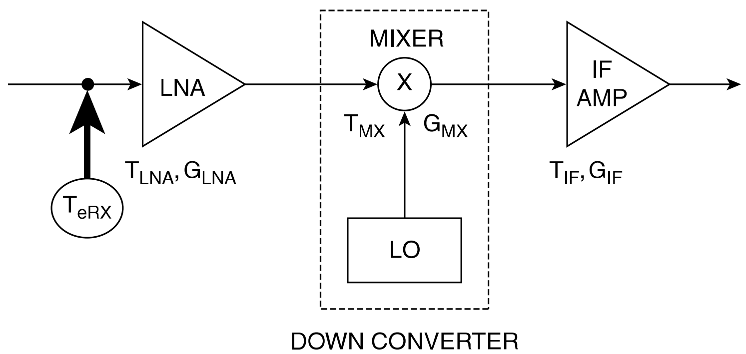

Figure 5.16 The organisation of a receiver.

Example:

Low noise amplifier (LNA): $T_{\mathrm{LNA}} = 150\,\mathrm{K}, G_{\mathrm{LNA}} = 50\,\mathrm{dB}$

Mixer: $T_{\mathrm{MX}} = 850\,\mathrm{K}, G_{\mathrm{MX}} = -10\,\mathrm{dB}\ (L_{\mathrm{MX}} = 10\,\mathrm{dB})$

IF amplifier: $T_{\mathrm{IF}} = 400\,\mathrm{K}, G_{\mathrm{IF}} = 30\,\mathrm{dB}$

Hence:

$$\begin{aligned} T_{\mathrm{eRX}} &= 150 + 850/10^5 + 400/10^5 10^{-1} \\ &= 150\,\mathrm{K} \end{aligned}$$

Notice the benefit of the high gain of the low noise amplifier which limits the noise temperature T_{eRX} of the receiver to that of the low noise amplifier T_{LNA}.

5.5.3 Noise temperature of an antenna

An antenna picks up noise from radiating bodies within the radiation pattern of the antenna. The noise output from the antenna is a function of the direction in which it is pointing, its radiation pattern and the state of the surrounding environment. The antenna is assumed to be a noise source characterised by a noise temperature called the noise temperature of the antenna T_A (K).

Let $T_{\mathrm{b}}(\theta, \varphi)$ be the brightness temperature of a radiating body located in a direction (θ, φ), where the gain of the antenna has a value $G(\theta, \varphi)$. The noise temperature of the antenna is obtained by integrating the contributions of all the radiating bodies within the radiation pattern of the antenna. The noise temperature of the antenna is thus:

$$T_{\mathrm{A}} = (1/4\pi)\iint T_{\mathrm{b}}(\theta, \varphi)G(\theta, \varphi)\sin\theta\,\mathrm{d}\theta\,\mathrm{d}\varphi \quad (\mathrm{K}) \tag{5.27}$$

There are two cases to be considered:

— a satellite antenna (the uplink);
— an earth station antenna (the downlink).

5.5.3.1 *Noise temperature of a satellite antenna (uplink)*

The noise captured by the antenna is noise from the earth and from outer space. The beamwidth of a satellite antenna is equal to or less than the angle of view of the earth from the satellite, that is 17.5° for a geostationary satellite. Under these conditions, the major contribution is that from the earth. For a beamwidth $\theta_{3\,\mathrm{dB}}$ of 17.5°, the antenna noise temperature depends on the frequency and the orbital position of the satellite (see Figure 5.17). For a smaller width (a spot beam), the temperature depends on the frequency and the area covered; the continents radiate more noise than the oceans as shown in Figure 5.18. For a preliminary design, the value 290 K can be taken as a conservative value.

5.5.3.2 *Noise temperature of an earth station antenna (the downlink)*

The noise captured by the antenna consists of noise from the sky and noise due to radiation from the earth. Figure 5.19 shows the situation.

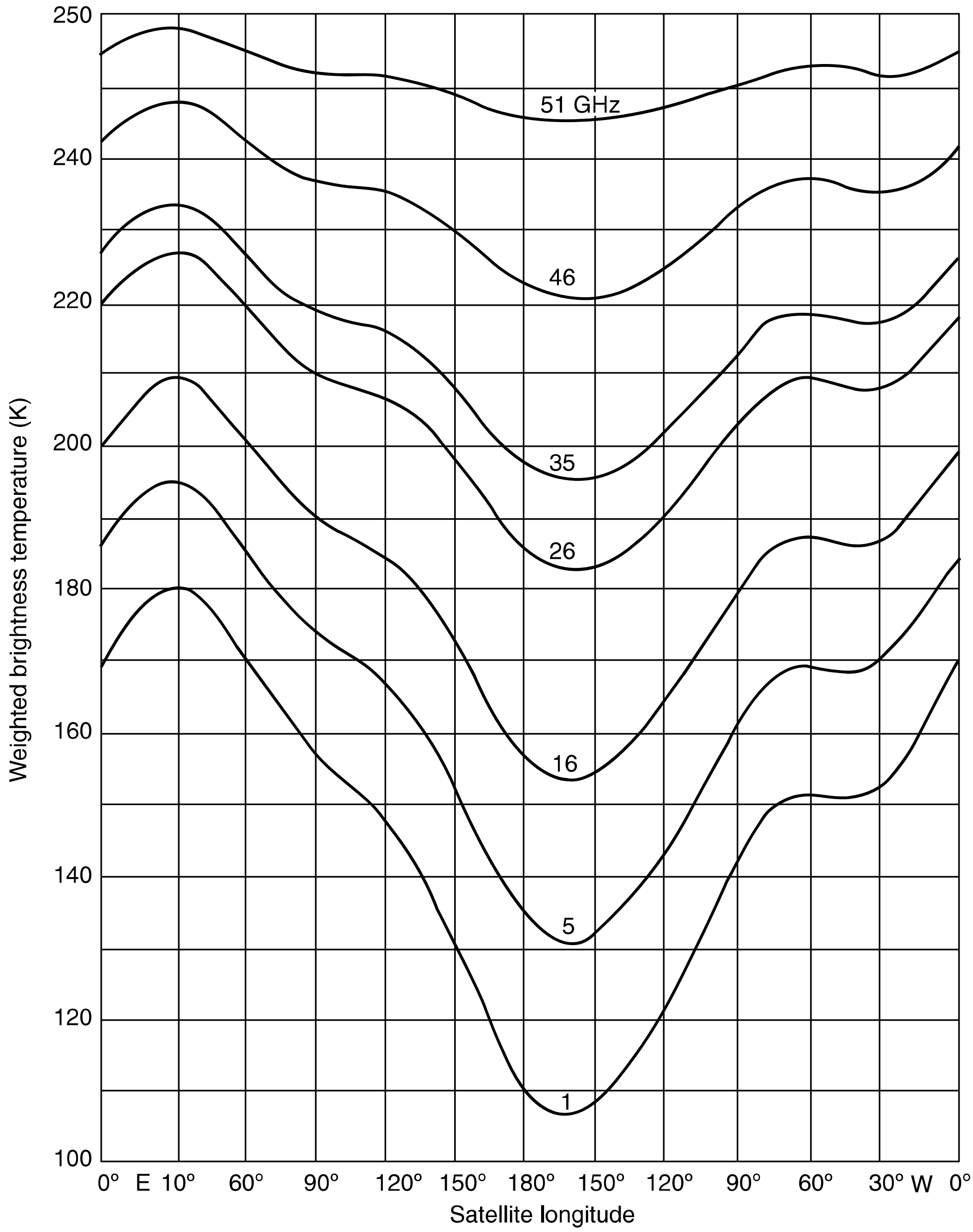

Figure 5.17 Satellite antenna noise temperature for global coverage as a function of frequency and orbital position. (From [NJ0-85]. Reproduced by permission of the American Geophysical Union.)

5.5.3.2.1 'Clear sky' conditions

At frequencies greater than 2 GHz, the greatest contribution is that of the non-ionised region of the atmosphere which, being an absorbent medium, is a noise source. In the absence of meteorological formations (conditions described as 'clear sky'), the antenna noise temperature contains contributions due to the sky and the surrounding ground (Figure 5.19a).

The sky noise contribution is determined from equation (5.27), where $T_b(\theta, \varphi)$ is the brightness temperature of the sky in the direction (θ, φ). In practice, only that part of the sky in the direction of the antenna boresight contributes to the integral as the gain has a high value only in that direction. As a consequence, the noise contribution of the clear sky T_{SKY} can be assimilated with the

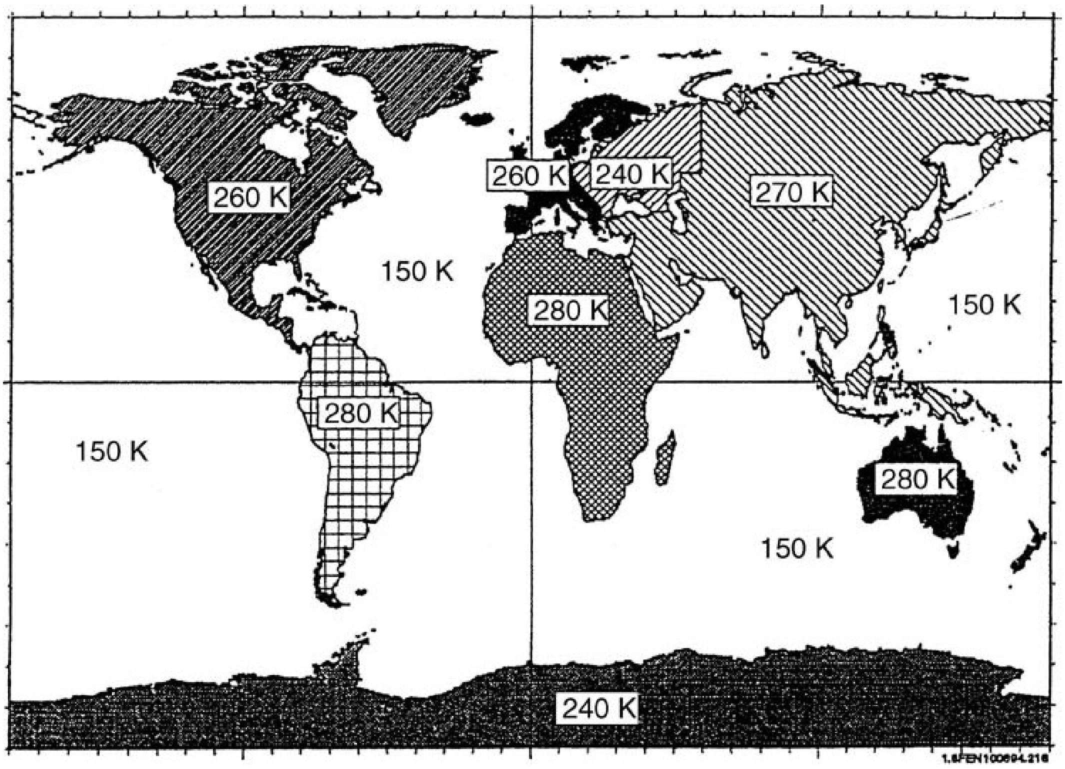

Figure 5.18 The ESA/EUTELSAT model of the earth's brightness temperature at Ku band. (From [FEN-95]. Reproduced by permission.)

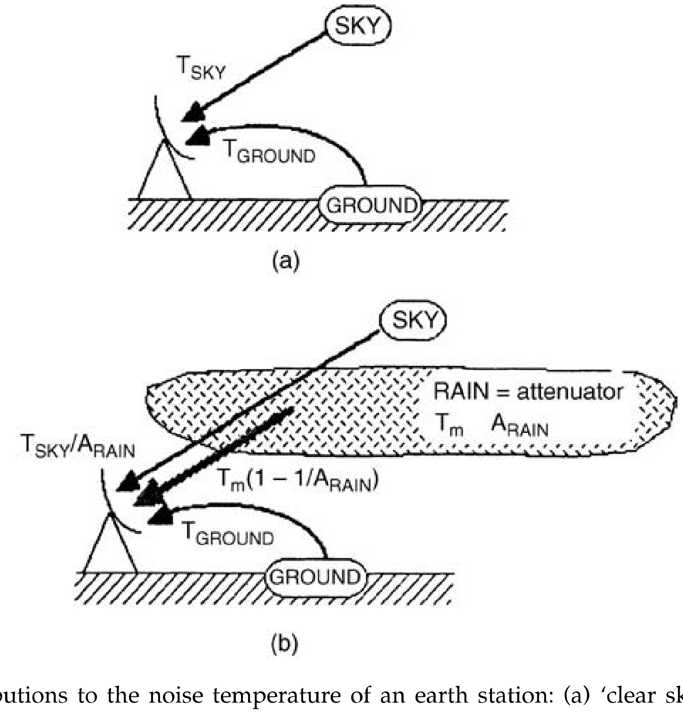

Figure 5.19 Contributions to the noise temperature of an earth station: (a) 'clear sky' conditions and (b) conditions of rain.

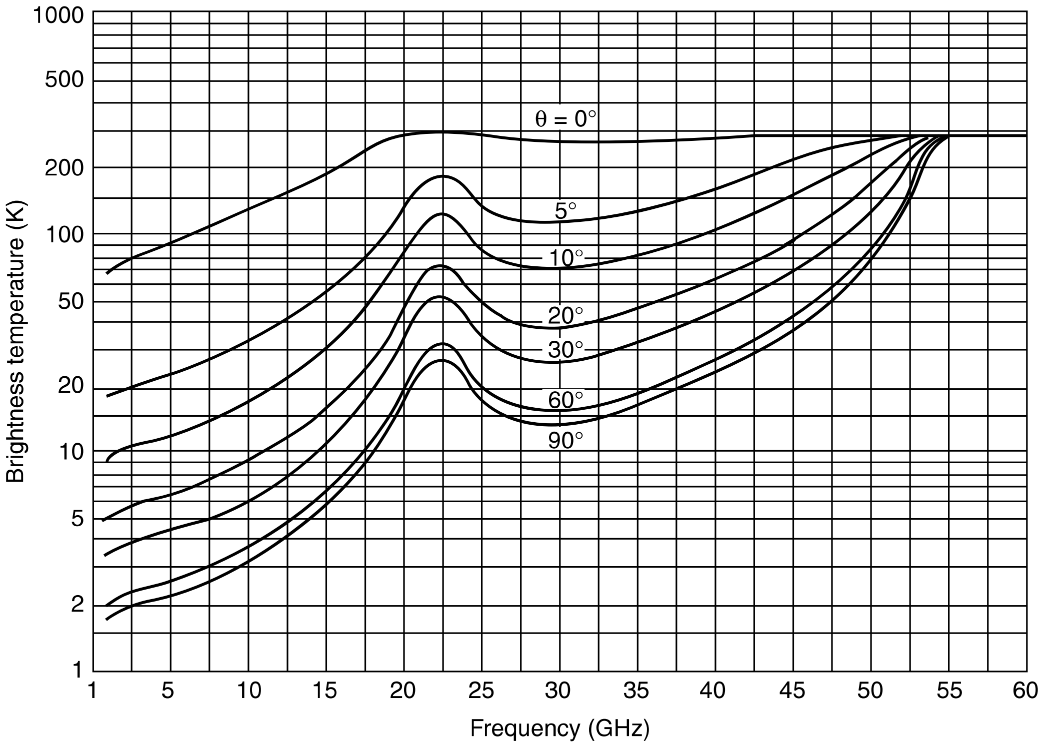

Figure 5.20 Brightness temperature of clear sky as a function of frequency and elevation angle E for mean atmospheric humidity (7.5 g/cm^3 humidity at ground level) and standard temperature and pressure conditions at ground level. (From CCIR Rep 720–2. Reproduced by permission of the ITU.)

brightness temperature for the angle of elevation of the antenna. Figure 5.20 shows the clear sky brightness temperature as a function of frequency and elevation angle.

Radiation from the ground in the vicinity of the earth station is captured by the side lobes of the antenna radiation pattern and partly by the main lobe when the elevation angle is small. The contribution of each lobe is determined by $T_i = G_i(\Omega_i/4\pi)T_G$, where G_i is the mean gain of the lobe of solid angle Ω_i and T_G the brightness temperature of the ground. The sum of these contributions yields the value T_{GROUND}. The following can be taken as a first approximation (CCIR Rep. 390):

$-T_G = 290$ K for lateral lobes whose elevation angle E is less than $-10°$
$-T_G = 150$ K for $-10° < E < 0°$
$-T_G = 50$ K for $0° < E < 10°$
$-T_G = 10$ K for $10° < E < 90°$

The antenna noise temperature is thus given by:

$$T_A = T_{SKY} + T_{GROUND} \quad \text{(K)} \tag{5.28}$$

To this noise may be added that of individual sources which are located in the vicinity of the antenna boresight. For a radio source of apparent angular diameter α and noise temperature T_n at the frequency considered and measured at ground level after attenuation by the atmosphere, the additional noise temperature ΔT_A for an antenna of beamwidth $\theta_{3\,dB}$ is given by:

$$\begin{aligned} \Delta T_A &= T_n(\alpha/\theta_{3\,dB})^2 \quad \text{if} \quad \theta_{3\,dB} > \alpha \quad \text{(K)} \\ \Delta T_A &= T_n \quad\quad\quad\quad\;\; \text{if} \quad \theta_{3\,dB} < \alpha \quad \text{(K)} \end{aligned} \tag{5.29}$$

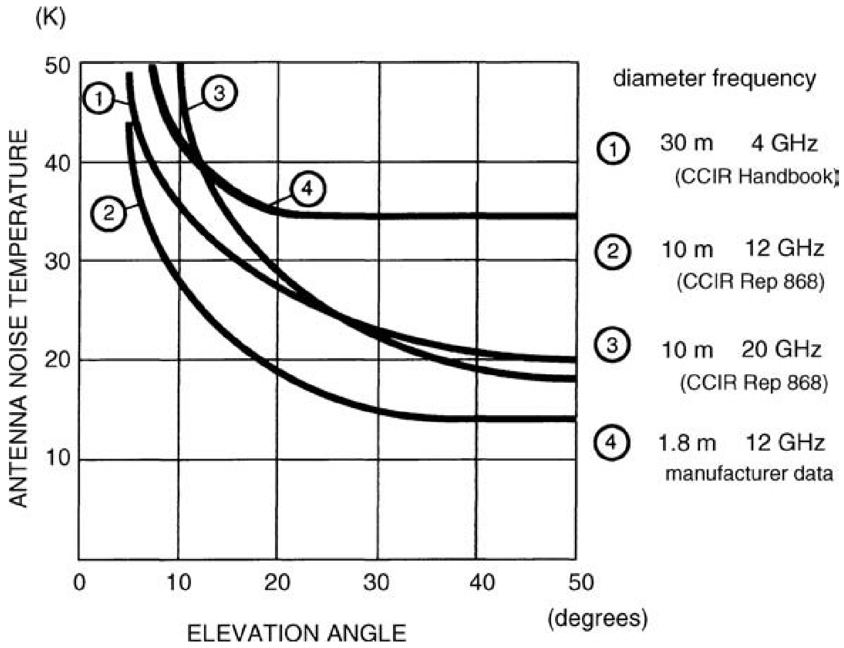

Figure 5.21 Typical values of antenna noise temperature T_A as a function of elevation angle E., Curve 1 : diameter = 30 m, frequency = 4 GHz (From [ITU-85]. Reproduced by permission of the ITU). Curve 2 : diameter = 10 m, frequency = 12 GHz. Curve 3 : diameter = 10 m, frequency = 20 GHz. Curve 4 : diameter = 1.8 m, frequency = 12 GHz (Alcatel Telspace).

For earth stations pointing towards a geostationary satellite, only the sun and the moon needto be considered. The sun and the moon have an apparent angular diameter of 0.5°. There is an increase of noise temperature when these heavenly bodies are aligned with the earth station pointing towards the satellite. This particular geometrical configuration can be predicted. To be more specific, at 12 GHz, a 13 m antenna undergoes a noise temperature increase due to the sun, at a time of quiet sun, by an amount ΔT_A equal to 12 000 K. The conditions of occurrence and the value of ΔT_A as a function of the antenna diameter and frequency are discussed in detail in Chapters 2 and 8. For the moon, the increase is at most 250 K at 4 GHz (CCIR Rep. 390).

Figure 5.21 shows the variation of antenna noise temperature T_A as a function of elevation angle E for various types of antenna at different frequencies in a clear sky (CCIR Rep. 868 and [ITU-85]). It can be seen that the antenna noise temperature decreases as the elevation angle increases.

5.5.3.2.2 Conditions of rain

The antenna noise temperature increases during the presence of meteorological formations, such as clouds and rain (Figure 5.19b), which constitute an absorbent, and consequently emissive, medium. Using equation (5.23), the antenna noise temperature becomes:

$$T_A = T_{SKY}/A_{RAIN} + T_m(1-1/A_{RAIN}) + T_{GROUND} \quad (K) \tag{5.30}$$

where A_{RAIN} is the attenuation and T_m the mean thermodynamic temperature of the formations in question. For T_m, a value of 275 K can be assumed ([THO-83] and CCIR Rep. 564).

5.5.3.2.3 Conclusion

In conclusion, the antenna noise temperature T_A, is a function of:

— the frequency,
— the elevation angle,
— the atmospheric conditions (clear sky or rain).

Consequently, the figure of merit of an earth station must be specified for particular conditions of frequency, elevation angle and atmospheric conditions.

5.5.4 System noise temperature

Consider the receiving equipment shown in Figure 5.22. This consists of an antenna connected to a receiver. The connection (feeder) is a lossy one and is at a thermodynamic temperature T_F (which is close to $T_0 = 290$ K). It introduces an attenuation L_{FRX}, which corresponds to a gain $G_{FRX} = 1/L_{FRX}$ and is less than 1 ($L_{FRX} \geq 1$). The effective input noise temperature T_e of the receiver is T_{eRX}.

The noise temperature may be determined at two points as follows:

— at the antenna output, before the feeder losses, temperature T_1;
— at the receiver input, after the losses, temperature T_2.

The noise temperature T_1 at the antenna output is the sum of the noise temperature of the antenna T_A and the noise temperature of the subsystem consisting of the feeder and the receiver in cascade. The noise temperature of the feeder is given by equation (5.23). From equation (5.24) the noise temperature of the subsystem is $(L_{FRX}-1)T_F + T_{eRX}/G_{FRX}$. Adding the contribution of the antenna, considered as a noise source, this becomes:

$$T_1 = T_A + (L_{FRX}-1)T_F + T_{eRX}/G_{FRX} \quad \text{(K)} \tag{5.31}$$

Now consider the receiver input. This noise must be attenuated by a factor L_{FRX}. Replacing G_{FRX} by $1/L_{FRX}$, one obtains the noise temperature T_2 at the receiver input:

$$T_2 = T_1/L_{FRX} = T_A/L_{FRX} + T_F(1-1/L_{FRX}) + T_{eRX} \quad \text{(K)} \tag{5.32}$$

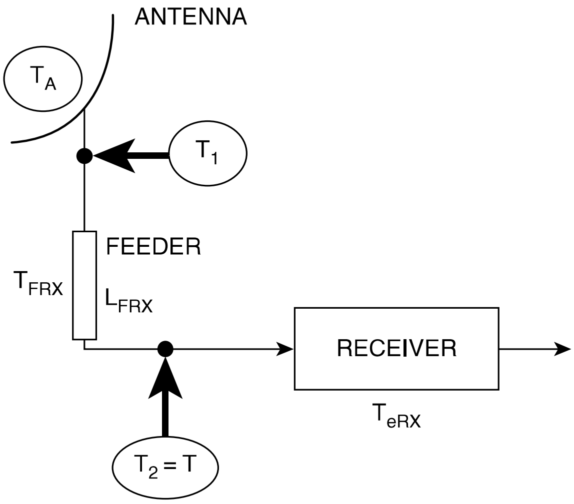

Figure 5.22 A receiving system: T is the system noise temperature at the receiver input.

This noise temperature T_2, which takes account of the noise generated by the antenna and the feeder together with the receiver noise, is called the *system noise temperature* T at the receiver input. Notice that measurement of noise at the considered point would reflect only the noise contribution upstream of this point. Actually, the system noise temperature takes into account all sources of noise within the receiving equipment.

5.5.5 System noise temperature: Example

Consider the receiving system of Figure 5.22 with the following values:

— Antenna noise temperature: $T_A = 50\,\mathrm{K}$;
— Thermodynamic temperature of the feeder: $T_F = 290\,\mathrm{K}$;
— Effective input noise temperature of the receiver: $T_{eRX} = 50\,\mathrm{K}$.

The system noise temperature at the receiver input will be calculated for two cases: (1) no feeder loss between the antenna and the receiver and (2) feeder loss $L_{FRX} = 1$ dB. Using equation (5.31), $T = T_A/L_{FRX} + T_F(1-1/L_{FRX}) + T_{eRX}$:

For case (1), $T = 50 + 50 = 100\,\mathrm{K}$

For case (2), $T = 50/10^{0.1} + 290(1-1/10^{0.1}) + 50 = 39.7 + 59.6 + 50 = 149.3\,\mathrm{K}$ or around 150 K

Notice the influence of the feeder loss; it reduces the antenna noise but makes its own contribution to the noise and this finally causes an increase in the system noise temperature. The contribution of an attenuation to the noise can quickly be estimated using the following rule: every attenuation of 0.1 dB upstream of the receiver makes a contribution to the system noise temperature at the receiver input of $290(1-1/10^{0.01}) = 6.6\,\mathrm{K}$ or around 7 K. To realise a receiving system with a low noise temperature, it is imperative to avoid losses upstream of the receiver.

5.5.6 Conclusion

At the receiver input, all sources of noise in the link contribute to the system noise temperature T. Those sources include the noise captured by the antenna and generated by the feeder, which can actually be measured at the receiver input, plus the noise generated downstream in the receiver, which is modelled as a fictitious source of noise at the receiver input, treating the receiver as noiseless.

The noise superimposed on the received carrier power has a power spectral density given by:

$$N_0 = kT \quad (\mathrm{W/Hz}) \tag{5.33}$$

where k is the Boltzmann constant ($k = 1.379 \times 10^{-23}\,\mathrm{J/K} = -228.6\,\mathrm{dBJ/K}$).

5.6 INDIVIDUAL LINK PERFORMANCE

The link performance is evaluated as the ratio of the received carrier power, C, to the noise power spectral density, N_0, and is quoted as the C/N_0 ratio, expressed in hertz. One can evaluate the link performance using other ratios besides C/N_0; for instance:

— C/T represents the carrier power over the system noise temperature; expressed in units of watts per Kelvin (W/K), it is given by $C/T = (C/N_0)k$, where k is the Boltzmann constant.
— C/N represents the carrier power over the noise power; dimensionless, it is given by $C/N = (C/N_0)(1/B_N)$, where B_N is the receiver noise bandwidth.

5.6.1 Carrier power to noise power spectral density ratio at receiver input

The power received at the receiver input, as given by equation (5.19), is that of the carrier. Hence

$$C = P_{RX}$$

The noise power spectral density at the same point is $N_0 = kT$, where T is given by equation (5.32). Hence:

$$\begin{aligned} C/N_0 = {} & [P_{TX}G_{Tmax}/L_T L_{FTX})(1/L_{FS}L_A)(G_{Rmax}/L_R L_{FRX} L_{POL})] \\ & /[T_A/L_{FRX} + T_F(1-1/L_{FRX}) + T_{eRX}](1/k) \quad \text{(Hz)} \end{aligned} \tag{5.34}$$

This expression can be interpreted as follows:

$$\begin{aligned} C/N_0 = {} & \text{(transmitter EIRP) (1/path loss)} \\ & \times\text{(composite receiving gain/noise temperature)} \times (1/k) \quad \text{(Hz)} \end{aligned} \tag{5.35}$$

C/N_0 can also be expressed as a function of the power flux density Φ:

$$C/N_0 = \Phi(\lambda^2/4\pi)\text{(composite receiving gain/noise temperature)}\,(1/k) \quad \text{(Hz)} \tag{5.36}$$

where $\Phi = \text{(transmitter EIRP)}/(4\pi R^2) \quad (\text{W/m}^2)$

Finally, it can be verified that evaluation of C/N_0 is independent of the point chosen in the receiving chain as long as the carrier power and the noise power spectral density are calculated at the same point.

