CHAPTER 14

Fixed point microwave communications

14.1 Introduction

The two most important public service developments in fixed point radio communications over the last 40 years are the installation of national microwave relay networks and international satellite communications. In both cases these systems originally used analogue modulation and frequency division multiplexing and represented simply an increase in capacity, improvement in quality, or increase in convenience, over the traditional wired PSTN. Their subsequent development, however, has been in favour of digital modulation and time division multiplexing. The proliferation of services which can be offered using digital transmission (via ISDN, Chapter 19) has been at least as important in this respect as improvements in digital technology.

14.2 Terrestrial microwave links

Many wideband point-to-point radio communications links employ microwave carriers in the 1 to 20 GHz frequency range. These are used principally by PTTs to carry national telephone and television signals. Their extensive use can be gauged by the number of antenna masts now seen carrying microwave dishes at the summits of hills located between densely populated urban areas. Figure 14.1 shows the main trunk routes of the UK microwave link network. The following points can be made about microwave links:

1. Microwave energy does not follow the curvature of the earth, or diffract easily over mountainous terrain, in the way that MW and LW transmissions do. Microwave transmissions are, therefore, restricted essentially to line of sight (LOS) links.
2. Microwave transmissions are particularly well suited to point-to-point communications since narrow beam, high gain, antennas of reasonable size can be easily designed. (At 2 GHz the wavelength is 0.15 m and a 10-wavelength reflector, i.e. a 1.5 m dish, is still practical. Antenna gains are thus typically 30 to 50 dB.)

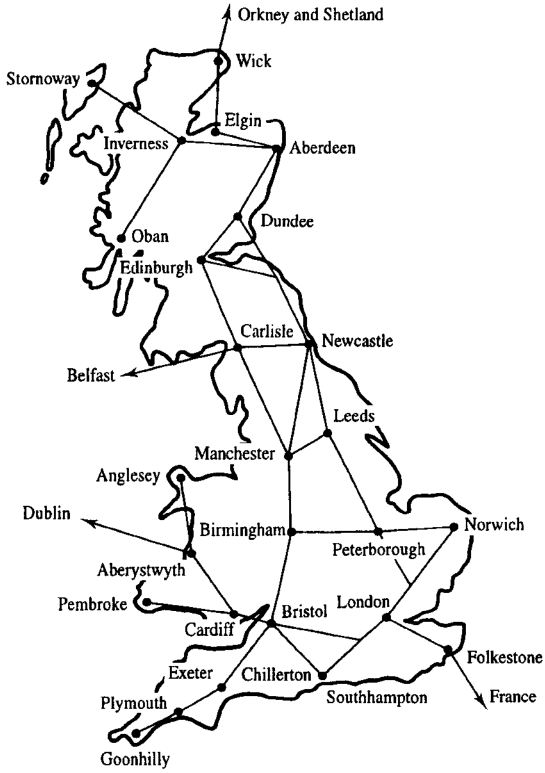


Figure 14.1 *The UK microwave communications wideband distribution network.*

3. At about 1 GHz circuit design techniques change from using lumped to distributed elements. Above 20 GHz it becomes difficult and/or expensive to generate reasonable amounts of microwave power.

Antennas are located on high ground to avoid obstacles such as large buildings or hills, and repeaters are used every 40 to 50 km to compensate for path losses. (On a 6 GHz link with a hop distance of 40 km the free space path loss, given by equation (12.71(c)), is 140 dB, see Example 12.6. With transmit and receive antenna gains of 40 dB, however, the basic transmission loss, P_T/C, reduces to 60 dB. This provides sufficient received power for such links to operate effectively.)

Frequencies are allocated in the UK by the Radiocommunication Agency. The principal frequency bands in current use are near 2 GHz, 4 GHz, 6 GHz, 11 GHz and 18 GHz. There are also allocations near 22 GHz and 28 GHz. (The frequency bands listed here are those which are allocated mainly to common carriers. Other allocations for more general use do exist.)

14.2.1 Analogue systems

In the UK microwave links were widely installed during the 1960s for analogue FDM telephony. In these systems each allocated frequency band is subdivided into a number of (approximately) 30 MHz wide radio channels. Figure 14.2 shows how the 500 MHz band, allocated at 4 GHz, is divided into 16 separate channels with 29.65 MHz centre spacings. Each radio channel supports an FDM signal (made up of many individual SSB voice signals) which is frequency modulated onto a carrier. This could be called an SSB/FDM/FM system. Adjacent radio channels use orthogonal antenna polarisations, horizontal (H) and vertical (V), to reduce crosstalk. The allocated band is split into a low and high block containing eight radio channels each, Figure 14.2. One block is used for transmission and the other for reception.

In the superheterodyne receiver a microwave channel filter, centred on the appropriate radio channel, extracts that channel from the block as shown in Figure 14.3. The signal is mixed down to an intermediate frequency (IF) for additional filtering and amplification. The microwave link can accommodate traffic simultaneously in all 16 channels. It is customary, however, not to transmit a given data signal on the same radio channel over

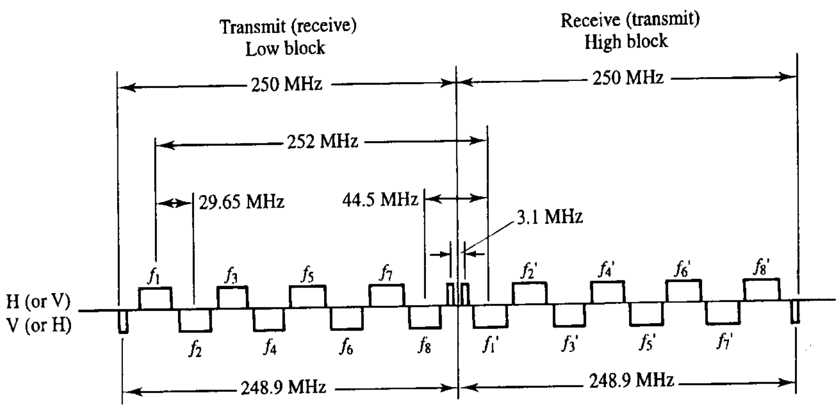


Figure 14.2 *Splitting of a microwave frequency allocation into radio channels.*

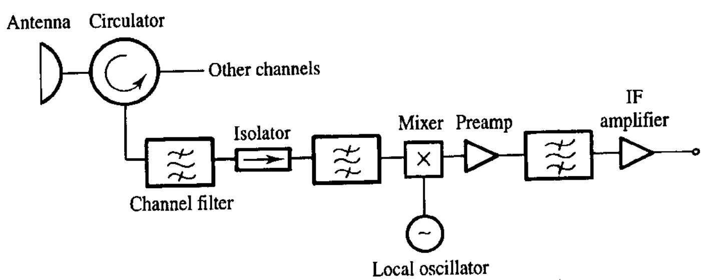


Figure 14.3 *Extraction (dropping) of a single radio channel in a microwave repeater.*

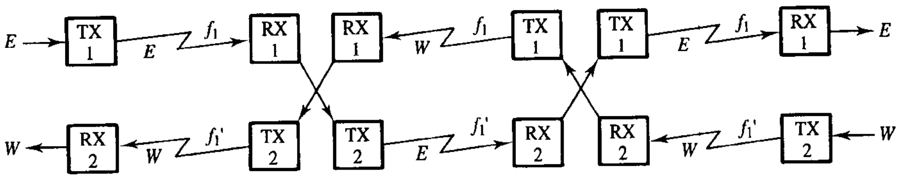


Figure 14.4 *Frequency allocations on adjacent repeaters.*

successive hops. Thus in Figure 14.4, for example, f_1 is used for the first hop but f_1', Figure 14.2, is used on the second hop. Under anomalous propagation conditions, temperature inversions can cause ducting to occur with consequent low loss propagation over long distances, Figure 14.5. This may cause overreaching, with signals being received not only at the next repeater but also at the one after that. Moving the traffic to a different radio channel on the next hop thus ensures that interference arising via this mechanism is essentially uncorrelated with the received signal.

14.2.2 Digital systems

The first digital microwave (PSTN) links were installed in the UK in 1982 [Harrison]. They operated with a bit rate of 140 Mbit/s at a carrier frequency of 11 GHz using QPSK modulation. In more recent systems there has been a move towards 16- and 64-QAM (see Chapter 11). Figure 14.6 shows a block diagram of a typical microwave digital radio terminal. (Signal pre-distortion shown in the transmitter can be used to compensate for distortion introduced by the high power amplifier.)

The practical spectral efficiency of 4 to 5 bit/s/Hz, which 64-QAM systems offer, means that the 30 MHz channel can support a 140 Mbit/s multiplexed telephone traffic signal. Figure 14.7 is a schematic of a digital regenerative repeater, assuming DPSK modulation, for a single 30 MHz channel. Here the circulator and channel filter access the part of the microwave spectrum where the signal is located. With the ever increasing demand for high capacity transmission, digital systems have a major advantage over the older analogue systems in that they can operate satisfactorily at a much lower carrier-to-noise ratio, Figure 11.21. There is also strong interest in even higher level modulation schemes, for example 1024-QAM, to increase the capacity of the traditional 30 MHz channel still further.

Microwave radio links at 2 and 18 GHz are also being applied, at low modulation rates, in place of copper wire connections, in rural communities for implementing the local loop exchange connection [Harrison]. (These are configured as point-to-multipoint

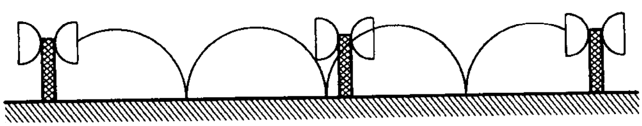


Figure 14.5 *Schematic illustration of ducting causing overreaching.*

links, not unlike the multidrop systems of section 18.2.2. Furthermore the channel access scheme can use derivatives of the digital cellular systems of section 15.4.)

14.2.3 LOS link design

The first-order design problem for a microwave link, whether analogue or digital, is to ensure adequate clearance over the underlying terrain. Path clearance is affected by the following factors:

1. Antenna heights.
2. Terrain cover.
3. Terrain profile.
4. Earth curvature.
5. Tropospheric refraction.

The last of these, tropospheric refraction, occurs because the refractive index, n, of the troposphere depends on its temperature, T, pressure, P, and water vapour partial pressure, e. Since significant changes in P, T and e make only small differences to n, the tropospheric refractive index is usually characterised by the related quantity, refractivity, defined by $N = (n - 1)10^6$.

A good model, [ITU-R, Rec. 453], relating N to P, T and e is:

$$N = 77.6 \frac{P}{T} + 3.73 \times 10^5 \frac{e}{T^2} \tag{14.1}$$

where P and e are in millibars and T is in K, as in Chapter 12. Although refractivity is dimensionless the term 'N units' is usually appended to its numerical value.

Equation (14.1) and the variation of P, T and e with altitude result in an approximately exponential decrease of N with height, h, i.e.:

$$N(h) = N_s \, e^{-h/h_0} \tag{14.2}$$

where N_s is surface refractivity and h_0 is a scale height. ITU-R Rec. 369 defines a standard reference troposphere, Figure 14.8, as one for which $N_s = 315$ and $h_0 = 7.35$ km. Over the first kilometre of height this is usually approximated as a linear height dependence with refractivity gradient given by:

$$\begin{aligned} \left.\frac{dN}{dh}\right|_{first\ km} &= N(1) - N(0) \\ &= -40 \ \ (N \text{ units/km}) \end{aligned} \tag{14.3}$$

The vertical gradient in refractive index causes microwave energy to propagate not in straight lines but along approximately circular arcs [Kerr] with radius of curvature, r, given by:

$$\frac{1}{r} = -\frac{1}{n}\frac{dn}{dh}\cos\alpha \tag{14.4}$$

Here α is the grazing angle of the ray to the local horizontal plane, Figure 14.9. For

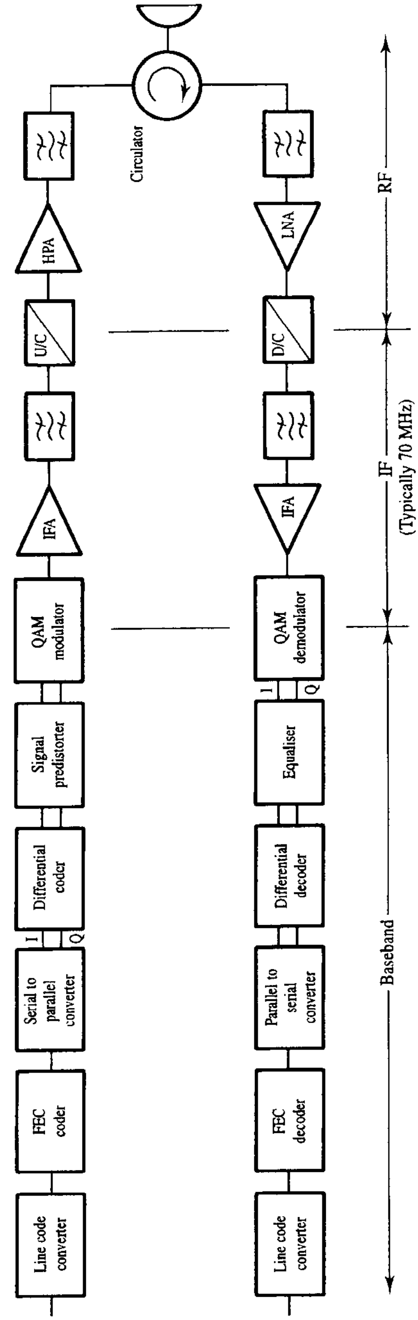


Figure 14.6 *Block diagram of a typical microwave digital radio terminal.*

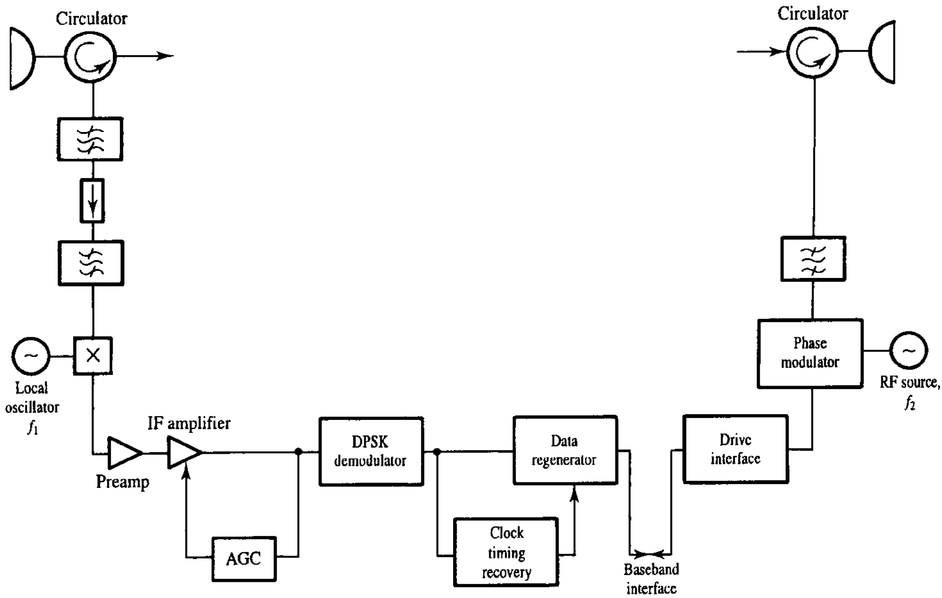


Figure 14.7 *Digital DPSK regenerative repeater for a single 30 MHz radio channel.*

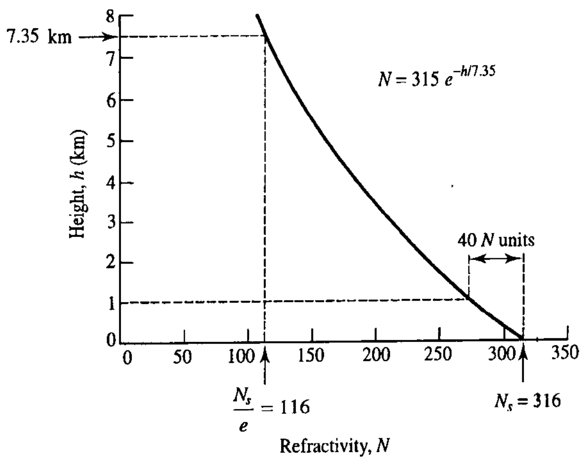


Figure 14.8 *ITU-R standard refractivity profile.*

terrestrial LOS links α is small, i.e. cos $\alpha \approx 1$, and for a standard troposphere with $n \approx 1$ and $dn/dh = -40 \times 10^{-6}$ the radius of curvature is 25,000 km. Microwave energy under these conditions thus bends towards the earth's surface but with a radius of curvature much larger than that of the earth itself, Figure 14.10. (The earth's mean radius, a, may be taken to be 6371 km.) There are two popular ways in which ray curvature is

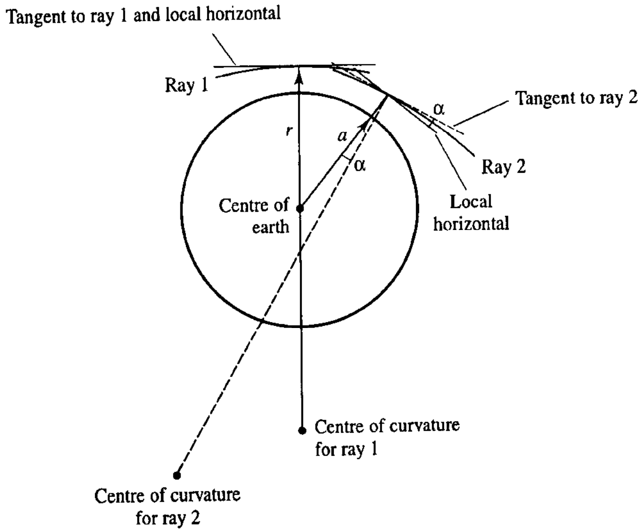


Figure 14.9 *Illustration of circular paths for rays in atmosphere with vertical n-gradient ($\alpha = 0$ for ray 1, $\alpha \neq 0$ for ray 2). Geometry distorted for clarity.*

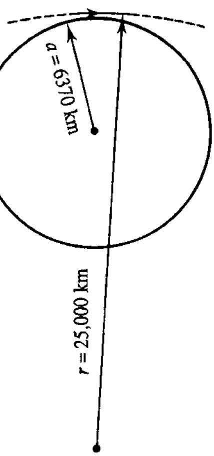


Figure 14.10 *Relative curvatures of earth's surface and ray path in a standard atmosphere.*

accounted for in path profiling. One subtracts the curvature of the ray (equation (14.4)) from that of the earth giving (for $n = 1$ and $\alpha = 0$):

$$\frac{1}{a_e} = \frac{1}{a} + \frac{dn}{dh} \tag{14.5}$$

This effectively decreases the curvature of both ray and earth until the ray is straight,

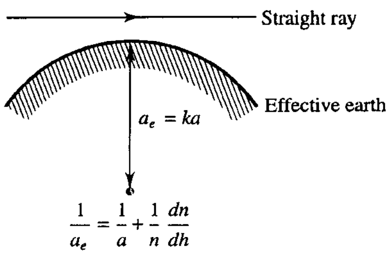


Figure 14.11 *Straight ray model. (Note n ≈ 1.)*

Figure 14.11, resulting in an effective earth radius, a_e. This may be called the straight ray model. Alternatively the negative of this transformation can be applied, i.e.:

$$\frac{1}{r_e} = -\left(\frac{1}{a} + \frac{dn}{dh}\right) \tag{14.6}$$

which decreases the curvature of both ray and earth until the earth is flat, Figure 14.12, resulting in an effective radius of curvature, r_e, for the ray. This may be called the flat earth model. For the straight ray model the ratio of effective earth radius to actual earth radius is called the k factor, i.e.:

$$k = \frac{a_e}{a} \tag{14.7}$$

which, using equation (14.5) and the definition of refractivity, is given by:

$$k = \frac{1}{1 + a\dfrac{dN}{dh} \times 10^{-6}} \tag{14.8}$$

k factor thus represents an alternative way of expressing refractivity lapse rate, dN/dh. In the lowest kilometre of the standard ITU-R troposphere $k = 4/3$. Table 14.1 shows the relationship between k and dN/dh and Figure 14.13 shows the characteristic curvatures of rays under standard, sub-refracting, super-refracting and ducting conditions, each drawn (schematically) on a $k = 4/3$ earth profile. For standard meteorological conditions path profiles can be drawn on special $k = 4/3$ earth profile paper.