Equation (5.35) for C/N_0 introduces three factors:

— EIRP, which characterises the transmitting equipment:
— 1/L, which characterises the transmission medium;
— the composite receiving gain/noise temperature, which characterises the receiving equipment; it is called the *figure of merit*, or *G/T*, of the receiving equipment.

By examining equation (5.34) it can be seen that the figure of merit *G/T* of the receiving equipment is a function of the antenna noise temperature T_A and the effective input noise temperature T_{eRX} of the receiver. These magnitudes will now be quantified.

In conclusion, equation (5.34) boils down to:

$$C/N_0 = \text{(EIRP)}(1/L)(G/T)(1/k) \quad \text{(Hz)} \tag{5.37}$$

5.6.2 Clear sky uplink performance

Figure 5.23 shows the geometry of the uplink. It is assumed that the transmitting earth station is on the edge of the 3 dB coverage of the satellite receiving antenna.

The data are as follows:

— Frequency: $f_U = 14\,\text{GHz}$
— For the earth station (ES):
  Transmitting amplifier power: $P_{TX} = 100\,\text{W}$
  Loss between amplifier and antenna: $L_{FTX} = 0.5\,\text{dB}$
  Antenna diameter: $D = 4\,\text{m}$
  Antenna efficiency: $\eta = 0.6$
  Maximum pointing error: $\theta_T = 0.1°$
— Earth station–satellite distance: $R = 40\,000\,\text{km}$
— Atmospheric attenuation: $L_A = 0.3\,\text{dB}$ (typical value for attenuation by atmospheric gases at this frequency for an elevation angle of 10°)

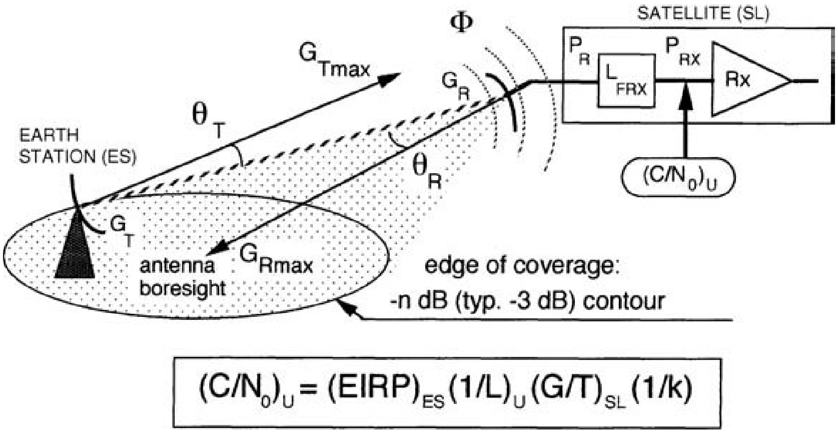

Figure 5.23 The geometry of an uplink.

— For the satellite (SL):
  Receiving beam half power angular width: $\theta_{3\,dB} = 2^\circ$
  Antenna efficiency: $\eta = 0.55$
  Receiver noise figure: $F = 3$ dB
  Loss between antenna and receiver: $L_{FRX} = 1$ dB
  Thermodynamic temperature of the connection: $T_F = 290$ K
  Antenna noise temperature: $T_A = 290$ K

To calculate the EIRP of the earth station:

$$(\text{EIRP})_{ES} = (P_{TX}G_{Tmax}/L_T L_{FTX}) \quad (\text{W}) \tag{5.38}$$

with:

$$\begin{aligned} P_{TX} &= 100\,\text{W} = 20\,\text{dBW} \\ G_{Tmax} &= \eta(\pi D/\lambda_U)^2 = \eta(\pi D f_U/c)^2 = 0.6\,[\pi \times 4 \times (14 \times 10^9)/(3 \times 10^8)]^2 = 206\,340 \\ &= 53.1\,\text{dBi} \\ L_T(\text{dB}) &= 12(\theta_T/\theta_{3\,dB})^2 = 12(\theta_T D f_U/70c)^2 = 0.9\,\text{dB} \\ L_{FTX} &= 0.5\,\text{dB} \end{aligned}$$

Hence:

$$(\text{EIRP})_{ES} = 20\,\text{dBW} + 53.1\,\text{dB} - 0.9\,\text{dB} - 0.5\,\text{dB} = 71.7\,\text{dBW}$$

To calculate the attenuation on the upward path (U):

$$L_U = L_{FS}L_A \tag{5.39}$$

with:

$$\begin{aligned} L_{FS} &= (4\pi R/\lambda_U)^2 = (4\pi R f_U/c)^2 = 5.5 \times 10^{20} = 207.4\,\text{dB} \\ L_A &= 0.3\,\text{dB} \end{aligned}$$

Hence:

$$L_U = 207.4\,\text{dB} + 0.3\,\text{dB} = 207.7\,\text{dB}$$

To calculate the figure of merit G/T of the satellite (SL):

$$(G/T)_{SL} = (G_{Rmax}/L_R L_{FRX} L_{POL})/[T_A/L_{FRX} + T_F(1-1/L_{FRX}) + T_{eRX}] \quad (K^{-1}) \tag{5.40}$$

with:

$$\begin{aligned} G_{R\,max} &= \eta(\pi D/\lambda_U)^2 = \eta(\pi.70/\theta_{3\,dB})^2 = 0.55(\pi.70/2)^2 = 6650 = 38.2\,\text{dBi} \\ L_R &= 12(\theta_R/\theta_{3\,dB})^2 \end{aligned}$$

As the earth station is on the edge of the 3 dB coverage area, $\theta_R = \theta_{3\,dB}/2$ and $L_R = 3$ dB.

$$\begin{aligned} \text{Assume } L_{POL} &= 0\,\text{dB} \\ L_{FRX} &= 1\,\text{dB} \\ \text{Given} \quad T_A &= 290\,\text{K} \\ T_F &= 290\,\text{K} \\ T_{eRX} &= (F-1)T_0 = (10^{0.3}-1)290 = 290\,\text{K} \end{aligned}$$

Hence:

$$\begin{aligned} (G/T)_{SL} &= 38.2-3-1-10\log[290/10^{0.1} + 290(1-1/10^{0.1}) + 290] \\ &= 6.6\,\text{dBK}^{-1} \end{aligned}$$

Notice that when the thermodynamic temperature of the feeder between the antenna and the satellite receiver is close to the antenna noise temperature, which is the case in practice, the uplink system noise temperature at the receiver input is $T_U \approx T_F + T_{eRX} \approx 290 + T_{eRX}$. It is, therefore, needlessly costly to install a receiver with a very low noise figure on board a satellite.

To calculate the ratio C/N_0 for the uplink:

$$(C/N_0)_U = (\text{EIRP})_{ES}(1/L_U)(G/T)_{SL}(1/k) \quad (\text{Hz}) \tag{5.41}$$

Hence:

$$(C/N_0)_U = 71.7\,\text{dBW} - 207.7\,\text{dB} + 6.6\,\text{dBK}^{-1} + 228.6\,\text{dBW/HzK} = 99.2\,\text{dBHz}$$

Figure 5.24 summarises the variations in power level throughout the path.

5.6.3 Clear sky downlink performance

Figure 5.25 shows the geometry of the downlink. It is assumed that the receiving earth station is located on the edge of the 3 dB coverage area of the satellite receiving antenna. The data are as follows:

— Frequency: $f_D = 12$ GHz
— For the satellite (SL):
  Transmitting amplifier power: $P_{TX} = 10$ W
  Loss between amplifier and antenna: $L_{FTX} = 1$ dB
  Transmitting beam half power angular width: $\theta_{3\,dB} = 2°$
  Antenna efficiency: $\eta = 0.55$
— Earth station–satellite distance: $R = 40\,000$ km
— Atmospheric attenuation: $L_A = 0.3$ dB (typical attenuation by atmospheric gases at this frequency for an elevation angle of 10°)

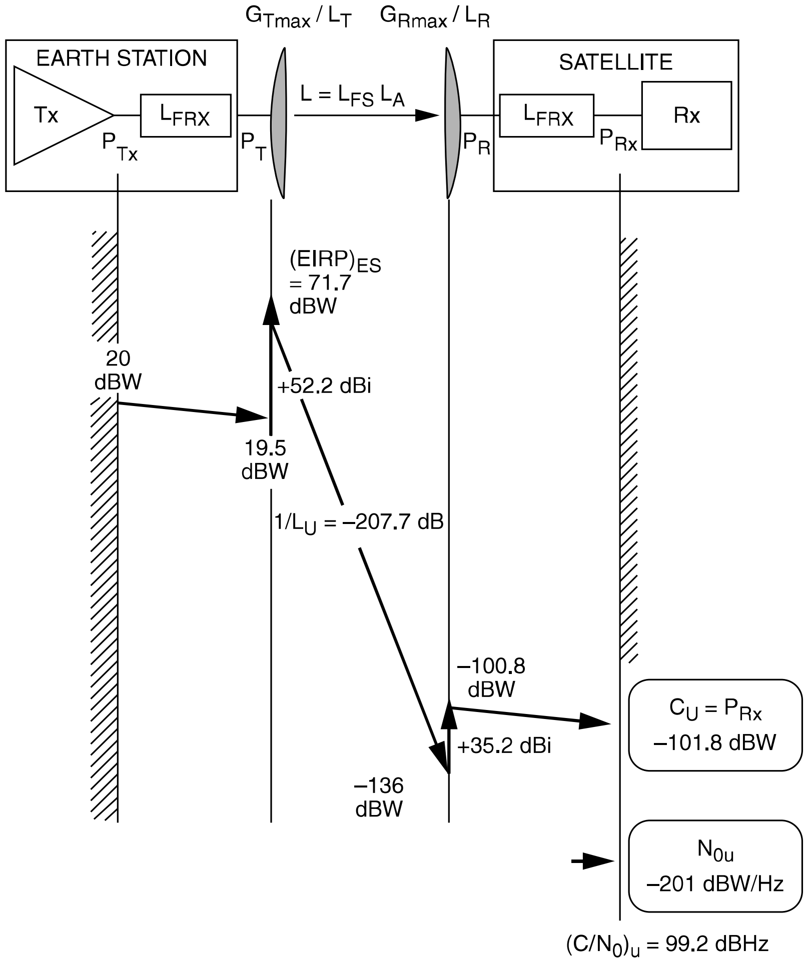

Figure 5.24 Variations in power for the clear sky uplink.

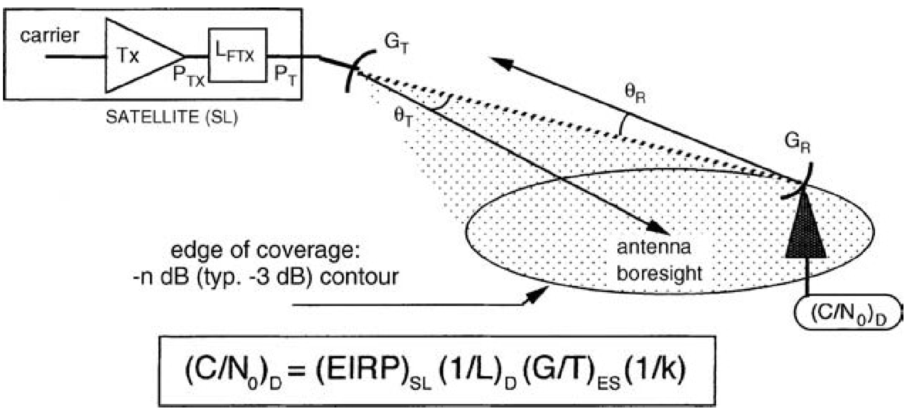

Figure 5.25 The geometry of a downlink.

— For the earth station (ES):
  Receiver noise figure: $F = 1$ dB
  Loss between antenna and receiver: $L_{FRX} = 0.5$ dB
  Thermodynamic temperature of the feeder: $T_F = 290$ K

Antenna diameter: $D = 4\,\text{m}$
Antenna efficiency: $\eta = 0.6$
Maximum pointing error: $\theta_R = 0.1^\circ$
Ground noise temperature: $T_{\text{GROUND}} = 45\,\text{K}$

To calculate the EIRP of the satellite:

$$(\text{EIRP})_{\text{SL}} = P_{\text{TX}} G_{\text{Tmax}} / L_{\text{T}} L_{\text{FTX}} \quad (\text{W}) \tag{5.42}$$

with:

$$\begin{aligned} P_{\text{TX}} &= 10\,\text{W} = 10\,\text{dBW} \\ G_{\text{Tmax}} &= \eta(\pi D/\lambda_{\text{D}})^2 = \eta(\pi 70/\theta_{3\,\text{dB}})^2 = 0.55(\pi 70/2)^2 = 6650 = 38.2\,\text{dBi} \\ L_{\text{T}}(\text{dB}) &= 3\,\text{dB (earth station on edge of coverage)} \\ L_{\text{FTX}} &= 1\,\text{dB} \end{aligned}$$

Hence:

$$(\text{EIRP})_{\text{SL}} = 10\,\text{dBW} + 38.2\,\text{dBi} - 3\,\text{dB} - 1\,\text{dB} = 44.2\,\text{dBW}$$

To calculate the attenuation on the downlink (D):

$$L_{\text{D}} = L_{\text{FS}} L_{\text{A}} \tag{5.43}$$

with:

$$\begin{aligned} L_{\text{FS}} &= (4\pi R/\lambda_{\text{D}})^2 = (4\pi R f_{\text{D}}/c)^2 = 4.04 \times 10^{20} = 206.1\,\text{dB} \\ L_{\text{A}} &= 0.3\,\text{dB} \end{aligned}$$

Hence:

$$L_{\text{D}} = 206.1\,\text{dB} + 0.3\,\text{dB} = 206.4\,\text{dB}$$

To calculate the figure of merit G/T of the earth station in the satellite direction:

$$(G/T)_{\text{ES}} = (G_{\text{Rmax}}/L_{\text{R}} L_{\text{FRX}} L_{\text{POL}})/T_{\text{D}} \quad (K^{-1})$$

T_{D} is the downlink system noise temperature at the receiver input given by:

$$T_{\text{D}} = T_{\text{A}}/L_{\text{FRX}} + T_{\text{F}}(1 - 1/L_{\text{FRX}}) + T_{\text{eRX}}$$

and:

$$\begin{aligned} G_{\text{Rmax}} &= \eta(\pi D/\lambda_{\text{D}})^2 = \eta(\pi D f_{\text{D}}/c)^2 = 0.6(\pi \times 4 \times 12 \times 10^9/3 \times 10^8)^2 \\ &= 151\,597 = 51.8\,\text{dBi} \\ L_{\text{R}}(\text{dB}) &= 12(\theta_{\text{R}}/\theta_{3\,\text{dB}})^2 = 12(\theta_{\text{R}} D f_{\text{D}}/70c)^2 = 0.6\,\text{dB} \\ L_{\text{FRX}} &= 0.5\,\text{dB} \\ L_{\text{POL}} &= 0\,\text{dB} \end{aligned}$$

$T_{\text{A}} = T_{\text{SKY}} + T_{\text{GROUND}}$ with $T_{\text{SKY}} = 20\,\text{K}$ (see Figure 5.20 for $f = 12\,\text{GHz}$ and $E = 10^\circ$) and $T_{\text{GROUND}} = 45\,\text{K}$, from which $T_{\text{A}} = 65\,\text{K}$

$$\begin{aligned} T_{\text{F}} &= 290\,\text{K} \\ T_{\text{eRX}} &= (F - 1)T_0 = (10^{0.1} - 1)290 = 75\,\text{K} \end{aligned}$$

Hence:

$$T_D = 65/10^{0.05} + 290(1-1/10^{0.05}) + 75 = 164.5\,\text{K}$$

then:

$$\begin{aligned}(G/T)_{ES} &= 51.8-0.6-0.5-10\log[65/10^{0.05}+290(1-1/10^{0.05})+75] \\ &= 28.5\,\text{dBK}^{-1}\end{aligned}$$

To calculate the ratio C/N_0 for the downlink:

$$(C/N_0)_D = (\text{EIRP})_{SL}(1/L_D)(G/T)_{ES}(1/k) \quad (\text{Hz}) \tag{5.44}$$

Hence:

$$\begin{aligned}(C/N_0)_D &= 44.2\,\text{dBW} - 206.4\,\text{dB} + 28.5\,\text{dBK}^{-1} + 228.6\,\text{dBW/HzK} \\ &= 94.9\,\text{dBHz}\end{aligned}$$

Figure 5.26 summarises the variations of power level throughout the path.

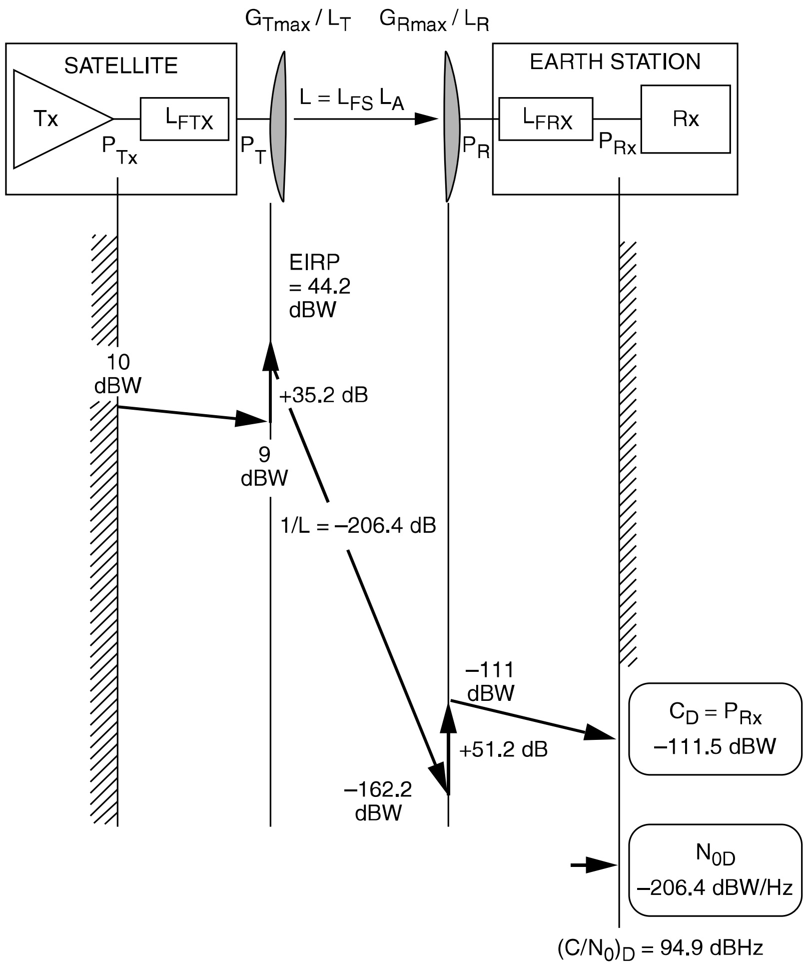

Figure 5.26 Variations in power for the clear sky downlink.

5.7 INFLUENCE OF THE ATMOSPHERE

On both the up- and downlinks, the carrier passes through the atmosphere. Recall that the range of frequencies concerned is from 1 to 30 GHz. From the point of view of wave propagation at these frequencies, only two regions of the atmosphere have an influence—the troposphere and the ionosphere. The troposphere extends practically from the ground to an altitude of 15 km. The ionosphere is situated between around 70 and 1000 km. The regions where their influence is maximum are in the vicinity of the ground for the troposphere and at an altitude of the order of 400 km for the ionosphere.

The influence of the atmosphere has been mentioned previously in order to introduce the losses L_A due to atmospheric attenuation into equation (5.14) and in connection with antenna noise temperature. However, other phenomena can occur. Their nature and significance is now explained.

The predominant effects are those caused by absorption and depolarisation due to tropospheric precipitation (rain and snow). Dry snow has little effect. Although wet snowfalls can cause greater attenuation than the equivalent rainfall rate, this situation is rare and has little effect on attenuation statistics. Effects are particularly significant for frequencies greater than 10 GHz. The occurrence of rain is defined by the percentage of time during which a given rainfall rate is exceeded. Low rainfall rates with negligible effects correspond to high percentages of time (typically 20%); these are described as *clear sky* conditions. High rainfall rate, with significant effects, correspond to small percentages of time (typically 0.01%); these are described as *rain* conditions. These effects can degrade the quality of the link below an acceptable threshold. The availability of a link is thus directly related to the rainfall rate time statistics. In view of their importance, the effects of precipitation are presented first. The effects of other phenomena are examined later.

5.7.1 Impairments caused by rain

The intensity of precipitation is measured by the rainfall rate R expressed in mm/h. The temporal precipitation statistic is given by the cumulative probability distribution which indicates the annual percentage p (%) during which a given value of rainfall rate R_p (mm/h) is exceeded. In the absence of precipitation data for the location of the earth station involved in the link, data from Figure 5.27 (ITU-R Rec. P.837) can be used. For instance, in Europe (Figure 5.27b) a rainfall rate of $R_{0.01}$ ($p = 0.01\%$ corresponds to 53 minutes per year) is around 30 mm/h with the exception of some Mediterranean regions where the occurrence of storms (heavy precipitation for a short time interval) leads to a value of $R_{0.01} = 50$ mm/h. In equatorial regions, $R_{0.01} = 120$ mm/h (Central America or South East Asia, for example). Rain causes attenuation and depolarisation.

5.7.1.1 Attenuation

The value of attenuation due to rain A_{RAIN} is given by the product of the specific attenuation γ_{R} (dB/km) and the effective path length of the wave in the rain L_{e} (km), that is:

$$A_{\mathrm{RAIN}} = \gamma_R L_{\mathrm{e}} \quad (\mathrm{dB}) \tag{5.45}$$

The value of γ_{R} depends on the frequency and intensity R_p (mm/h) of the rain. The result is a value of attenuation which is exceeded during the percentage of time p. Determination of A_{RAIN} proceeds in several steps:

(1) Determine the rainfall rate $\mathrm{R}_{0.01}$ exceeded for 0.01% of the time of an average year, where the earth station is located on the maps in Figure 5.27.

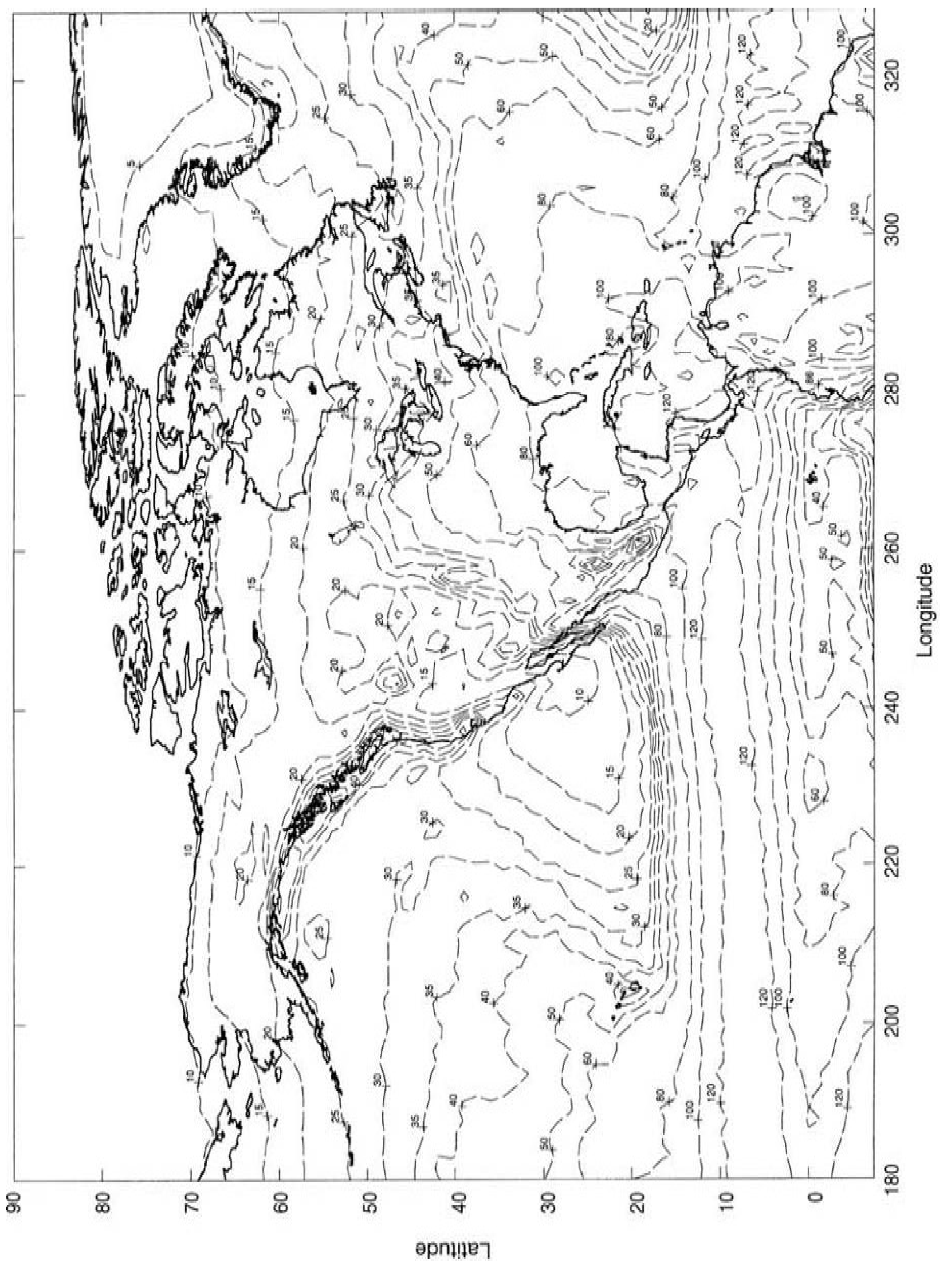

Figure 5.27 Rain intensity exceeded for more than 0.01% of the average year (From ITU-R Rec. P.618-7. Reproduced by permission of the ITU).

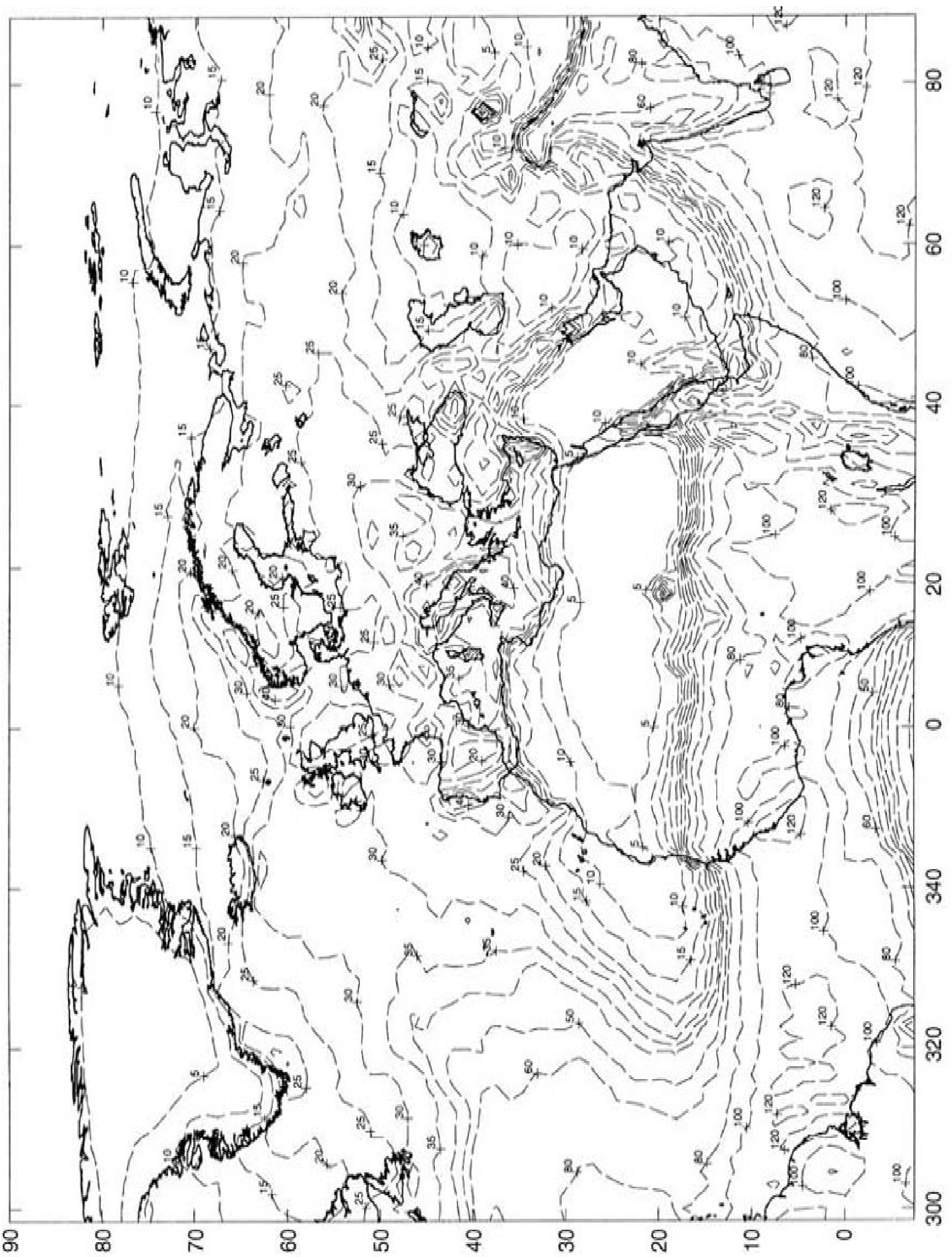

Figure 5.27 (*Continued.*)

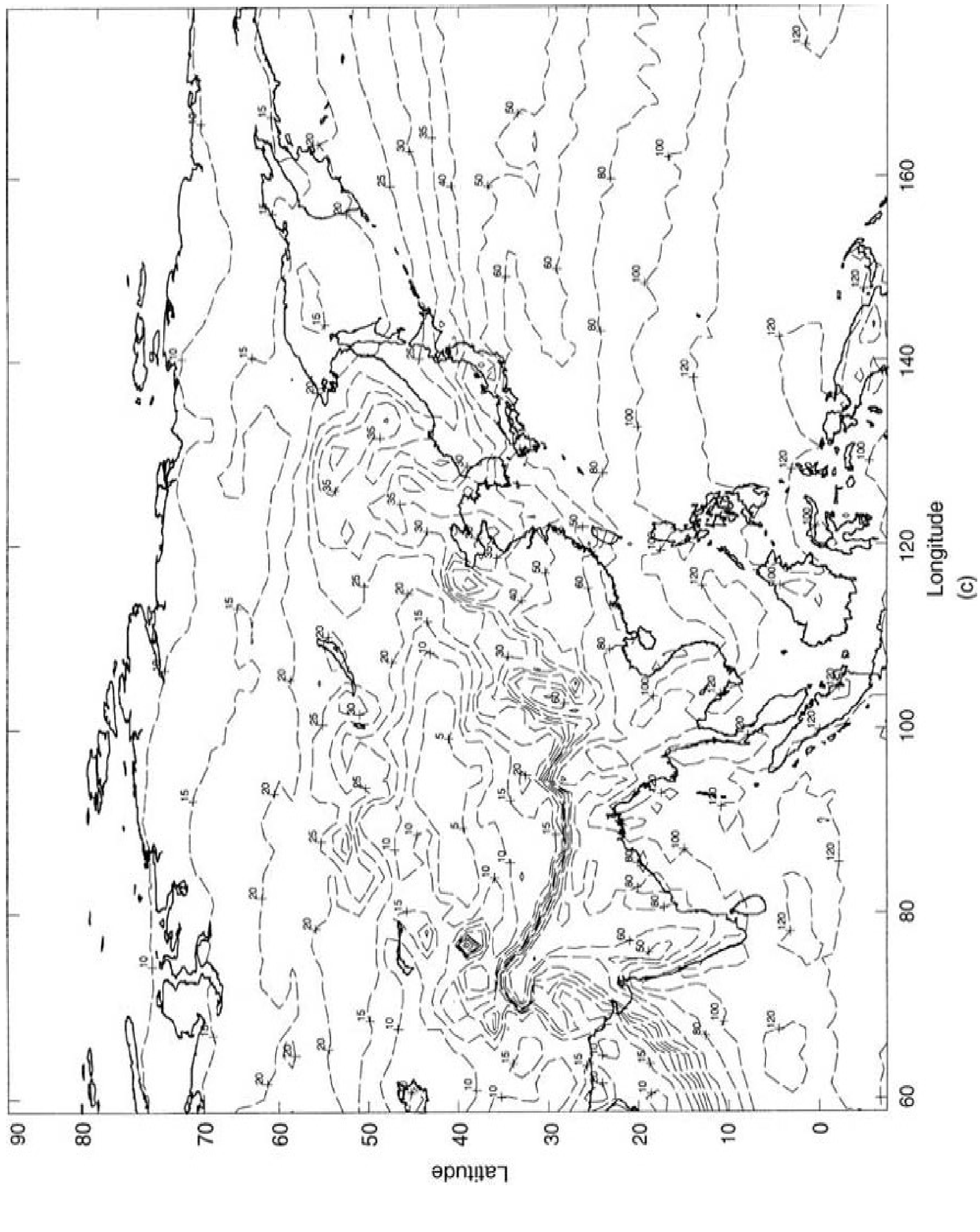

Figure 5.27 (*Continued.*)

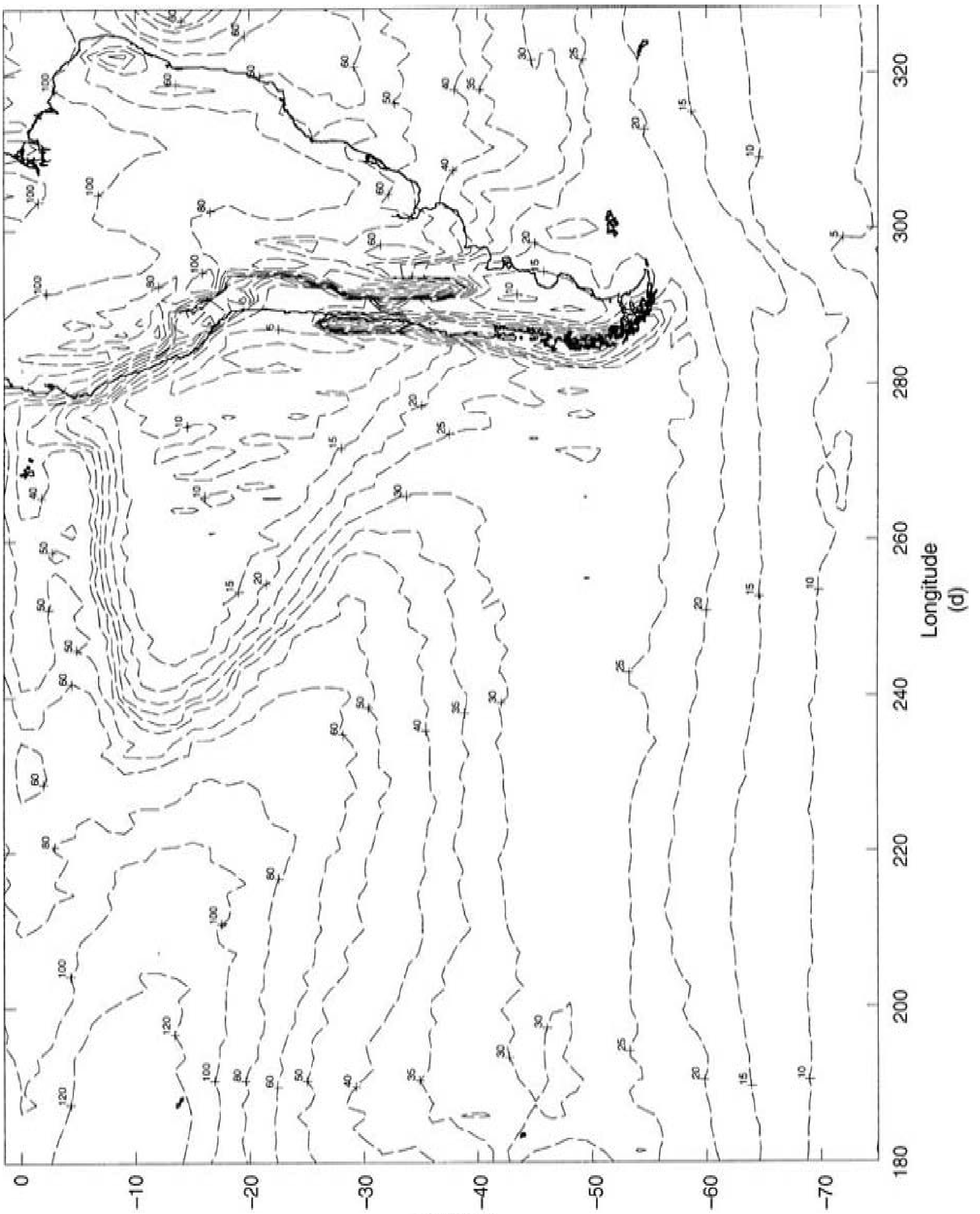

Figure 5.27 (*Continued.*)

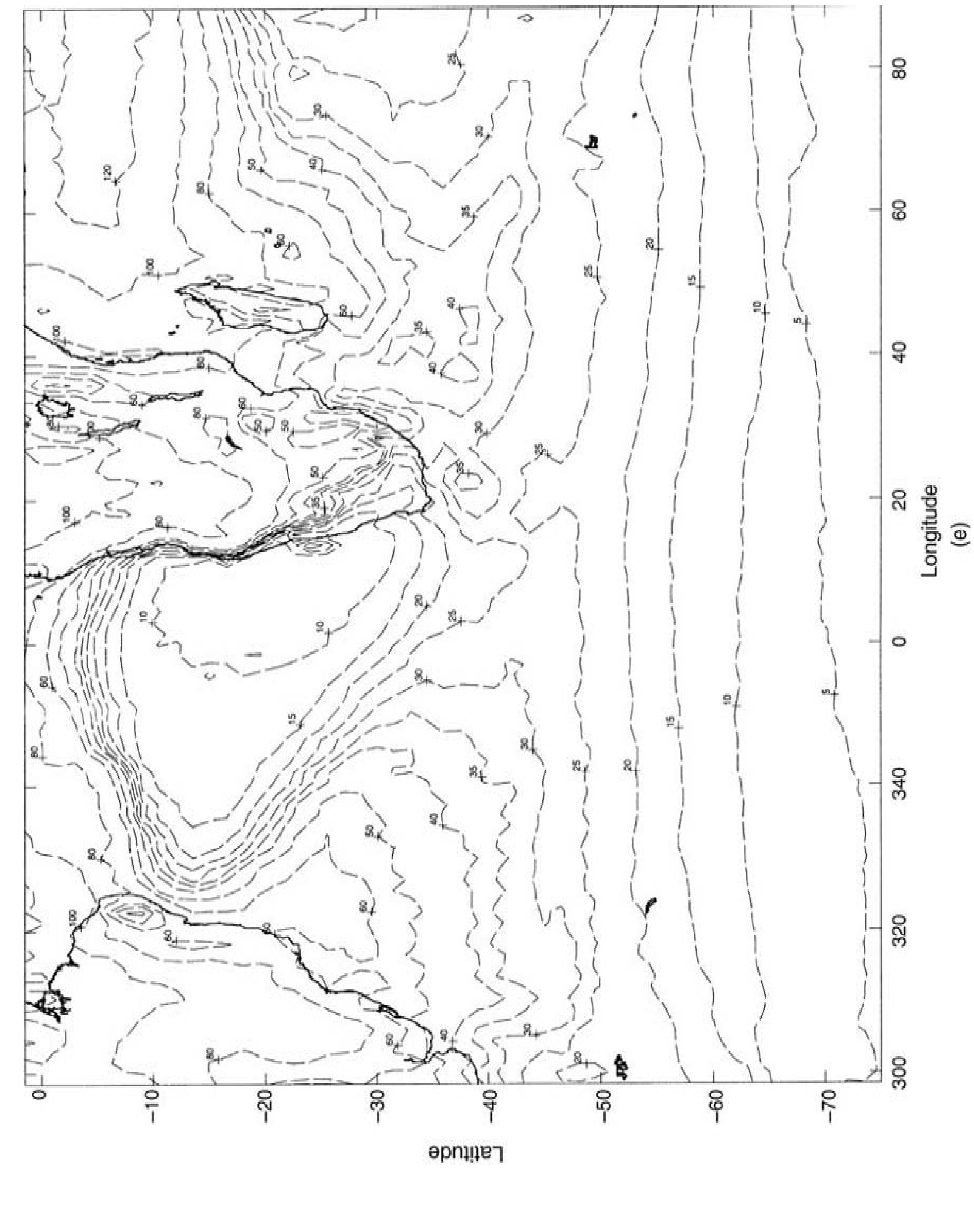

Figure 5.27 (*Continued.*)

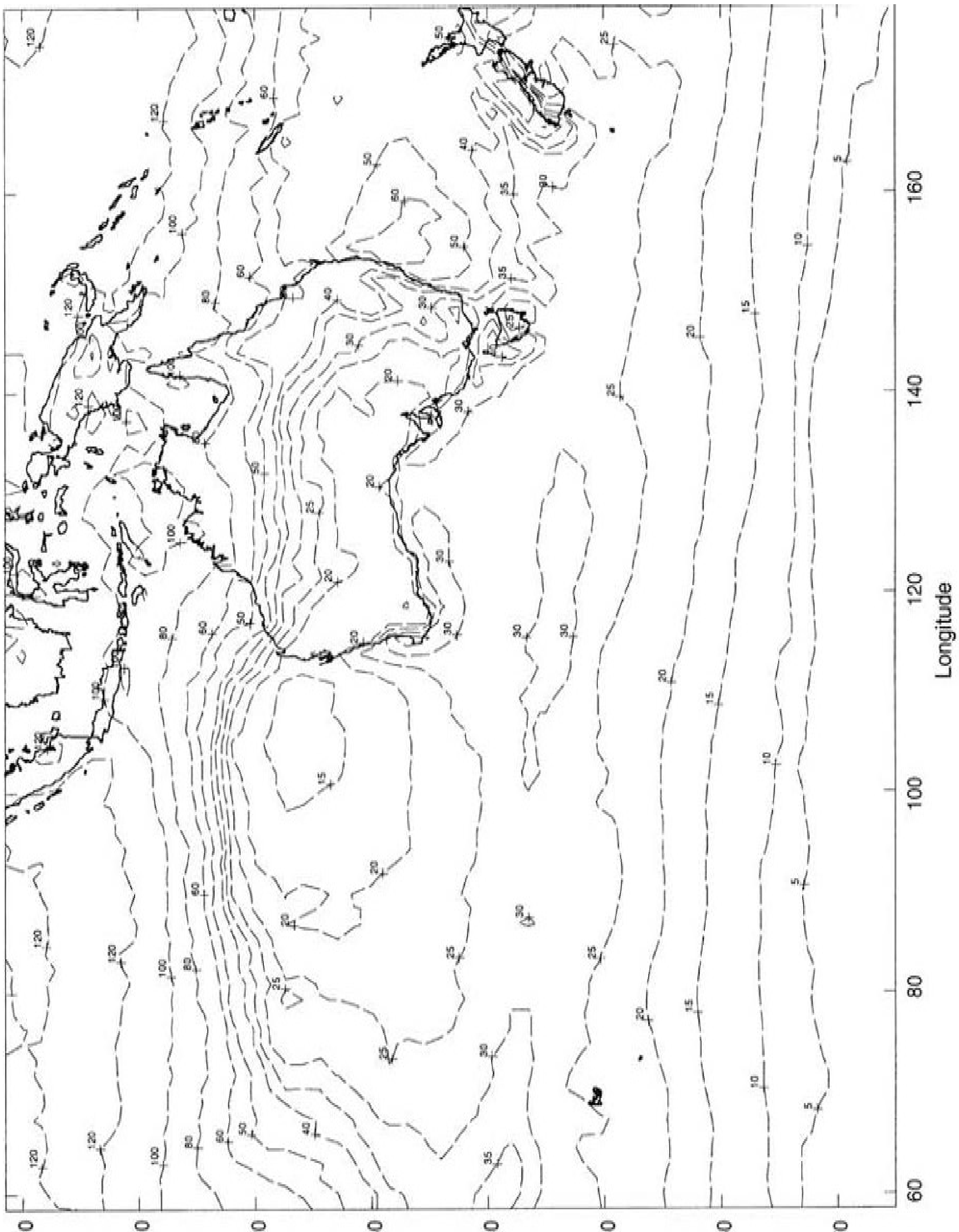

Figure 5.27 (*Continued.*)

(2) Calculate the effective height of the rain h_R as given in (ITU-R Rec. P.839-3)

$$h_R(\text{km}) = h_0 + 0.36\,\text{km}$$

where h_0 is the mean 0°C isotherm height above mean sea level, given in Figure 5.28.

(3) Compute the slant-path length, L_s, below the rain height:

$$L_s = \frac{h_R - h_s}{\sin E} \quad (\text{km})$$

where h_s (km) is the earth station height above mean sea level and E is the satellite elevation angle. This is valid for $E \geq 5°$.

(4) Calculate the horizontal projection, L_G, of the slant-path length:

$$L_G = L_s \cos E \quad (\text{km})$$

(5) Obtain the specific attenuation, γ_R, as a function of $R_{0.01}$ and frequency from Table 5.1. These values are derived from the frequency-dependent coefficients given in (ITU-R Rec. P.838):

$$\gamma_R = k(R_{0.01})^{\alpha} \quad (\text{dB/km})$$

where

$$\begin{aligned} k &= [k_H + k_V + (k_H - k_V)\cos^2 E \cos 2\tau]/2 \\ \alpha &= [k_H\alpha_H + k_V\alpha_V + (k_H\alpha_H - k_V\alpha_V)\cos^2 E \cos 2\tau]/2\text{k} \end{aligned}$$

and E is the elevation angle and τ is the polarisation tilt angle relative to the horizontal ($\tau = 45°$ for circular polarisation). For a rapid but approximate estimate of γ_R, one can use the nomogram of Figure 5.29; for circular polarisation, take the mean value of the attenuation obtained for each linear polarisation.

(6) Calculate the horizontal reduction factor, $r_{0.01}$, for 0.01% of the time:
(enter L_G in km, γ_R in dB/km, f in GHz)

$$r_{0.01} = \left[1 + 0.78\sqrt{L_G\gamma_R/f} - 0.38(1 - e^{-2L_G})\right]^{-1}$$

(7) Calculate the vertical adjustment factor, $v_{0.01}$ for 0.01% of the time:

$$\zeta = \tan^{-1}\left(\frac{h_R - h_s}{L_G r_{0.01}}\right) \quad (\text{degrees})$$

$$L_R(\text{km}) = \begin{cases} L_G r_{0.01}/\cos E & \text{for } \zeta > E \\ (h_R - h_s)/\sin E & \text{otherwise} \end{cases}$$

$$\chi = \begin{cases} 36 - |\text{latitude}| & \text{if} |\text{latitude}| < 36° \\ 0 & \text{otherwise} \end{cases}$$

$$v_{0.01} = \left[1 + \sqrt{\sin E}\left(31(1 - e^{-(E/(1+\chi))})\frac{\sqrt{L_R\gamma_R}}{f^2} - 0.45\right)\right]^{-1}$$

(8) The effective path length is:

$$L_E = L_R v_{0.01} \quad (\text{km})$$

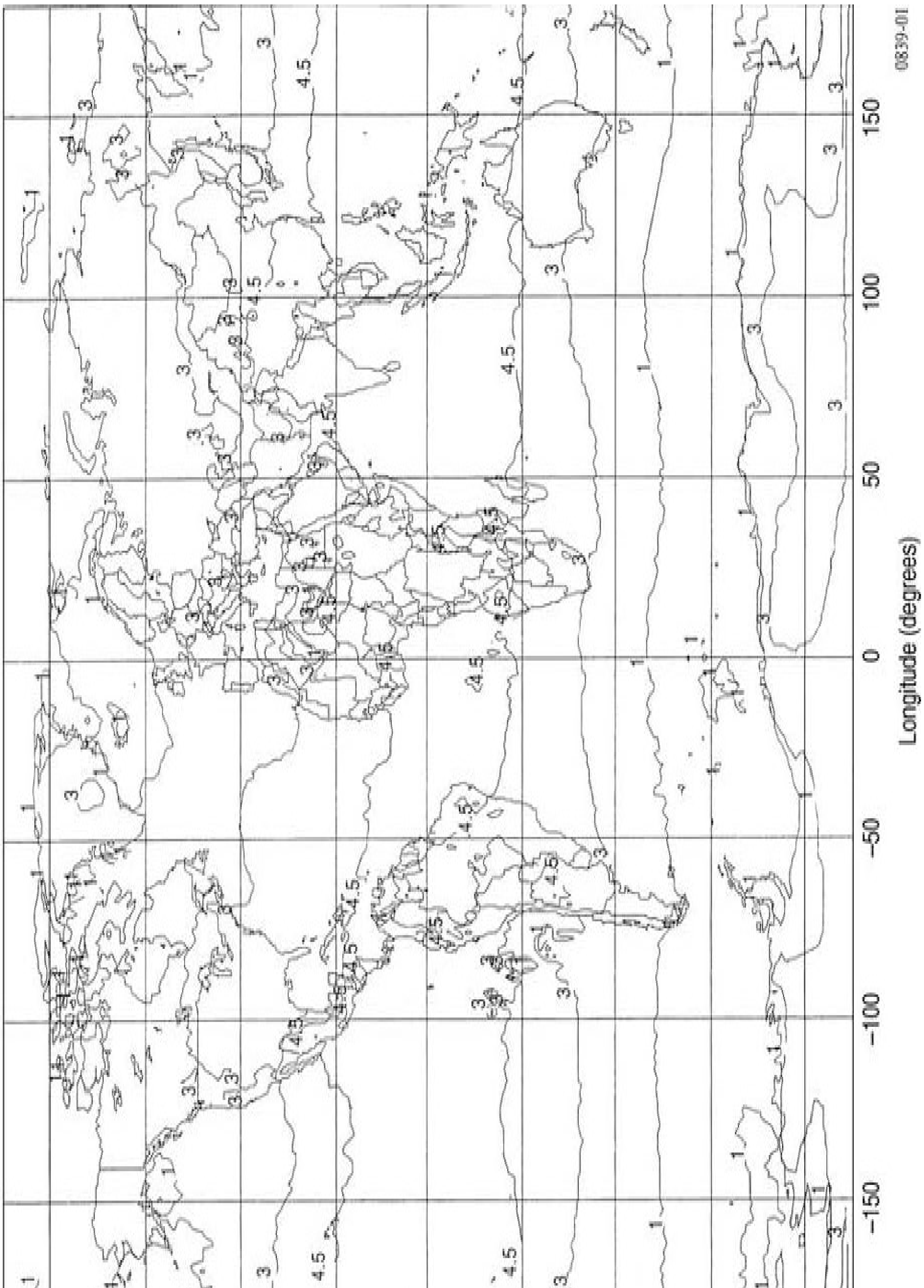

Figure 5.28 Yearly average 0°C isotherm height above mean sea level (km).

Table 5.1 Values and interpolation formulas for frequency-dependent coefficients k_H, k_V, α_H and α_V (log is base 10, i.e. $\log 10 = 1$)

Frequency	Coefficients
$f = 1\,\text{GHz}$	$k_H = 0.000\,038\,7$ $k_V = 0.000\,035\,2$ $\alpha_H = 0.912$ $\alpha_V = 0.880$
$1\,\text{GHz} \le f \le 2\,\text{GHz}$	$k_H = 3.870 \times 10^{-5} f_{\text{GHz}}^{1.9925}$ $k_V = 3.520 \times 10^{-5} f_{\text{GHz}}^{1.9710}$ $\alpha_H = 0.1694 \log f_{\text{GHz}} + 0.9120$ $\alpha_V = 0.1428 \log_{f\text{GHz}} + 0.8800$
$f = 2\,\text{GHz}$	$k_H = 0.000\,154$ $k_V = 0.000\,138$ $\alpha_H = 0.963$ $\alpha_V = 0.923$
$2\,\text{GHz} \le f \le 4\,\text{GHz}$	$k_H = 3.649 \times 10^{-5}\, f_{\text{GHz}}^{2.0775}$ $k_V = 3.222 \times 10^{-5}\, f_{\text{GHz}}^{2.0985}$ $\alpha_H = 0.5249 \log f_{\text{GHz}} + 0.8050$ $\alpha_V = 0.5049 \log f_{\text{GHz}} + 0.7710$
$f = 4\,\text{GHz}$	$k_H = 0.000\,650$ $k_V = 0.000\,591$ $\alpha_H = 1.121$ $\alpha_V = 1.075$
$4\,\text{GHz} \le f \le 6\,\text{GHz}$	$k_H = 2.199 \times 10^{-5} f_{\text{GHz}}^{2.4426}$ $k_V = 2.187 \times 10^{-5} f_{\text{GHz}}^{2.3780}$ $\alpha_H = 1.0619 \log_{f\text{GHz}} + 0.4816$ $\alpha_V = 1.0790 \log f_{\text{GHz}} + 0.4254$
$f = 6\,\text{GHz}$	$k_H = 0.001\ 75$ $k_V = 0.001\ 55$ $\alpha_H = 1.308$ $\alpha_V = 1.265$
$6\,\text{GHz} \le f \le 7\,\text{GHz}$	$k_H = 3.202 \times 10^{-6} f_{\text{GHz}}^{3.5181}$ $k_V = 3.041 \times 10^{-6} f_{\text{GHz}}^{3.4791}$ $\alpha_H = 0.3585 \log f_{\text{GHz}} + 1.0290$ $\alpha_V = 0.7021 \log f_{\text{GHz}} + 0.7187$
$f = 7\,\text{GHz}$	$k_H = 0.003\,01$ $k_V = 0.002\,65$ $\alpha_H = 1.332$ $\alpha_V = 1.312$
$7\,\text{GHz} \le f \le 8\,\text{GHz}$	$k_H = 7.542 \times 10^{-6} f_{\text{GHz}}^{3.0778}$ $k_V = 7.890 \times 10^{-6} f_{\text{GHz}}^{2.9892}$ $\alpha_H = -0.0862 \log f + 1.4049$ $\alpha_V = -0.0345 \log f + 1.3411$
$f = 8\,\text{GHz}$	$k_H = 0.004\,54$ $k_V = 0.003\,95$ $\alpha_H = 1.327$ $\alpha_V = 1.310$

(*continued*)

Table 5.1 (*Continued*)

Frequency	Coefficients
$8\,\text{GHz} \le f \le 10\,\text{GHz}$	$k_H = 2.636 \times 10^{-6} f_{\text{GHz}}^{3.5834}$ $k_V = 2.102 \times 10^{-6} f_{\text{GHz}}^{3.6253}$ $\alpha_H = -0.5263 \log f_{\text{GHz}} + 1.8023$ $\alpha_V = -0.4747 \log f_{\text{GHz}} + 1.7387$
$f = 10\,\text{GHz}$	$k_H = 0.0101$ $k_V = 0.008\,87$ $\alpha_H = 1.276$ $\alpha_V = 1.264$
$10\,\text{GHz} \le f \le 12\,\text{GHz}$	$k_H = 3.949 \times 10^{-6} f_{\text{GHz}}^{3.4078}$ $k_V = 2.785 \times 10^{-6} f_{\text{GHz}}^{3.5032}$ $\alpha_H = -0.7451 \log f_{\text{GHz}} + 2.0211$ $\alpha_V = -0.8083 \log f_{\text{GHz}} + 2.0723$
$f = 12\,\text{GHz}$	$k_H = 0.0188$ $k_V = 0.0168$ $\alpha_H = 1.217$ $\alpha_V = 1.200$
$12\,\text{GHz} \le f \le 15\,\text{GHz}$	$k_H = 1.094 \times 10^{-5} f_{\text{GHz}}^{2.9977}$ $k_V = 7.718 \times 10^{-6} f_{\text{GHz}}^{3.0929}$ $\alpha_H = -0.6501 \log f_{\text{GHz}} + 1.9186$ $\alpha_V = -0.7430 \log f_{\text{GHz}} + 2.0018$
$f = 15\,\text{GHz}$	$k_H = 0.0367$ $k_V = 0.0335$ $\alpha_H = 1.154$ $\alpha_V = 1.128$
$15\,\text{GHz} \le f \le 20\,\text{GHz}$	$k_H = 4.339 \times 10^{-5} f_{\text{GHz}}^{2.4890}$ $k_V = 3.674 \times 10^{-5} f_{\text{GHz}}^{2.5167}$ $\alpha_H = -0.4402 \log f_{\text{GHz}} + 1.6717$ $\alpha_V = -0.5042 \log f_{\text{GHz}} + 1.7210$
$f = 20\,\text{GHz}$	$k_H = 0.0751$ $k_V = 0.0691$ $\alpha_H = 1.099$ $\alpha_V = 1.065$
$20\,\text{GHz} \le f \le 25\,\text{GHz}$	$k_H = 8.951 \times 10^{-5} f_{\text{GHz}}^{2.2473}$ $k_V = 3.674 \times 10^{-5} f_{\text{GHz}}^{2.2041}$ $\alpha_H = -0.3921 \log f_{\text{GHz}} + 1.6092$ $\alpha_V = -0.3612 \log f_{\text{GHz}} + 1.5349$
$f = 25\,\text{GHz}$	$k_H = 0.124$ $k_V = 0.1113$ $\alpha_H = 1.061$ $\alpha_V = 1.030$
$25\,\text{GHz} \le f \le 30\,\text{GHz}$	$k_H = 8.779 \times 10^{-5} f_{\text{GHz}}^{2.2533}$ $k_V = 1.143 \times 10^{-4} f_{\text{GHz}}^{2.1424}$ $\alpha_H = -0.5052 \log f_{\text{GHz}} + 1.7672$ $\alpha_V = -0.3789 \log f_{\text{GHz}} + 1.5596$