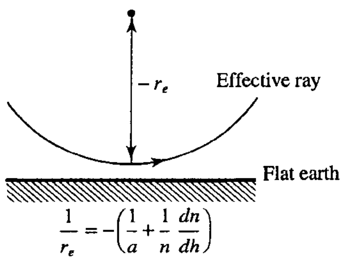


Figure 14.12 *Flat earth model (n≈1 and negative radius indicates ray is concave upwards).*

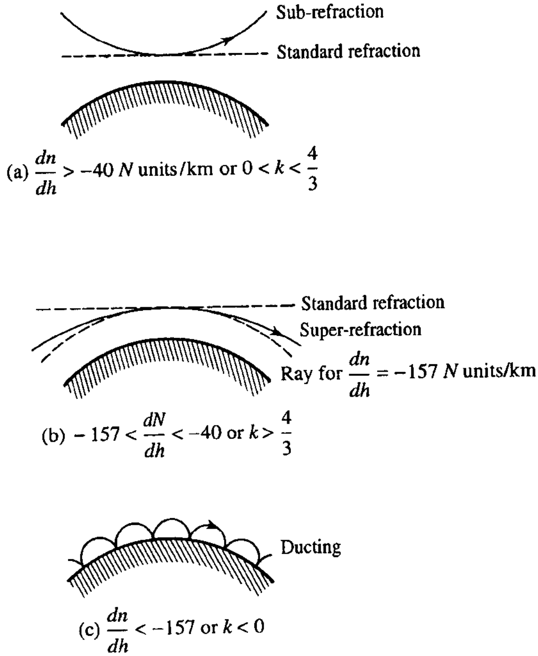


Figure 14.13 *Characteristic ray trajectories drawn with respect to a k = 4/3 earth radius.*

Table 14.1 *Equivalent values of refractivity lapse rate and k factor.*

$\frac{dN}{dh}$ (N units/km)	157	78	0	−40	−100	−157	−200	−300
k factor	$\frac{1}{2}$	$\frac{2}{3}$	1	$\frac{4}{3}$	2.75	∞	−3.65	−1.09

Sub-refraction Super-refraction Ducting

<------------------------>|<----------------->|<------------------------>

EXAMPLE 14.1

Show that the line-of-sight range over a smooth spherical earth is given by $L = \sqrt{2ka}\,(\sqrt{h_T} + \sqrt{h_R})$ where k is a k-factor, a is earth radius, h_T is transmit antenna height and h_R is receive antenna height.

Consider the geometry shown in Figure 14.14.

$$L_1^2 = (h_T + a_e)^2 - a_e^2$$

$$= h_T^2 + 2\,h_T\,a_e$$

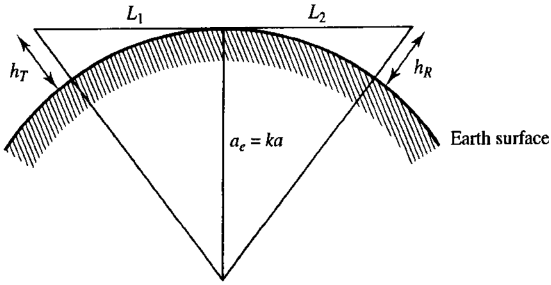


Figure 14.14 *Geometry for maximum range LOS link over a smooth, spherical, earth.*

Similarly:

$$L_2^2 = h_R^2 + 2\ h_R\ a_e$$

Since $h_T << a_e$ and $h_R << a_e$ then:

$$L_1^2 \approx 2\ h_T\ a_e$$

$$L_2^2 \approx 2\ h_R\ a_e$$

Therefore:

$$L = L_1 + L_2 = \sqrt{2\ a_e}\left(\sqrt{h_T} + \sqrt{h_R}\right)$$

Fresnel zones and path profiling

Fresnel zones are defined by the intersection of Fresnel ellipsoids with a plane perpendicular to the LOS path. The ellipsoids in turn are defined by the loci of points, Figure 14.15, which give an excess path length, ΔL, over the direct path of:

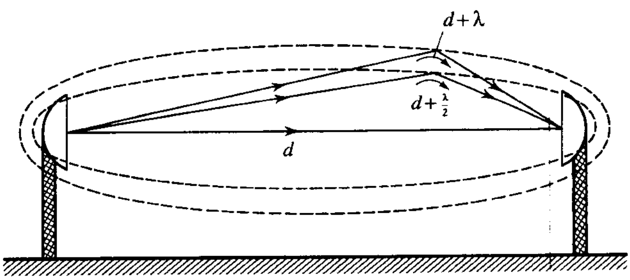


Figure 14.15 *Fresnel ellipsoids.*

$$\Delta L = n \frac{\lambda}{2} \quad \text{(m)} \tag{14.9}$$

The first, second, and third Fresnel radii are illustrated in Figure 14.16. For points not too near either end of the link (i.e. for $r_n \ll d_1$ and $r_n \ll d_2$) the nth Fresnel zone radius, r_n, is given by:

$$r_n \approx \sqrt{n \frac{\lambda\, d_1\, d_2}{d_1 + d_2}} \tag{14.10}$$

where d_1 and d_2 are distance from transmitter and receiver respectively. With personal computers it is now practical to plot ray paths for any k-factor on a 4/3 earth radius profile and, making due allowance for terrain cover, find the minimum clearance along the path as a fraction of r_1. (The required clearance of a link is often specified as a particular fraction of the first Fresnel zone radius under meteorological conditions corresponding to a particular k factor.) This design exercise is called path profiling. The principles of path profiling, however, are simple and best illustrated by describing the process as it might be implemented manually:

1. A terrain profile (ignoring earth curvature) is plotted on Cartesian graph paper choosing any suitable vertical and horizontal scales.
2. For all points on the path profile likely to give rise to poor clearance (e.g. local maxima, path midpoint, etc.) the height of the earth's bulge, h_B, is calculated using:

$$h_B = \frac{d_1\, d_2}{2a} \quad \text{(m)} \tag{14.11}$$

   (h_B is the height of the, smooth, earth's surface above the straight line connecting the points at sea level below transmit and receive antennas.)
3. The effect of tropospheric refraction on clearance is calculated using h_B and k factor to give:

$$h_{TR} = h_B(k^{-1} - 1) \quad \text{(m)} \tag{14.12(a)}$$

4. The required Fresnel zone clearance (in metres) is calculated using:

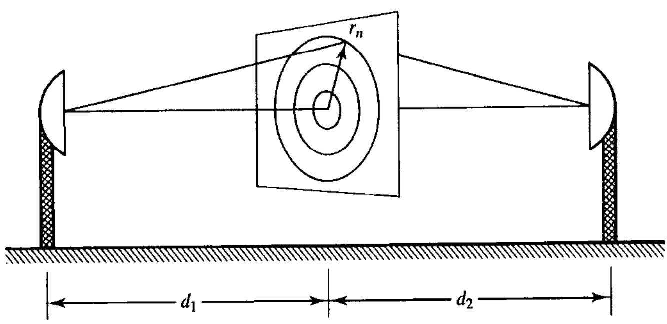


Figure 14.16 *Fresnel zones.*

$$h_{FZC} = f\, r_1 \quad \text{(m)} \tag{14.12(b)}$$

where r_1 is given by equation (14.10) and f is typically 1.0 for $k = 4/3$. (In the UK $f = 0.6$ for $k = 0.7$ is often used.)

5. An estimate is obtained for the terrain cover height, h_{TC}.
6. $h_{TC} + h_{TR} + h_B + h_{FZC}$ is plotted on the terrain path profile.
7. A straight line passing through the highest of the plotted points then allows appropriate antenna heights at each end of the link to be established, Figure 14.17.

There will usually be a trade-off between the height of transmit and receive antennas. Normally these heights would be chosen to be as equal as possible in order to minimise the overall cost of towers. If one or more points of strong reflection occur on the path, however, the antenna heights may be varied to shift these points away from areas of high reflectivity (such as regions of open water).

If adequate Fresnel zone clearance cannot be guaranteed under all conditions then diffraction may occur leading to signal fading. If the diffracting obstacle can be modelled as a knife edge, Figure 14.18, the diffraction loss can be found from Figure 14.19. The parameter v in this figure is called the Fresnel diffraction parameter and is given by:

$$v = \sqrt{\frac{2}{\lambda}\frac{d_1 + d_2}{d_1\, d_2}}\; h \tag{14.13}$$

where d_1 and d_2 locate the knife edge between transmitter and receiver, and h is the

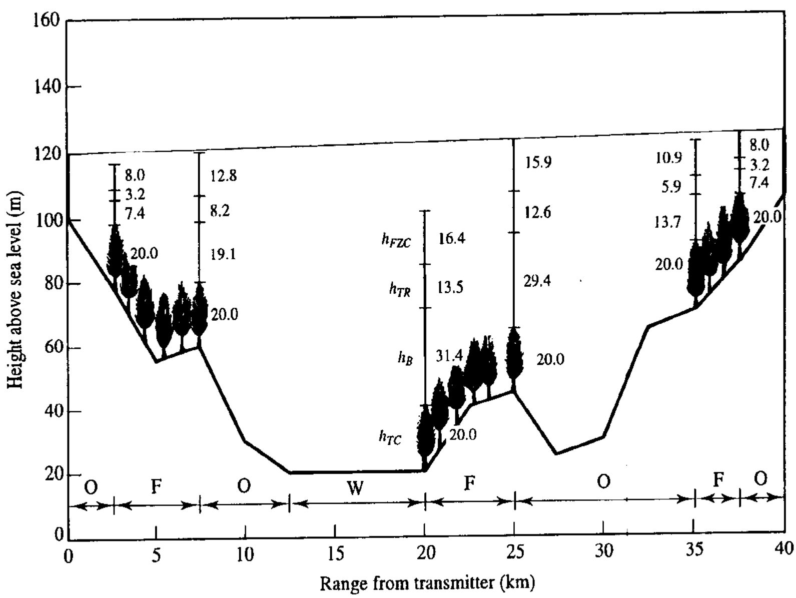


Figure 14.17 *Path profile for hypothetical 4 GHz LOS link designed for 0.6 Fresnel zone (FZ) clearance when k = 0.7 (O = open ground, F = forested region, W = water).*

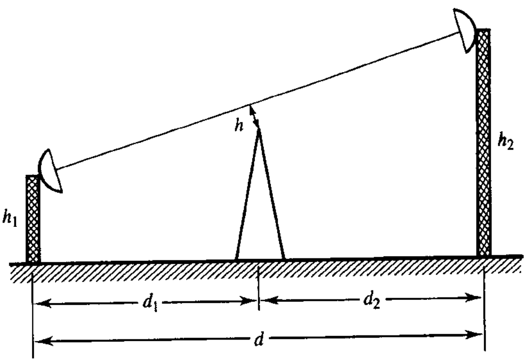


Figure 14.18 *Definition of clearance, h, for knife edge diffraction.*

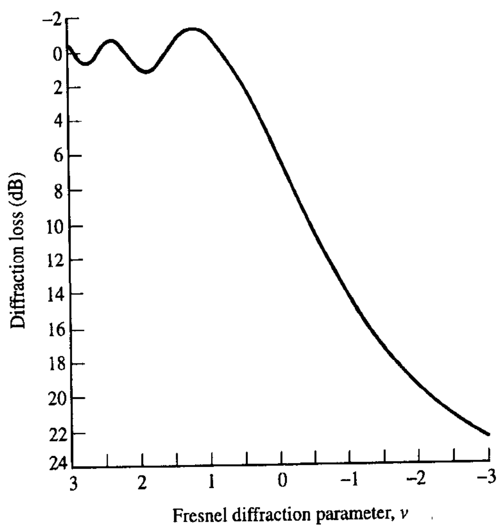


Figure 14.19 *Diffraction loss over a knife-edge (negative loss indicates a diffraction gain).*

minimum clearance measured perpendicularly from the LOS path over the diffracting edge. (If the obstacle blocks the LOS path then h is negative.) If the obstacle cannot be modelled as a knife edge then the diffraction loss will generally be greater. A model for cylindrical edge diffraction, given in [Doughty and Maloney], takes the form of a knife edge loss plus a curvature loss and a correction factor.

EXAMPLE 14.2

A 6 GHz microwave link operating over flat ground is 36 km long. 15 km from the transmitter a long building with a pitched roof is oriented perpendicularly to the LOS path. The height of the building to the vertex of the roof is 30 m. How high must the transmitter and receiver antenna towers be if, under standard refraction conditions ($k = 4/3$): (i) first Fresnel zone clearance is to be maintained and (ii) diffraction fading is not to exceed 8 dB?

(i)

$$\lambda = \frac{3 \times 10^8}{6 \times 10^9} = 0.05 \text{ m}$$

$$h_B = \frac{d_1 d_2}{2a} = \frac{15000 \times 21000}{2 \times 6371000} = 24.72 \text{ m}$$

$$h_{TR} = h_B (k^{-1} - 1) = 24.72 \left(\frac{3}{4} - 1 \right) = -6.18 \text{ m}$$

$$h_{FZC} = f r_1 = 1.0 \sqrt{\frac{\lambda d_1 d_2}{d_1 + d_2}} = \sqrt{\frac{0.05 \times 15000 \times 21000}{36000}} = 20.92 \text{ m}$$

Required antenna heights are given by:

$$h_T = h_R = h_B + h_{TR} + h_{FZC} = 24.72 - 6.18 + 20.92$$
$$= 39.46 \text{ m}$$

(ii) If diffraction loss is not to exceed 8 dB then, from Figure 14.19, $v = -0.25$. Rearranging equation (14.13):

$$h = \frac{v}{\sqrt{\frac{2}{\lambda} \frac{d_1 + d_2}{d_1 d_2}}} = \frac{-0.25}{\sqrt{\frac{2}{0.05} \frac{36000}{15000 \times 21000}}} = -3.70 \text{ m}$$

Required antenna heights are now given by:

$$h_T = h_R = h_B + h_{TR} + h = 24.72 - 6.18 - 3.70 = 14.84 \text{ m}$$

14.2.4 Other propagation considerations for terrestrial links

Once a LOS link has been profiled a detailed link budget would normally be prepared. A first-order free space signal budget is described in Chapter 12 (section 12.4.2). The actual transmitter power required, however, will be greater than that implied by this calculation since, even in the absence of diffraction, some allowance must be made for atmospheric absorption, signal fading and noise enhancements. For terrestrial microwave links with good clearance the principal mechanisms causing signal reductions are:

1. Background gaseous absorption.
2. Rain fading.
3. Multipath fading.

The principal sources of noise enhancement are:

1. Thermal radiation from rain.

2. Interference caused by precipitation scatter and ducting.
3. Crosstalk caused by cross-polarisation.

In addition to reducing the CNR (and/or carrier to interference ratio) signal fading processes also have the potential to cause distortion of the transmitted signal. For narrowband transmissions this will not usually result in a performance degradation beyond that due to the altered CNR. For sufficiently wideband digital transmissions, however, such distortion may result in extra BER degradation.

Background gaseous absorption

Figure 14.20 shows the nominal specific attenuation, $\gamma(f)$ (dB/km), of the atmosphere due to its gaseous constituents for a horizontal path at sea level. (Separate, and more detailed, curves for the dry air and water vapour constituents are given in [ITU-R, Rec. 676].) The rather peaky nature of this curve is the result of molecular resonance effects (principally due to water vapour and oxygen). For normal applications it is adequate simply to multiply $\gamma(f)$ by the path length, L, and include the resulting attenuation in the detailed link budget.

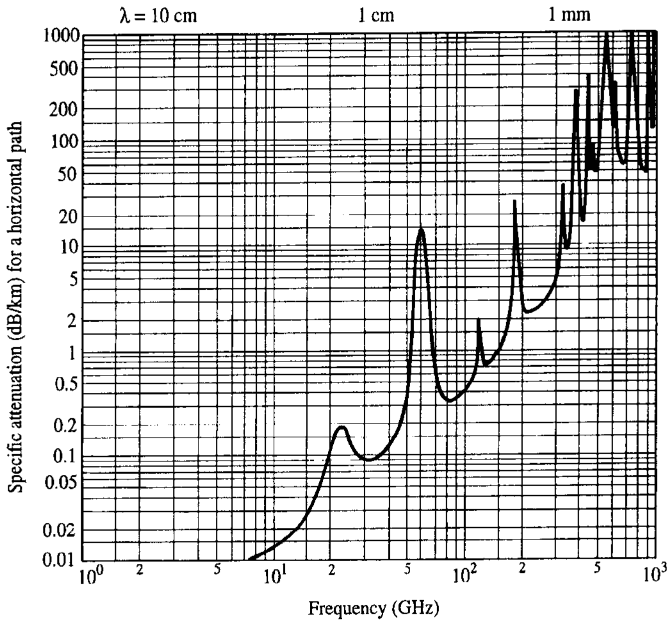


Figure 14.20 *Specific attenuation due to gaseous constituents for transmissions through a standard atmosphere (20°C, pressure one atmosphere, water vapour content 7.5 g/m^3). (Source: ITU-R Handbook of Radiometeorology, 1996, with permission of ITU.)*

Rain fading

Significant rain intensity (measured in mm/h) occurs only for small percentages of time. The specific attenuation which is present during such rain is, however, large for frequencies above a few GHz. When fading of this sort occurs the BER can degrade due to a reduction in signal level and an increase in antenna noise temperature.