(*continued*)

Table 5.1 (*Continued*)

Frequency	Coefficients
$f = 30\,\text{GHz}$	$k_H = 0.187$ $k_V = 0.167$ $\alpha_H = 1.021$ $\alpha_V = 1.000$
$30\,\text{GHz} \leq f \leq 35\,\text{GHz}$	$k_H = 1.009 \times 10^{-4} f_{\text{GHz}}^{2.2124}$ $k_V = 1.075 \times 10^{-4} f_{\text{GHz}}^{2.1605}$ $\alpha_H = -0.6274 \log f_{\text{GHz}} + 1.9477$ $\alpha_V = -0.5527 \log f_{\text{GHz}} + 1.8164$
$f = 35\,\text{GHz}$	$k_H = 0.263$ $k_V = 0.233$ $\alpha_H = 0.979$ $\alpha_V = 0.963$
$35\,\text{GHz} \leq f \leq 40\,\text{GHz}$	$k_H = 1.304 \times 10^{-4} f_{\text{GHz}}^{2.1402}$ $k_H = 1.163 \times 10^{-4} f_{\text{GHz}}^{2.1383}$ $\alpha_H = -0.6898 \log f_{\text{GHz}} + 2.0440$ $\alpha_H = -0.5863 \log f_{\text{GHz}} + 1.8683$
40 GHz	$k_H = 0.350$ $k_V = 0.310$ $\alpha_H = 0.939$ $\alpha_V = 0.929$

(9) The predicted attenuation exceeded for 0.01% of an average year is obtained from:

$$A_{0.01} = \gamma_R L_E \quad (\text{dB})$$

(10) The estimated attenuation to be exceeded for other percentages of an average year, in the range 0.001% to 5%, is determined from the attenuation to be exceeded for 0.01% for an average year:

$$\beta = \begin{cases} 0 & \text{if } p \geq 1\% \text{ or } |\text{latitude}| \geq 36^\circ \\ -0.005(|\text{latitude}| - 36) & \text{if } p < 1\%, |\text{latitude}| < 36^\circ,\ E \geq 25^\circ \\ -0.005(|\text{latitude}| - 36) + 1.8 - 4.25 \sin E & \text{otherwise} \end{cases}$$

$$A_p = A_{0.01} \left(\frac{p}{0.01}\right)^{-(0.655 + 0.033 \ln p - 0.045 \ln A_{0.01} - \beta(1-p)\sin E)} \quad (\text{dB})$$

It is sometimes required to estimate the attenuation exceeded during a percentage p_w of any month (that is the worst month). The corresponding annual percentage is given (for $1.9 \times 10^{-4} < p_w < 7.8$) by

$$p = 0.3(p_w)^{1.15} \quad (\%) \tag{5.46}$$

Performance objectives are often stipulated for $p_w = 0.3\%$ (Section 3.2). This corresponds to an annual percentage $p = 0.075\%$.

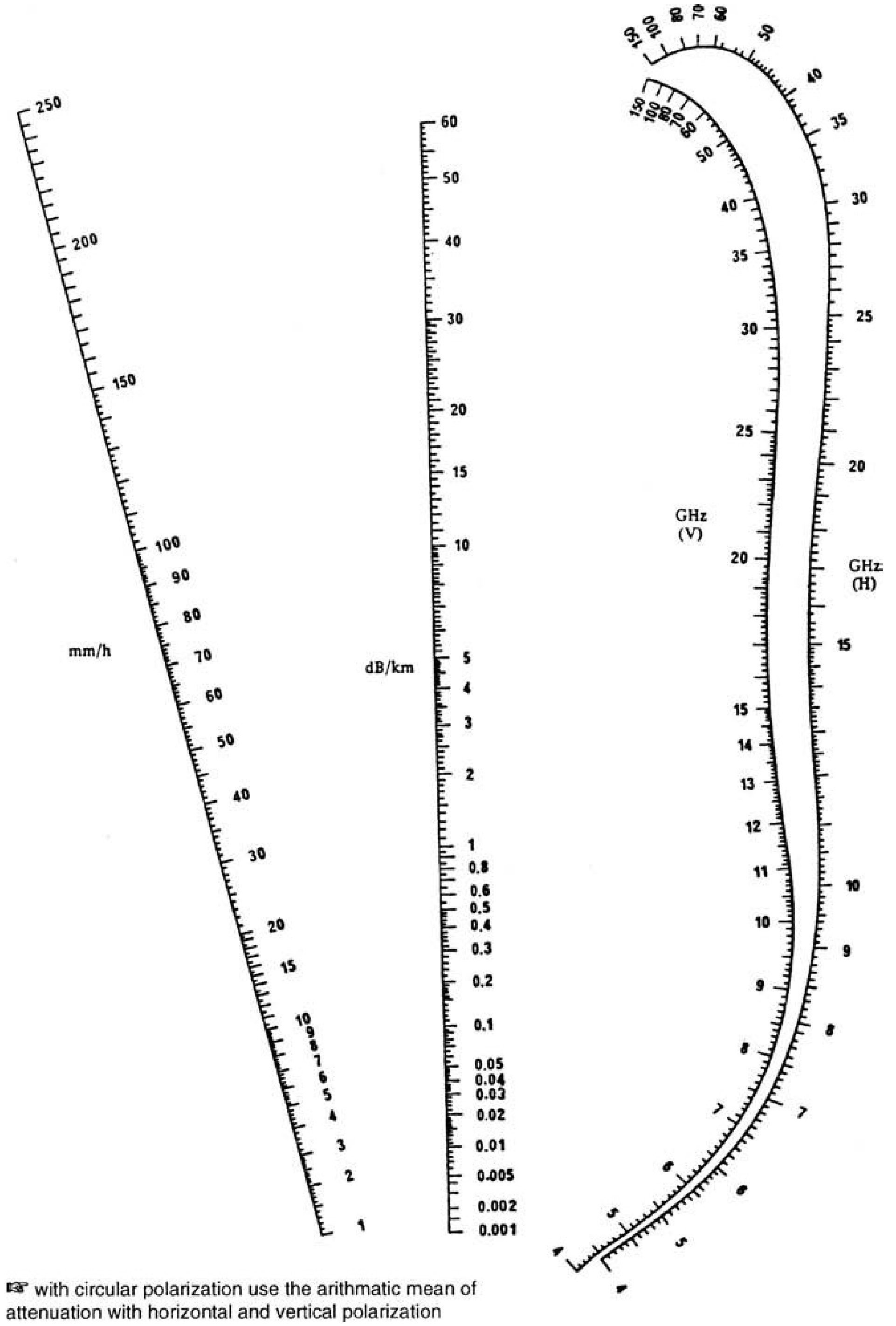

Figure 5.29 Nomogram for determination of the specific attenuation γ_R as a function of the frequency (GHz) and rain density R (mm/h) (CCIR Rep. 721. Reproduced by permission of the ITU.)

Typical values of attenuation due to rain exceeded for 0.01% of an average year can be deduced from the previous procedure for regions where the rainfall rate $R_{0.01}$ exceeded for 0.01% of an average year is in the range from 30 to 50 mm/h. This gives typically around 0.1 dB at 4 GHz, 5–10 dB at 12 GHz, 10–20 dB at 20 GHz and 25–40 dB at 30 GHz.

5.7.1.2 Depolarisation

Cross-polarisation was identified in Section 5.2.3 as energy transfer from one polarisation to an orthogonal polarisation. Section 5.2.3 considered imperfect isolation between waves in two orthogonal polarisation states transmitted or received by a given antenna. Now consider cross-polarisation caused by wave depolarisation due to rain and ice clouds.

Rain introduces depolarisation as a result of differential attenuation and differential phase shift between two orthogonal characteristic polarisations. These effects originate in the non-spherical shape of raindrops. A commonly adopted model for a falling raindrop is an oblate spheroid with its major axis inclined to the horizontal and where deformation depends upon the radius of a sphere of equal volume. Assume the angles of inclination vary randomly in space and time.

Statistics for the cross-polarisation discrimination $\mathrm{XPD_{rain}}$ due to rain can be derived from rain attenuation statistics, i.e. the value of attenuation (termed 'copolar attenuation'), $A_{\mathrm{RAIN}}(p)$, which is exceeded during an annual percentage p for the considered polarisation.

Ice clouds, where high-altitude ice crystals are in a region close to the 0°C isotherm, are also a cause of cross-polarisation. However, in contrast to rain, this effect is not accompanied by attenuation.

The cross-polarisation discrimination XPD (p) not exceeded for p% of the time is given by:

$$\mathrm{XPD}(p) = \mathrm{XPD_{rain}} - C_{\mathrm{ice}} \quad \mathrm{(dB)} \tag{5.47}$$

where $\mathrm{XPD_{rain}}$ is the cross-polarisation discrimination due to rain and C_{ice} is the contribution of ice clouds, respectively given by:

$$\begin{aligned} \mathrm{XPD_{rain}} &= C_f - C_A + C_\tau + C_\theta + C_\sigma \quad \mathrm{(dB)} \\ C_{\mathrm{ice}} &= \mathrm{XPD_{rain}}(0.3 + 0.1 \log p)/2 \quad \mathrm{(dB)} \end{aligned}$$

where:

$$\begin{aligned} C_f &= 30 \log f \\ C_A &= V(f) \log A_{\mathrm{RAIN}}(p) \\ V(f) &= 12.8 f^{0.19} \quad \text{for} \quad 8 \le f \le 20 \quad \mathrm{(GHz)} \\ V(f) &= 22.6 \quad \text{for} \quad 20 \le f \le 35 \quad \mathrm{(GHz)} \\ C_\tau &= -10 \log[1 - 0.484(1 + \cos 4\tau)] \end{aligned}$$

where f is the frequency (GHz) and τ is the tilt angle of the linearly polarised electric field vector with respect to the horizontal (for circular polarisation, use $\tau = 45°$).

$$C_\theta = -40 \log(\cos E) \quad \text{for} \quad E \le 60°$$

where E is the elevation angle.

$$C_\sigma = 0.0052\sigma^2$$

where σ is the standard deviation of the raindrop inclination angle distribution, expressed in degrees; σ takes the values 0°, 5°, 10° and 15° for $p = 1\%$, 0.1%, 0.01% and 0.001% of the time, respectively.

Equation (5.47) is in agreement with long-term measurements for $8\,\mathrm{GHz} \le f \le 35\,\mathrm{GHz}$ and elevation angle $E \le 60°$. For lower frequencies, down to 4 GHz, one can calculate $\mathrm{XPD_1}(p)$ at frequency $f_1 (8\,\mathrm{GHz} \le f_1 \le 30\,\mathrm{GHz})$ according to equation (5.47) and derive $\mathrm{XPD_2}(p)$ at frequency

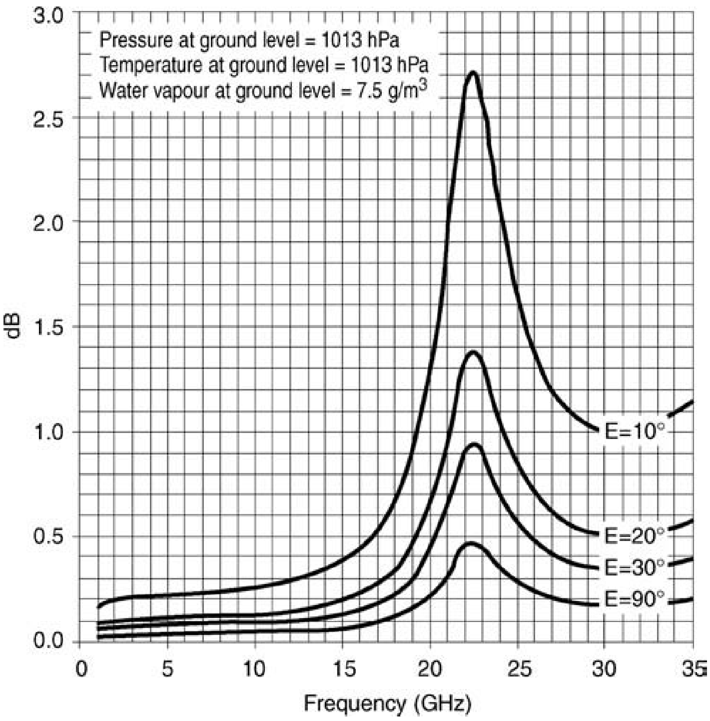

Figure 5.30 Attenuation due to atmospheric gases as a function of frequency and elevation angle E for a standard atmosphere with water vapour content at ground level of 7.5 g/m^3.

$f_2(4\,\text{GHz} \le f_2 \le 8\,\text{GHz})$ from the following semi-empirical formula:

$$\text{XPD}_2(p) = \frac{\text{XPD}_1(p) - 20\log[f_2[1-0.484(1+\cos 4\tau_2)]^{0.5}}{f_1[1-0.484(1+\cos 4\tau_1)]^{0.5}]}$$

where τ_1 and τ_2 are the respective polarisation tilt angles at frequencies f_1 and f_2.

5.7.2 Other impairments

5.7.2.1 Attenuation by atmospheric gases

Attenuation due to gas in the atmosphere depends on the frequency, the elevation angle, the altitude of the station and the water vapour concentration. Figure 5.30 displays the attenuation for a standard atmosphere. The attenuation is negligible at frequencies less than 10 GHz and does not exceed 3 dB at 22.24 GHz (the frequency corresponding to a water vapour absorption band) for mean atmospheric humidity and elevation angles greater than 10°.

5.7.2.2 Attenuation due to rain, fog or ice clouds

Attenuation due to rain clouds or fog can be calculated according to (ITU-R Rec. P.840). The specific attenuation γ_C is calculated as:

$$\gamma_C = KM \quad (\text{dB/km}) \tag{5.48}$$

where an approximate value for K is $K = 1.2 \times 10^{-3} f^{1.9}$ (dB/km)/(g/m^3), f is expressed in GHz from 1 GHz to 30 GHz, K in (dB/km)/(g/m^3) and M = water concentration, of the cloud or fog (g/m^3).

Attenuation due to rain clouds and fog is usually small compared with that due to rain. This attenuation, however, is observed for a greater percentage of the time. For an elevation angle $E = 20°$, the attenuation exceeded for 1% of the year due to rain clouds is of the order of 0.2 dB at 12 GHz, 0.5 dB at 20 GHz and 1.1 dB at 30 GHz in North America and Europe, and 0.8 dB at 12 GHz, 2.1 dB at 20 GHz and 4.5 dB at 30 GHz in South East Asia. For thick fog ($M = 0.5$ g/m^3) the attenuation is of the order of 0.4 dB/km at 30 GHz. Attenuation due to ice clouds is smaller still.

5.7.2.3 Attenuation by sandstorms

The specific attenuation (dB/km) is inversely proportional to the visibility and depends strongly on the humidity of the particles. At 14 GHz it is of the order of 0.03 dB/km for dry particles and 0.65 dB/km for particles of 20% humidity. If the path length is 3 km, the attenuation can reach 1 to 2 dB.

5.7.2.4 Scintillation

Scintillation is a variation of the amplitude of received carrier caused by variations of the refractive index of the troposphere and the ionosphere. The peak-to-peak amplitude of these variations, at Ku-band and medium latitudes, can exceed 1 dB for 0.01% of the time. The troposphere and the ionosphere have different refractive indices. The refractive index of the troposphere decreases with altitude, is a function of meteorological conditions and is independent of frequency. That of the ionosphere depends on frequency and the electronic content of the ionosphere. Both are subject to rapid local fluctuations. The effect of refraction is to cause curvature of the trajectory of the wave, variation of wave velocity and hence propagation time. The most troublesome scintillation is ionospheric scintillation; it is greater when the frequency is low and the earth station is close to the equator.

5.7.2.5 The Faraday rotation

The ionosphere introduces a rotation of the plane of polarisation of a linearly polarised wave. The angle of rotation is inversely proportional to the square of the frequency. It is a function of the electronic content of the ionosphere and consequently varies with time, the season and the solar cycle. The order of magnitude is several degrees at 4 GHz. The result, for a small percentage of the time, is an attenuation $L_{POL}(\text{dB}) = -20 \log(\cos \Delta\psi)$ of the copolar carrier (see Section 5.4.4.4) where $\Delta\psi$ is the polarisation mismatch angle due to Faraday rotation, and the appearance of a cross-polarised component which reduces the value of cross-polarisation discrimination XPD. The value of XPD is given by $\text{XPD (dB)} = -20 \log(\tan \Delta\psi)$. For the case of a rotation $\Delta\psi = 9°$ at a frequency of 4 GHz, this gives $L_{POL} = 0.1$ dB and XPD = 16 dB. As seen from the earth station, the planes of polarisation rotate in the same direction on the uplink and the downlink. It is therefore not possible to compensate for Faraday rotation by rotating the feed system of the antenna, if the antenna is used for both transmission and reception.

5.7.2.6 Multipath contributions

When the earth station antenna is small, and hence has a beam with a large angular width, the received carrier can be the result of a direct path and contributions of significant amplitude can be received after reflection on the ground or surrounding obstacles (multipath). In the case of destructive combination (phase opposition), a large attenuation is observed. This effect is weak when the earth station is equipped with an antenna which is sufficiently directional to eliminate multipath contributions. Multipath effects become predominant for mobile communications with non-directional terminal antennas.

5.7.3 Link impairments—relative importance

At low frequencies (less than 10 GHz), the attenuation L_A is generally small and the principal cause of degradation of the link is cross-polarisation. This is caused by the ionosphere and by high-altitude ice crystals in the troposphere. At higher frequencies, the phenomena of attenuation and cross-polarisation are both observed. These are caused essentially by atmospheric gases, rainfall and other hydrometeors.

Statistically, these phenomena become greater when a short percentage of time is considered. The availability of the link increases when these effects can be compensated. Compensation techniques are available and are discussed in Section 5.8.

5.7.4 Link performance under rain conditions

5.7.4.1 Uplink performance

In the presence of rain, propagation attenuation is greater due to the attenuation A_{RAIN} caused by rain in the atmosphere. This is in addition to the attenuation due to gases in the atmosphere (0.3 dB). A typical value of attenuation due to rain for an earth station situated in a temperate climate (for example, in Europe) can be considered to be $A_{RAIN} = 10$ dB. Such an attenuation would not be exceeded, at a frequency of 14 GHz, for more than 0.01% of an average year. This gives $L_A = 0.3\,\text{dB} + 10\,\text{dB} = 10.3\,\text{dB}$.

Hence:

$$L_U = 207.4\,\text{dB} + 10.3\,\text{dB} = 217.7\,\text{dB}$$

Referring to the example of Section 5.6.2, the uplink performance under rain conditions becomes:

$$(C/N_0)_U = 71.7\,\text{dBW} - 217.7\,\text{dB} + 6.6\,\text{dBK}^{-1} + 228.6\,\text{dBW/Hz K} = 89.2\,\text{dBHz}$$

The ratio $(C/N_0)_U$ for the uplink would be greater than the value calculated in this way for 99.99% of an average year.

5.7.4.2 Downlink performance

Referring now to the example of Section 5.6.3, $A_{RAIN} = 7$ dB is taken as the typical value of attenuation due to rain for an earth station situated in a temperate climate (for example, in Europe) which will not be exceeded, at a frequency of 12 GHz, for more than 0.01% of an average year; this gives $L_A = 0.3\,\text{dB} + 7\,\text{dB} = 7.3\,\text{dB}$. Hence, $L_D = 206.1 + 7.3\,\text{dB} = 213.4\,\text{dB}$. The antenna noise temperature is given by:

$$T_A = T_{SKY}/A_{RAIN} + T_m(1 - 1/A_{RAIN}) + T_{GROUND} \quad \text{(K)} \tag{5.49}$$

Taking

$$\begin{aligned} T_{\mathrm{m}} &= 275\,\mathrm{K} \\ T_A &= 20/10^{0.7} + 275(1-1/10^{0.7}) + 45 = 269\,\mathrm{K} \\ T_D &= 269/10^{0.05} + 290(1-1/10^{0.05}) + 75 = 346\,\mathrm{K} \end{aligned}$$

Hence

$$\begin{aligned} (G/T)_{\mathrm{ES}} &= 51.8-0.6-0.5-10\log[269/10^{0.05} + 290(1-1/10^{0.05}) + 75] \\ &= 25.3\,\mathrm{dBK}^{-1} \end{aligned}$$

To calculate the ratio C/N_0 for the downlink:

$$(C/N_0)_{\mathrm{D}} = (\mathrm{EIRP})_{\mathrm{SL}}(1/L_{\mathrm{D}})(G/T)_{\mathrm{ES}}(1/k) \quad (\mathrm{Hz})$$

Hence:

$$(C/N_0)_{\mathrm{D}} = 44.2\,\mathrm{dBW} - 213.4\,\mathrm{dB} + 25.3\,\mathrm{dBK}^{-1} + 228.6\,\mathrm{dBW/HzK} = 84.7\,\mathrm{dBHz}$$

The ratio $(C/N_0)_{\mathrm{D}}$ for the downlink would be greater than the value calculated in this way for 99.99% of an average year.

5.7.5 Conclusion

The quality of the link between a transmitter and a receiver can be characterised by the ratio of the carrier power to the noise power spectral density C/N_0. This is a function of the transmitter EIRP, the receiver figure of merit *G/T* and the properties of the transmission medium. In a satellite link between two earth stations, two links must be considered—the uplink, characterised by the ratio $(C/N_0)_{\mathrm{U}}$, and the downlink, characterised by the ratio $(C/N_0)_{\mathrm{D}}$. The propagation conditions in the atmosphere affect the uplink and downlink differently; rain reduces the value of the ratio $(C/N_0)_{\mathrm{U}}$ by decreasing the value of received power C_{U} while it reduces the value of $(C/N_0)_{\mathrm{D}}$ by reducing the value of received power C_{D} and increasing the downlink system noise temperature. Denoting the resulting degradation by $\Delta(C/N_0)$ gives:

$$\Delta(C/N_0)_{\mathrm{U}} = \Delta C_{\mathrm{U}} = (A_{\mathrm{RAIN}})_{\mathrm{U}} \qquad (\mathrm{dB}) \tag{5.50}$$

$$\Delta(C/N_0)_{\mathrm{D}} = \Delta C_{\mathrm{D}} - \Delta(G/T) = (A_{\mathrm{RAIN}})_{\mathrm{D}} + \Delta T \quad (\mathrm{dB}) \tag{5.51}$$

5.8 MITIGATION OF ATMOSPHERIC IMPAIRMENTS

5.8.1 Depolarisation mitigation

The method of compensation relies on modification of the polarisation characteristics of the earth station (see Chapter 8). Compensation is achieved as follows:

— for the uplink, by correcting the polarisation of the transmitting antenna by anticipation so that the wave arrives matched to the satellite antenna;
— for the downlink, by matching the antenna polarisation to that of the received wave.

Compensation can be automatic; the signals transmitted by the satellite must be made available (as beacons) so that the effects of the propagating medium can be detected and the required control signal deduced.

5.8.2 Attenuation mitigation

The mission specifies a value of the ratio C/N_0 greater than or equal to $(C/N_0)_{\text{required}}$ during a given percentage of the time, equal to $(100-p)\%$. For example, 99.99% of the time implies $p = 0.01\%$. As seen in Section 5.7.5, the attenuation A_{RAIN} due to rain causes a reduction of the ratio C/N_0 given by:

$$(C/N_0)_{\text{rain}} = (C/N_0)_{\text{clear sky}} - A_{\text{RAIN}}(\text{dB}) \quad (\text{dBHz}) \tag{5.52}$$

for an uplink and:

$$(C/N_0)_{\text{rain}} = (C/N_0)_{\text{clear sky}} - A_{\text{RAIN}}(\text{dB}) - \Delta(G/T) \quad (\text{dBHz}) \tag{5.53}$$

for a downlink.

$\Delta(G/T) = (G/T)_{\text{clear sky}} - (G/T)_{\text{rain}}$ represents the reduction (in dB) of the figure of merit of the earth station due to the increase of noise temperature.

For a successful mission, one must have $(C/N_0)_{\text{rain}} = (C/N_0)_{\text{required}}$; this can be achieved by including a margin M(p) in the clear sky link budget with M(p) defined by:

$$\begin{aligned} M(p) &= (C/N_0)_{\text{clear sky}} - (C/N_0)_{\text{required}} \\ &= (C/N_0)_{\text{clear sky}} - (C/N_0)_{\text{rain}} \quad (\text{dB}) \end{aligned} \tag{5.54}$$

The value of A_{RAIN} to be used is a function of the time percentage p. It increases as p decreases.

Making provision for a margin M(p) in the clear sky link requirement implies an increase of the EIRP which requires a higher transmitting power. For high attenuations, which are encountered for a small percentage of the time and at the highest frequencies (see Section 5.7.1.1), the extra power necessary can exceed the capabilities of the transmitting equipment and other solutions must then be considered: site diversity and adaptivity.

5.8.3 Site diversity

High attenuations are due to regions of rain of small geographical extent. Two earth stations at two distinct locations can establish links with the satellite which, at a given time t, suffer attenuations of $A_1(t)$ and $A_2(t)$ respectively; $A_1(t)$ is different from $A_2(t)$ as long as the geographical separation is sufficient. The signals are thus routed to the link less affected by attenuation. On this link, the attenuation is $A_{\text{D}}(t) = \min[A_1(t), A_2(t)]$. The mean attenuation for a single location is defined as $A_{\text{M}}(t) = [A_1(t) + A_2(t)]/2$; all values are in dB.

Two concepts are useful to quantify the improvement provided by location diversity (ITU-R Rec. PN.618):

— the diversity gain;
— the diversity improvement factor.

5.8.3.1 *Diversity gain $G_D(p)$*

This is the difference (in dB) between the mean attenuation at a single location $A_{\text{M}}(p)$, exceeded for a time percentage p, and the attenuation with diversity $A_{\text{D}}(p)$ exceeded for the same time percentage p. Hence, for a downlink, for example, the required margin $M(p)$ at a given location is obtained from:

$$M(p) = A_{\mathrm{RAIN}} + \Delta(G/T) \quad \text{(dB)} \tag{5.55}$$

With site diversity, the required margin becomes:

$$M(p) = A_{\mathrm{RAIN}} + \Delta(G/T) - G_{\mathrm{D}}(p) \quad \text{(dB)} \tag{5.56}$$

5.8.3.2 Diversity improvement factor F_{D}

This is the ratio between the percentage of time p_1 during which the mean attenuation at a single site exceeds the value A dB and the percentage of time p_2 during which the attenuation with diversity exceeds the same value A dB.

Figure 5.31 shows the relationship between p_2 and p_1 as a function of the distance between the two locations. These curves can be modelled by the following relations:

$$p_2 = (p_1)^2(1+\beta^2)/(p_1 + 100\beta^2) \tag{5.57}$$

with $\beta^2 = 10^{-4}d^{1.33}$, when the distance $d > 5$ km.

Site diversity also provides protection against scintillation and cross-polarisation.

5.8.4 Adaptivity

Adaptivity involves the variation of certain parameters of the link for the duration of the attenuation in such a way as to maintain the required value for the ratio C/N_0. Several approaches can be envisaged as follows [CAS-98]:

— Assignment of an additional resource, which is normally kept in reserve, to the link affected by attenuation. This additional resource can be:
  - an increase in transmission time (such as an unoccupied frame time slot in the case of TDMA multiple access, see Chapter 6) with or without the use of error correcting codes;
  - use of a frequency band at a lower frequency which is less affected by the attenuation;
  - use of higher EIRP on the uplink.

— Reduction of capacity. In the case of digital transmission, using forward error correction coding within the imposed bandwidth reduces the required value of C/N_0, at the cost of a reduced information bit rate R_b (see Section 4.2.7.2). The reduction in the required value of C/N_0 provides a margin equal to the C/N_0 reduction. This can be used for an overall link via a transparent satellite (see Section 5.9) or for the uplink or downlink of a regenerative satellite (see Section 5.10).

5.8.5 Cost–availability trade-off

A low unavailability (0.01% of the time, for example) corresponds to a high availability (99.99%, for the example considered). If only the effects of the propagating medium are considered as the cause of unavailability, the accepted value of unavailability represents the percentage of time p during which a given attenuation can be exceeded. When p is small (that is, the required availability

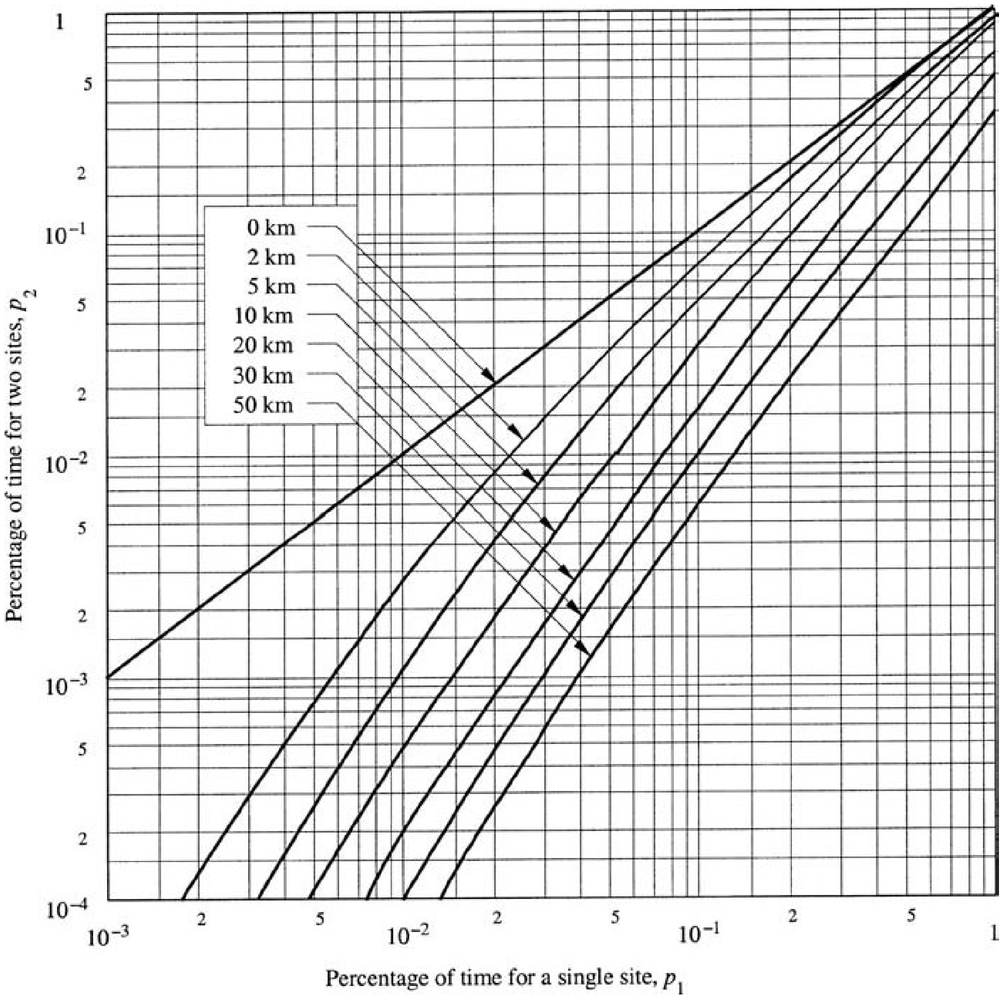

Figure 5.31 Relationship between percentage of time with and without diversity for the same attenuation (earth satellite paths).

is high), the value of this attenuation is high. Since the methods used to compensate for attenuation become more costly as the attenuation increases, the specified availability has a marked effect on system cost. A typical trend is shown in Figure 5.32.

5.9 OVERALL LINK PERFORMANCE WITH TRANSPARENT SATELLITE

Section 5.6 presents the individual link performance in terms of C/N_0. This section discusses the expression for the overall station-to-station link performance; that is, the link involving one uplink and one downlink via a transparent satellite (no on-board demodulation and remodulation). Up to now, noise on the uplink and on the downlink has been considered to be thermal noise only. In practice, one has to account for interference noise originating from other carriers in the considered frequency bands and intermodulation noise resulting from multicarrier operation of non-linear amplifiers. The overall link performance is discussed (Section 5.9.2) without intermodulation or interference, then expressions are introduced considering interference and finally intermodulation.

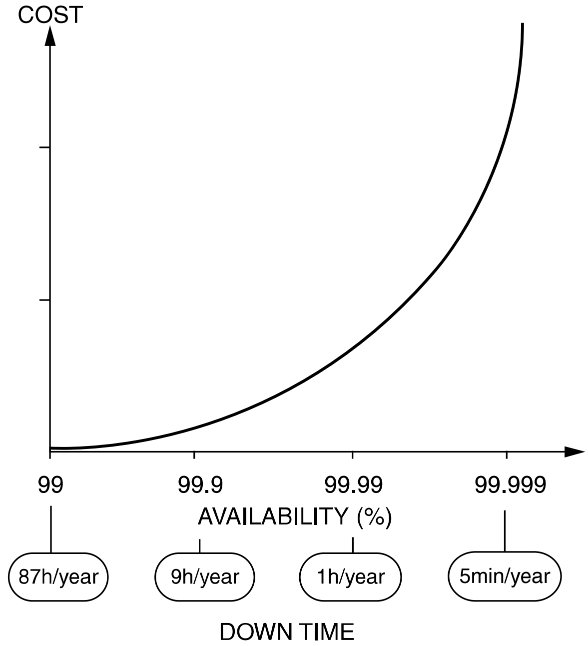

Figure 5.32 Link cost as a function of availability.

The following notation is used:

— $(C/N_0)_U$ is the uplink carrier power-to-noise power spectral density ratio (Hz) at the satellite receiver input, considering no other noise contribution than the uplink system thermal noise temperature T_U.
— $(C/N_0)_D$ is the downlink carrier power-to-noise power spectral density ratio (Hz) at the earth station receiver input, considering no other noise contribution than the downlink system thermal noise temperature T_D.
— $(C/N_0)_I$ is the carrier power-to-interference noise power spectral density ratio (Hz) at the input of the considered receiver.
— $(C/N_0)_{IM}$ is the carrier power-to-intermodulation noise power spectral density ratio (Hz) at the output of the considered non-linear amplifier.
— $(C/N_0)_T$ is the overall carrier power-to-noise power spectral density ratio (Hz) at the earth station receiver input.

5.9.1 Characteristics of the satellite channel

Figure 5.33 shows a *transparent* payload, where carriers are power amplified and frequency downconverted. Due to technology power limitations, the overall bandwidth is split into several sub-bands, the carriers in each sub-band being amplified by a dedicated power amplifier. The amplifying chain associated with each sub-band is called a *satellite channel*, or transponder. The satellite channel amplifies one or several carriers. Here is some more notation:

— C_U is the considered carrier power at the satellite receiver input; at saturation, it is denoted $(C_U)_{sat}$;
— P_{in} is the power at the input to the satellite channel amplifier (i = input, n = number of carriers in the channel);

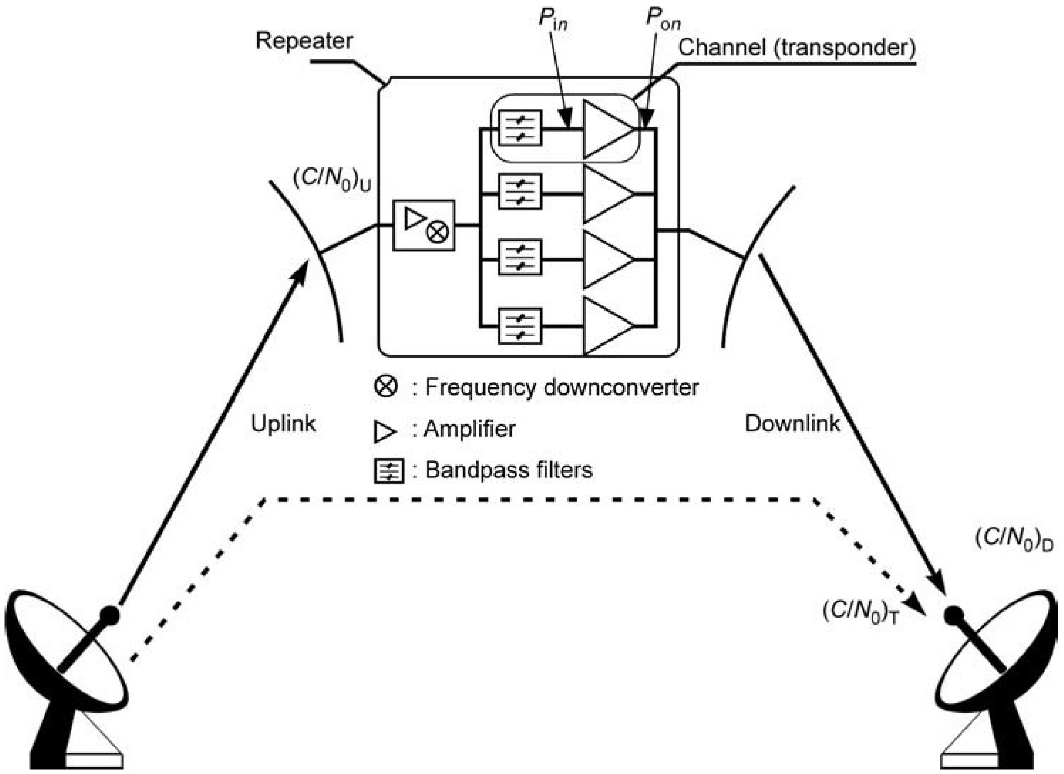

Figure 5.33 Overall station-to-station link for a transparent satellite.

— P_{on} is the power at the output of the satellite channel amplifier (o = output, n = number of carriers in the channel);
— $n = 1$ corresponds to a single-carrier operation of the satellite channel;
— $(P_{i1})_{sat}$ is the power at the input to the satellite channel amplifier at saturation in single-carrier operation;
— $(P_{o1})_{sat}$ is the power at the output of the satellite channel amplifier at saturation in single-carrier operation.

Saturation refers to the operation of the amplifier at maximum output power in single-carrier operation (Section 9.2.1.2). The satellite operator provides characteristic values of a satellite channel in terms of flux density at saturation, Φ_{sat}, and EIRP at saturation, $EIRP_{sat}$.

5.9.1.1 *Satellite power flux density at saturation*

Power flux density was introduced in Section 5.3.2. This flux is provided by the transmit earth station and is considered at the satellite receive antenna (Section 5.4.2). Its nominal value to drive the satellite channel amplifier at saturation is given by:

$$\Phi_{sat,\ nom} = \frac{(P_{i1})_{sat}}{G_{FE}} \frac{L_{FRX}}{G_{Rmax}} \frac{4\pi}{\lambda_U^2} \quad (W/m^2) \tag{5.58}$$

where G_{FE} is the front end gain from the input to the satellite receiver to the input to the satellite channel amplifier; L_{FRX} is the loss from the output of the satellite receive antenna to the input of the satellite receiver; and G_{Rmax} is the satellite receive antenna maximum gain (at boresight).

The formula assumes that the transmit earth station is located at the centre of the satellite receive coverage (satellite antenna boresight).

In practice, the flux to be provided from a given earth station to drive the satellite channel amplifier to saturation depends on the location of the transmit earth station within the satellite coverage and the polarisation mismatch of the satellite receiving antenna with respect to the uplink carrier polarisation. Assuming that the receive satellite antenna gain in the direction of the transmit earth station experiences a gain fallout L_R with respect to the maximum gain, and a polarisation mismatch loss L_{POL}, the actual flux density to be provided by the transmit earth station is larger than or equal to:

$$\Phi_{sat} = \Phi_{sat,\ nom} L_R L_{POL} = \frac{(P_{i1})_{sat}}{G_{FE}} \frac{L_{FRX}}{G_{Rmax}} \frac{4\pi}{\lambda_U^2} L_R L_{POL} \quad (\mathrm{W/m^2})$$

5.9.1.2 Satellite EIRP at saturation

EIRP was introduced in Section 5.3.1. The satellite EIRP at saturation and boresight, $\mathrm{EIRP}_{sat,\ max}$, relates to the satellite channel amplifier output power at saturation, $(P_{o1})_{sat}$, as follows:

$$\mathrm{EIRP}_{sat,\ max} = \frac{(P_{o1})_{sat}}{L_{FTX}} G_{Tmax} \quad (W) \tag{5.59}$$

where L_{FTX} is the loss from the output of the power amplifier to the transmit antenna, and G_{Tmax} is the satellite transmit antenna maximum gain (at boresight).

In practice, the satellite EIRP_{sat} which conditions the available carrier power at a given earth station receiver input is reduced by the transmit satellite antenna gain fallout L_T (the gain fallout is defined in the direction of the receiving earth station, with respect to the maximum gain) when the earth station is not located at the centre of transmit coverage (satellite antenna boresight):

$$\mathrm{EIRP}_{sat} = \frac{\mathrm{EIRP}_{sat,max}}{L_T} = \frac{(P_{01})_{sat}}{L_{FTX}} \frac{G_{Tmax}}{L_T} = \frac{(P_{01})_{sat}}{L_{FTX}} G_T \quad (\mathrm{W}) \tag{5.60}$$

5.9.1.3 Satellite repeater gain

The satellite repeater gain, G_{SR}, is the power gain from the satellite receiver input to the satellite channel amplifier output. At saturation, it is called $G_{SR\ sat}$.

$$G_{SR} = G_{FE} G_{CA} \tag{5.61}$$

where G_{FE} is the front end gain (from satellite receiver input to satellite channel amplifier input) and G_{CA} is the satellite channel amplifier gain.

5.9.1.4 Input and output back-off

In practice, the satellite channel power amplifier is not always operated at saturation, and it is convenient to determine the operating point Q of the satellite channel amplifier determined by the input power $(P_{in})_Q$ and the output power $(P_{on})_Q$. It is convenient to normalise these quantities with respect to $(P_{i1})_{sat}$ and $(P_{o1})_{sat}$ respectively. This defines the input back-off (IBO) and the output back-off (OBO):

$$\mathrm{IBO} = (P_{\mathrm{i}n})_{\mathrm{Q}}/(P_{\mathrm{i1}})_{\mathrm{sat}} \tag{5.62}$$

$$\mathrm{OBO} = (P_{\mathrm{o}n})_{\mathrm{Q}}/(P_{\mathrm{o1}})_{\mathrm{sat}} \tag{5.63}$$

From now on, the operating power value is denoted without the Q subscript.

5.9.1.5 Carrier power at satellite receiver input

The carrier power required at the satellite receiver input to drive the satellite channel amplifier to operate at the considered operating point Q is given by:

$$C_{\mathrm{U}} = \frac{(P_{\mathrm{i}n})_{\mathrm{Q}}}{G_{\mathrm{FE}}} = \frac{P_{\mathrm{i}n}}{G_{\mathrm{FE}}} = \mathrm{IBO}\frac{(P_{\mathrm{i}n})_{\mathrm{sat}}}{G_{\mathrm{FE}}} \quad (\mathrm{W}) \tag{5.64}$$

The carrier power can also be expressed as a function of the satellite channel amplifier output power:

$$C_{\mathrm{U}} = \mathrm{IBO}\frac{P_{\mathrm{o}n}}{G_{\mathrm{FE}}G_{\mathrm{CA}}} = \mathrm{IBO}\frac{(P_{01})_{\mathrm{sat}}}{G_{\mathrm{FE}}(G_{\mathrm{CA}})_{\mathrm{sat}}} \quad (\mathrm{W}) \tag{5.65}$$

where $(G_{\mathrm{CA}})_{\mathrm{sat}}$ is the satellite amplifier gain *at saturation*. Finally, C_{U} can be expressed as:

$$C_{\mathrm{U}} = \mathrm{IBO}(C_{\mathrm{U}})_{\mathrm{sat}} \quad (\mathrm{W}) \tag{5.66}$$

where

$$(C_{\mathrm{U}})_{\mathrm{sat}} = \frac{(P_{\mathrm{i1}})_{\mathrm{sat}}}{G_{\mathrm{FE}}} = \frac{(P_{\mathrm{o1}})_{\mathrm{sat}}}{G_{\mathrm{FE}}(G_{\mathrm{CA}})_{\mathrm{sat}}}$$

is the carrier power required at the satellite receiver input to drive the satellite channel amplifier at saturation. $(C_{\mathrm{U}})_{\mathrm{sat}}$ can also be expressed as a function of Φ_{sat}:

$$(C_{\mathrm{U}})_{\mathrm{sat}} = \Phi_{\mathrm{sat}}\frac{G_{\mathrm{Rmax}}}{L_{\mathrm{FRX}}}\frac{\lambda_{\mathrm{U}}^2}{4\pi}\frac{1}{L_{\mathrm{R}}\,L_{\mathrm{POL}}} \quad (\mathrm{W}) \tag{5.67}$$

or

$$(C_{\mathrm{U}})_{\mathrm{sat}} = \Phi_{\mathrm{sat,\ nom}}\frac{G_{\mathrm{Rmax}}}{L_{\mathrm{FRX}}}\frac{\lambda_{\mathrm{U}}^2}{4\pi}$$

Note that the input back-off IBO can also be expressed as the ratio of the power flux density Φ required to operate the satellite channel amplifier at the considered operating point to the satellite power flux density at saturation:

$$\mathrm{IBO} = \frac{C_{\mathrm{U}}}{(C_{\mathrm{U}})_{\mathrm{sat}}} = \frac{\Phi}{\Phi_{\mathrm{sat}}}$$

5.9.2 Expression for $(C/N_0)_T$

5.9.2.1 Expression for $(C/N_0)_T$ without interference from other systems or intermodulation

The power of the carrier received at the input of the earth station receiver is C_D. The noise at the input of the earth station receiver corresponds to the sum of the following:

— the downlink system noise considered in isolation ($T_D = T_2$, given by equation (5.32)) which defines the ratio C/N_0 for the downlink $(C/N_0)_D$ and can be calculated as in the example of Section 5.6.3 with $(N_0)_D = kT_D$;
— the uplink noise retransmitted by the satellite.

Hence:

$$(N_0)_T = (N_0)_D + G(N_0)_U \quad \text{(W/Hz)} \tag{5.68}$$

where $G = G_{SR}G_TG_R/L_{FTX}L_DL_{FRX}$ is the total power gain between the satellite receiver input and the earth station receiver input. G takes into account the satellite repeater gain G_{SR} from the input to the satellite receiver to the output of the satellite channel amplifier; the gain G_T/L_{FTX} of the satellite transmit antenna including the gain fallout and the loss L_{FTX} from the output of the power amplifier to the transmit antenna; the downlink path loss L_D and the receiving station composite gain G_R/L_{FRX}. This gives

$$\begin{aligned}(C/N_0)_T^{-1} &= (N_0)_T/C_D \\ &= [(N_0)_D + G(N_0)_U]/C_D = (N_0)_D/C_D + (N_0)_U/G^{-1}C_D \quad (\text{Hz}^{-1})\end{aligned} \tag{5.69}$$

In the above expression, the term $G^{-1}C_D$ represents the carrier power at the satellite receiver input. Hence $(N_0)_U/G^{-1}C_D = (C/N_0)_U^{-1}$. Finally:

$$(C/N_0)_T^{-1} = (C/N_0)_U^{-1} + (C/N_0)_D^{-1} \quad (\text{Hz}^{-1}) \tag{5.70}$$

In this expression:

$$\begin{aligned}(C/N_0)_U &= (P_{i1})/(N_0)_U = \text{IBO}(P_{i1})_{sat}/(N_0)_U \\ &= \text{IBO}(P_{o1})_{sat}/G_{SR\,sat}(N_0)_U \\ &= \text{IBO}(C/N_0)_{U\,sat} \quad \text{(Hz)} \\ (C/N_0)_D &= \text{OBO}(\text{EIRP}_{sat})_{SL}(1/L_D)(G/T)_{ES}(1/k) \\ &= \text{OBO}(C/N_0)_{D\,sat} \quad \text{(Hz)}\end{aligned}$$

$(C/N_0)_{U\,sat}$ and $(C/N_0)_{D\,sat}$ are the values of C/N_0 for the uplink and downlink when the satellite channel operates at saturation. L_D represents the attenuation on the downlink and is given by equation (5.14) and $(G/T)_{ES}$, the figure of merit of the earth station in the satellite direction.

5.9.2.2 Expression for $(C/N_0)_T$ taking account of interference

Interference is the unwanted power contribution of other carriers in the frequency band occupied by the wanted carrier. A given link may suffer interference from other satellite links or from terrestrial systems operating in the same frequency bands as the considered link. Indeed, most of the frequency bands allocated to space radiocommunications are also allocated on a shared basis to

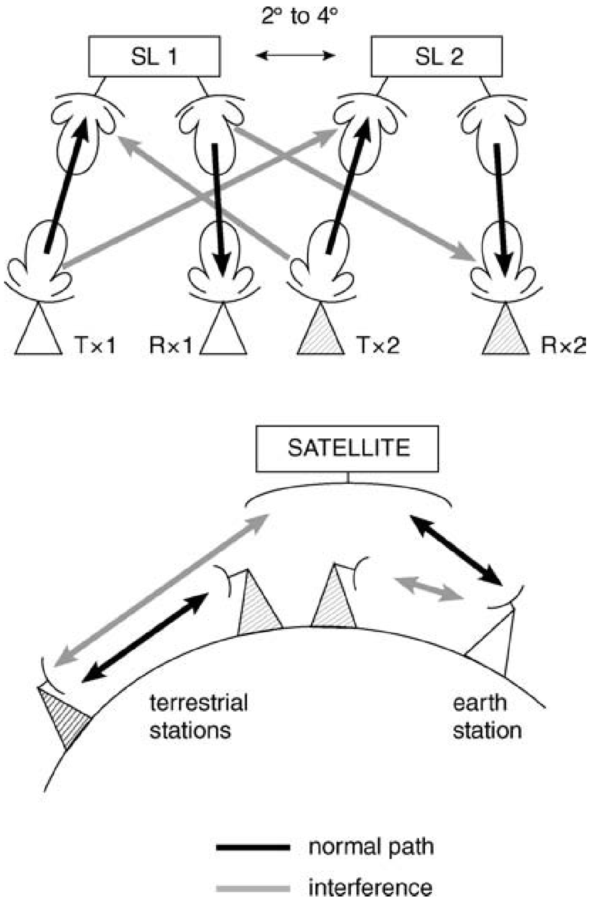

Figure 5.34 The geometry of interference between systems.

terrestrial radiocommunications. To facilitate this sharing, a number of provisions have been introduced into the *Radiocommunication Regulations*, Articles S 21 and S 9, and a coordination procedure has been instituted between earth and terrestrial stations (*Radiocommunication Regulations*, Article 11).

Four types of interference between systems can be distinguished:

— a satellite interfering with a terrestrial station;
— a terrestrial station interfering with a satellite;
— an earth station interfering with a terrestrial station;
— a terrestrial station interfering with an earth station.

Figure 5.34 illustrates the geometry associated with these forms of interference.

The carriers emitted by other systems are superimposed on the wanted carrier of the station-to-station link at two levels:

— at the input of the satellite repeater on the uplink;
— at the input of the earth station receiver on the downlink

The effect of interference is similar to an increase of the thermal noise on the link affected by interference. It is allowed for in the equations in the form of an increase of the spectral density:

$$N_0 = (N_0)_{\text{without interference}} + (N_0)_\text{I} \quad (\text{W/Hz}) \tag{5.71}$$

where $(N_0)_I$ represents the increase of the noise power spectral density due to interference. A ratio $(C/N_0)_I$ which expresses the signal power in relation to the spectral density of the interference can be associated with $(N_0)_I$; these are $(C/N_0)_{I,U}$ for the uplink and $(C/N_0)_{I,D}$ for the downlink. This leads to modification of equation (5.70), replacing $(C/N_0)_U$ and $(C/N_0)_D$ by the following expressions:

$$(C/N_0)_U^{-1} = [(C/N_0)_U^{-1}]_{\text{without interference}} + (C/N_0)_{I,U}^{-1} \quad (\text{Hz}^{-1})$$

$$(C/N_0)_D^{-1} = [(C/N_0)_D^{-1}]_{\text{without interference}} + (C/N_0)_{I,D}^{-1} \quad (\text{Hz}^{-1}) \tag{5.72}$$

The total expression becomes:

$$(C/N_0)_T^{-1} = (C/N_0)_U^{-1} + (C/N_0)_D^{-1} + (C/N_0)_I^{-1} \quad (\text{Hz}^{-1}) \tag{5.73}$$

where $(C/N_0)_U$ and $(C/N_0)_D$ are the values appearing in equation (5.70) and:

$$(C/N_0)_I^{-1} = (C/N_0)_{I,U}^{-1} + (C/N_0)_{I,D}^{-1} \quad (\text{Hz}^{-1}) \tag{5.74}$$

5.9.2.3 *Expression for $(C/N_0)_T$ taking account of intermodulation and interference*

When several carriers are amplified in a non-linear amplifier, the output is not only the amplified carriers but also the intermodulation products, which appear as power at frequencies that are linear combinations of the input carrier frequencies (Section 6.5.4). Some of these intermodulation products fall in the bandwidth of the considered carrier and act as noise with spectral density $(N_0)_{IM}$. The ratio of the carrier power to the intermodulation noise spectral density is $(C/N_0)_{IM}$.

Intermodulation noise is added to the other sources of noise analysed in this chapter. Equation (5.74) for the carrier power-to-noise power spectral density ratio for the overall station-to-station link $(C/N_0)_T$ is modified as follows:

$$(C/N_0)_T^{-1} = (C/N_0)_U^{-1} + (C/N_0)_D^{-1} + (C/N_0)_I^{-1} + (C/N_0)_{IM}^{-1} \quad (\text{Hz}^{-1}) \tag{5.75}$$

with:

$$(C/N_0)_{IM}^{-1} = (C/N_0)_{IM,\,U}^{-1} + (C/N_0)_{IM,\,D}^{-1}$$

where $(C/N_0)_{IM,\,U}^{-1}$ and $(C/N_0)_{IM,\,D}$ correspond to the generation of intermodulation noise in the transmitting earth station and the satellite repeater channel, respectively.

In this case, the expressions for the ratios $(C/N_0)_U$, $(C/N_0)_D$ and $(C/N_0)_{IM}$ are to be used with values of input and output back-off IBO and OBO for operation of the amplifier in multicarrier mode with carriers of *equal power*. The output power of the amplifier is shared among the carriers, the thermal noise and the intermodulation noise to which the interference noise for the channel is added.