For signals with normal bandwidths the fading can be considered to be flat and the distortion negligible. In this case the attenuation exceeded for a particular percentage of time is all that is required. This can be estimated as follows:

1. The point rain rate, R_P (mm/h), exceeded for the required time percentage is found either from local meteorological records or by using an appropriate model.
2. The equivalent uniform (or line) rain rate, R_L, is found using curves such as those in Figure 14.21. This accounts for the non-uniform spatial distribution of rain along the path.
3. The specific attenuation, γ_R, for the line rain rate of interest may be estimated using Figure 14.22, or more accurately, using a formula of the form:

   $$\gamma_R = k\, R_L^{\alpha} \text{ (dB/km)} \tag{14.14}$$

   where the regression coefficients k and α are given in [ITU-R, Rec. 838] for a particular polarisation and frequency of interest.
4. The total path attenuation, A (dB), exceeded for the required percentage of time is then found using:

   $$A = \gamma_R L \text{ (dB)} \tag{14.15}$$

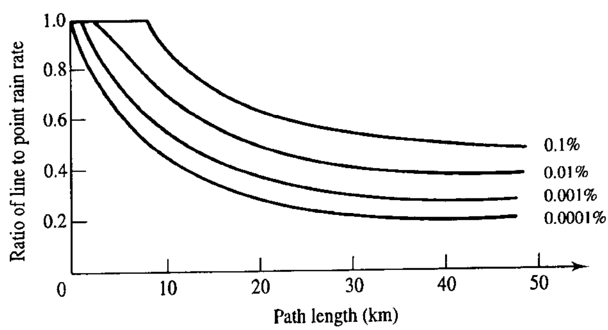


Figure 14.21 *Relationship between point and line rain rates as a function of hop length and percentage time point rain rate is exceeded (source: Hall and Barclay, 1989, reproduced with permission of Peter Peregrinus).*

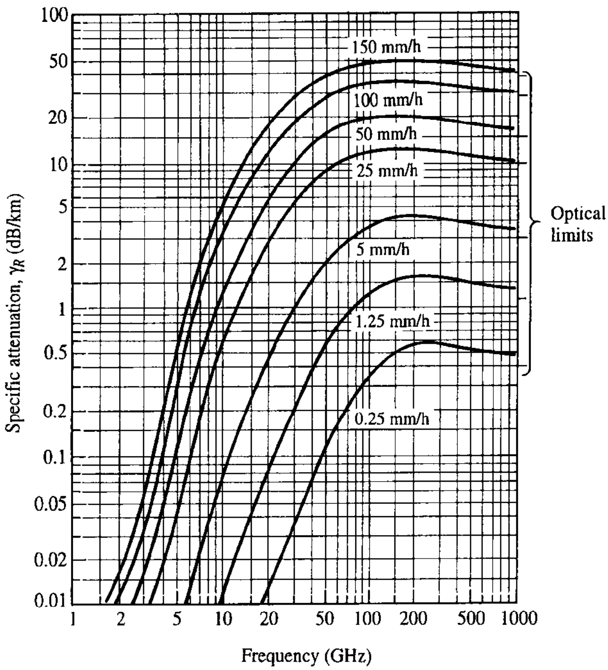


Figure 14.22 *Specific attenuation due to rain (curves derived on the basis of spherical raindrops). (Source: ITU-R Handbook of Radiometeorology, 1996, reproduced with the permission of ITU.)*

Multipath fading on LOS links

Although broadband measurements have shown that several (identifiably discrete) propagation paths can occur on LOS links the consensus at present is that, in practice, the channel can be adequately modelled with only three paths [Rummler, 1979]. The impulse response of such a three-ray model is (to within a multiplicative constant):

$$h(t) = \delta(t) + \alpha\delta(t - T_1) + \beta\delta(t - T_2) \tag{14.16}$$

where T_1 and T_2 define the relative delays between the rays, and α and β define the relative strengths of the rays. The frequency response, $H(f)$, corresponding to equation (14.16) is:

$$H(f) = 1 + \alpha e^{-j2\pi fT_1} + \beta e^{-j2\pi fT_2} \tag{14.17}$$

If $T_1 << 1/B$, where B is the bandwidth of the transmitted signal, then fT_1 is approximately constant over the range $f_c \pm B/2$ (where f_c is the carrier frequency). Equation (14.17) can be written as:

$$H(f) = ae^{-j\theta}\left[1 + \frac{\beta}{a}\, e^{-j(2\pi fT_2 - \theta)}\right] \tag{14.18}$$

where:

$$ae^{-j\theta} = 1 + \alpha e^{-j2\pi fT_1} \tag{14.19(a)}$$

$$a = \sqrt{1 + \alpha^2 + 2\alpha \cos(2\pi fT_1)} \tag{14.19(b)}$$

$$\theta = \tan^{-1}\left(\frac{\alpha \sin 2\pi fT_1}{1 + \alpha \cos 2\pi fT_1}\right) \tag{14.19(c)}$$

The factor $e^{-j\theta}$ in equation (14.18) represents an overall phase shift caused by the first delayed ray which can be ignored.[1] Furthermore, since a notch occurs in the channel's amplitude response at:

$$f(= f_o) = (\theta \pm \pi)/2\pi T_2 \quad \text{(Hz)} \tag{14.20}$$

then equation (14.18) can be rewritten as:

$$H(f) = a\left[1 - be^{-j2\pi(f-f_o)T_2}\right] \tag{14.21}$$

where $b = \beta/a$. If $b < 1$ then equation (14.21) represents a minimum phase frequency response. If unity and b in equation (14.21) are interchanged then the amplitude response, $|H(f)|$, remains unchanged but the phase response becomes non-minimum. Apart from an extra overall phase shift of $-2\pi T_2 f + 2\pi T_2 f_o + \pi$ (the first term representing pure delay, the second term representing intercept distortion and the third term representing signal inversion) the non-minimum phase frequency response can be modelled by changing the sign of the exponent in equation (14.21) which can then be written as:

$$H(f) = a\left[1 - be^{\pm j2\pi(f - f_o)T_2}\right] \tag{14.22}$$

to represent both minimum and non-minimum phase conditions [CCIR, Report 718]. Equation (14.22) has four free parameters (a, b, f_o and T_2) which is excessive, in the sense that if $T_2 < (6B)^{-1}$ then normal channel measurements are not accurate enough to determine all four parameters uniquely. One parameter can therefore be fixed without significantly degrading the formula's capacity to match measured channel frequency responses. The parameter normally fixed is T_2 and the value chosen for it is often set using the rule $T_2 = (6B)^{-1}$. Using this rule, joint statistics of the quantities:

$$A = 10\log_{10}\left[a^2(1 + b^2)\right] \tag{14.23(a)}$$

$$B = 10\log_{10}(2a^2 b) \tag{14.23(b)}$$

[1] A frequency response with $\phi(f) = \theta$, where θ is a non-zero constant, has zero delay distortion (since $d\phi/df = 0$) but non-zero intercept distortion. Intercept distortion, however, results only in a change in the 'phase' relationship between a wave packet carrier and its envelope, the shape of the envelope remaining unchanged. Providing such distortion changes slowly compared to the time interval between phase training sequences it will have no significant effect on the performance of a PSK communications system.

have been determined experimentally for several specific links [CCIR, Report 338]. A and B appear to be well described by a bivariate Gaussian random variable. The mean, variance and correlation of this distribution for three different links are shown in Table 14.2.

Table 14.2 *Measured statistics of multipath channel parameters (after CCIR, Report 338).*

Path length	*Frequency*	*Bandwidth*	$\bar{A}$	$\bar{B}$	σ_A	σ_B	ρ
37 km	11 GHz	55 MHz	−7.25	−5.5	6.5	6.5	0.45
50 km	11 GHz	55 MHz	−8.25	−3.0	9.0	8.5	0.75
42 km	6 GHz	25 MHz	−24.00	−14.5	7.5	7.5	0

Equations (14.23(a) and (b)) can be inverted to allow the joint distribution of a and b to be derived.

Multipath fading is a potentially severe problem for wideband digital links. Full transversal filter equalisation, which can compensate for both the amplitude and the phase distortions in the transmission medium, is therefore generally desirable. Current digital repeater and receiver equipment has considerable sophistication, often incorporating equalisers which remove distortion arising not only from propagation effects but also from other sources such as non-ideal filters. ([ITU-R, Rec. 530] details methods for predicting fading due to multipath effects in combination with other clear air mechanisms.)

Mechanisms of noise enhancement

The presence of loss on a propagation path will increase the noise temperature of a receiving antenna due to thermal radiation. Rain is the most variable source of such loss

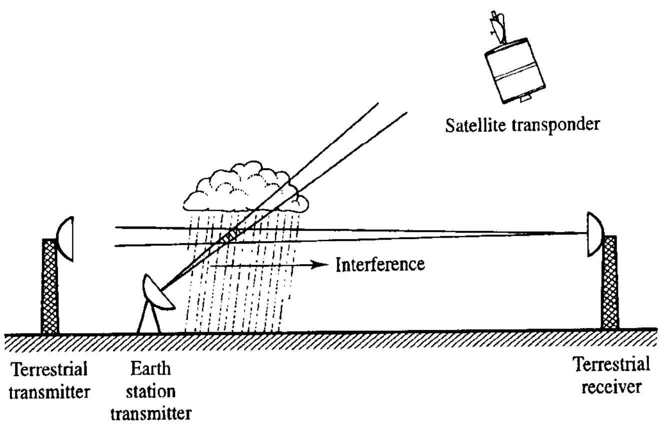


Figure 14.23 *Hydrometeor scatter causing interference between co-frequency systems.*

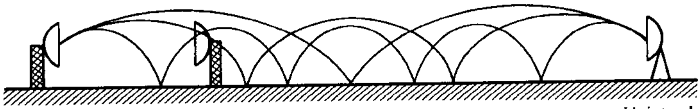


Figure 14.24 *Interference caused by ducting.*

along most microwave paths. During a severe fade the aperture temperature (see Chapter 12) of the antenna, T_A, will approach the physical temperature of the rain producing the loss. A simple model relating T_A to fade depth is:

$$T_A = T_M(1 - \alpha) \tag{14.24}$$

where α is the fade depth expressed as a fraction of unfaded power (i.e. $\alpha = 10^{-\text{atten(dB)}/10}$) and T_M is the absorption temperature of the medium. T_M is not identical to the medium's physical temperature since, in addition to direct thermal radiation, it accounts for scattering and other effects. T_M has been related empirically to surface air temperature, T_S, by [Freeman]:

$$T_M = 1.12\, T_S - 50 \quad \text{(K)} \tag{14.25}$$

where T_S is expressed in K. Equations (14.24) and (14.25) are only appropriate, of course, for values of attenuation which result in aperture temperatures in excess of the clear sky values indicated by Figure 12.25 (i.e. approximately 100 K for terrestrial links between 1 and 10 GHz). Clear sky aperture temperature increases rapidly with frequency above 10 GHz and, for terrestrial links, can be assumed to be 290 K at 20 GHz and above for most purposes (see later, Figure 14.30).

A quite different mechanism of noise enhancement also occurs due to rain, namely precipitation scatter. Here the existence of a common volume between the transmitting antenna of one system and the receiving antenna of a nominally independent system has the potential to couple energy between the two, Figure 14.23. Such coupled energy represents interference but it is thought that its effects are conservatively modelled by an equal amount of thermal noise power. Meteorological ducting conditions can also cause anomalous, long distance, propagation of microwave signals resulting in interference or crosstalk between independent systems, Figure 14.24. Detailed modelling of precipitation scatter, ducting and methods to reduce the interference caused by them has recently been the subject of intensive study [COST 210].

EXAMPLE 14.3

A 7 GHz terrestrial LOS link is 40 km long and operates with good ground clearance (several Fresnel zones) in a location which experiences a rain rate of 25 mm/h or greater for 0.01% of time. The link has a bandwidth of 1.0 MHz, antenna gains of 30.0 dB and an overall receiver noise figure of 5.0 dB. Assuming the ratio of line to point rain rate, R_L/R_p, is well modelled by Figure 14.21, for this location, estimate the transmitter power required to ensure a CNR of 30.0 dB is achieved or exceeded for 99.99% of time.

From Figure 14.21:

$$R_{L,0.01} = 0.4\, R_{p,0.01} = 0.4 \times 25 = 10 \text{ mm/h}$$

Using Figure 14.22, specific attenuation, $\gamma_R = 0.2$ dB/km, i.e.:

$$A_{R,0.01} = \gamma_R\, L = 0.2 \times 40 = 8.0 \text{ dB}$$

Assuming a surface temperature of 290 K, equation (14.25) gives:

$$T_M = 1.12\, T_S - 50 = 1.12 \times 290 - 50 = 275 \text{ K}$$

From equation (14.24):

$$T_A = T_M \left(1 - 10^{-\frac{A}{10}}\right) = 275 \left(1 - 10^{-\frac{8}{10}}\right) = 231 \text{ K}$$

Equivalent noise temperature of receiver:

$$T_e = (f - 1)\, 290 = \left(10^{\frac{5}{10}} - 1\right) 290 = 627 \text{ K}$$

Total system noise temperature:

$$T_{syst} = T_A + T_e = 231 + 627 = 858 \text{ K}$$

Noise power:

$$N = kTB = 1.38 \times 10^{-23} \times 858 \times 1 \times 10^{6} = 1.18 \times 10^{-14} \text{ W} = -139.3 \text{ dBW}$$

Received carrier power:

$$\begin{aligned} C &= P_T + G_T - \text{FSPL} + G_R - A_R \\ &= P_T + 30.0 - 20 \log_{10} \left(\frac{4\pi\, 40 \times 10^3}{0.0429} \right) + 30.0 - 8.0 \\ &= P_T - 89.4 \text{ dBW} \end{aligned}$$

$$\frac{C}{N} = P_T - 89.4 - (-139.3) = P_T + 49.9 \text{ dB}$$

$$P_T = \frac{C}{N} - 49.9 = 30.0 - 49.9 = -19.9 \text{ dBW or } 10.1 \text{ dBm}$$

(Note that the calculation shown above has been designed to illustrate the application of the preceding material and probably contains spurious precision. In particular a simpler estimate for T_A of 290 K results in an increase in P_T of only 0.3 dB.)

Cross-polarisation and frequency reuse

It is possible to double the capacity of a microwave link by using orthogonal polarisations for independent co-frequency channels. For QPSK systems this combines with the advantage of orthogonal inphase and quadrature signalling to give a four-fold increase in transmission capacity over a simple BPSK, single polarisation, system. Unusual propagation conditions along the radio path can give rise to polarisation changes (called cross-polarisation) which results in potential crosstalk at the receiver. The use of corrugated horns and scalar feeds in the design of microwave antennas reduces antenna induced cross-polarisation during refractive bending or multipath conditions. Similarly the use of vertical and horizontal linear polarisations minimises rain induced cross-polarisation which occurs when the angle between the symmetry axis of falling rain drops and the electric field vector of the signal is other than 0° or 90°.

14.3 Fixed point satellite communications

The use of satellites is one of the three most important developments in telecommunications over the past 40 years. (The other two are cellular radio and the use of optical fibres.) Geostationary satellites, which are essentially motionless with respect to points on the earth's surface and which first made satellite communications commercially feasible, were proposed by the scientist and science fiction writer Arthur C. Clarke. The geostationary orbit lies in the equatorial plane of the earth, is circular and has the same sense of rotation as the earth, Figure 14.25. Its orbital radius is 42,164 km and since the earth's mean equatorial radius is 6,378 km its altitude is 35,786 km. (For simple calculations of satellite range from a given earth station, the earth is assumed to be spherical with radius 6,371 km.) There are other classes of satellite orbit which have advantages over the geostationary orbit for certain applications. These include highly

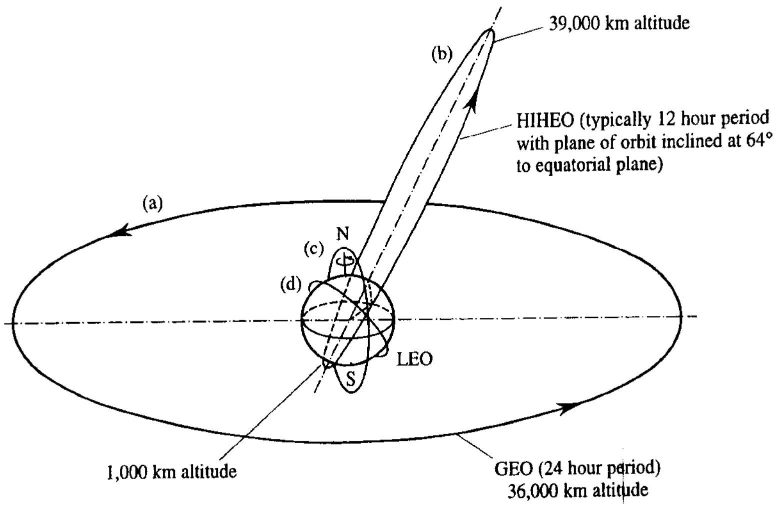


Figure 14.25 *Selection of especially useful satellite orbits: (a) geostationary (GEO); (b) highly inclined highly elliptical (HIHEO); (c) polar orbit; and (d) low earth (LEO).*

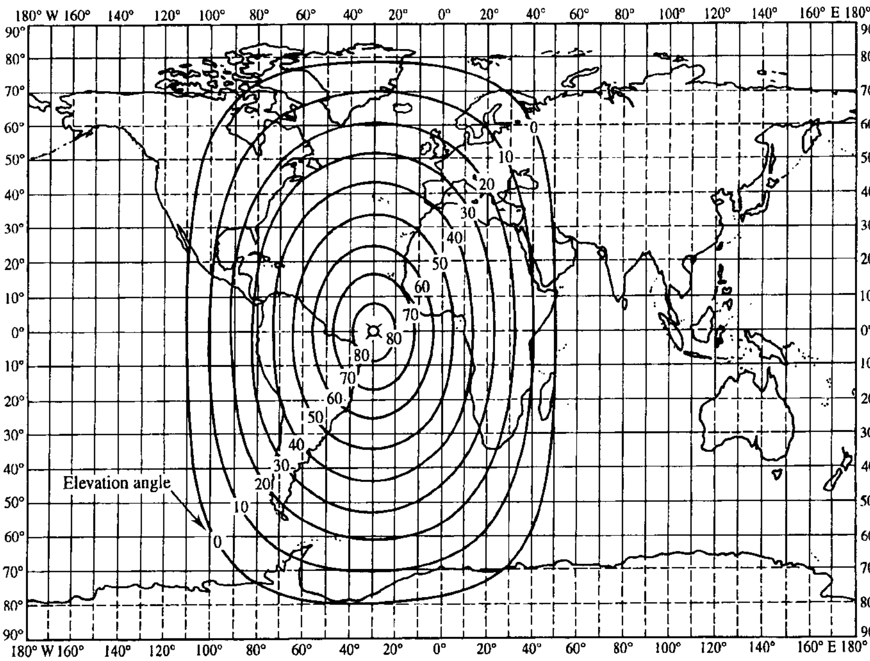


Figure 14.26 *Coverage areas as a function of elevation angle for a satellite with global beam antenna (from CCIR Handbook, 1988, reproduced with the permission of ITU).*

inclined highly elliptical (HIHE) orbits, polar orbits and low earth orbits (LEOs), Figure 14.25. For fixed point communications the geostationary orbit is the most commercially important, for the following reasons:

1. Its high altitude means that a single satellite is visible from a large fraction of the earth's surface (42% for elevation angles > 0° and 38% for elevation angles > 5°). Figure 14.26 shows the coverage area as a function of elevation angle for a geostationary satellite with a global beam antenna. (Elevation angles < 10° are not recommended, and angles < 5° are not used, because of the severe scintillation and fading of the signal, and high antenna noise temperature, which occur due to the large thickness of atmosphere traversed by the propagation path.)
2. No tracking of the satellite by earth station antennas is necessary.
3. No handover from one satellite to another is necessary since the satellite never sets.
4. Three satellites give almost global coverage, Figure 14.27. (The exception is the polar regions with latitudes > 81° for elevation angles > 0° and latitudes > 77° for elevation angles > 5°.)
5. No Doppler shifts occur in the received carrier.