Denoting by P_{in} and P_{on} respectively the input and output and power of one carrier among the n amplified ones, input and output back-off are defined as follows:

— input back-off *per carrier*:
- IBO_1 = single carrier input power/single carrier input power at saturation = $P_{i1}/(P_{i1})_{sat}$
  or, in dB:
- IBO_1 (dB) = 10 log $\{P_{i1}/(P_{i1})_{sat}\}$

— output back-off *per carrier*:
- OBO_1 = single carrier output power/single carrier output power at saturation = $P_{o1}/(P_{o1})_{sat}$

or, in dB:
- IBO_1 (dB) = 10 $\log\{P_{o1}/(P_{o1})_{sat}\}$

— total input back-off:
- IBO_t = sum of all input carrier power/single carrier input power at saturation = $\Sigma P_{in}/(P_{i1})_{sat}$

or, in dB:
- IBO_t (dB) = 10 $\log\{\Sigma P_{in}/(P_{i1})_{sat}\}$

— total output back-off:
- OBO_t = sum of all output carrier power/single carrier output power at saturation = $\Sigma P_{on}/(P_{o1})_{sat}$

or, in dB:
- OBO_t (dB) = 10 $\log\{\Sigma P_{on}/(P_{o1})_{sat}\}$

With n equally powered carriers:

— $IBO_1 = IBO_t/n$ or, in dB, IBO_1 (dB) = IBO_t (dB) − 10 log n
— $OBO_1 = OBO_t/n$ or, in dB, OBO_1 (dB) = OBO_t (dB) − 10 log n

If the carriers at the amplifier input are of *unequal power*, the power at the amplifier output is shared unequally between carriers and noise. Therefore the amplifier does not have equal power gain for all carriers and a *capture effect* can arise: carriers of high power acquire more power than carriers of low power. For carriers of high power, the value of the ratio is greater than that given by equation (5.75). For carriers of low power, it is smaller. Generation of intermodulation products is also observed between noise on the uplink and the carriers; this effect can be taken into account in the form of an increase in the noise temperature at the channel input.

5.9.2.4 *Influence of back-off*

Figure 5.35 shows the variation of each of the terms in equation (5.75) as a function of input back-off IBO assuming the equivalent interference noise to be negligible. Because of the opposite direction of variation of the term $(C/N_0)_{IM}$ compared to that of the ratios $(C/N_0)_U$ and $(C/N_0)_D$, the value of $(C/N_0)_T$ passes through a maximum for a non-zero value of back-off. Two effects are, therefore, observed which are consequences of using the same repeater channel to amplify several carriers:

— The total power at the output of the channel is less than that which would exist in the absence of back-off.
— The useful power per carrier is reduced by allocation of part of the total power to intermodulation products.

5.9.3 Overall link performance for a transparent satellite without interference or intermodulation

It is required to establish a satellite link between two earth stations (Figure 5.33), assumed to be located at the centre of the satellite antenna's coverage. The data are as follows:

— Uplink frequency: f_U = 14 GHz.
— Downlink frequency: f_D = 12 GHz.
— Downlink path loss: L_D = 206 dB.

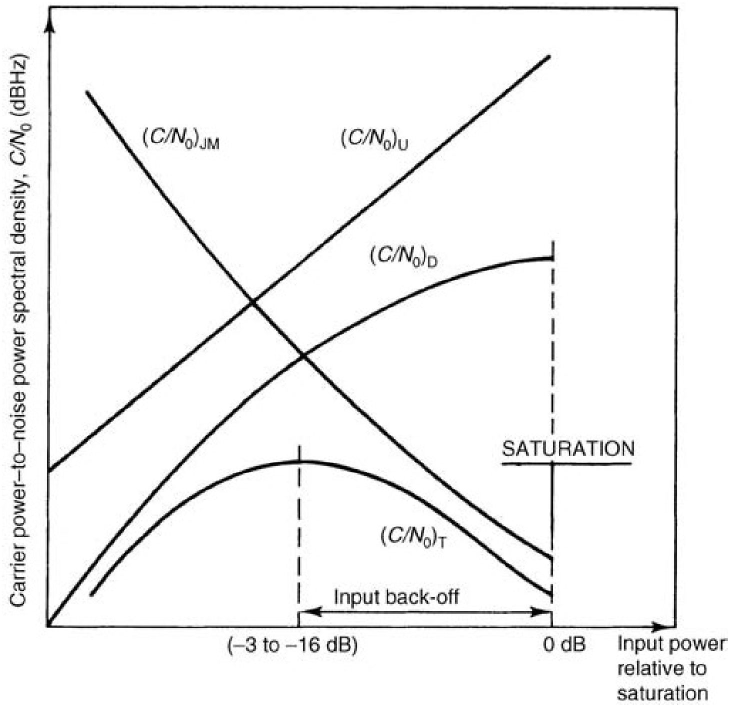

Figure 5.35 Variation of $(C/N_0)_U$, $(C/N_0)_D$, $(C/N_0)_{IM}$ and $(C/N_0)_T$ as a function of input back-off IBO.

— For the satellite (SL):
  – Power flux density required to saturate the satellite channel amplifier:

$$(\Phi_{\mathrm{sat,\,nom}})_{\mathrm{SL}} = -90\,\mathrm{dBW/m^2}$$

  – Satellite receiving antenna gain at boresight: $G_{\mathrm{Rmax}} = 30\,\mathrm{dBi}$
  – Satellite figure of merit at boresight: $(G/T)_{\mathrm{SL}} = 3.4\,\mathrm{dBK^{-1}}$
  – Satellite channel amplifier characteristic (single carrier operation) modelled by:

$$\mathrm{OBO(dB)} = \mathrm{IBO(dB)} + 6 - 6\exp[\mathrm{IBO(dB)}/6]$$

  – Satellite effective isotropic radiated power at saturation in the direction of the considered receiving earth station (i.e. at boresight of the satellite transmitting antenna)

$$(\mathrm{EIRP_{sat}})_{\mathrm{SL}} = 50\,\mathrm{dBW}$$

  – Satellite transmitting antenna gain at boresight: $G_{\mathrm{Tmax}} = 40\,\mathrm{dBi}$

The following losses are considered:

— Satellite reception and transmission feeder losses: $L_{\mathrm{FRX}} = L_{\mathrm{FTX}} = 0\,\mathrm{dB}$
— Satellite antenna polarisation mismatch loss $L_{\mathrm{POL}} = 0\,\mathrm{dB}$
— Satellite antenna depointing losses: $L_{\mathrm{R}} = L_{\mathrm{T}} = 0\,\mathrm{dB}$ (earth stations at boresight)
— For the earth station (ES): Figure of merit of earth station in satellite direction $(G/T)_{\mathrm{ES}} = 25\,\mathrm{dBK^{-1}}$

It is assumed that there is no interference.

5.9.3.1 *Satellite repeater gain at saturation $G_{SR\ sat}$*

$$G_{\text{SR sat}} = (P_o^1)_{\text{sat}}/(C_{\text{U}})_{\text{sat}}$$

where $(C_{\text{U}})_{\text{sat}}$ is the carrier power required at the satellite receiver input to drive the satellite channel amplifier at saturation. From equation (5.60):

$$(P_{\text{o1}})_{\text{sat}} = (\text{EIRP}_{\text{sat}})_{\text{SL}} L_T L_{\text{FTX}}/G_{\text{Tmax}} \quad (\text{W})$$

Hence:

$$(P_{\text{o1}})_{\text{sat}} = 50\,\text{dBW} - 40\,\text{dBi} = 10\,\text{dBW} = 10\,\text{W}$$

From equation (5.67)

$$(C_{\text{U}})_{\text{sat}} = (\Phi_{\text{sat}})_{\text{SL}} G_{\text{Rmax}}/L_{\text{FRX}} L_{\text{R}} L_{\text{POL}} (4\pi/\lambda_{\text{U}}^2) \quad (\text{W})$$

hence:

$$\begin{aligned}(C_{\text{U}})_{\text{sat}} &= -90\,\text{dBW/m}^2 + 30\,\text{dBi} - 44.4\,\text{dBm}^2 = -104.4\,\text{dBW} = 36\,\text{pW}\\ G_{\text{SR sat}} &= (P_{o1})_{\text{sat}}/(C_{\text{U}})_{\text{sat}} = 10\,\text{dBW} - (-104.4\,\text{dBW}) = 114.4\,\text{dB}\end{aligned}$$

5.9.3.2 *Calculation of C/N_0 for the up- and downlinks and the overall link when the repeater operates at saturation*

$$(C/N_0)_{\text{U sat}} = (C_U)_{\text{sat}}/kT_{\text{U}} = (C_U)_{\text{sat}}(G/T)_{\text{SL}}/(kG_{\text{Rmax}}/L_R L_{FRX} L_{POL})$$

$$(C/N_0)_{\text{U sat}} = -104.4 + 3.4 - (-228.6) - 30 = 97.6\,\text{dBHz}$$

$$(C/N_0)_{\text{D sat}} = (\text{EIRP}_{\text{sat}})_{\text{SL}}(1/L_{\text{D}})(G/T)_{\text{ES}}(1/k) \quad (\text{Hz})$$

$$(C/N_0)_{\text{D sat}} = 50 - 206 + 25 - (-228.6) = 97.6\,\text{dBHz}$$

$$(C/N_0)^{-1}_{\text{T sat}} = (C/N_0)^{-1}_{\text{U sat}} + (C/N_0)^{-1}_{\text{D sat}} \quad (\text{Hz}^{-1})$$

$$(C/N_0)_{\text{T sat}} = 94.6\,\text{dBHz}$$

5.9.3.3 *Calculation of the input and output back-off to achieve $(C/N_0)_T = 80$ dBHz and the corresponding values of $(C/N_0)_U$ and $(C/N_0)_D$*

One must have:

$$(C/N_0)^{-1}_{\text{U}} + (C/N_0)^{-1}_{\text{D}} = 10^{-8}\,\text{Hz}^{-1}$$

Hence:

$$\text{IBO}^{-1}(C/N_0)^{-1}_{\text{U sat}} + \text{OBO}^{-1}(C/N_0)^{-1}_{\text{D sat}} = 10^{-8}\,\text{Hz}^{-1}$$

This gives:

$$10^{-\text{IBO(dB)}/10} + 10^{-\text{OBO(dB)}/10} = 10^{1.76}$$

with:

$$\mathrm{OBO(dB)} = \mathrm{IBO(dB)} + 6 - 6\exp(\mathrm{IBO(dB)}/6)$$

Numerical solution gives:

$$\begin{aligned} \mathrm{IBO} &= -16.4\,\mathrm{dB} \\ \mathrm{OBO} &= -10.8\,\mathrm{dB} \end{aligned}$$

Hence:

$$(C/N_0)_{\mathrm{U}} = \mathrm{IBO}(C/N_0)_{\mathrm{U\,sat}} = -16.4\,\mathrm{dB} + 97.6\,\mathrm{dBHz} = 81.2\,\mathrm{dBHz}$$
$$(C/N_0)_{\mathrm{D}} = \mathrm{OBO}(C/N_0)_{\mathrm{D\,sat}} = -10.8\,\mathrm{dB} + 97.6\,\mathrm{dBHz} = 86.8\,\mathrm{dBHz}$$

5.9.3.4 Value of $(C/N_0)_{\mathrm{T}}$ under rain conditions causing an attenuation of 6 dB on the uplink

The attenuation of 6 dB on the uplink reduces the input back-off by 6 dB. The new value of IBO becomes:

$$\mathrm{IBO(dB)} = -16.4\,\mathrm{dB} - 6\,\mathrm{dB} = -22.4\,\mathrm{dB}$$

The new value of output back-off corresponding to this is:

$$\mathrm{OBO(dB)} = \mathrm{IBO(dB)} + 6 - 6\exp(\mathrm{IBO(dB)}/6) = -16.5\,\mathrm{dB}$$

Hence:

$$(C/N_0)_{\mathrm{U}} = \mathrm{IBO}(C/N_0)_{\mathrm{U\,sat}} = -22.4\,\mathrm{dB} + 97.6\,\mathrm{dBHz} = 75.2\,\mathrm{dBHz}$$
$$(C/N_0)_{\mathrm{D}} = \mathrm{OBO}(C/N_0)_{\mathrm{D\,sat}} = -16.5\,\mathrm{dB} + 97.6\,\mathrm{dBHz} = 81.1\,\mathrm{dBHz}$$

and, from equation (5.70)

$$(C/N_0)_{\mathrm{T}} = 74.2\,\mathrm{dBHz}$$

To regain the required value $(C/N_0)_{\mathrm{T}} = 80$ dBHz, it is necessary to increase the $(\mathrm{EIRP})_{\mathrm{ES}}$ of the transmitting earth station by 6 dB.

5.9.3.5 Value of $(C/N_0)_{\mathrm{T}}$ under rain conditions causing an attenuation of 6 dB on the downlink with a reduction of 2 dB in the figure of merit of the earth station due to the increase of antenna noise temperature

The value of $(C/N_0)_{\mathrm{D}}$ reduces by 8 dB, hence: $(C/N_0)_{\mathrm{D}} = 86.8\,\mathrm{dBHz} - 8\,\mathrm{dB} = 78.8$ dBHz. From which:$(C/N_0)_{\mathrm{T}} = 76.8$ dBHz.

To regain the required value $(C/N_0)_{\mathrm{T}} = 80$ dBHz, it is necessary to increase the $(\mathrm{EIRP})_{\mathrm{ES}}$ of the transmitting earth station in such a way that the value of IBO satisfies the equation:

$$\mathrm{IBO}^{-1}(C/N_0)^{-1}_{\mathrm{U\,sat}} + \mathrm{OBO}^{-1}(C/N_0)^{-1}_{\mathrm{D\,sat}} = 10^{-8}\,\mathrm{Hz}^{-1}$$

in which:

$$(C/N_0)_{U\,sat} = 97.6\,\text{dBHz}$$
$$(C/N_0)_{D\,sat} = 97.6\,\text{dBHz} - 8\,\text{dB} = 89.6\,\text{dBHz}$$

This gives:

$$\text{IBO} = -13\,\text{dB}$$
$$\text{OBO} = -7.7\,\text{dB}$$

It is necessary to increase the $(\text{EIRP})_{ES}$ of the earth station transmission by $-13\,\text{dB} - (-16.4\,\text{dB}) = 3.4\,\text{dB}$.

Hence:

$$(C/N_0)_U = \text{IBO}(C/N_0)_{U\,sat} = -13\,\text{dB} + 97.6\,\text{dBHz} = 84.6\,\text{dBHz}$$
$$(C/N_0)_D = \text{OBO}(C/N_0)_{D\,sat} = -7.7\,\text{dB} + 89.6\,\text{dBHz} = 81.9\,\text{dBHz}$$

5.10 OVERALL LINK PERFORMANCE WITH REGENERATIVE SATELLITE

Figure 5.36 shows the difference between a regenerative satellite repeater and a transparent one. With the regenerative repeater, baseband signals, which have modulated the uplink carrier, are available at the output of the demodulator and these signals are used (possibly after processing not shown in the figure) to modulate the downlink carrier. Hence the change of frequency from uplink to downlink which is obtained by mixing with the local radio-frequency oscillator in a transparent satellite is obtained in this case by modulation of a new carrier.

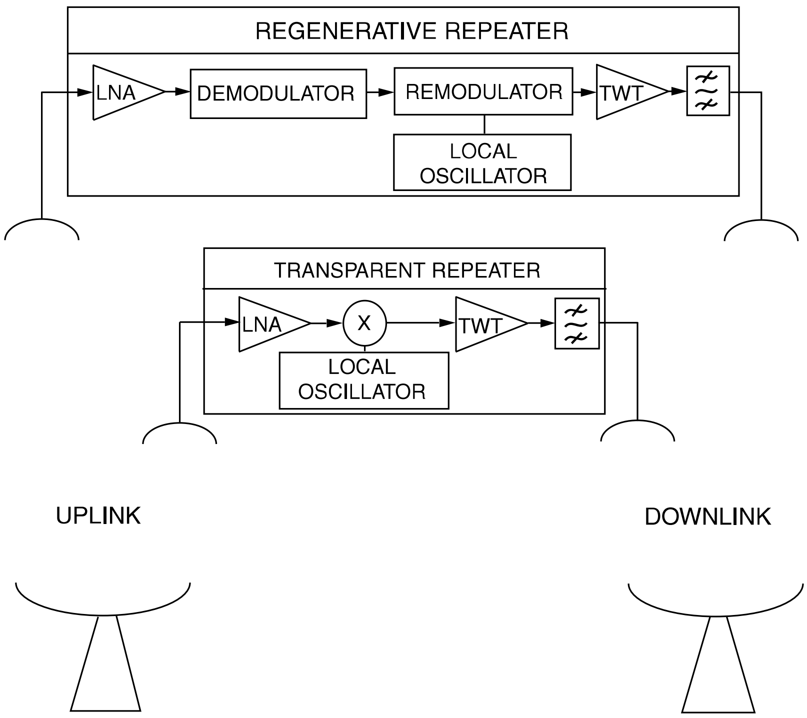

Figure 5.36 Organisation of a regenerative repeater and a transparent repeater.

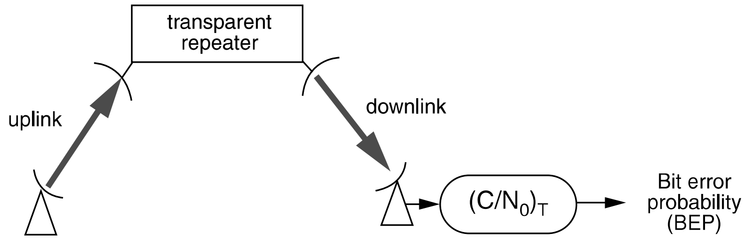

Figure 5.37 Link by transparent repeater.

5.10.1 Linear satellite channel without interference

It is assumed that the probability of error at the output of the demodulator is that given by theory (Table 4.4); that is, there is no degradation due to filtering or non-linearities.

5.10.1.1 Link with transparent repeater

The performance of the link (Figure 5.37) is specified in terms of the bit error probability (BEP) at the output of the earth station demodulator. BEP is a function of the ratio $(E/N_0)_\mathrm{T}$ given by equation (4.10) in Section 4.2.6.2 and recalled here:

$$(E/N_0)_\mathrm{T} = (C/N_0)_T/R_\mathrm{c} \tag{5.76}$$

where R_c is the carrier data rate and $(C/N_0)_\mathrm{T}$ is the ratio of carrier power to noise spectral density of the station-to-station link given by equation (5.70) and recalled here:

$$(C/N_0)_\mathrm{T}^{-1} = (C/N_0)_\mathrm{U}^{-1} + (C/N_0)_\mathrm{D}^{-1} \tag{5.77}$$

Defining $(E/N_0)_\mathrm{U} = (C/N_0)_U/R_\mathrm{c}$ and $(E/N_0)_\mathrm{D} = (C/N_0)_\mathrm{D}/R_\mathrm{c}$ and using equations (5.76) and (5.77) gives:

$$(E/N_0)_\mathrm{T}^{-1} = (E/N_0)_\mathrm{U}^{-1} + (E/N_0)_\mathrm{D}^{-1} \tag{5.78}$$

5.10.1.2 Link with regenerative repeater

The performance of the link (Figure 5.38), specified in terms of the bit error probability (BEP), is expressed as the probability of having an error on the uplink (measured by $\mathrm{BEP_U}$) and no error on

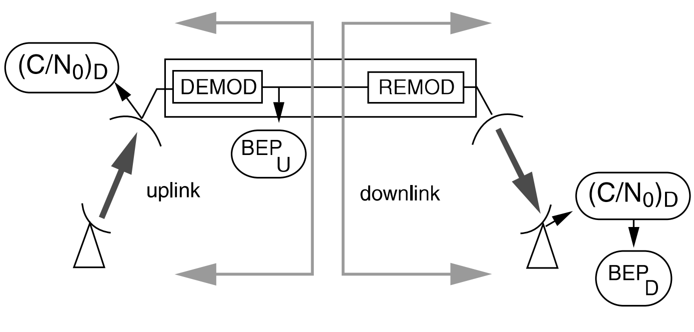

Figure 5.38 Link by regenerative repeater.

the downlink ($1-\text{BEP}_\text{D}$) or no error on the uplink ($1-\text{BEP}_\text{U}$) and an error on the downlink (BEP_D), hence:

$$\text{BEP} = \text{BEP}_\text{U}(1-\text{BEP}_\text{D}) + (1-\text{BEP}_\text{U})\text{BEP}_\text{D} \tag{5.79}$$

As BEP_U and BEP_D are small compared with 1, this becomes:

$$\text{BEP} = \text{BEP}_\text{U} + \text{BEP}_\text{D} \tag{5.80}$$

BEP_U is a function of $(E/N_0)_\text{U}$ and BEP_D is a function of $(E/N_0)_\text{D}$.

5.10.1.3 Comparison at constant bit error probability (BEP)

The value of the bit error probability is given as follows:

— For a transparent repeater the value of $(E/N_0)_\text{T}$ is determined by the BEP specified for the link. The required performance is obtained for a set of values of $(E/N_0)_\text{U}$ and $(E/N_0)_\text{D}$ combined using equation (5.78). This is shown by curve A of Figure 5.39 for an error probability of 10^{-4} and QPSK modulation with coherent demodulation.

— For a regenerative repeater, by combining BEP_U and BEP_D from equation (5.80) with the constraint $\text{BEP} = \text{constant} = 10^{-4}$ and deducing the corresponding pairs of values of $(E/N_0)_\text{U}$ and $(E/N_0)_\text{D}$, curves B and C of Figure 5.39 are obtained. These curves correspond to QPSK modulation with coherent demodulation (curve B) and differential demodulation (curve C) on the uplink. On the downlink, demodulation is coherent in both cases. The ratio $\alpha = (E/N_0)_\text{U}/(E/N_0)_\text{D}$ is used as a parameter.

Comparing curve A and curve B, it can be seen that the regenerative repeater provides a reduction of 3 dB in the value required for E/N_0 for the uplink and the downlink when the links are identical ($\alpha = 0$ dB). This is explained by the fact that the regenerative repeater does not transmit the amplified uplink noise along with the signal on the downlink, unlike a transparent repeater.

However, for very different values of E/N_0 this advantage disappears. For example, for α greater than 12 dB the two curves join; in this case the uplink noise is negligible and the performance of the overall link reduces in both cases to that of the downlink.

Curve C indicates that by dimensioning the uplink for a value of $(E/N_0)_\text{U}$ greater than that of $(E/N_0)_\text{D}$ by about 4 dB, on-board differential demodulation can be used without degrading the global performance and this is simpler to realise than coherent demodulation.

5.10.2 Non-linear satellite channel without interference

This corresponds more closely to a real system since a real channel is non-linear and band limited; the combination of non-linearities and filtering introduces a performance degradation of the demodulator which increases as the chain of non-linearities and filters increases. This is the case for a link with a transparent repeater (two non-linearities and filters—on transmission at the earth station and at the transponder). With a regenerative repeater, separation of the up- and downlinks means that there is now only one non-linearity and filter per link. Figure 5.40 shows the results obtained by means of computer simulation [WAC-81] for the case where $(E/N_0)_\text{U}$ is at least 12 dB greater than $(E/N_0)_\text{D}$. This figure and other results show that, contrary to the conclusions from a linear analysis, the regenerative repeater can provide between 2 and 5 dB reduction in E/N_0 with respect to a transparent repeater even when the ratio $\alpha = (E/N_0)_\text{U}/(E/N_0)_\text{D}$ is large.

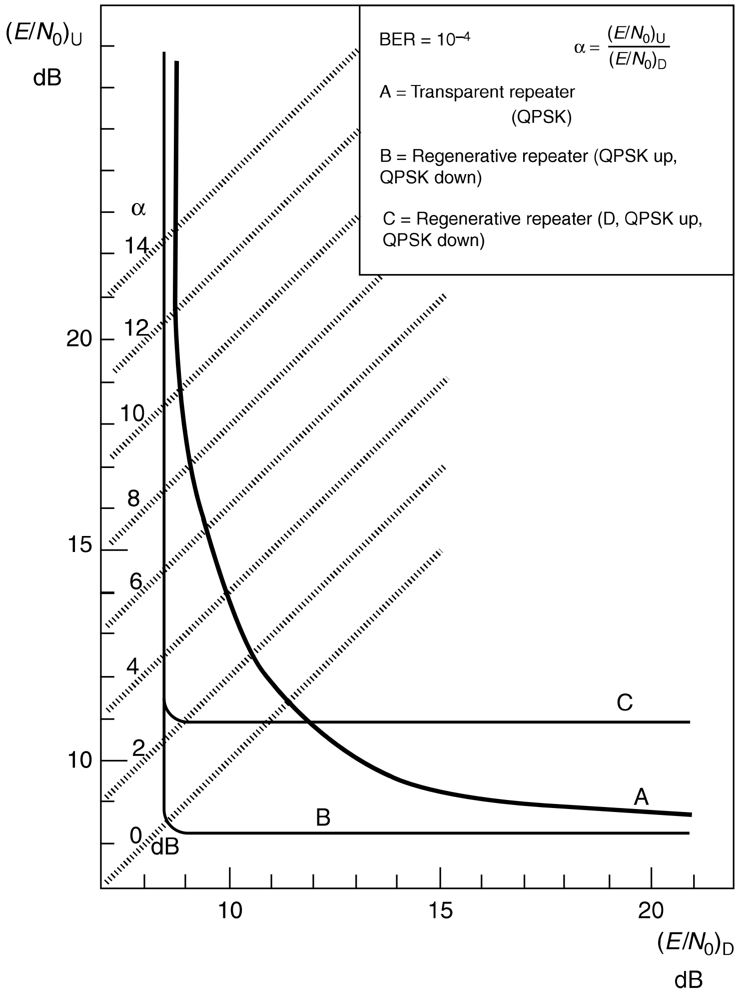

Figure 5.39 Comparison of station-to-station links by transparent repeater and regenerative repeater for the same bit error probability (BEP $= 10^{-4}$) (linear channel): (A) transparent repeater, (B) regenerative repeater with QPSK modulation and coherent demodulation on the uplink and the downlink and (C) regenerative repeater with QPSK modulation and differential demodulation on the uplink, and coherent demodulation on the downlink.

5.10.3 Non-linear satellite channel with interference

5.10.3.1 Link with transparent repeater

The value of $(C/N_0)_T$ depends on the value of $(C/N_0)_{T \text{ without interference}}$ in the absence of interference and $(C/N_0)_I$ due to interference on the up- and downlinks (see Section 5.9.2). More precisely:

$$(C/N_0)_T^{-1} = (C/N_0)_{T_{\text{without interference}}}^{-1} + (C/N_0)_I^{-1} \tag{5.81}$$

where:

$$(C/N_0)_{T_{\text{without interference}}}^{-1} = (C/N_0)_U^{-1} + (C/N_0)_D^{-1}$$
$$(C/N_0)_I^{-1} = (C/N_0)_{I,U}^{-1} + (C/N_0)_{I,D}^{-1}$$

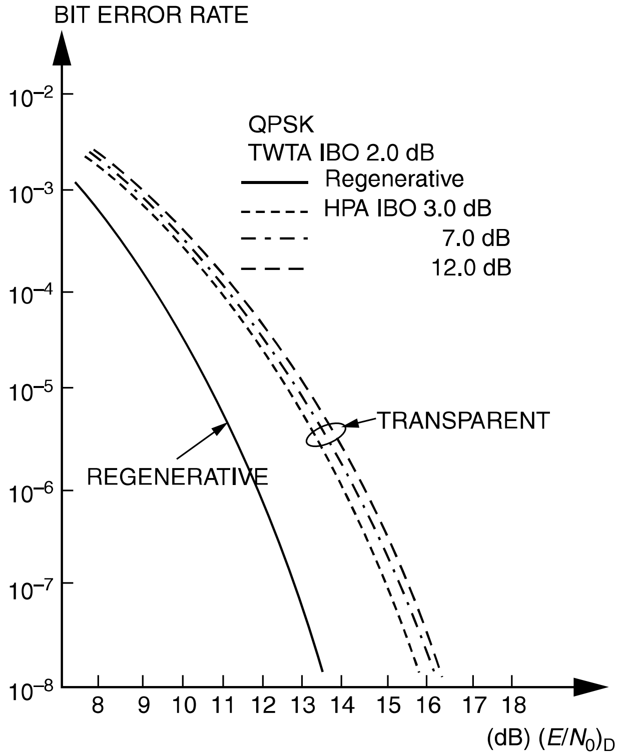

Figure 5.40 BEP as a function of $(E/N_0)_D$ for a link with a transparent repeater and for a link with a regenerative repeater in the absence of interference, for the case where $\alpha = (E/N_0)_U/(E/N_0)_D$ is high (larger than 12 dB). TWTA IBO: input back-off of the on-board travelling wave tube; HPA IBO: input back-off of the earth station transmitting amplifier ([WAC-81] ©1981 IEEE. Reproduced by permission.)