The following advantages apply to geostationary satellites but may also apply, to a greater or lesser extent, to some communication satellites in non-geostationary orbits:

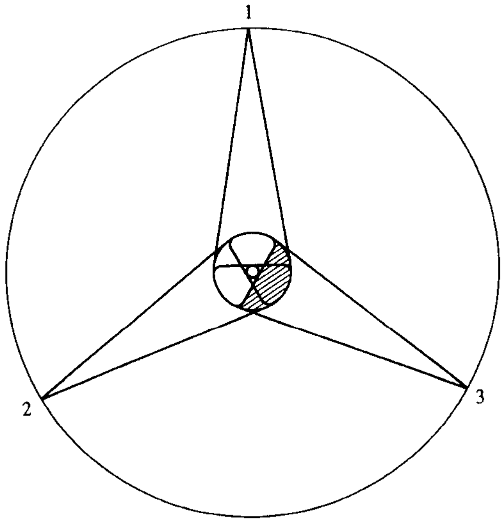


Figure 14.27 *Global coverage (excepting polar regions) from 3 geostationary satellites. (Approximately to scale, innermost circle represents 81° parallel.)*

1. The communications channel can be either broadcast or point-to-point.
2. New communication network connections can be made simply by pointing an antenna at the satellite. (For non-geostationary satellites this is not entirely trivial since tracking and/or handover are usually necessary.)
3. The cost of transmission is independent of distance.
4. Wide bandwidths are available, limited at present only by the speed of the transponder electronics and receiver noise performance.

Despite their very significant advantages, geostationary satellites do suffer some disadvantages. These include:

1. Polar regions are not covered (i.e. latitudes > 77° for elevation angles > 5°).
2. High altitude means large FSPL (typically 200 dB).
3. High altitude results in long propagation delays (approximately 1/8 s for uplink and 1/8 s for downlink).

The latter disadvantage means that inadequately suppressed echoes from subscriber receiving equipment arrive at the transmitting subscriber 0.5 s after transmission. Both the 0.25 s delay between transmission and reception and the 0.5 s delay between transmission and echo can be disturbing to telephone users. This means echo suppression or cancellation equipment, which may use techniques [Mulgrew and Grant] not dissimilar to the equalisers of section 8.5, is almost always required.

14.3.1 Satellite frequency bands and orbital spacing

Figure 14.28 shows the principal European frequency bands allocated to fixed point satellite services. The 6/4 GHz (C-band) allocation is now fairly congested and new systems are being implemented at 14/11 GHz (Ku-band). 30/20 GHz (Ka-band) systems are currently being investigated. The frequency allocation at 12 GHz is mainly for direct broadcast satellites (DBS). Inter-satellite crosslinks use the higher frequencies as here there is no atmospheric attenuation. The higher of the two frequencies allocated for a satellite communications system is invariably the uplink frequency. This is because the satellite has limited antenna size and a high antenna noise temperature (typically 290 K). The gain of the satellite receiving antenna (and therefore the satellite *G/T*) is maximised by using the higher frequency on the uplink.

The reason why two frequencies are necessary at all (one for the uplink and one for the downlink) is that the isolation between the satellite transmit and receive antennas is finite. Since the satellite transponder has enormous gain there would be the possibility of positive feedback and oscillation if a frequency offset were not introduced.

Although the circumference of a circle of radius 42,000 km is large, the number of satellites which can be accommodated in the geostationary orbit is limited by the need to illuminate only one satellite when transmitting signals from a given earth station. If other satellites are illuminated then interference may result. For practical antenna sizes 4° spacing is required between satellites in the 6/4 GHz bands. Since narrower beamwidths are achievable in the 14/11 GHz band, 3° spacing is permissible here and in the 30/20 GHz band spacing can approach 1°.

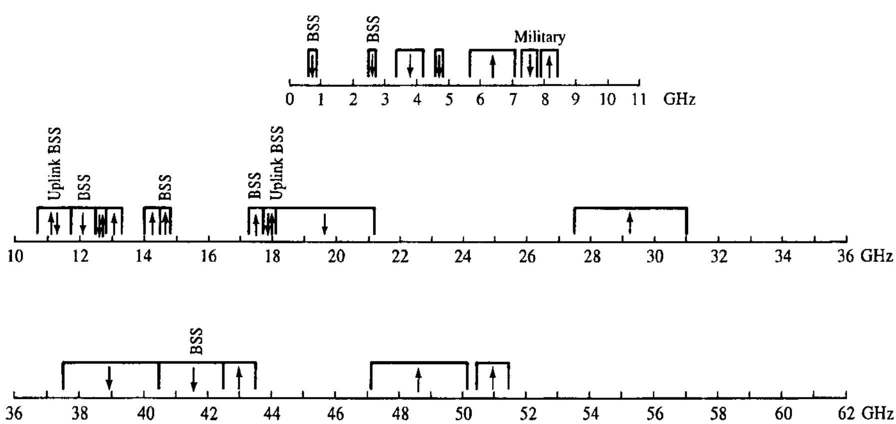


Figure 14.28 *Approximate uplink (↑) and dowlink (↓) allocations for region 1 (Europe, Africa, former USSR, Mongolia) fixed satellite, and broadcast satellite (BSS), services.*

14.3.2 Earth station look angles and satellite range

Figure 14.29 shows the geometry of an earth station (E) and geostationary satellite (S). Some careful trigonometry shows that the earth station antenna elevation angle, α, the azimuth angle, β, and the satellite range from the earth station, R_{ES}, are given by:

$$\alpha = \tan^{-1}[(\cos\gamma - 0.15127)/\sin\gamma] \tag{14.26}$$

$$\beta = \pm\cos^{-1}[-\tan\theta_E/\tan\gamma] \tag{14.27}$$

$$R_{ES} = 23.188 \times 10^6 \sqrt{3.381 - \cos\gamma}\ \text{(m)} \tag{14.28}$$

where γ, the angle subtended by the satellite and earth station at the centre of the earth, is given by:

$$\gamma = \cos^{-1}[\cos\theta_E \cos(\phi_E - \phi_S)] \tag{14.29}$$

θ_E is the earth station latitude (positive north, negative south), ϕ_E and ϕ_S are the earth station and satellite longitude respectively (positive east, negative west).

Notice that azimuth, β, is defined clockwise (eastwards) from north. Negative β therefore indicates an angle anticlockwise (westwards) from north. The negative sign is taken in equation (14.27) if the earth station is east of the satellite and the positive sign is

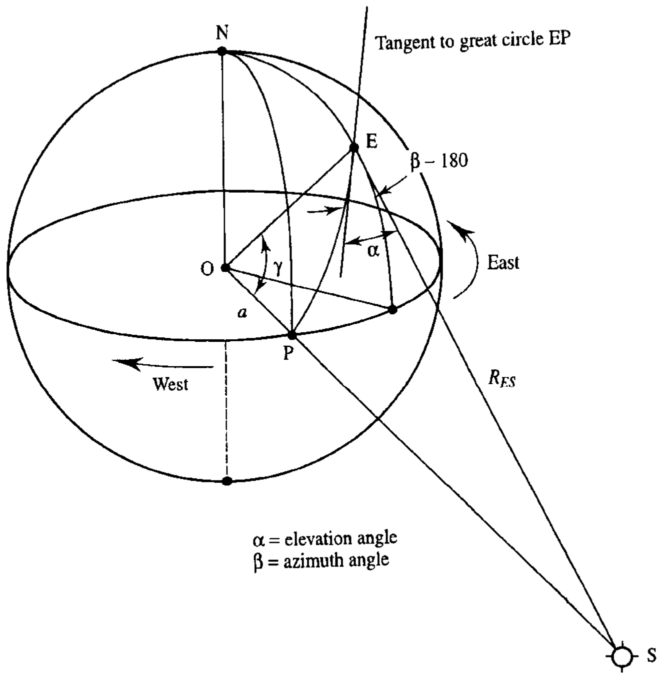


Figure 14.29 *Geostationary satellite geometry.*

taken otherwise.

EXAMPLE 14.4
Find the look angles and range to a geostationary, Indian Ocean, satellite located at 60.0° E as seen from Edinburgh in the UK (55.95° N, 3.20° W).

$$\begin{aligned}
\gamma &= \cos^{-1}[\cos\theta_E \cos(\phi_E - \phi_S)] \\
&= \cos^{-1}[\cos(55.95°)\cos(-3.20° - 60.0°)] \\
&= 75.38° \ \ (= 1.316 \text{ rad}) \\
\alpha &= \tan^{-1}[(\cos\gamma - 0.15127)/\sin\gamma] \\
&= \tan^{-1}[(\cos(1.316) - 0.15127)/\sin(1.316)] \\
&= 0.1038 \text{ rad } (= 5.95°) \\
\beta &= \pm\cos^{-1}[-\tan\theta_E/\tan\gamma] \\
&= \pm\cos^{-1}[-\tan(55.95°)/\tan(75.38°)] \\
&= \pm 112.7°
\end{aligned}$$

Since the earth station is west of the satellite the upper (positive) sign is taken, i.e. the satellite azimuth is +112.7° (eastwards from North). Now from equation (14.28):

$$\begin{aligned}
R_{ES} &= 23.188 \times 10^6 \sqrt{3.381 - \cos\gamma} \\
&= 23.188 \times 10^6 \sqrt{3.381 - \cos(1.316)} \\
&= 4102 \times 10^4 \text{ m} \\
&= 41{,}020 \text{ km}
\end{aligned}$$

14.3.3 Satellite link budgets

On the uplink the satellite antenna, which looks at the warm earth, has a noise temperature of approximately 290 K. The noise temperature of the earth station antenna, which looks at the sky, is usually in the range 5 to 100 K for frequencies between 1 and 10 GHz, Figure 12.25. For frequencies above 10 GHz resonance effects of water vapour and oxygen molecules, at 22 GHz and 60 GHz respectively, become important, Figure 14.30. For wideband multiplex telephony, the front end of the earth station receiver is also cooled so that its equivalent noise temperature, T_e, section 12.3.1, is very much smaller than 290 K. This therefore achieves a noise floor that is very much lower than −174 dBm/Hz.

The gain, G, of the cooled low noise amplifier boosts the received signal so that the following amplifiers in the receiver can operate at room temperature. Traditional operating frequencies for satellite communication systems were limited at the low end to

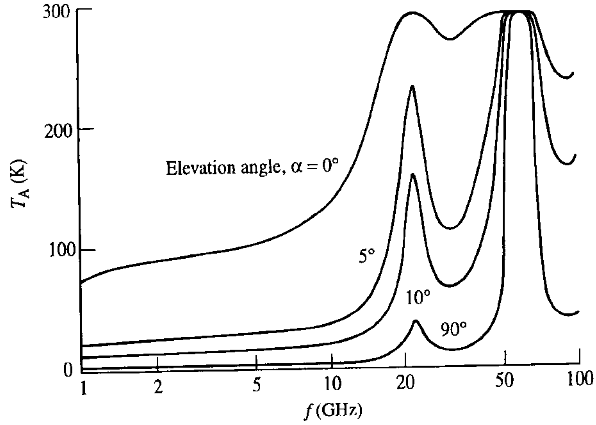


Figure 14.30 *Antenna aperture temperature, T_A, in clear air (pressure one atmosphere, surface temperature 20°C, surface water vapour concentration 10 g/m^3). (Source: ITU-R Handbook of Radiometeorology, 1996, reproduced with the permission of the ITU.)*

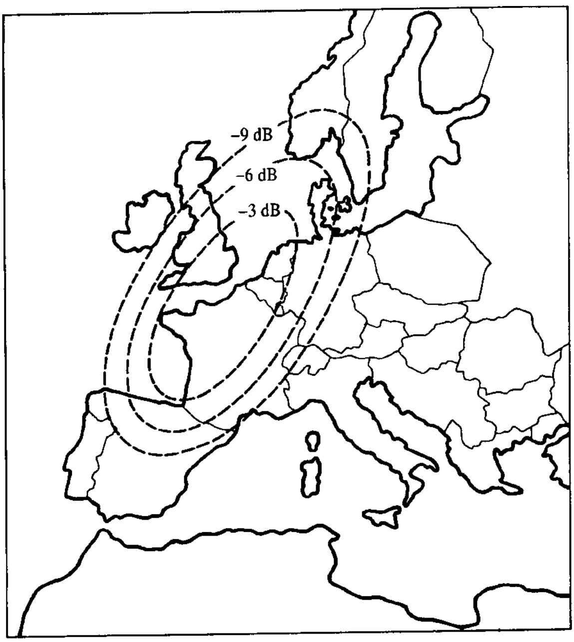
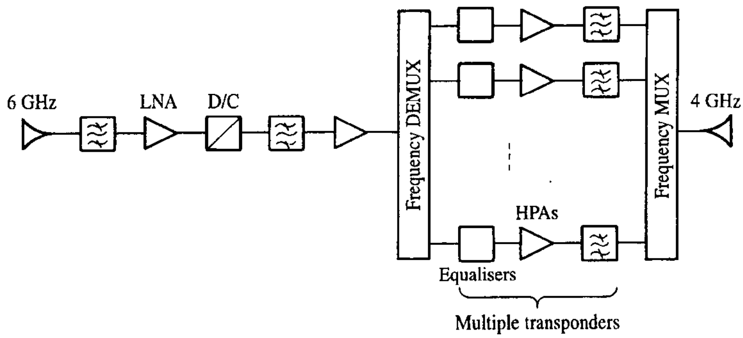


Figure 14.31 *Contours of EIRP with respect to EIRP on antenna boresight.*

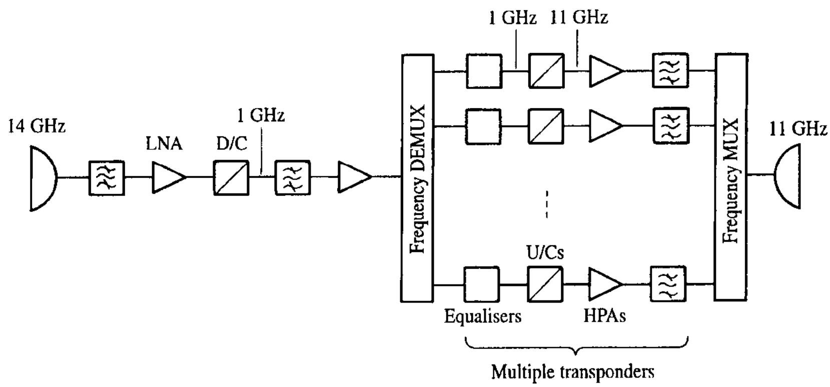
>1 GHz by galactic noise and at the high end to <15 GHz by atmospheric thermal noise and rain attenuation, Figure 12.25. However in recent years a more detailed physical and statistical understanding of rain fading, and other hydrometeor effects, has made operation in higher frequency bands practical.

The satellite transponder is the critical component in a satellite link as its transmitter is invariably power limited by the onboard power supply (i.e. solar cell area and battery capacity). The downlink usually has the worst power budget and this often constrains performance. EIRP is also dependent on satellite antenna design and in particular the earth 'footprint'. Spot beam antennas covering only a small part of the earth, e.g. Figure 14.31, have higher gain giving EIRP values of 30 to 40 dBW or greater.

In addition to being power limited the satellite transponder must be as small and light as possible due to the large launching costs, per kg of weight and m^3 of space. This means that the transponder's high power amplifier (HPA), Figure 14.32, must be operated



(a)



(b)

Figure 14.32 *Simplified block diagram of satellite transponders: (a) single conversion C-band; (b) double conversion Ku-band (redundancy not shown).*

at as high an output power as possible to maintain adequate downlink CNR. As a consequence the transponder is operated in its non-linear region near saturation, resulting in amplitude to amplitude (AM/AM) and amplitude to phase (AM/PM) conversion, Figure 14.33. Intermodulation products (IPs), arising due to mixing of nominally independent signals, simultaneously present in the transponder, can be severe, effectively reducing the overall CNR. Some *back-off* from the transponder saturating input and output power is therefore necessary. (For digital satellite systems the IP problem and resulting need for back-off is usually less serious than for analogue systems, section 14.3.5.) Input and output back-off (BO_i and BO_o respectively) are shown in Figure 14.33. Typical values of back-off are a few dB, BO_i being somewhat greater than BO_o. The transmitted power for the satellite downlink and received power for the satellite uplink during clear sky conditions are therefore given respectively by:

$$P_T = P_{o\,sat} - \mathrm{BO}_o \tag{14.30}$$

$$C = P_{i\,sat} - \mathrm{BO}_i \tag{14.31}$$

where $P_{o\,sat}$ and $P_{i\,sat}$ are the saturated transponder output and input powers. The uplink CNR can therefore be calculated using:

$$\left(\frac{C}{N}\right)_u = W_s - \mathrm{BO}_i + 10\log_{10}\left(\frac{\lambda^2}{4\pi}\right) + \left(\frac{G}{T}\right)_s + 228.6 - 10\log_{10} B \tag{14.32(a)}$$

where W_s is the power density (dBW m^{-2}) at the satellite receiving antenna required to saturate the transponder, $10\log_{10}(\lambda^2/4\pi)$ is the effective area of an isotrope (dBm^2), $(G/T)_s$ is the satellite G/T (dB K^{-1}), 228.6 ($= -10\log_{10} k$) is Boltzmann's constant (dBW Hz^{-1}K^{-1}), and 10 $\log_{10} B$ is bandwidth (dB Hz). Note that the transmitting earth station EIRP, EIRP_e, and the received power density at the satellite are related by:

$$W_s - \mathrm{BO}_i = \mathrm{EIRP}_e - \text{spreading loss} - L_{Au} \quad (\text{dBW m}^{-2}) \tag{14.32(b)}$$

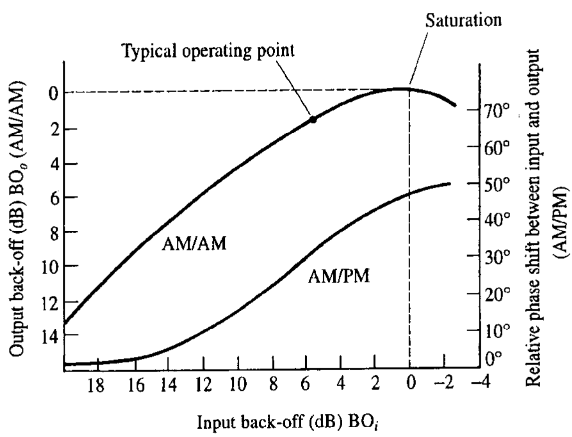


Figure 14.33 *Amplitude and phase characteristic for typical satellite transponder TWT amplifier.*

where L_{Au} is uplink atmospheric attenuation. The equivalent formula, to equation (14.32(a)), for the downlink is:

$$\left(\frac{C}{N}\right)_d = \mathrm{EIRP}_s - \mathrm{BO}_o + \left(\frac{G}{T}\right)_e - \mathrm{FSPL}_d + 228.6 - 10\log_{10} B - L_{Fd} - L_{Ad} \quad (14.33)$$

where EIRP_s is the saturated satellite EIRP, $(G/T)_e$ is the earth station G/T, L_{Fd} represents the downlink fixed losses (such as earth station transmission line loss) and L_{Ad} is the downlink atmospheric attenuation.

Table 14.3 *Typical 4/6 GHz satellite link budget.*

Uplink (6 GHz)		
Saturation flux density		−72.2 dBW m^{-2}
Input back-off, BO_i		−5.8 dB
Satellite antenna gain, G_R	23.1 dB	
Satellite system noise temperature, T_{syst}	27.6 dBK	
Satellite G/T	−4.5 dB/K	−4.5 dB/K
Effective area of isotrope		−37.0 dB m^2
Minus Boltzmann's constant		228.6 dBW Hz^{-1}K^{-1}
Minus transponder bandwidth		−75.6 dB Hz
Resulting clear sky CNR_u		33.5 dB

Downlink (4 GHz)		
Saturated transponder output power, P_o	7.0 dBW	
Satellite antenna gain, G_T	22.5 dB	
EIRP_s	29.5 dBW	29.5 dBW
Output back-off, BO_o		−3.2 dB
FSPL, $20\log_{10}(4\pi R/\lambda)$		−196.1 dB
Earth station antenna gain, G_R	58.3 dB	
Clear sky earth station noise temperature, T_{syst}	18.8 dBK	
Clear sky G/T	39.5 dB/K	39.5 dB/K
Minus Boltzmann's constant		228.6 dBW Hz^{-1}K^{-1}
Minus transponder bandwidth		−75.6 dB Hz
Atmospheric attenuation, L_{Ad}		−0.8 dB
Fixed losses, L_{Fd}		−2.0 dB
Resulting clear sky CNR_d		19.9 dB

In satellite multiplex telephony systems, large earth station antennas with kW transmitters give very large EIRP_e (typically 90 dBW) and the uplink exhibits good noise, or E_b/N_0, performance. The transmitter power on the satellite is typically restricted by battery and solar cell capacity to 10 to 100 W (and hence EIRP_s is typically restricted to 30 to 50 dBW). The downlink often, therefore, limits overall CNR and BER performance.

A typical, clear sky, link budget for a 6/4 GHz satellite communications system is shown in Table 14.3.

As illustrated schematically in Figure 14.34 the received downlink carrier power, C_d, for a transparent transponder, is given by:

$$C_d = C_u G L_d \quad \text{(W)} \tag{14.34}$$

where C_u is the *received* uplink power, G is the operating gain of the transponder and L_d represents *all* downlink losses. Furthermore, the total received downlink noise power, N, can be expressed by:

$$N = N_u G L_d + N_d \quad \text{(W)} \tag{14.35}$$

where N_u is the noise power contributed by the uplink and N_d is the noise power contributed by the downlink. Thus:

$$\begin{aligned} \frac{N}{C_d} &= \frac{N_u G L_d + N_d}{C_u G L_d} \\ &= \frac{N_u}{C_u} + \frac{N_d}{C_d} \end{aligned} \tag{14.36}$$

The overall CNR for the satellite link can therefore be written as:

$$\frac{C}{N} = \frac{1}{\left(\frac{C}{N}\right)_u^{-1} + \left(\frac{C}{N}\right)_d^{-1}} \tag{14.37}$$

(dropping the subscript d from the received signal power, C, to conform to convention).

More accurate calculations of satellite systems performance must account for the effects of intermodulation products and interference on uplink and/or downlink. Since noise, intermodulation and interference processes are independent, equation (14.37) can be extended as follows:

$$\frac{C}{N} = \frac{1}{\left(\frac{C}{N}\right)_u^{-1} + \left(\frac{C}{N}\right)_d^{-1} + \left(\frac{C}{N}\right)_{IP}^{-1} + \left(\frac{C}{I}\right)_u^{-1} + \left(\frac{C}{I}\right)_d^{-1}} \tag{14.38}$$

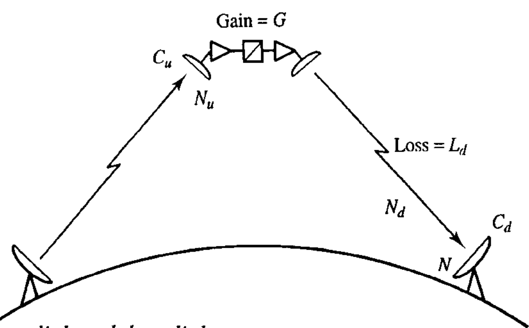


Figure 14.34 *CNRs on uplink and downlink.*

where $(C/N)_{IP}$ is the carrier to intermodulation noise ratio and C/I is the carrier to interference ratio. For digital satellite systems (section 14.3.6) in which only a single carrier is present in the transponder at any given time, then intermodulation noise is absent and $(C/N)_{IP}^{-1}$ is zero. For the example link budget in Table 14.3 the overall clear sky CNR, assuming no intermodulation products and neglecting interference, would be:

$$\frac{C}{N} = \frac{1}{(\text{antilog}_{10}\, 3.35)^{-1} + (\text{antilog}_{10}\, 1.99)^{-1}} \tag{14.39}$$

$$= \frac{1}{(2239)^{-1} + (97.72)^{-1}} = \frac{1}{0.045 \times 10^{-2} + 1.023 \times 10^{-2}}$$

$$= \frac{1}{1.068 \times 10^{-2}}$$

$$= 93.63 = 19.7 \text{ dB}$$

Notice that here, as in many cases, the CNR performance of the system is dominated by the downlink.

The clear sky carrier to noise ratio calculated above allows only for gaseous background attenuation in the term L_A. In practice a satellite system link budget must also account for fading of the signal and enhancement of the noise (both due, mainly, to the sporadic presence of rain along the propagation path).

Rain induced specific attenuation (in dB/km) varies with frequency, Figure 14.22. The effective length of the propagation path subject to a given rain event depends on elevation angle and climatic factors. *Typical* fade margins included in link budgets to account for 99.99% of meteorological conditions are given in Table 14.4 for the different satellite frequency bands. The detailed calculation of gaseous background attenuation and rain margins is described in section 14.3.4.

Table 14.4 *Typical values for fade margin in different frequency bands.*

Elevation angle	*C*	*Ku*	*Ka*
10°	2.0 dB	8 dB	15 dB
30°	1.0 dB	6 dB	10 dB
90°	0.7 dB	5 dB	8 dB

EXAMPLE 14.5

An 11.7 GHz satellite downlink operates from geosynchronous orbit with 25 W of transmitter output power connected to a 20 dB gain antenna with 2 dB feeder losses. The earth station is at a range of 38,000 km from the satellite and uses a 15 m diameter receive antenna, with 55% efficiency, feeding a low noise (cooled) amplifier which results in a receiver system noise temperature of 100 K. If an E_b/N_0 of 20 dB is required for adequate BER performance what maximum bit rate can be accommodated using BPSK modulation, assuming performance is limited by the downlink and atmospheric attenuation can be neglected?

Transmitter power = 14 dBW
EIRP = 14 + 20 − 2 = 32 dBW = 62 dBm
Free space loss = 20 log10 $(4\pi 38 \times 10^6)/0.0256$ = 205.4 dB

Receiver antenna $G_R = \frac{4\pi}{\lambda^2}\frac{\pi d^2}{4}\,0.55 = 62.7$ dB

Received power level = 62 − 205.4 + 62.7 dBm = −80.7 dBm
Receiver noise at 100 K noise temperature = −178.6 dBm/Hz
If $BT_o = 1.0$ then $E_b/N_0 = C/N$.
Available margin for E_b/N_0 and modulation bandwidth = 178.6 − 80.7 = 97.9 dB Hz.
If $E_b/N_0 = 20$ dB then the margin for modulation = 77.9 dB Hz = 61.6 MHz.
With BPSK at 1 bit/s/Hz then the modulation rate can be 61.6 Mbit/s.

Alternatively:
If $G_R = 62.7$ dB and $T_{syst} = 100$ K then $G/T = 42.7$ dB/K
Then radiated power at receiver antenna = + 62 − 205.4 dBm = −143.4 dBm
Power at receiver input = − 143.4 + 42.7 dBm/K = −100.7 dBm/K
Boltzmann's constant = − 198.6 dBm/Hz/K
Difference = 97.9 dBHz
Allowing 20 dB for acceptable E_b/N_0 leaves 77.9 dBHz which will support a 61.6 Mbit/s BPSK symbol rate.

14.3.4 Slant path propagation considerations

The discussion of satellite link budgets in section 14.3.3 referred to atmospheric effects which must be accounted for to achieve adequate system availability. The principal effects which contribute to changes in signal level on earth–space paths from that expected for free space propagation are:

1. Background atmospheric absorption.
2. Rain fading.
3. Scintillation.

The principal mechanisms of noise and interference enhancement are:

1. Sun transit.
2. Rain enhancement of antenna temperature.
3. Interference caused by precipitation scatter and ducting.
4. Crosstalk caused by cross-polarisation.

Background gaseous absorption

Gaseous absorption on slant path links can be described by $A = \gamma L$ but with L replaced by effective path length in the atmosphere, L_{eff}. L_{eff} is less than the physical path length in the atmosphere due to the decreasing density of the atmosphere with height. In practice the total attenuation, $A(f)$, is usually calculated using curves of zenith attenuation, Figure 14.35, and a simple geometrical dependence on elevation angle, α, i.e.:

$$A(f) = \frac{A_{zenith}(f)}{\sin \alpha} \tag{14.40}$$

A_{zenith} is the one-way total zenith attenuation and depends on both frequency and surface pressure (reflecting the height of the earth station above sea level). Curve A is for a dry atmosphere and curve B includes the effect of water vapour at a concentration which is typical of temperate climates. (The scale height of the water vapour concentration is 2 km.) Correction factors have been derived which can be used with Figure 14.35 and equation (14.40) to find slant-path gaseous attenuation for other surface pressures and water vapour densities [Freeman].

Rain fading

The same comments can be made for rain fading on slant paths as those which have already been made for terrestrial paths. The slant-path geometry, however, means that the calculation of effective path length depends not only on the horizontal structure of the rain but also on its vertical structure.

One model [ITU-R, Rec. 618] for predicting the rain fading exceeded for a given percentage of time therefore incorporates the following formula which derives an

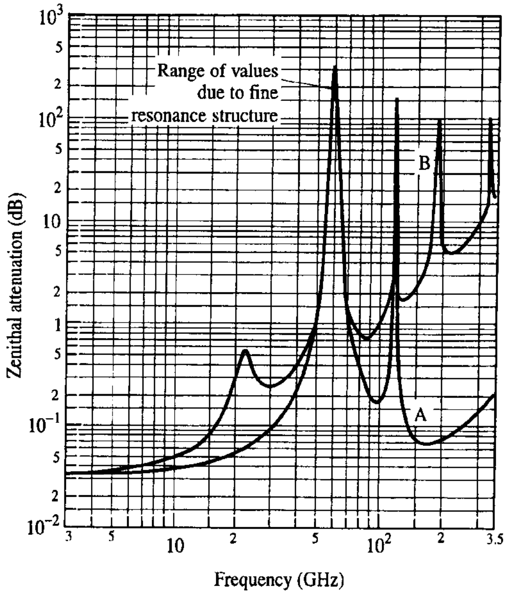


Figure 14.35 *Total ground level zenith attenuation (15°C, 1013 mb) for, A, a dry atmosphere and, B, with a surface water vapour content of 7.5 g/m³ decaying exponentially with height. (Source: ITU-R Rec. 676, 1995, reproduced with the permission of the ITU.)*

effective rain height from earth station latitude, i.e.:

$$h_R \text{ (km)} = \begin{cases} 3.0 + 0.028\phi, & 0 < \phi < 36° \\ 4.0 - 0.075(\phi - 36), & \phi \geq 36° \end{cases} \tag{14.41}$$

The slant path length below the rain height is found from Figure 14.36, i.e.:

$$L_s = \frac{h_R - h_E}{\sin \alpha} \text{ (km)} \tag{14.42}$$

where h_E is the height of the earth station and α is the slant path elevation angle. (A more accurate formula which takes account of earth curvature is used for $\alpha < 5°$.) The ground projection of L_s is calculated using:

$$L_G = L_s \cos \alpha \text{ (km)} \tag{14.43}$$

and a path length reduction formula appropriate for an exceedance value of 0.01% of time is applied, i.e.:

$$r_{0.01} = \frac{1}{1 + L_G(e^{0.015\,R_{0.01}})/35} \tag{14.44}$$

(For $R_{0.01} > 100$ mm/h then 100 mm/h is used instead of $R_{0.01}$.) The one-minute rain rate exceeded for 0.01% of time, $R_{0.01}$, is estimated, preferably from local meteorological data, and the corresponding 0.01% specific attenuation is calculated using equation (14.14). (Alternatively a first order estimate of $\gamma_{R\,0.01}$ can be made by interpolating the curves of Figure 14.22.) The total path rain attenuation exceeded for 0.01% of time is then given by:

$$A_{0.01} = \gamma_{R\,0.01} L_s r_{0.01} \text{ (dB)} \tag{14.45}$$

The attenuation exceeded for some other time percentage, p (between 0.001% and 1.0%), can be estimated using the empirical scaling law:

$$A_p = A_{0.01} 0.12 p^{-(0.546 + 0.043 \log_{10} p)} \text{ (dB)} \tag{14.46}$$

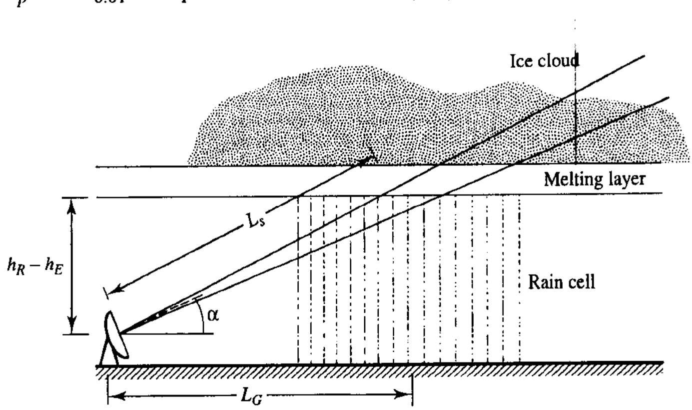


Figure 14.36 *Slant path geometry.*

The slant path attenuation prediction method described here is different from that described in section 14.2.4 (for terrestrial links) in that the former uses an actual rain rate and a path length reduction factor whilst the latter uses an actual path length and a rain rate reduction factor. Clearly both types of model are equivalent in that they take account of the non-uniform distribution of rain along an extended path and either approach can be applied to either type of link. In particular the rain rate prediction method described for satellite links can also be applied to terrestial links [ITU-R, Rec. 530] if L_G and L_s are replaced by the actual terrestial path length, L.

Scintillation

Scintillation refers to the relatively small fluctuations (usually less than, or equal to, a few dB peak to peak) of received signal level due to the inhomogeneous and dynamic nature of the atmosphere. Spatial fluctuations of electron density in the ionosphere and fluctuations of temperature and humidity in the troposphere result in non-uniformities in the atmospheric refractive index. As the refractive index structure changes and/or moves across the slant-path (with, for example, the mean wind velocity) these spatial variations are translated to time variations in received signal level. The fluctuations occur typically on a time scale of a few seconds to several minutes. Scintillation, unlike rain fading, can result in signal enhancements as well as fades. The CNR is degraded, however, during the fading part of the scintillating signal and as such has the potential to degrade system performance. Whilst severe fading is usually dominated by rain and occurs for only small percentages of time the less severe fading due to scintillation occurs for large percentages of time and may be significant in the performance of low-margin, low-availability, systems such as VSATs (see section 14.3.9). At very low elevation angles multipath propagation due to reflection from, and/or refraction through, stable atmospheric layers may occur. Distinguishing between severe scintillation and multipath propagation in this situation may, in practice, be difficult however. Scintillation intensity is sensitively dependent on elevation angle, increasing as elevation angle decreases.