By putting $E/N_0 = C/N_0/R_c$, the relations between the values of E/N_0 can be deduced from these equations:

$$(E/N_0)_T^{-1} = (E/N_0)_{T_{\text{without interference}}}^{-1} + (E/N_0)_I^{-1} \tag{5.82}$$

From the curve of Figure 5.40 (which implies that α is large), BEP $= 10^{-4}$ with QPSK modulation requires that $(E/N_0)_T = 11$ dB. The upper curve of Figure 5.41 shows the relation between $(E/N_0)_I$ and $(E/N_0)_{T_{\text{without interference}}}$ obtained from equation (5.82) for this case.

5.10.3.2 Link with regenerative repeater

Considering that $\alpha = (E/N_0)_U/(E/N_0)_D$ is high, the bit error probability (BEP) of the station-to-station link is defined by the downlink BEP. A BEP of 10^{-4} requires, from Figure 5.40, $(E/N_0)_D = (E/N_0)_T = 9$ dB. The lower curve of Figure 5.41 shows the relation between $(E/N_0)_I$ and $(E/N_0)_{T_{\text{without interference}}}$ for this case.

Comparison of the two curves of Figure 5.41 shows that, for a given link quality (BEP $= 10^{-4}$), the ratio $(E/N_0)_I$ is less for a link with a regenerative repeater. This implies that the required link performance is obtained in spite of a higher level of interference. This is a useful advantage in the context of multibeam satellites, which possibly face higher levels of interference compared to single-beam satellites (Section 5.11.2).

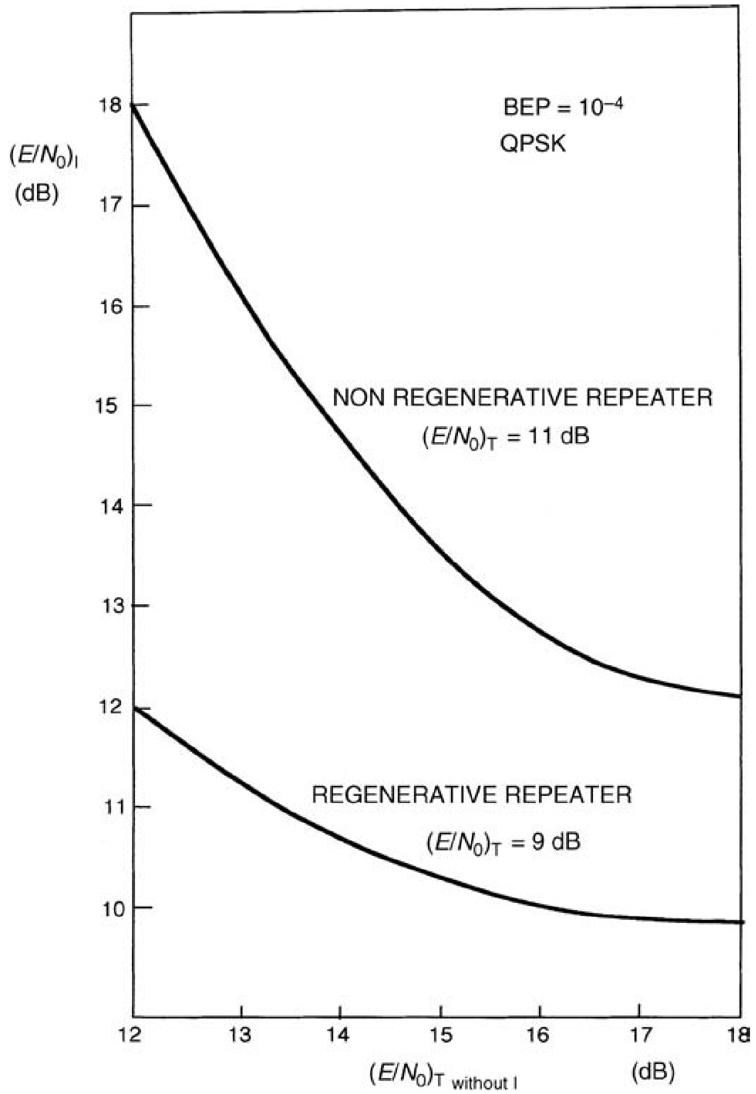

Figure 5.41 Permissible interference level; comparison of links with regenerative and transparent repeaters.

5.11 LINK PERFORMANCE WITH MULTIBEAM ANTENNA COVERAGE VS MONOBEAM COVERAGE

From the previous sections, it can be noticed that the overall radio-frequency link quality depends on the gain of the satellite antenna. From equation (5.7), $G_{max} = 29000/(\theta_d)^2$, it can be seen that the satellite antenna gain is constrained by its beamwidth, whatever the frequency at which the link is operated. So the antenna gain is imposed by the angular width of the antenna beam covering the zone to be served (see distinction between *coverage zone* and *service zone* in Sections 9.7 and 9.8). If the service zone is covered using a single antenna beam, this is referred to as 'single beam coverage'.

Single beam antenna coverage displays one of these characteristics:

— The satellite may provide coverage of the whole region of the earth which is visible from the satellite (global coverage) and thus permit long-distance links to be established, for example from one continent to another. In this case, the gain of the satellite antenna is limited by its beamwidth as imposed by the coverage. For a geostationary satellite, global coverage implies a 3 dB beamwidth of 17.5° and consequently an antenna gain of no more than G_{max} (dBi) = 10 log (29000) − 20 log (17.5) = 20 dBi.

— The satellite may provide coverage of only part of the earth (a region or country) by means of a narrow beam (a zone or spot beam), with 3dB beamwidth of the order of 1° to a few degrees. One thus benefits from a higher antenna gain due to the reduced antenna beamwidth, but the satellite cannot service earth stations situated outside this reduced coverage, and some of the earth stations that could be serviced with a global coverage are left out. These earth stations can be reached only by terrestrial links or by other satellites linked to the considered one by intersatellite links.

With single beam antenna coverage, it is therefore necessary to choose between either extended coverage providing service with reduced quality to geographically dispersed earth stations, or reduced coverage providing service with improved quality to geographically concentrated earth stations.

Multibeam antenna coverage allows these two alternatives to be reconciled. Satellite extended coverage may be achieved by means of the juxtaposition of several narrow beam coverages, each beam providing an antenna gain which increases as the antenna beamwidth decreases (reduced coverage per beam). The link performance improves as the number of beams increases; the limit is determined by the antenna technology, whose complexity increases with the number of beams, and the mass. The complexity originates in the more elaborate satellite antenna technology (multibeam antennas, see Chapter 9) and the requirement to provide on-board interconnection of the coverage areas, so as to ensure within the satellite payload routing of the various carriers that are uplinked in different beams to any wanted destination beam (see Chapter 7).

5.11.1 Advantages of multibeam coverage

In Figure 5.42a, a satellite provides global coverage with a single beam of beamwidth $\theta_{3\,dB} = 17.5°$ and, in Figure 5.42b, the satellite supports spot beams with beamwidth $\theta_{3\,dB} = 1.75°$ with a consequently reduced coverage. In both cases, all earth stations in the satellite network are within the satellite coverage.

5.11.1.1 Impact on earth segment

The expression for $(C/N_0)_U$ for the uplink is given by (see Section 5.6.2):

$$(C/N_0)_U = (\mathrm{EIRP})_{\mathrm{station}}(1/L_U)(G/T)_{\mathrm{satellite}}(1/k) \quad (\mathrm{Hz}) \tag{5.83}$$

Assuming that the noise temperature at the satellite receiver input is $T_{\mathrm{satellite}} = 800\ \mathrm{K} = 29$ dBK and is independent of the beam coverage (this is not rigorously true but satisfies a first approximation), let $L_U = 200$ dB and neglect the implementation losses. Equation (5.83) becomes (all terms in dB):

$$\begin{aligned}(C/N_0)_U &= (\mathrm{EIRP})_{\mathrm{station}} - 200 + (G_R)_{\mathrm{satellite}} - 29 + 228.6\\ &= (\mathrm{EIRP})_{\mathrm{station}} + (G_R)_{\mathrm{satellite}} - 0.4 \quad (\mathrm{dBHz})\end{aligned} \tag{5.84}$$

where $(G_R)_{\mathrm{satellite}}$ is the gain of the satellite receiving antenna in the direction of the transmitting earth station. This relation is represented in Figure 5.43 for the two cases considered:

— Global coverage ($\theta_{3\,dB} = 17.5°$) which implies $(G_R)_{\mathrm{satellite}} = 29\,000/(\theta_{3\,dB})^2 \approx 20$ dBi.
— Spot beam coverage ($\theta_{3\,dB} = 1.75°$) which implies $(G_R)_{\mathrm{satellite}} = 29\,000/(\theta_{3\,dB})^2 \approx 40$ dBi.

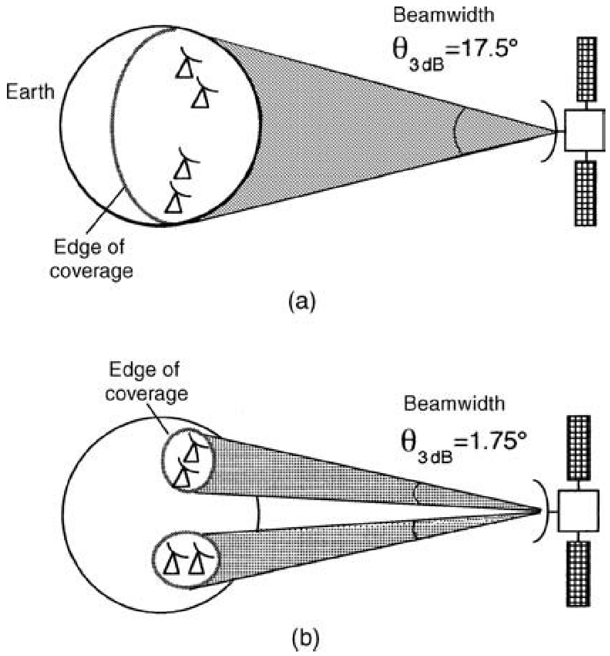

Figure 5.42 (a) Global coverage and (b) coverage by several narrow beams.

The expression for $(C/N_0)_D$ for the downlink is given by:

$$(C/N_0)_D = (\mathrm{EIRP})_{\mathrm{satellite}}(1/L_D)\,(G/T)_{\mathrm{station}}(1/k) \quad (\mathrm{Hz}) \tag{5.85}$$

Assume that the power of the carrier transmitted by the satellite is $P_T = 10\,\mathrm{W} = 10\,\mathrm{dBW}$. Let $L_U = 200\,\mathrm{dB}$ and neglect the implementation losses. Equation (5.85) becomes (all terms in dB):

$$\begin{aligned}(C/N_0)_D &= 10{-}200 + (G_T)_{\mathrm{satellite}} + (G/T)_{\mathrm{satation}} + 228.6 \\ &= (G_T)_{\mathrm{satellite}} + (G/T)_{\mathrm{station}} + 38.6 \quad (\mathrm{dBHz})\end{aligned} \tag{5.86}$$

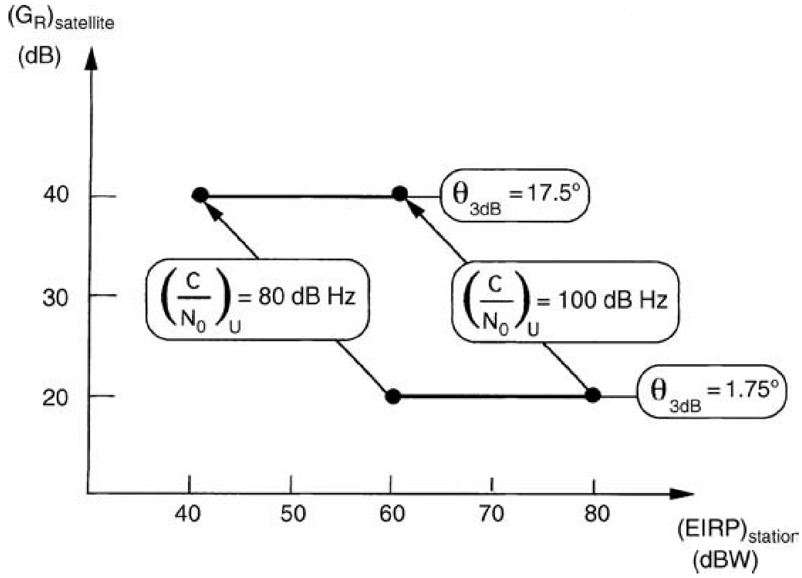

Figure 5.43 Comparison of the EIRP values required for an earth station in the case of global coverage ($\theta_{3\,\mathrm{dB}} = 17.5°$) and in the case of a spot beam ($\theta_{3\,\mathrm{dB}} = 1.75°$).

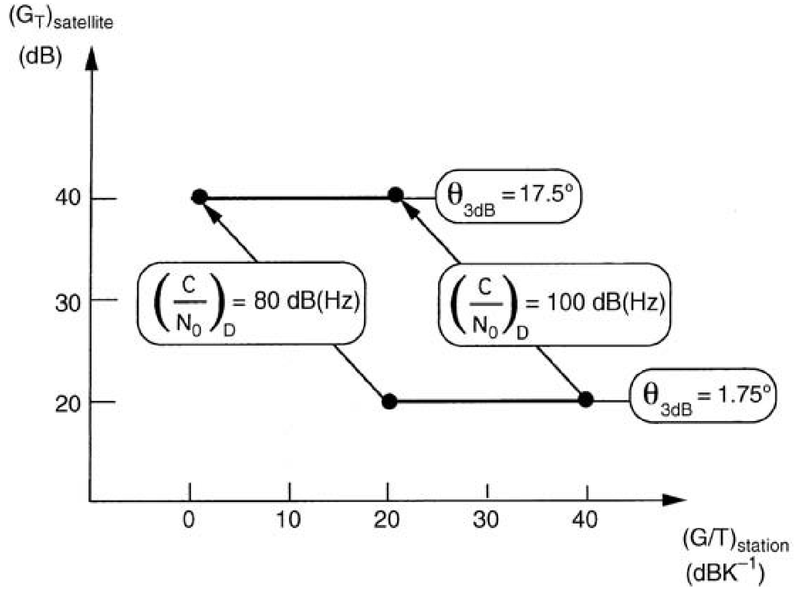

Figure 5.44 Comparison of the required values of factor of merit G/T for an earth station in the case of global coverage ($\theta_{3\,\text{dB}} = 17.5°$) and in the case of a spot beam ($\theta_{3\,\text{dB}} = 1.75°$).

This relation is represented in Figure 5.44 for the two cases considered:

— Global coverage ($\theta_{3\,\text{dB}} = 17.5°$) which implies $(G_T)_{\text{satellite}} = 29\,000/(\theta_{3\,\text{dB}})^2 \approx 20$ dBi.
— Spot beam coverage ($\theta_{3\,\text{dB}} = 1.75°$) which implies $(G_T)_{\text{satellite}} = 29\,000/(\theta_{3\,\text{dB}})^2 \approx 40$ dBi.

In Figures 5.43 and 5.44 the oblique arrows indicate the reduction in $(\text{EIRP})_{\text{station}}$ and $(G/T)_{\text{station}}$ when changing from a satellite with global coverage to a multibeam satellite with coverage by several spot beams. In this case, the multibeam satellite permits an economy of size, and hence cost, of the earth segment. For instance, a 20 dB reduction of $(\text{EIRP})_{\text{station}}$ and $(G/T)_{\text{station}}$ may result in a tenfold reduction of the antenna size (perhaps from 30 m to 3 m) with a cost reduction for the earth station (perhaps from a few million Euros to a few 10 000 Euros). If an identical earth segment is retained (a vertical displacement towards the top), an increase of C/N_0 is achieved which can be transferred to an increase of capacity, if sufficient bandwidth is available, at constant signal quality (in terms of bit error rate).

5.11.1.2 Frequency re-use

Frequency re-use consists of using the same frequency band several times in such a way as to increase the total capacity of the network without increasing the allocated bandwidth. An example has been seen in Section 5.2.3 of frequency re-use by orthogonal polarisation. In the case of a multibeam satellite, the isolation resulting from antenna directivity can be exploited to re-use the same frequency band in separate beam coverages. Figure 5.45 illustrates the principle of frequency re-use by orthogonal polarisation (Figure 5.45a) and the principle of re-use by angular beam separation (Figure 5.45b). A beam is associated with a given polarisation and a given coverage. In both cases, the bandwidth allocated to the system is B. The system uses this bandwidth B centred on the frequency f_U for the uplink and on the frequency f_D for the downlink. In the case of re-use by orthogonal polarisation, the bandwidth B is used twice only. In the case of re-use by angular separation, the bandwidth B can be re-used for as many beams as the permissible interference level allows. Both types of frequency re-use can be combined.

The frequency re-use factor is defined as the number of times that the bandwidth B is used. In theory, a multibeam satellite with M single-polarisation beams, each being allocated the

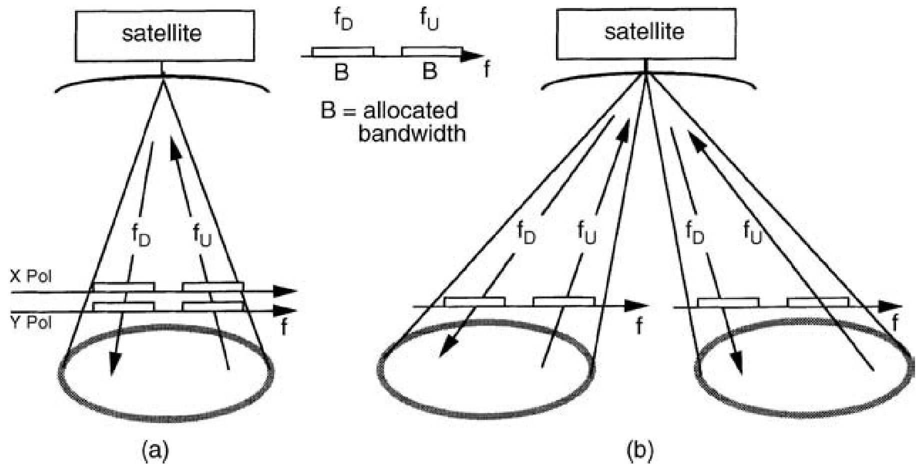

Figure 5.45 Frequency re-use with two beams ($M = 2$) by (a) orthogonal polarisation and (b) angular separation of the beams in a multibeam satellite system.

bandwidth B, and which combines re-use by angular separation and re-use by orthogonal polarisation may have a frequency re-use factor equal to $2M$. This signifies that it can claim the capacity which would be offered by a single beam satellite with single polarisation using a bandwidth of $M \times B$. In practice, the frequency re-use factor depends on the configuration of the service area which determines the coverage before it is provided by the satellite. If the service area consists of several widely separated regions (for example, urban areas separated by extensive rural areas), it is possible to re-use the same band in all beams. The frequency re-use factor can then attain the theoretical value of M. Figure 5.46 shows an example of multibeam coverage.

As the beam coverages are contiguous, the same frequency band cannot be used from one beam coverage to the next. In this example, the bandwidth allocated is divided into three equal separate sub-bands and each is used in beam coverages (1, 2 and 3) with sufficient angular separation from each other. The equivalent bandwidth, in the absence of re-use by orthogonal polarisation, has a value given by $6 \times (B/3) + 4 \times (B/3) + 3 \times (B/3) = 4.3\,B$ for $M = 13$ beams. The frequency re-use factor is then 4.3 instead of 13. With re-use by orthogonal polarisation within each beam coverage, the number of beams would be $M = 26$ and the frequency re-use factor would be 8.6.

5.11.2 Disadvantages of multibeam coverage

5.11.2.1 Interference between beams

Figure 5.47 illustrates interference generation within a multibeam satellite system, sometimes called *self-interference*. The allocated bandwidth B is divided into two sub-bands B_1 and B_2. The figure shows three beams. Beams 1 and 2 use the same band B_1. Beam 3 uses band B_2.

On the uplink (Figure 5.47a), the carrier at frequency f_{U1} of bandwidth B_1 transmitted by the beam 2 earth station is received by the antenna defining beam 1 in its side lobe with a low but non-zero gain. The spectrum of this carrier superimposes itself on that of the carrier of the same frequency emitted by the beam 1 earth station which is received in the main lobe with the maximum antenna gain. The carrier of beam 2 therefore appears as interference noise in the spectrum of the carrier of beam 1. This noise is called *co-channel interference* (CCI). Furthermore, part of the power of the carrier at frequency f_{U2} emitted by the earth station of beam 3 is introduced as a result of imperfect filtering of the IMUX filters defining the satellite channels (see Chapter 9) in the channel occupied by carrier f_{U1}. In this case, it consists of *adjacent channel interference* (ACI) analogous to that encountered in connection with frequency division multiple access in Section 6.5.3.

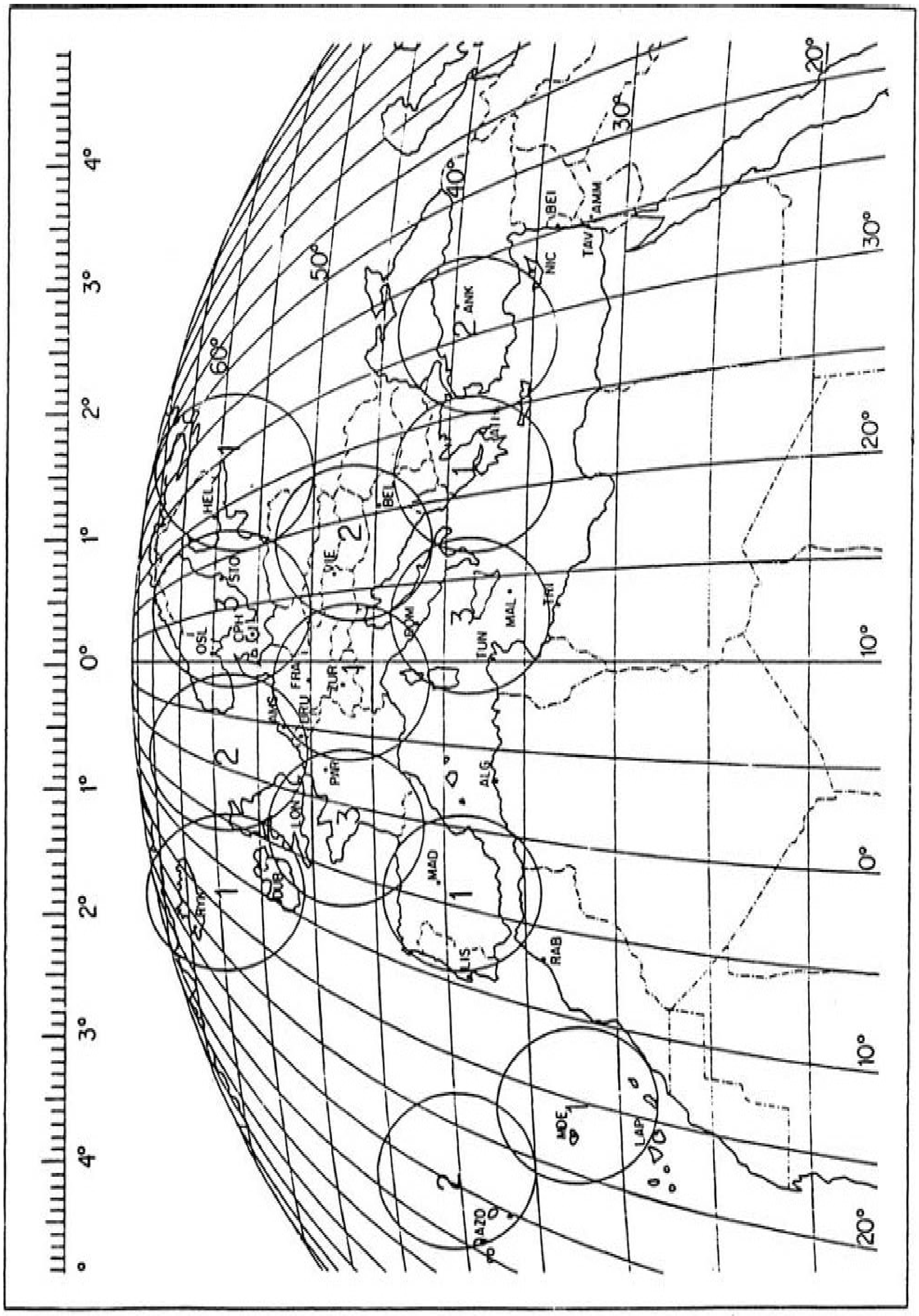

Figure 5.46 European coverage by a multibeam satellite system [LOP-82]. (Reproduced by permission of the European Space Agency.)

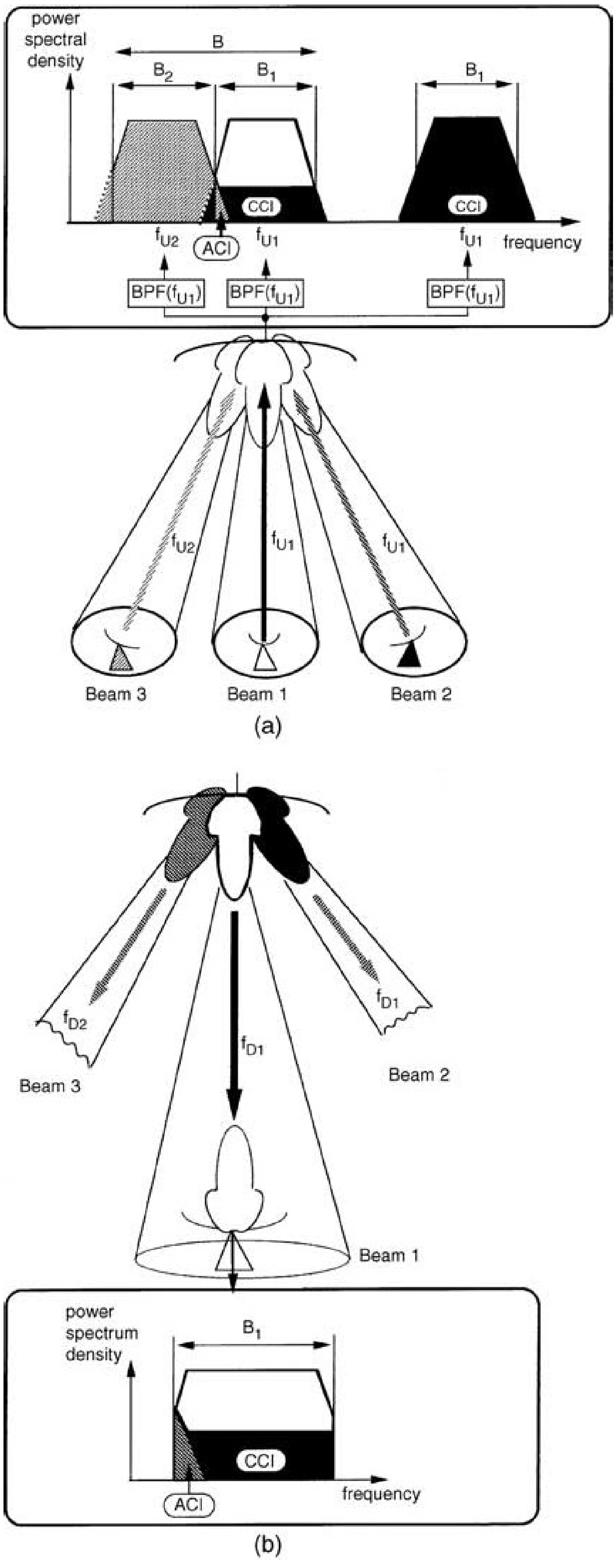

Figure 5.47 Self-interference between beams in a multibeam satellite system: (a) uplink and (b) downlink.

On the downlink (Figure 5.47b), the beam 1 earth station receives the carrier at frequency f_{D1} emitted with maximum gain in the antenna lobe defining beam 1. Downlink interference originates from the following contributions of power spectral density superimposed on the spectrum of this carrier:

— the spectra of the uplink adjacent channel and co-channel interference noise retransmitted by the satellite;
— the spectrum of the carrier at the same frequency f_{D1} emitted with maximum gain in beam 2 and with a small but non-zero gain in the direction of the beam 1 station. This represents additional co-channel interference (CCI).