Mechanisms of noise enhancement

Excess thermal noise arising from rain, precipitation scatter, ducting and cross-polarisation may all affect satellite systems in essentially the same way as terrestrial systems (see section 14.2.4). Rain induced cross-polarisation, however, is usually more severe on slant-path links since the system designer is not free to choose the earth station's polarisation. Furthermore, since the propagation path continues above the rain height, tropospheric ice crystals may also contribute to cross-polarisation. Earth–space links employing full frequency reuse (i.e. orthogonal polarisations for independent co-frequency carriers) may therefore require adaptive cross-polar cancellation devices to maintain satisfactory isolation between carriers.

Sun transit refers to the passage of the sun through the beam of a receiving earth station antenna. The enormous noise temperature of the sun effectively makes the system unavailable for the duration of this effect. Geostationary satellite systems suffer sun transit for a short period each day around the spring and vernal equinoxes. (Other celestial noise sources can also cause occasional increases in earth station noise temperature.)

System availability constraints

The propagation effects described above will degrade a system's CNR below its clear sky level for a small, but significant, fraction of time. In order to estimate the constraints which propagation effects put on system availability (i.e. the fraction of time that the CNR exceeds its required minimum value) the clear sky CNR must be modified to account for these propagation effects.

In principle, since received signal levels fluctuate due to variations in gaseous absorption and scintillation, these effects must be combined with the statistics of rain fading to produce an overall fading cumulative distribution, in order to estimate the CNR exceeded for a given percentage of time. Gaseous absorption and scintillation give rise to relatively small fade levels compared to rain fading (at least at the large time percentage end of the fading CD) and it is therefore often adequate, for traditional high availability systems, to treat gaseous absorption as constant (as in the link budget of section 14.3.3) and neglect scintillation altogether. (This approach may not be justified in the case of VSATs (see section 14.3.9), for which transmit power, G/T, and consequent availability are all low.) Once the uplink and downlink fade levels for the required percentage of time have been established then the CNRs can be modified as described below.

The uplink CNR exceeded for $100 - p\%$ of time (where typically $100 - p\% = 99.99\%$, i.e. $p = 0.01\%$), $(C/N)_{u,\,100-p}$, is simply the clear sky carrier to noise ratio, $(C/N)_u$, reduced by the fade level exceeded for $p\%$ of time, $F_u(p)$, i.e.:

$$\left(\frac{C}{N}\right)_{u,\,100-p} = \left(\frac{C}{N}\right)_u - F_u(p) \quad \text{(dB)} \tag{14.47}$$

The uplink noise is not increased by the fade since the attenuating event is localised to a small fraction of the receiving satellite antenna's coverage area. (Even if this were not so the temperature of the earth behind the event is essentially the same as the temperature of the event itself.)

If uplink interference arises from outside the fading region then the uplink carrier to interference ratio exceeded for $100 - p\%$ of time will also be reduced by $F_u(p)$, i.e.:

$$\left(\frac{C}{I}\right)_{u,\,100-p} = \left(\frac{C}{I}\right)_u - F_u(p) \quad \text{(dB)} \tag{14.48}$$

In the absence of uplink fading (or the presence of uplink power control to compensate uplink fades) the downlink CNR exceeded for $100 - p\%$ of time is determined by the downlink fade statistics alone. $(C/N)_d$, however, is reduced not only by downlink carrier fading (due to downlink attenuating events) but also by enhanced antenna noise temperature (caused by thermal radiation from the attenuating medium in the earth station's normally cold antenna beam), i.e.:

$$\left(\frac{C}{N}\right)_{d,\,100-p} = \left(\frac{C}{N}\right)_d - F_d(p) - \left(\frac{N_{faded}}{N_{clear\ sky}}\right)_{dB} \quad \text{(dB)} \tag{14.49}$$

where $N_{faded} = k(T_{ant} + T_e)B$, see Chapter 12.

A simple, but conservative, estimate of $(C/N)_{d,\,100-p}$ can be made by assuming, in the 'clear sky' link budget, a (worst case) antenna noise temperature which is equal to the physical temperature of the lossy medium (typically 290 K) and ignoring $(N_{faded}/N_{clear\ sky})_{dB}$ in equation (14.49).

Downlink carrier to interference ratio is usually unaffected by a downlink fade since both the wanted and interfering signals are equally attenuated, i.e.:

$$\left(\frac{C}{I}\right)_{d,\,100-p} = \left(\frac{C}{I}\right)_d \quad \text{(dB)} \tag{14.50}$$

From a system design point of view fade *margins* can be incorporated into the satellite uplink and downlink budgets such that under clear sky conditions the system operates with the correct back-off but with excess uplink and downlink CNR (over those required for adequate overall CNR) of $F_u(p)$ and $F_d(p)$ respectively. Assuming fading does not occur simultaneously on uplink and downlink this ensures that an adequate overall CNR will be available for $100 - 2p\%$ of time. More accurate estimates of the system performance limits imposed by fading would require joint statistics of uplink and downlink attenuation, consideration of changes in back-off produced by uplink fades (including consequent *improvement* in intermodulation noise), allowance for possible cross-polarisation induced crosstalk, hydrometeor scatter and other noise and interference enhancement effects.

EXAMPLE 14.6

A 30 GHz receiving earth station is located near Bradford (54° N, 2° W). It has an overall receiver noise figure of 5.0 dB and a free space link budget which yields a CNR of 35.0 dB. (The free space budget does not account for background gaseous attenuation of the carrier but does allow for normal atmospheric noise.) The station is at a height of 440 m above sea level and the elevation angle to the satellite is 29°. If the one-minute rain rate exceeded for 0.01% of time at the earth station is 28 mm/h estimate the CNR exceeded for 99.9% of time assuming that the CNR is downlink limited and uplink power control is used to compensate all uplink fades.

From Figure 14.35 $A_{zenith} = 0.23$ dB. (This is strictly the value for an earth station at sea level, but since Figure 14.20 shows a specific attenuation *at sea level* of only 0.09 dB/km then the error introduced is less than $0.09 \times 0.44 = 0.04$ dB and can therefore be neglected.)

Clear sky slant path attenuation from equation (14.40):

$$A = \frac{A_{zenith}}{\sin\alpha} = \frac{0.23}{\sin 29^\circ} = 0.5 \text{ dB}$$

$$\left.\frac{C}{N}\right|_{clear\ sky} = \left.\frac{C}{N}\right|_{free\ space} - A$$

$$= 35.0 - 0.5 = 34.5 \text{ dB}$$

From equation (14.41) rain height is:

$$h_R = 4.0 - 0.075\ (54 - 36) = 2.65\ \text{km}$$

Slant path length in rain (equation (14.42)) is:

$$L_s = \frac{2.65 - 0.44}{\sin 29°} = 4.56\ \text{km}$$

Ground projection of path in rain (equation (14.43)):

$$L_G = 4.56 \cos 29° = 3.99\ \text{km}$$

Path length reduction factor (equation (14.44)):

$$r_{0.01} = \left[1 + \frac{3.99\ e^{(0.015 \times 28)}}{35}\right]^{-1} = 0.852$$

Using Figure 14.22 specific attenuation, γ_R, for 30 GHz at 28 mm/h can be estimated to be 5.3 dB/km. (A more accurate ITU-R model using the formula aR^b gives 5.6 dB/km for horizontal polarisation and 4.7 dB/km for vertical polarisation.)

From equation (14.45):

$$A_{0.01} = 5.3 \times 4.56 \times 0.852 = 20.6\ \text{dB}$$

Using equation (14.46):

$$A_{0.1} = 20.6 \times 0.12 \times 0.1^{-(0.546 + 0.043 \log_{10} 0.1)} = 7.9\ \text{dB}$$

Using equations (14.24) and (14.25) and assuming a surface temperature of 290 K:

$$T_A = [(1.12 \times 290) - 50]\ [1 - 10^{-\frac{7.9}{10}}] = 230\ \text{K}$$

Ratio of effective noise powers under faded and clear sky conditions is:

$$\begin{aligned}
\frac{N_{faded}}{N_{clear\ sky}} &= \frac{T_{syst\ faded}}{T_{syst\ clear\ sky}} \\
&= \frac{T_{A\ faded} + (f - 1)\ 290}{T_{A\ clear\ sky} + (f - 1)\ 290} \\
&= \frac{230 + (10^{\frac{5}{10}} - 1)\ 290}{50 + (10^{\frac{5}{10}} - 1)\ 290} \\
&= \frac{230 + 627}{50 + 627} \\
&= 1.266\ =\ 1.0\ \text{dB}
\end{aligned}$$

(Note that the noise enhancement in this case is within the uncertainty introduced by estimating γ_R from Figure 14.22.) From equation (14.49):

$$\left(\frac{C}{N}\right)_{d,\ 99.9} = 34.5 - 7.9 - 1.0\ =\ 25.6\ \text{dB}$$

14.3.5 Analogue FDM/FM/FDMA trunk systems

Figure 14.37 shows a schematic diagram of a large, traditional, earth station. Such an earth station would be used mainly for fixed point-to-point international PSTN communications. The available transponder bandwidth (typically 36 MHz) is subdivided into several transmission bands (typically 3 MHz wide) each allocated to one of the participating earth stations, Figure 14.38. All the signals transmitted by a given earth station, irrespective of their destination, occupy that earth station's allocated transmission band. Individual SSB voice signals arriving from the PSTN at an earth station are frequency division multiplexed (see Figure 5.12) into a position in the earth station's transmission band which depends on the voice signal's destination. Thus all the signals arriving for transmission at earth station 2 and destined for earth station 6 are multiplexed into sub-band 6 of transmission band 2. The FDM signal, consisting of all sub-bands, is then frequency modulated onto the earth station's IF carrier. The FDM/FM signal is subsequently upconverted (U/C) to the 6 GHz RF carrier, amplified (to attain the required EIRP) and transmitted.

A receiving earth station demodulates the carriers from *all* the other earth stations in the network. (Each earth station therefore requires $N - 1$ receivers where N is the number of participating earth stations.) It then filters out the sub-band of each transmission band designated to itself and discards all the other sub-bands. The sub-band signals are then demultiplexed, the resulting SSB voice signals demodulated if necessary (i.e. translated back to baseband) and interfaced once again with the PSTN. This method

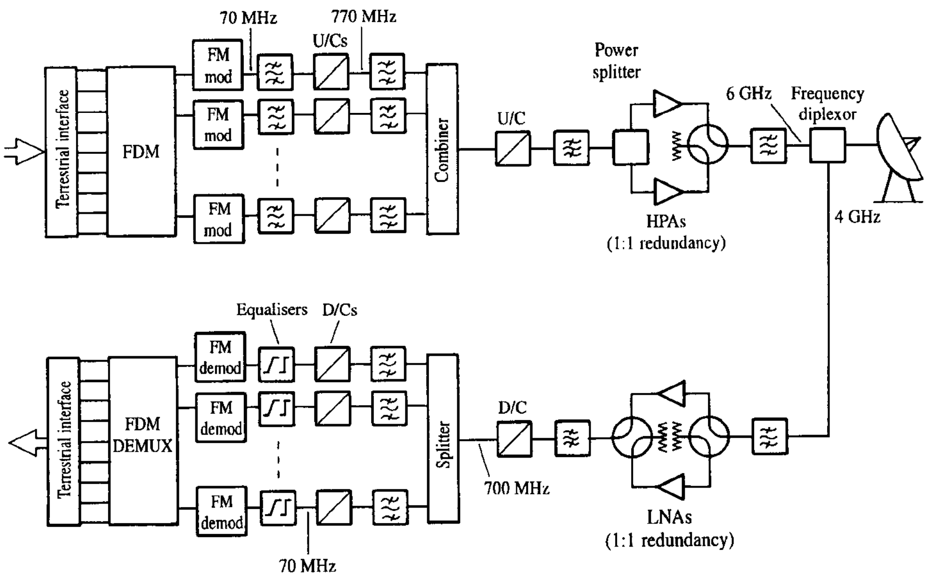


Figure 14.37 *Simplified block diagram of a traditional FDM/FM/FDMA earth station (only HPA/LNA redundancies shown).*

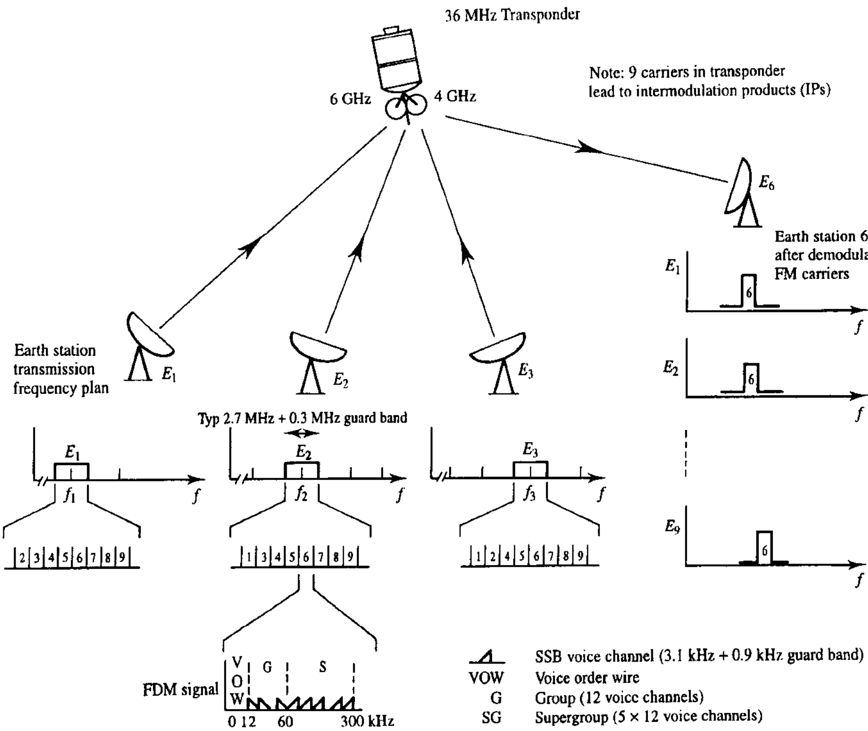


Figure 14.38 *Illustration of MCPC FDM/FM/FDMA single transponder satellite network and frequency plan for the transponder (with nine participating earth stations).*

of transponder resource sharing between earth stations is called frequency division multiple access (FDMA).

When assessing the SNR performance of FDM/FM/FDMA voice systems the detection gain of the FM demodulator must be included. Assuming that operation is at a CNR above threshold, this gain is given by [Pratt and Bostian]:

$$\frac{(S/N_b)_{wc}}{C/N} = \left(\frac{\Delta f_{RMS}}{f_M}\right)^2 \frac{B}{b} \tag{14.51}$$

where:

$(S/N_b)_{wc}$ is the SNR of the worst-case voice channel (see Figure 14.39),
Δf_{RMS} is the RMS frequency deviation of the FM signal,
f_M is the maximum frequency of the modulating (FDM) signal,
B is the bandwidth of the modulated (FM) signal,
b is the bandwidth of a single voice channel (typically 3.1 kHz).

The quantities needed to apply equation (14.51) can be estimated using:

$$f_M = 4.2 \times 10^3 N \quad \text{(Hz)} \tag{14.52}$$

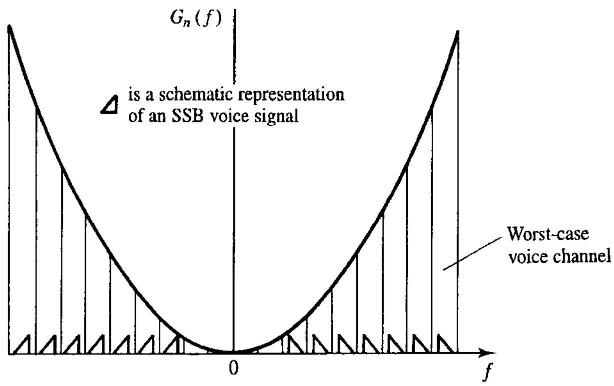


Figure 14.39 *Parabolic noise power spectral density after FM demodulation.*

where N is the number of voice channels in the FDM signal,

$$B = 2(\Delta f + f_M) \quad \text{(Hz)} \tag{14.53}$$

where Δf is the peak frequency deviation of the FM signal, and

$$\Delta f = \begin{cases} 3.16\ \Delta f_{RMS}, & \text{for } N > 24 \\ 6.5\ \Delta f_{RMS}, & \text{for } N \le 24 \end{cases} \tag{14.54}$$

(The FDM signal is usually amplitude limited to ensure the peak to RMS frequency deviation ratio is ≤ 3.16. Since the FDM signal is essentially a sum of many independent voice signals its pdf will be Gaussian, $\Delta f/\Delta f_{RMS} = 3.16$ corresponding to the 0.1% extreme value, Figure 4.22. Clipping will therefore take place for approximately 0.2% of time.) The RMS frequency deviation is set by adjusting the FM modulator constant, K (Hz/V), of the earth station transmitter, i.e.:

$$\Delta f_{RMS} = K\sqrt{\langle g^2(t) \rangle} \tag{14.55}$$

where $g(t)$ is the modulating (FDM) signal. In practice the modulator constant, K, is set using a 1 kHz, 0 dBm test tone as the modulating signal. The test tone RMS frequency deviation, Δf_{RMS}^T, can be related to the required FDM signal frequency deviation, Δf_{RMS}, by:

$$\Delta f_{RMS} = l\ \Delta f_{RMS}^T \tag{14.56}$$

where:

$$20 \log_{10} l = \begin{cases} -1 + 4 \log_{10} N, & 12 \le N \le 240 \\ -15 + 10 \log_{10} N, & N > 240 \end{cases} \tag{14.57}$$

In addition to the FM detection gain given by equation (14.51) an extra SNR gain can be obtained by using a pre-emphasis network prior to modulation and a de-emphasis network after demodulation. Using the ITU pre-emphasis/de-emphasis standards the pre-emphasis SNR gain is 4 dB. Finally, the combined frequency response of a telephone earpiece and the subscriber's ear matches the spectrum of the voice signal better than the spectrum of the noise. This results in a further (if partly subjective) improvement in SNR. This improvement is accounted for by what is called the *psophometric weighting* and has a numerical value of 2.5 dB. Since many voice channels are modulated (as a single FDM signal) onto a single carrier, FDM/FM/FDMA is often referred to as a multiple channel per carrier (MCPC) system. MCPC is efficient providing each earth station is heavily loaded with traffic.

For lightly loaded earth stations MCPC suffers the following disadvantages:

1. Expensive FDM equipment is necessary.
2. Channels cannot be reconfigured easily and must therefore be assigned on essentially a fixed basis.
3. Each earth station carrier is transmitted irrespective of traffic load. This means that full transponder power is consumed even if little or no traffic is present.
4. Even under full traffic load, since an individual user speaks for only about 40% of time, significant transponder resource is wasted.

An alternative to MCPC for lightly loaded earth stations is a single channel per carrier (SCPC) system. In this scheme each voice signal is modulated onto its own individual carrier and each voice carrier is transmitted only as required. This saves on transponder power at the expense of a slightly increased bandwidth requirement. This scheme might be called FM/FDM/FDMA in contrast to the FDM/FM/FDMA process used by MCPC systems. The increased bandwidth per channel requirement over MCPC makes it an uneconomical scheme for traditional point-to-point international trunk applications. The fact that the channels can be demand assigned (DA) as traffic volumes fluctuate, and that the carrier can be switched on (i.e. voice activated) during the 35-40% of active speech time typical of voice signals (thus saving 4 dB of transponder power) makes SCPC superior to MCPC for systems with light, or highly variable, traffic.

Another type of SCPC system dispenses with FM entirely. Compatible single sideband systems simply translate the FDM signal (comprising many SSB voice signals) directly to the RF transmission band (using amplitude modulation). This is the most bandwidth efficient system of all and is not subject to a threshold effect as FM systems are. Compatible single sideband does not, however, have the large SNR detection gain that both FDM/FM/FDMA and FM/FDM/FDMA systems have.

14.3.6 Digital TDM/PSK/TDMA trunk systems

Time division multiplex access (TDMA) is an alternative to FDMA for transponder resource sharing between earth stations. Figure 14.40 illustrates the essential TDMA principle. Each earth station is allocated a time slot (in contrast to an FDMA frequency slot) within which it has sole access to the entire transponder bandwidth. The earth station time slots, or bursts, are interleaved on the uplink, frequency shifted, amplified, and retransmitted by the satellite to all participating earth stations. One earth station

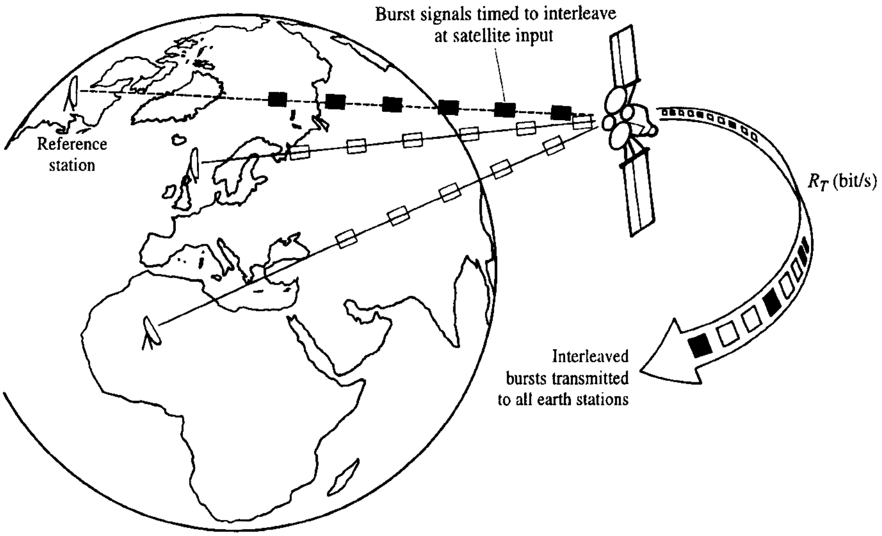


Figure 14.40 *Principal of time division multiplex accessing (TDMA).*

periodically transmits a reference burst in addition to its information burst in order to synchronise the bursts of all the other earth stations in the TDMA system. Time division multiplexing and digital modulation are obvious techniques to use in conjunction with TDMA. In order to minimise AM/PM conversion in the non-linear transponder, Figure 14.33, constant envelope PM is attractive. MPSK is therefore used in preference to MQAM, for example (see Chapter 11). Since some filtering of the PSK signal prior to transmission is necessary (for spectrum management purposes) even MPSK envelopes are not, in fact, precisely constant. QPSK signals, for instance, have envelopes which fall to zero when both inphase and quadrature symbols change simultaneously, Figure 11.31. Offset QPSK (OQPSK) reduces the maximum envelope fluctuation to 3 dB by offsetting inphase and quadrature symbols by half a symbol period (i.e. one information bit period), Figure 11.31. Chapter 11 discusses bandpass modulation (including OQPSK) in detail. Figure 14.41 shows a schematic diagram of a TDM/PSK/TDMA earth station.

For digital satellite systems having only a single carrier present in the transponder at any one time then intermodulation products are absent and $(C/N)_{IP}^{-1}$ in equation (14.38) is zero. Recall (Chapter 11) that the quantity E_s/N_0 is related to C/N by:

$$\frac{\langle E_s \rangle}{N_0} = \frac{C}{N} BT_o \tag{14.58}$$

where T_o is the symbol period (i.e. the reciprocal of the baud rate, R_s) and BT_o depends on the particular digital modulation scheme and filtering employed. Equation (14.58) can also be expressed in terms of bit energy, E_b, and bit duration, T_b, i.e.:

$$\frac{E_b}{N_0} = \frac{C}{N} BT_b \tag{14.59}$$

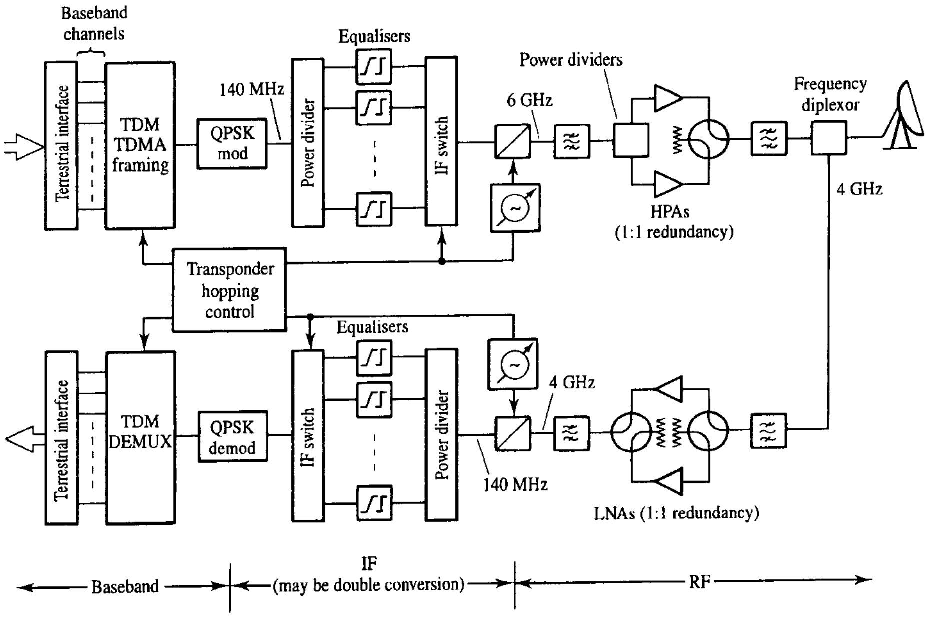


Figure 14.41 *Simplified block diagram of traditional TDM/QPSK/TDMA earth station. (Only HPA/LNA redundancies are shown.)*

For QPSK modulation, which is the primary TDM/PSK/TDMA modulation standard currently used by INTELSAT, the E_b/N_0 required to support a given P_b performance is found using equation (11.47):

$$P_b = \frac{1}{2}\,\mathrm{erfc}\left(\frac{E_b}{N_0}\right)^{1/2} \tag{14.60}$$

and the required CNR is then found using equations (11.6) and (11.7):

$$\frac{C}{N} = \frac{E_b/T_b}{N_0 B} = \frac{E_b R_b}{N_0 B} \tag{14.61}$$

or in decibels:

$$\left(\frac{C}{N}\right)_{dB} = \left(\frac{E_b}{N_0}\right)_{dB} - (BT_b)_{dB} \tag{14.62}$$

(For minimum bandwidth, ISI free, filtering such that the transmitted signal occupies the DSB Nyquist bandwidth then BT_b = 1.0 (or 0 dB) and C/N = E_b/N_0.) In practice an implementation margin of a few decibels would be added after using equations (14.60) and (14.61) to allow for imperfect modulation, demodulation, etc.

A typical frame structure showing the TDMA slots allocated to different earth stations is shown in Figure 14.42. Two reference bursts are often included (provided by different

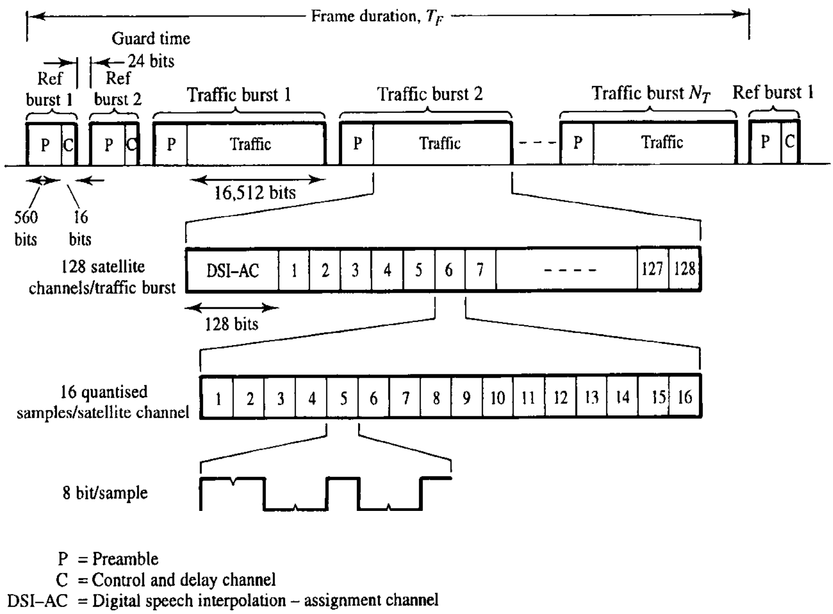


Figure 14.42 *Typical TDMA frame structure. (DSI-AC time slot is discussed in section 14.3.7.)*

earth stations) so that the system can continue to function in the event of losing one reference station due, for example, to equipment failure. Typically the frame period T_F is of the order of 5 ms.

The traffic bursts each consist of a preamble followed by the subscriber traffic. The preamble, which might typically be 280 QPSK symbols long, is used for carrier recovery, symbol timing, frame synchronisation and station identification. In addition it supports voice and data channels (voice order wires) to enable operations and maintenance staff at different earth stations in the network to communicate without using traffic slots. Reference bursts have the same preamble as the traffic bursts followed by control and delay signals (typically 8 symbols in duration) which ensure that the TDMA bursts from participating earth stations are timed to interleave correctly at the transponder input.

Subscriber traffic is subdivided into a number (typically 128) of *satellite channels*, Figure 14.42. In conventional preassigned (PA) systems the satellite channels are predivided into groups, each group being assigned to a given destination earth station.

Each satellite channel carries (typically) 128 bits (64-QPSK symbols) representing 16 consecutive 8-bit PCM samples from a single voice channel. For a conventional 8 kHz PCM sampling rate this corresponds to $16 \times 125 \mu s = 2$ ms of voice information. In this case the frame duration would therefore be limited in length to 2 ms so that the next frame could convey the next 2 ms of each voice channel. Channels carrying non-voice, high data rate, information are composed of multiple voice channels. Thus, for example, a 320 kbit/s signal would require five 64 kbit/s voice channels.

Frames may be assembled into master frames as shown in Figure 14.43. The relative lengths of information bursts from each earth station can then be varied from master frame to master frame depending on the relative traffic loads at each station. This would represent a simple demand assigned system.

The frame efficiency, η_F, of a TDM/PSK/TDMA system is equal to the proportion of frame bits which carry revenue earning traffic, i.e.:

$$\eta_F = \frac{b_F - b_o}{b_F} \tag{14.63}$$

where b_F is the total number of frame bits and b_o is the number of overhead (i.e. non-revenue earning) bits. The total number of frame bits is given by:

$$b_F = R_T T_F \tag{14.64}$$

where R_T is the TDMA bit rate and T_F is the frame duration. The number of overhead bits per frame can be calculated using:

$$b_o = N_R b_R + N_T(b_p + b_{AC}) + (N_R + N_T)b_G \tag{14.65}$$

where N_R is the number of participating reference stations, N_T is the number of participating traffic stations, b_R is the number of bits in a reference burst, b_P is the number of preamble bits (i.e. all bits excluding traffic bits) in a traffic burst, b_{AC} is the number of digital speech interpolation–assignment channel bits per traffic burst and b_G is the number of guard bits per reference or traffic burst.

Typically η_F is about 90% for a TDMA system with 15 to 20 participating earth stations and a frame period of several milliseconds.

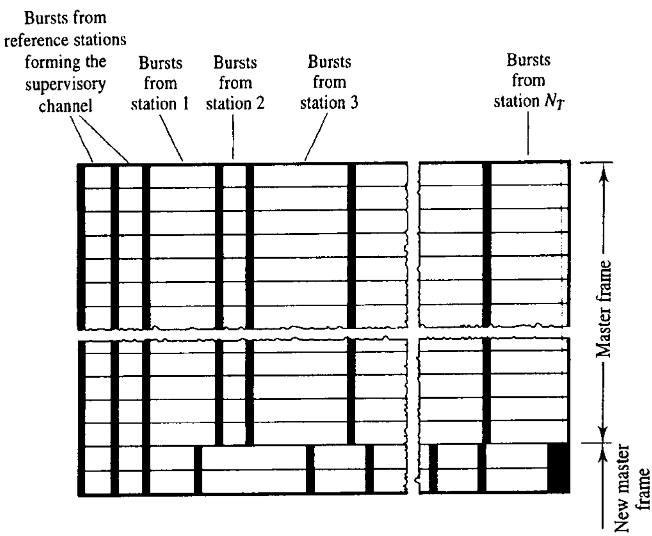


Figure 14.43 *TDMA master frame structure.*

The number of voice channels which a TDMA system can support is called its voice channel capacity, χ. This can be calculated from:

$$\chi = \frac{R_i}{R_v} \tag{14.66}$$

where R_i is the information bit rate and R_v is the bit rate of a single voice channel.

The information bit rate is given by:

$$R_i = \frac{b_F - b_o}{T_F} = \eta_F R_T \quad \text{(bit/s)} \tag{14.67}$$

and the bit rate of a single voice channel is given by:

$$R_v = f_s n \quad \text{(bit/s)} \tag{14.68}$$

where f_s is the sampling rate (usually 8 kHz) and n is the number of PCM bits per sample (typically 8), Chapter 6. (For ADPCM, R_v is 32 kbit/s, Table 9.3.) Finally, the average number of voice channels *per earth station access* $\bar{\chi}_A$ is:

$$\bar{\chi}_A = \frac{\chi}{N} \tag{14.69}$$

where N is the number of accesses per frame. (If all earth station only access once per frame then N is also, of course, the number of participating earth stations.)

The primary TDM/PSK/TDMA modulation standard used by INTELSAT is summarised in Table 14.5.

Table 14.5 *INTELSAT TDM/PSK/TDMA modulation standard.*

Modulation	QPSK
Nominal symbol rate	60.416 Mbaud
Nominal bit rate	120.832 Mbit/s
Encoding	Absolute (i.e. not differential phase)
Phase ambiguity resolution	Using unique word in preambles

EXAMPLE 14.7

The frame length of the pure TDMA system illustrated in Figure 14.42 is 2.0 ms. If the QPSK symbol rate is 60.136 Mbaud and all traffic bursts are of equal length determine: (i) the maximum number of earth stations which the system can serve and (ii) the frame efficiency.

$$R_T = 2 \times \text{QPSK baud rate} = 2 \times 60.136 \times 10^6 = 1.20272 \times 10^8 \text{ bit/s}$$

$$b_F = R_T T_F = 1.20272 \times 10^8 \times 2.0 \times 10^{-3} = 240544 \text{ bits}$$

Overhead calculation:

$$b_o = N_R b_R + N_T (b_p + b_{AC}) + (N_R + N_T) b_G$$

Now for a 560 + 16 bit reference, b_R, 560 bit preamble, b_P, 24 bit guard interval, b_G and 128 bit DSI assignment channel:

$$b_o = 2\,(560 + 16) + (560 + 128)\,N_T + (2 + N_T)\,24$$

$$= 1200 + 712\,N_T$$

$$b_F - b_o = 240544 - (1200 + 712\,N_T) = 239344 - 712\,N_T$$

(i)

$$N_T = \frac{b_F - b_o}{16512 - 128} = \frac{239344 - 712\,N_T}{16384}$$

i.e.:

$$N_T\,(16384 + 712) = 239344$$

Therefore: $N_T = 14$, i.e. a maximum of 14 earth stations may participate.

(ii)

$$\eta_F = \frac{b_F - b_o}{b_F}$$

$$= \frac{240544 - (1200 + 712 \times 14)}{240544} = 0.954$$

i.e. frame efficiency is 95.4%.

14.3.7 DA-TDMA, DSI and random access systems

Preassigned TDMA (PA-TDMA) risks the situation where, at a certain earth station, all the satellite channels assigned to a given destination station are occupied whilst free capacity exists in channels assigned to other destination stations. Demand assigned TDMA (DA-TDMA) allows the reallocation of satellite channels in the traffic burst as the relative demand between earth stations varies. In addition to demand assignment of satellite channels within the earth station's traffic burst DA-TDMA may also allow the number traffic bursts per frame, and/or the duration of the traffic bursts, allocated to a given earth station to be varied.

Digital speech interpolation (DSI) is another technique employed to maximise the use made of available transponder capacity. An average speaker engaged in conversation actually talks for only about 35% of the time. This is because for 50% of time he, or she, is passively listening to the other speaker and for 30% of the remaining 50% of time there is silence due to pauses and gaps between phrases and words. DSI systems automatically detect when speech is present in the channel, and during speech absences reallocate the channel to another user. The inevitable clipping at the beginning of speech which occurs as the channel is being allocated is sufficiently short for it to go unnoticed.