The effect of self-interference appears as an increase in thermal noise under the same conditions as interference noise between systems analysed in Section 5.9.2. It must be included in the term $(C/N_0)_I$ which appears in equation (5.75). Taking account of the multiplicity of sources of interference, which become more numerous as the number of beams increases, relatively low values of $(C/N_0)_I$ may be achieved and the contribution of this term impairs the performance in terms of $(C/N_0)_T$ of the total link. As modern satellite systems tend to re-use frequency as much as possible to increase capacity, self-interference noise in a multibeam satellite link may contribute up to 50% of the total noise.

5.11.2.2 Interconnection between coverage areas

A satellite payload using multibeam coverage must be in a position to interconnect all network earth stations and consequently must provide interconnection of coverage areas. The complexity of the payload is added to that of the multibeam satellite antenna subsystem which is already much more complex than that of a single beam satellite.

Different techniques, depending on the on-board processing capability (no processing, transparent processing, regenerative processing, etc.) and on the network layer, are considered for interconnection of coverage:

— interconnection by transponder hopping (no on-board processing);
— interconnection by on-board switching (transparent and regenerative processing);
— interconnection by beam scanning.

These solutions are discussed in Chapter 7.

5.11.3 Conclusion

Multibeam satellite systems make it possible to reduce the size of earth stations and hence the cost of the earth segment. Frequency re-use from one beam to another permits an increase in capacity without increasing the bandwidth allocated to the system. However, interference between adjacent channels, which occurs between beams using the same frequencies, limits the potential capacity increase, particularly as interference is greater with earth stations equipped with small antennas.

5.12 INTERSATELLITE LINK PERFORMANCE

Intersatellite links are links between satellites. Three types of intersatellite links can be considered:

— GEO to LEO links between geostationary (GEO) satellites and low earth orbit (LEO) satellites, also called interorbital links (IOLs);

— GEO to GEO links between geostationary satellites;
— LEO to LEO links between low earth orbit satellites.

Of course one could consider intersatellite links between satellites in any type of orbit, but the above configurations are those most considered in practice. The reader is referred to Section 7.5 for a discussion of the practical applications. Only the transmission aspects are presented here.

5.12.1 Frequency bands

Table 5.2 indicates the frequency bands allocated to intersatellite links by the *Radiocommunication Regulations*. These frequencies correspond to strong absorption by the atmosphere and have been chosen to provide protection against interference between intersatellite links and terrestrial systems. However, these bands are shared with other space services and the limitation on interference level is likely to impose constraints on the choice of the defining parameters of intersatellite links (CCIR Reports 451, 465, 874, 951). Table 5.2 also indicates the wavelengths envisaged for optical links. These result from the transmission characteristics of the components.

5.12.2 Radio-frequency links

The budget equations presented in Sections 5.1 to 5.6 can apply. Propagation losses reduce to free space losses since there is no passage through the atmosphere. Antenna pointing error can be maintained at around a tenth of the beamwidth and this leads to a pointing error loss of the order of 0.5 dB. The antenna temperature in the case of a GEO–GEO link, in the absence of solar conjunction, is of the order of 10 K. Table 5.3 indicates typical values for the terminal equipment. For practical applications, antenna dimensions are of the order of 1 to 2 m. Considering a frequency of 60 GHz and transmission and reception losses of 1 dB leads to:

— a receiver figure of merit G/T of the order of 25 to 29 dBK^{-1};
— a transmitter EIRP of the order of 72 to 78 dBW.

Because of the relatively wide beamwidth of the antenna (0.2° at 60 GHz for a 2 m antenna), establishing the link is not a problem. Each satellite orientates its receiving antenna in the direction of the transmitting satellite with a precision of the order of 0.1° to acquire a beacon signal which is subsequently used for tracking.

The development of high-capacity, radio-frequency intersatellite links between geostationary satellite systems implies re-use of frequencies from one beam to another. In view of the small angular separation of the satellites, it is preferable to use narrow beam antennas with reduced side lobes in order to avoid interference between systems. Consequently, and in view of the limited

Table 5.2 Frequency bands for intersatellite links

Intersatellite service	Frequency bands
Radio frequency	22.55–23.55 GHz
	24.45–24.75 GHz
	32–33 GHz
	54.25–58.2 GHz
Optical	0.8–0.9 μm (AlGaAs laser diode)
	1.06 μm (Nd:YAG laser diode)
	0.532 μm (Nd:YAG laser diode)
	10.6 μm (CO_2 laser)

Table 5.3 Typical values for terminal equipment of a radio-frequency intersatellite link

Frequency	Receiver noise factor	Transmitter power
23–32 GHz	3–4.5 dB	150 W
60 GHz	4.5 dB	75 W
120 GHz	9 dB	30 W

antenna size imposed by the launcher and the technical complexity of the deployable antennas, the use of high frequencies is indicated. The use of optical links may be usefully considered in this context.

5.12.3 Optical links

In comparison with radio links, optical links have specific characteristics which are briefly described here. For a more complete presentation, refer to [KAT-87; GAG-91, Chapter 10; IJSC-88; WIT-94; BEG-00].

5.12.3.1 Establishing a link

Two aspects should be indicated:

— The small diameter of the telescope is typically of the order 0.3 m. In this way, one is freed from congestion problems and aperture blocking of other antennas in the payload.
— The narrowness of the optical beam is typically 5 microradians. Notice that this width is several orders of magnitude less than that of a radio beam and this is an advantage for protection against interference between systems. But it is also a disadvantage since the beamwidth is much less than the precision of satellite attitude control (typically 0.1° or 1.75 mrad). Consequently an advanced pointing device is necessary; this is probably the most difficult technical problem.

There are three basic phases to optical communications:

— Acquisition: the beam must be as wide as possible in order to reduce the acquisition time. But this requires a high-power laser transmitter. A laser of lower mean power can be used which emits pulses of high peak power with a low duty cycle. The beam scans the region of space where the receiver is expected to be located. When the receiver receives the signal, it enters a tracking phase and transmits in the direction of the received signal. On receiving the return signal from the receiver, the transmitter also enters the tracking phase. The typical duration of this phase is 10 seconds.
— Tracking: the beams are reduced to their nominal width. Laser transmission becomes continuous. In this phase, which extends throughout the following, the pointing error control device must allow for movements of the platform and relative movements of the two satellites. In addition, since the relative velocity of the two satellites is not zero, a *lead-ahead* angle exists between the receiver line of sight and the transmitter line of sight. As demonstrated below, the lead-ahead angle is larger than the beamwidth and must be accurately determined.
— Communications: information is exchanged between the two ends.

5.12.3.2 Lead-ahead angle

Consider two satellites, S1 and S2, respectively moving with velocity vectors $\mathbf{V}_{S1}$ and $\mathbf{V}_{S2}$, whose components orthogonal to the line joining S1 and S2 at time t are respectively the two vectors represented in Figure 5.48 by $\mathbf{V}_{T1}$ and $\mathbf{V}_{T2}$.

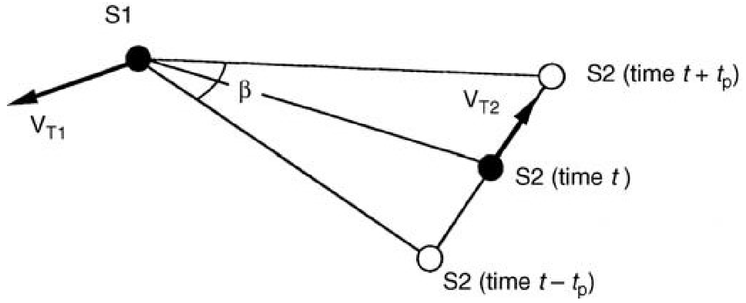

Figure 5.48 Lead-ahead angle for link between two satellites S1 and S2 with velocity vector components $\mathbf{V}_{T1}$ and $\mathbf{V}_{T2}$ in a plane perpendicular to the line joining S1 and S2 at time t; t_p is the propagation time of a photon from S1 to S2.

The propagation time of a photon from S1 to S2 is $t_p = d/c$, where d is the distance between the two satellites at time t and c is the speed of light ($c = 3 \times 10^8$ m/s).

The lead-ahead angle β is given by:

$$\beta = 2|\mathbf{V}_{T1} - \mathbf{V}_{T2}|/c \quad \text{(rad)} \tag{5.87}$$

where $|\mathbf{V}_{T1} - \mathbf{V}_{T2}|$ is the modulus of the difference vector $\mathbf{V}_{T1} - \mathbf{V}_{T2}$.

Two situations are now considered: intersatellite links between two geostationary satellites; interorbital links between a geostationary satellite and a low earth orbiting satellite.

5.12.3.2.1 GEO satellites separated by angle α

As both satellites are on the same circular orbit (Figure 5.49), the velocity vectors $\mathbf{V}_{S1}$ and $\mathbf{V}_{S2}$, which are tangential to the orbit, have equal modulus:

$$|\mathbf{V}_{S1}| = |\mathbf{V}_{S2}| = \omega(R_0 + R_E) = 3075 \text{ m/s}$$

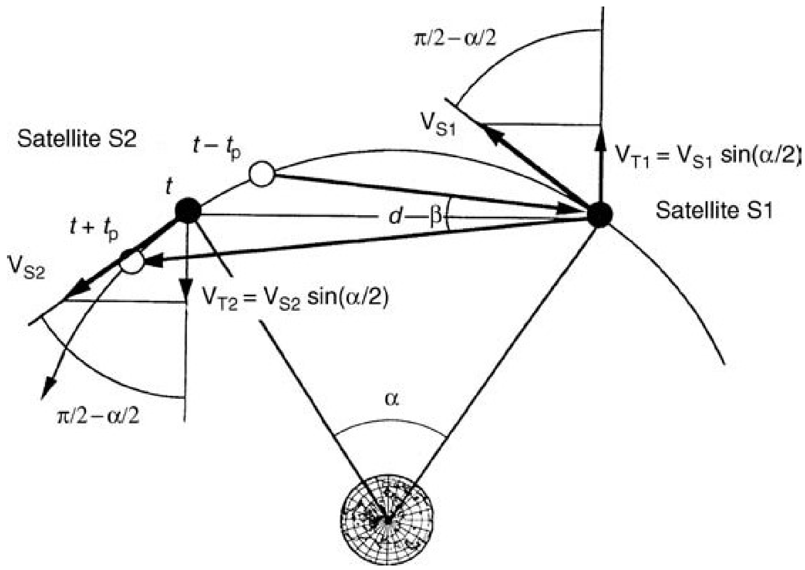

Figure 5.49 Lead-ahead angle for intersatellite links between two geostationary satellites.

where:

ω is the angular velocity of a geostationary satellite $= 7.293 \times 10^{-5}$ rad/s
R_0 is the altitude of a geostationary satellite $= 35\,786$ km
R_E is the earth radius $= 6378$ km

The component vectors $\mathbf{V}_{T1}$ and $\mathbf{V}_{T2}$, perpendicular to the line joining S1 and S2 at time t, both lie in the plane of the orbit and are opposite. They are at an angle $(\pi/2-\alpha/2)$ with respect to vectors $\mathbf{V}_{S1}$ and $\mathbf{V}_{S2}$. Therefore:

$$|\mathbf{V}_{T1}-\mathbf{V}_{T2}| = 2\omega(R_0 + R_E)\cos(\pi/2-\alpha/2) = 2\omega(R_0 + R_E)\sin(\alpha/2) \quad \text{(m/s)} \tag{5.88}$$

From equation (5.87):

$$\beta = 2|\mathbf{V}_{T1}-\mathbf{V}_{T2}|/c = 4\omega(R_0 + R_E)\sin(\alpha/2)/c \quad \text{(rad)} \tag{5.89}$$

Figure 5.50 displays the lead-ahead angle β as a function of the separation angle α between the two geostationary satellites. Note that, for a separation angle larger than 15°, the lead-ahead angle is larger than the beamwidth (typically 5 microradian). For instance, $\beta = 10.6$ microradian for $\alpha = 30°$, $\beta = 20.5$ microradian for $\alpha = 60°$, and $\beta = 35.5$ microradian for $\alpha = 120°$.

5.12.3.2.2 A GEO satellite and a LEO satellite with circular orbit

The relative velocity of the two satellites (Figure 5.51) varies with time, and so does the value of the lead-ahead angle. Its maximum value is obtained when the LEO satellite crosses the equatorial plane. Denoting as i the LEO satellite orbit inclination, then:

$$|\mathbf{V}_{T1}-\mathbf{V}_{T2}| = \{|\mathbf{V}_{S1}|^2 + |\mathbf{V}_{S2}|^2 - 2|\mathbf{V}_{S1}||\mathbf{V}_{S2}|\cos i\}^{1/2} \tag{5.90}$$

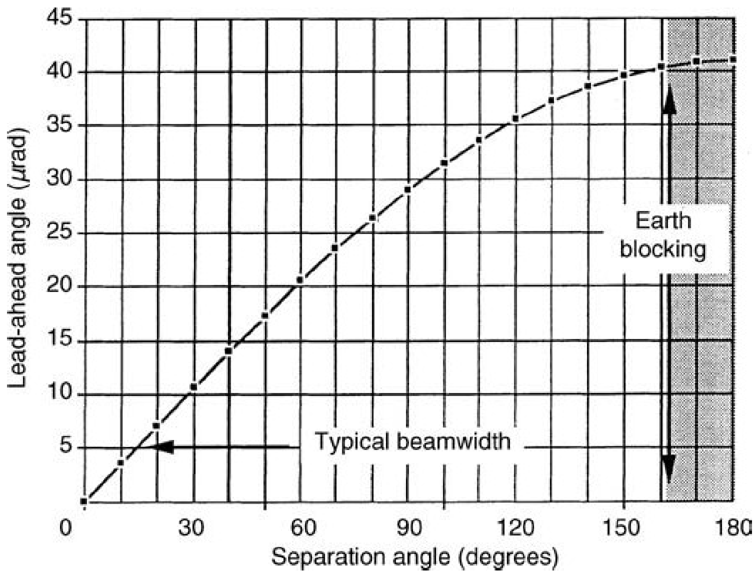

Figure 5.50 Lead-ahead angle as a function of the separation angle between two geostationary satellites.

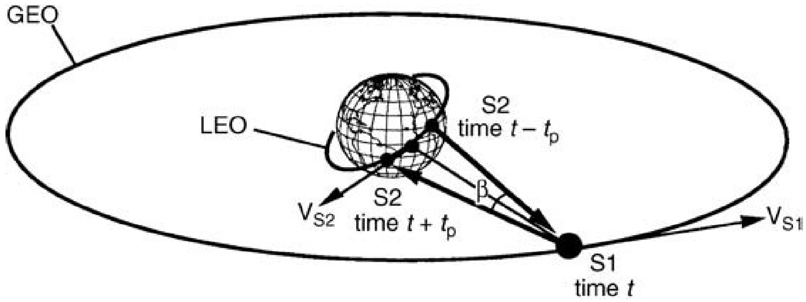

Figure 5.51 Lead-ahead angle at a GEO satellite for interorbital links between it and a LEO satellite.

where:

$$|\mathbf{V}_{S1}| = \omega_{GEO}(R_0 + R_E) = 3075\,\text{m/s}$$
$$|\mathbf{V}_{S2}| = \omega_{LEO}(h + R_E)$$

h is the LEO satellite altitude and $\omega_{LEO} + \mu^{1/2}(h + R_E)^{-3/2}$ is the LEO satellite angular rate ($\mu = 3.986 \times 10^{14}\,\text{m}^3/\text{s}^2$).

From equation (5.87), the lead-ahead angle is given by:

$$\begin{aligned}\beta &= 2|\mathbf{V}_{T1} - \mathbf{V}_{T2}|/c \\ &= (2/c)\{|\mathbf{V}_{S1}|^2 + |\mathbf{V}_{S2}|^2 - 2|\mathbf{V}_{S1}||\mathbf{V}_{S2}|\cos\, i\}^{1/2} \quad \text{(rad)}\end{aligned} \tag{5.91}$$

The lead-ahead angle is the same for the two satellites. Considering $i = 98.5°$ and $h = 800$ km, then $\beta = 57$ microradian. Note this value is even larger than for intersatellite links between two geostationary satellites.

5.12.3.3 Transmission

Laser sources operate in single and multi-frequency modes. In single frequency mode, spectral width varies between 10 kHz and 10 MHz. In multi-frequency mode, it is from 1.5 to 10 nm. The power emitted depends on the type of laser. Table 5.4 gives orders of magnitude.

Modulation can be internal or external. Internal modulation implies direct modification of the operation of the laser. External modulation is a modification of the light beam after its emission by the laser. The intensity, the frequency, the phase and the polarisation can be modulated. Phase and polarisation modulation are external. Intensity and frequency modulation can be internal or external. Polarisation modulation requires the presence of two detectors in the receiver, one for

Table 5.4 Typical values of transmitted power for lasers

Type of laser	Wavelength	Transmitted power
Solid state (laser diode)		
AlGaAs	0.8–0.9 μ	About 100 mW
InPAaGa	1.3–1.5 μ	About 100 mW
Nd:YAG	1.06 μ	0.5 to 1 W
Nd:YAG	0.532 μ	100 mW
Gas laser		
CO_2	10.6 μ	several tens of watts

each polarisation. Because of this, it is preferable to reserve polarisation for multiplexing of two channels.

The intensity distribution of a laser beam, as a function of angle with respect to the maximum intensity, follows a Gaussian law. The on-axis gain is given by:

$$G_{\mathrm{Tmax}} = 32/(\theta_{\mathrm{T}})^2 \tag{5.92}$$

where θ_{T} is the total beamwidth at $1/e^2$ where $e = 2.718$. The choice of θ_{T} depends on the pointing accuracy. With imprecise pointing, a large θ_{T} is better but gain is lost. If θ_{T} is reduced, there is benefit in gain but the pointing error loss increases. It can be shown that, if the pointing error is essentially an alignment error, the (maximum gain $\times$ pointing error loss) product is maximum when $\theta_T = 2.8\times$ (pointing error) [KAT-87, p. 51]. In general, for a pointing error of any kind, the beamwidth may be adapted to the pointing error.

In addition to losses due to pointing error, transmission losses and degradation of the wavefront in the emitting optics occur.

5.12.3.4 Transmission loss

Transmission loss reduces to the free space loss:

$$L = (\lambda/4\pi R)^2 \tag{5.93}$$

where λ is the wavelength and R is the distance between transmitter and receiver.

5.12.3.5 Reception

The receiving gain of the antenna is given by:

$$G_R = (\pi D_{\mathrm{R}}/\lambda)^2$$

where D_{R} is the effective diameter of the receiver antenna.

The receiver can be of a *direct detection* (Figure 5.52) or a *coherent detection* receiver (Figure 5.53). With direct detection, the incident photons are converted into electrons by a photodetector. The subsequent baseband electric current at the photodetected output is amplified then detected by a matched filter. With coherent detection, the optical signal field associated with the incident photons is mixed with the signal from a local laser. The resulting optical field is converted into a bandpass electric current by a photodetector and is subsequently amplified by an intermediate frequency amplifier. The demodulator detects the useful signal either by envelope detection or by coherent demodulation.

The receiving losses include optical transmission losses and, for coherent detection, losses associated with the degradation of the wavefront. (The quality of the wavefront is an important characteristic for optimum mixing of the received signal field and the local oscillator field at the

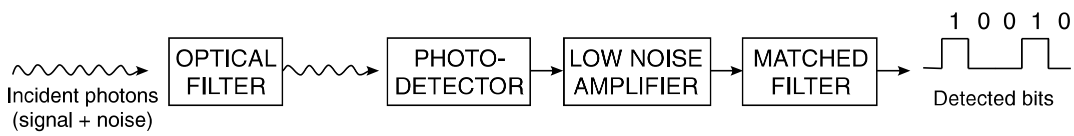

Figure 5.52 Optical direct detection receiver.

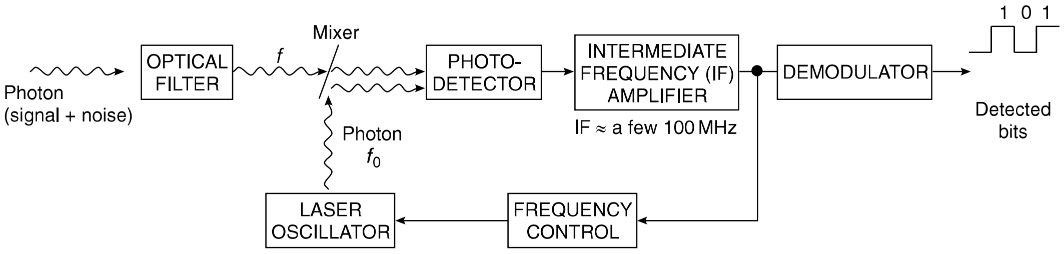

Figure 5.53 Optical coherent detection receiver.

photodetector front end.) Filtering, to reject out-of-band photons, also introduces losses, since the transmission coefficient reduces with bandwidth. A typical filter width is from 0.1 to 100 nm.

The signal-to-noise power ratio at the detector output depends on the type of detection. For direct detection (Figure 5.52):

$$S/N = I_{\mathrm{S\,dd}}^2 / i_{\mathrm{dd}}^2$$

The quantity $I_{\mathrm{S\,dd}}$ represents the signal current intensity:

$$I_{\mathrm{S\,dd}} = (P_{\mathrm{S}}/hf)\eta_p eG \quad (\mathrm{A}) \tag{5.94}$$

where:

P_{S} = useful optical signal power (W),
h = Planck's constant = 6.6×10^{-34} J/Hz,
f = laser frequency (Hz),
η_{p} = quantum efficiency of the photodetector, typically 0.8 for an avalanche photodetector (APD),
e = electron charge = 1.6×10^{-19} C,
G = photodetector gain, of the order of 50–300 for an avalanche photodetector (AP) and 10^4 to 10^6 for a vacuum tube photomultiplier.

Also (P_S/hf) represents the number of photons received per second, and $K = \eta_{\mathrm{p}} e/hf$ represents the sensitivity of the photodetector (A/W). Hence:.

$$I_{\mathrm{S\,dd}} = KGP_S \quad (\mathrm{A}) \tag{5.95}$$

The quantity i_{dd} represents the root mean square noise current intensity:

$$i_{\mathrm{dd}}^2 = i_{\mathrm{nS}}^2 + i_{\mathrm{nB}}^2 + i_{\mathrm{nD}}^2 + i_{\mathrm{nT}}^2 \quad (\mathrm{A}^2) \tag{5.96}$$

$i_{\mathrm{nS}}^2 = 2eKP_{\mathrm{S}}G^2 f(G)B_{\mathrm{N}}$ is the signal shot noise,
$i_{\mathrm{nB}}^2 = 2eKP_{\mathrm{n}}G^2 f(G)B_{\mathrm{N}}$ is the background shot noise,
$i_{\mathrm{nD}}^2 = 2ei_0 B_{\mathrm{N}}$ is the dark current shot noise,
$i_{\mathrm{nT}}^2 = N_0 B_{\mathrm{N}}$ is the thermal noise of the electronic amplifying circuits.

In these formulae, P_{n} is the received background optical noise power (W), $f(G)$ is a multiplying factor taking into account the noise generated in the photodetector by secondary electrons (typically $f(G) = a + bG$ where $a \approx 2$ and $b \approx 0.01$), i_0 is the dark current intensity (A), N_0 is the electronic amplifier thermal noise spectral density (A^2/Hz) and B_{N} is its noise bandwidth (Hz).

If all sources of noise besides the signal shot noise could be eliminated ($i_0 = 0$, $P_n = 0$, $N_0 = 0$, $f(G) = 1$), one would achieve the *quantum-limited* S/N value, given by:

$$(S/N)_{ql} = \eta_p P_S / 2hfB_N \quad (5.97)$$

For coherent detection (Figure 5.53):

$$S/N = I^2_{S\,cd}/(i^2_{dd} + i^2_{LO})$$

The quantity $I_{S\,cd}$ represents the signal current intensity:

$$I_{S\,cd} = KG\eta_m L_p (2P_S P_{LO})^{1/2} \quad (A) \quad (5.98)$$

where:

η_m = mixing efficiency,
L_P = loss due to polarisation mismatch.

The local oscillator power i_{dd} represents the root mean square noise current intensity for direct detection, as determined from equation (5.96); i_{LO} represents a supplementary source of noise, i.e. the local oscillator root mean square noise current intensity:

$$i^2_{LO} = 2eKP_{LO}G^2 f(G)B_{IF}$$

where B_{IF} is the noise bandwidth of the intermediate frequency amplifier (Hz).

The quantity P_{LO} can be increased to the point where i_{LO} is the predominant source of noise:

$$S/N \approx I^2_{S\,cd}/i^2_{LO} = \eta_m L_P \eta_p P_S / f(G) B_{IF} hf$$

Coherent detection confers a higher value of S/N than direct detection. In theory, one could achieve the quantum-limited S/N value, $(S/N)_{ql}$, given by equation (5.97), considering $\eta_m = 1$, $L_P = 1$, $f(G) = 1$ and $B_{IF} = 2B_N$. However, in the case of alignment error between the local oscillator and the beam signal, mixing efficiency is degraded. This type of detection cannot therefore be used both for acquisition and tracking. Unless high data rates are involved, there is no advantage in weight or power from using coherent detection techniques for communications along with a separate direct detection receiver for acquisition and tracking, compared to the situation where a direct detection receiver is used for both. For high data rates (typically greater than 1 Gbit/s), the power required for direct detection is excessive, and one may consider resorting to coherent detection for the communications function.

5.12.4 Conclusion

The choice between radio and optical links depends on the mass and power consumed. In general terms, it can be said that the advantage is with radio links for low throughputs (less than 1 Mbit/s). For high capacity links (several tens of Mbit/s) optical links command attention.

For a link involving one uplink, one or more optical intersatellite links and one downlink, the overall station-to-station link performance should be established in the same way as the overall link performance for regenerative satellites. Indeed the implementation of intersatellite links, be it at radio frequency or with optical technology, is mainly of interest when on-board demodulation is available, so as to provide flexible on-board switching.

REFERENCES

[BEG-00] D.L. Begey (2000) Laser cross links systems and technology, *IEEE Communications Magazine*, **38** (8), pp. 126–132.

[CAS-98] L. Castanet, J. Lemorton and M. Bousquet (1998) Fade mitigation techniques for new satcom services at Ku band and above: a review, *4th Ka band Utilisation Conference, Venice*, pp. 119–128.

[FEN-95] H.T. Fenech, B. Kasstan, A. Lindley, P.J.I. De Maagt and S. Badesi (1995) G/T predictions of communications satellites based on new earth brightness model, *International Journal of Satellite Communications*, **13**(5), pp. 367–376.

[GAG-91] R.M. Gagliardi (1991) *Satellite Communications*, 2nd edition, Van Nostrand Reinhold.

[IJSC-88] (1988) *International Journal of Satellite Communications*, **6**(2) Special Issue on Intersatellite links.

[ITU-85] ITU (1985) *CCIR Handbook on Satellite Communications*, p. 209.

[KAT-87] M. Kaitzman (Ed.) (1987) *Laser Satellite Communications*, Prentice-Hall.

[LOP-82] M. Lopriore, A. Saitto, G.K. Smith (1982) A unifying concept for future fixed satellite service payloads for Europe, *ESA Journal*, **6**(4), pp. 371–396.

[NJO-85] E.G. Njoku and E.K. Smith (1985) Microwave antenna temperature of the Earth from geostationary orbit, *Radio Science*, **20**(3), pp. 591–599.

[THO-83] R.W. Thorn, J. Thirlwell and D.J. Emerson (1983) Slant path radiometer measurements in the range 11–30 GHz at Martlesham Heath, England, *3rd International Conference of Antennas and Propagation, ICAP 83*, pp. 156–161.

[WAC-81] M. Washira, V. Arunachalam, K. Feher and G. Lo (1981) Performance of power and bandwidth efficient modulation techniques in regenerative and conventional satellite systems, *ICC 81, Denver*, pp. 37.2.1–37.2.5.

[WIT-94] M. Wittig (1994) Optical space communications, *Space Communications*, **12**(2).