Demand assigned systems require extra overhead in the TDMA frame structure to control the allocation of satellite channels and the relative number per frame, and lengths, of each earth station's traffic bursts. For systems with large numbers of earth stations each contributing short, bursty, traffic at random times then random access (RA) systems may use transponder resources more efficiently than DA systems. The earth stations of RA systems attempt to access the transponder (i.e. in the TDMA context, transmit bursts) essentially at will. There is the possibility of course, that the traffic bursts (usually called packets in RA systems) from more than one earth station will collide in the transponder causing many errors in the received data. Such collisions are easily detected, however, by both transmitting and receiving earth stations. After a collision all the transmitting earth stations wait for a random period of time before retransmitting their packets.

Many variations and hybrids of the multiple access techniques described here have been used, are being used, or have been proposed, for satellite communications systems. A more detailed and quantitative discussion of these techniques and their associated protocols can be found in [Ha].

14.3.8 Economics of satellite communications

The cost of a long distance point-to-point terrestrial voice circuit is about 2000 dollars p.a. Lease of a private, high quality, landline from New York to Los Angeles for FM broadcast use costs 13,000 dollars p.a. Satellite communication systems can provide the equivalent services at lower cost. The monthly lease for a video bandwidth satellite transponder is typically 50,000 to 200,000 dollars whilst the hourly rate for satellite TV programme transmission can be as low as 200 dollars. Access to digital audio programmes can be as low as 2,500 dollars per month for high quality satellite broadcasts. The advent of the space shuttle has greatly reduced satellite launch costs below the 30,000 dollars/kg of conventional rocket techniques making them much more competitive. (The figures given here reflect 1990 costs.)

14.3.9 VSAT systems

Satellite communication is predominantly used for international, point-to-point, multiplex telephony, fax and data traffic, making high EIRP transmissions necessary due to the wideband nature of the signal multiplex. For low data rates (2.4 to 64 kbit/s) with narrow signal bandwidth the reduced noise in the link budget allows the use of satellites with very small aperture (typically 1 m diameter antenna) terminals (VSATs) and modest power (0.1 to 10 W) earth station transmitters [Everett]. This has given rise to the development of VSAT low data rate networks, in which many remote terminals can, for example, access a central computer database. These systems use roof, or garden, located antennas which permanently face a geostationary satellite. VSAT systems are widely deployed in the USA and may have up to 10,000 VSAT earth station terminals in a single network. They are used for retail point-of-sale credit authorisation, cash transactions, reservations, stock control and other data transfer tasks. They are also now being expanded to give worldwide coverage in order to meet the telecommunications requirements of large international companies.

The use of non-geostationary satellites for mobile and personal communications with even lower antenna gains is discussed in Chapter 15.

14.3.10 Satellite-switched-TDMA and on-board signal processing

Satellites operating with small spot beams have high antenna gains. This implies either a low on-board power requirement or a large bandwidth and therefore high potential bit rate. If many spot beams with good mutual isolation are used, frequency bands can be reused thus increasing spectrum utilisation efficiency, Figure 14.44. Connectivity between a system's participating earth stations is potentially decreased, however, since a pair of earth stations in different spot beams can communicate only if their beams are connected. Satellite switched TDMA has the potential to re-establish complete connectivity between earth stations using a switching matrix onboard the satellite, Figure 14.45. The various sub-bursts (destined for different receiving stations) of a transmitting station's traffic burst can be directed by the matrix switch to the correct downlink spot beams. Furthermore, for areas with a sparse population of users, such that many fixed spot beams are uneconomic, the beams may be hopped from area to area and the uplink bursts from each earth station demodulated and stored. On-board signal processing is then used to reconfigure the uplink bursts into appropriately framed downlink bursts before the signals are remodulated and transmitted to the appropriate earth stations as the downlink spot beam is hopped. On-board demodulation and remodulation also has the normal advantage of digital communications, i.e. the uplink and downlink noise is decoupled. The NASA advanced communications satellite was used in the middle 1990s to evaluate these types of system.

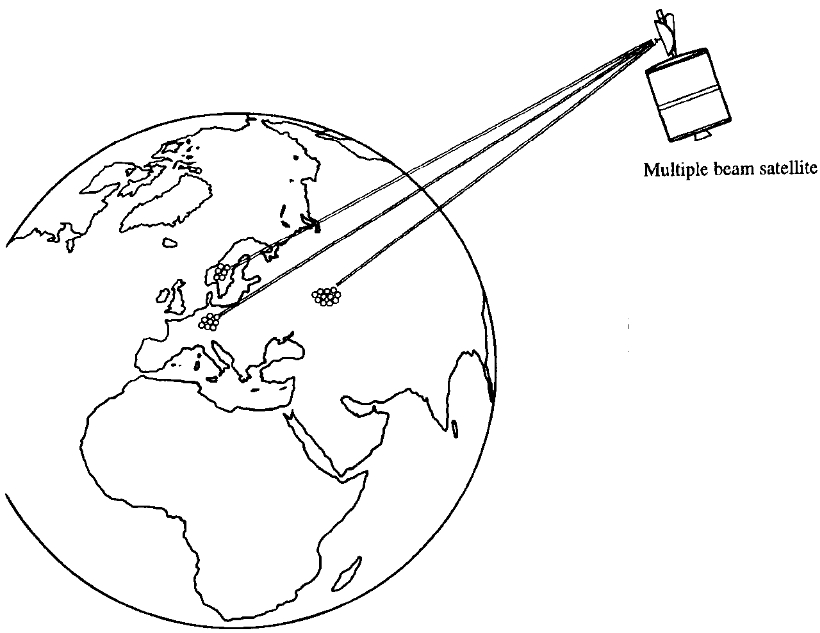


Figure 14.44 *Satellite switched time division multiplex access (SS-TDMA).*

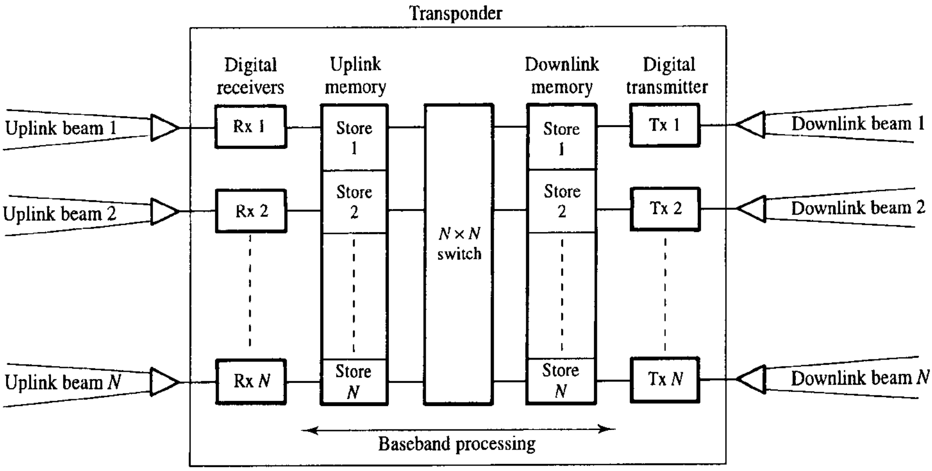


Figure 14.45 *SS-TDMA transponder.*

14.4 Summary

Point-to-point, terrestrial, microwave links now play a major part in the arterial trunk routes of the PSTN. Originally carrying SSB/FDM/FM telephony they now carry mixed services, principally using PCM/TDM/QAM. Typically, repeaters are spaced at 40 to 50 km intervals and the total microwave link bandwidth is split in to high and low frequency blocks, each of which contains eight 30 MHz channels. Adjacent channel isolation is improved by alternating vertical and horizontal polarisations across the transmission band. Correlation between the expected signal and interference due to overreaching caused by ducting, is avoided by alternating transmit and receive channels between high and low channel blocks on adjacent hops.

Proper path profiling of LOS microwave links ensures adequate clearance under specified propagation conditions. If clearance is less than the first Fresnel zone then serious diffraction losses may occur. Propagation paths may be conveniently plotted as straight rays over an earth with appropriately modified curvature, or as rays with modified curvature over a plane earth. Bulk refractive index conditions are described by k factor which is the ratio of effective earth radius to actual earth radius in the straight ray model. Standard propagation conditions correspond to a k factor of 4/3 (which represents a refractivity lapse rate of 40 N units/km). Positive $k < 4/3$ represents sub-refractive conditions and $k > 4/3$ represents super-refraction. Negative k represents ducting.

For accurate clear sky link budgets, background gaseous absorption must be accounted for in addition to free space path loss. Gaseous absorption is small below 10 GHz but rises rapidly with frequency due primarily to water vapour and oxygen resonance lines which occur at 22 GHz and 60 GHz respectively. Rain attenuation also increases rapidly with frequency and must be accounted for if links are to meet a specified availability. Multipath propagation may, like rain, result in deep signal fades but, unlike rain fading which is essentially flat, can also cause amplitude and phase distortion to wide band signals. Microwave paths which suffer from severe frequency selective fading can benefit significantly from the application of diversity techniques and/or the use of adaptive equalisation.

Hydrometeor scatter may result in interference between nominally independent cofrequency systems. Hydrometeor induced cross polarisation may also result in cross talk between the channels of systems using polarisation, in whole or in part, as the channel isolating mechanism.

Fixed service satellite communications makes use of the geostationary orbit. Satellites in this orbit are located above the earth's equator at a height of 36,000 km. They are stationary with respect to a point on the earth's surface and tracking is therefore unnecessary. A single geostationary satellite is visible from 42% of the earth's surface. With the exception of the polar regions three such satellites can give global coverage. The combined propagation delay of uplink and downlink is about 0.25 s making echo cancellation essential. FSPL is of the order of 200 dB making large transmitter powers, large antennas and sensitive receivers necessary for high data rate transmissions. Like terrestrial microwave paths gaseous absorption and rain fading must be considered when engineering an earth-space communications link. Scintillation on earth-space paths might also be important for low margin, low availability systems. Fixed point satellite services currently operate in the C and Ku bands although services using the Ka band will probably follow shortly.

The predominant multiplexing/modulation/multiple accessing technique used currently for international PSTN satellite telephony is FDM/FM/FDMA. Since satellite HPAs are operated near saturation the use of FDMA, which results in multiple FM carriers in the transponder, gives rise to large intermodulation products which must normally be accounted for in the overall CNR calculation. As with terrestrial LOS relay links, the trend in satellite communications is towards digital transmission. FDM/FM/FDMA systems will therefore gradually be replaced by TDM/PSK/TDMA systems. Intermodulation products will be of less concern since in (pure) TDMA systems only one carrier is present in the transponder at any particular time. TDMA systems require less back-off, therefore, than the equivalent FDMA systems. This means that constant envelope modulation is even more desirable to minimise AM/AM and AM/PM distortion. QPSK and OQPSK are attractive modulation techniques in these circumstances.

The bits transmitted by an earth station in a TDMA network are grouped into frames (and frames into master frames). Each frame carries overhead bits which serve various synchronisation, monitoring and control purposes. The frame efficiency is defined as the proportion of frame bits which carry (revenue earning) traffic.

For systems which have fluctuating traffic loads demand assignment may be used to reallocate capacity to those earth stations with highest demand. For systems with large numbers of lightly loaded earth stations, random access techniques may be appropriate.

VSAT systems can use modest transmitter powers and small antennas to transmit low bit rate (and therefore narrow bandwidth) data. Such technology is relatively inexpensive allowing commercial (and other) organisations to operate their own networks of earth stations. Switched spot-beam satellites with on-board signal processing, and satellite based mobile communications systems using both high and low earth orbits (HEOs and LEOs) are currently being developed.

14.5 Problems

14.1. Show that the effective height of the (smooth) earth's surface above a cord connecting two points on the surface, for a straight ray profile, is given (for practical microwave link geometries) by $d_1 d_2/(2ka)$, where k is k-factor and a is the earth radius. Hence confirm the correctness of equations (14.11) and (14.12(a)).

14.2. A 10.3 GHz microwave link with good ground clearance is 60 km long. The link's paraboloidal reflector antennas have a diameter of 2.0 m and are aligned to be correctly pointed during standard ($k = 4/3$) atmospheric conditions. What limiting k-factors can the link tolerate before 6 dB of power is lost from its signal budget due to refractive decoupling of the antennas? To what refractivity lapse rates do these k-factors correspond and in what regime (sub-refraction/super-refraction/ducting) do these lapse rates reside? (Assume that the antennas are mounted at precisely the same heights above sea level and that their 3 dB beamwidths are given by $1.2\lambda/D$ rad where λ is wavelength and D is antenna diameter.)

14.3. If the transmitting antenna in Problem 14.2 is 50 m high and is aligned under standard refractive index conditions such that its boresight is precisely horizontal, what must be the difference between transmit and receive antenna heights for the receive antenna to be located at the centre of the transmitted beam? [158.9 m]

14.4. Show that the radius of the nth Fresnel zone is given by equation (14.10).

14.5. A microwave LOS link operates at 10 GHz over a path length of 32 km. A minimum ground clearance of 50 m occurs at the centre of the path for a k-factor of 0.7. Find the Fresnel zone clearance of the path in terms of (a) the number of Fresnel zones cleared and (b) the fraction (or multiple) of first Fresnel zone radius. [10, 3.2]

14.6. A 6 GHz LOS link has a path length of 13 km. The height of both antennas above sea level at each site is 20 m. The only deviation of the path profile from a smooth earth is an ancient earth works forming a ridge 30 m high running perpendicular to the path axis 3 km from the receiver. If the transmitted power is 0 dBW, the transmit antenna has a gain of 30 dB and the receive antenna has an effective area of 4.0 m^2 estimate the power at the receive antenna terminals assuming knife edge diffraction. By how much would both antennas need to be raised if 0.6 first Fresnel zone clearance were required? (Assume knife edge diffraction and no atmospheric losses other than gaseous background loss.) [–74 dBW]

14.7. A terrestrial 20 GHz microwave link has the following specification:

Path length	35.0 km
Transmitter power	–3.0 dBW
Antenna gains	25.0 dB
Ground clearance	Good (no ground reflections or diffraction)
Receiver noise figure	3.0 dB
Receiver noise bandwidth	100 kHz

Estimate the effective CNR exceeded for 99.9% of time if the 1 minute rain rate exceeded for 0.1% of time is 10 mm/h. (Assume that Figure 14.21 applies to the climate in which the link operates and that the surface temperature is 290 K.) [30.4 dB]

14.8. A geostationary satellite is located at 35° W. What are its look angles and range from: (a) Bradford, UK (54° N, 2° W); Blacksburg, USA (37° N, 80° W); (c) Cape Town, South Africa (34° S, 18° E)? [22°, −140°, 39406 km; 27°, 121°, 38914 km, 22°, 292°, 39365 km]

14.9. A geostationary satellite sits 38,400 km above the earth's surface. The zenith pointing uplink is at 6 GHz and has an earth station power of 2 kW with $G_T = 61$ dB. If satellite antenna gain $G_R =$ 13 dB, calculate the power in mW at the satellite receiver. Calculate the effective carrier-to-noise ratio, given that the receiver bandwidth is 36 MHz, its noise figure F is 3 dB and the thermal noise level is −174 dBm/Hz. [6.2×10^{-7} mW, 33.3 dB]

14.10. A receiver for geostationary satellite transmission at 2 GHz has an equivalent noise temperature of 160 K and a bandwidth of 1 MHz. The receiving antenna gain is 35 dB and the antenna noise temperature is 50 K. If the satellite antenna gain is 10 dB and expected total path losses are 195 dB, what is the minimum required satellite transmitter power to achieve a 20 dB CNR at the output of the receiver? [24.6 dBW]

14.11. Ship–shore voice communications are conducted via an INTELSAT IV satellite in geostationary orbit at an equidistant range from ship and shore stations of 38,400 km. One 28 kbit/s voice channel is to be carried in each direction. The shipboard terminal contains a 2.24 m diameter parabolic antenna of 65% efficiency. An uncooled low-noise receiver tuned to 6040 MHz, having a bandwidth of 28 kHz and a system noise temperature of 150 K, is mounted directly behind the antenna. The satellite EIRP in the direction of the ship is 41.8 dBm. The shore based terminal has an antenna of gain 46 dB at the satellite–shore link frequency of 6414.6 MHz. The satellite EIRP in the direction of the shore is 31.8 dBm. The shore's receiver has a bandwidth of 28 kHz and a noise temperature of 142 K. Calculate the carrier-to-noise ratios in dB for reception at the ship and the shore from the two satellite downlinks. [15 dB, 10.1 dB]

14.12. A 14/11 GHz digital satellite, transparent transponder, communications link has the following specification:

Uplink saturating flux density	−83.2 dBW/m^2
Input back-off	8.0 dB
Satellite G/T	1.8 dB/K
Transponder bandwidth	36.0 MHz
Uplink range	41,000 km
Saturated satellite EIRP	45.0 dBW
Output back-off	2.8 dB
Downlink range	39,000 km
Earth station G/T	31.0 dB/K
Downlink earth station fixed losses	3.5 dB

If the uplink and downlink earth station elevation angles are 5.6° and 20.3° respectively estimate the overall link, clear sky, CNR. (Assume pure TDMA operation such that there is no intermodulation noise, and assume that interference is negligible.) What clear sky earth station EIRP is required on the uplink? [15.3 dB, 72.8 dBW]

14.13. If uplink power control is used to precisely compensate uplink fading in the system described in Problem 14.12 and the downlink earth station is located in Bradford (54° N, 2° W) at a height of 440 m, where the 1 minute rain rate exceeded for 0.01% of time is 25 mm/h, estimate the overall CNR exceeded for the following time percentages: (a) 0.1%; (b) 0.01%; (c) 0.001%. (Assume that under clear sky conditions one third of the total system noise can be attributed to the aperture temperature of the antenna and two thirds originates in the receiver.) What would be the maximum possible ISI free bit rate if the modulation is QPSK and what values of BER would you expect to be exceeded for 0.1, 0.01, and 0.001% of time? [12.8 dB; 7.1 dB; −3.9 dB, 459 error/s;

8.5 kerror/s; 35.7 Merror/s]

14.14. A particular satellite communication system has the following TDMA frame structure:

Single reference burst containing 88 bits.
Preamble to each traffic burst containing 144 bits.
Frame duration of 750 μs.
Guard time after each burst of 24 bits duration.
Overall TDMA bit rate of 90.389 Mbit/s.

10 earth stations are each allocated 2 traffic bursts per frame, and one station provides (in addition) the single reference burst. What is the frame efficiency assuming DSI is not employed? If the satellite were used purely for standard (64 kbit/s) PCM voice transmission, what would be the TDMA voice channel capacity of the system? How many consecutive samples from each voice channel must be transmitted per frame? [94.9%, 1340, 6]