Chapter

1

Fundamental Parameters

1.1 Overview

Communication satellite systems depend significantly on both space segment and user segment antenna designs. Space segment antennas must meet their performance requirements over their specified coverage areas with allowance for satellite attitude variations. User segment antennas likewise must meet their performance requirements while tracking the satellite in orbit. Antenna requirements depend on specific program needs, and a significant diversity of technology has been developed to accommodate the diverse objectives of individual programs. As a result, space segment antennas are the most diverse technology in the space segment, and specific designs for one application cannot be applied to other applications. User segment antenna hardware likewise exhibits a wide variety of antenna hardware ranging from small handheld technologies to much larger ground terminal antennas, which are often associated in the public's mind with communication satellite antenna systems. A review of the system parameters used to quantify antenna performance is presented as a basis for subsequent chapters.

1.2 Antenna Parameters

Antenna parameters must describe both the spatial characteristics and terminal interfaces with system electronics. The spatial characteristics specify the two-dimensional description of the antenna's sensitivity variations in a coordinate system embedded in the antenna. These spatial characteristics must also indicate the antenna's polarization properties that define the orientation of the electric field during one RF (radio frequency) cycle. The antenna's terminal impedance quantifies the interface relations with system electronics. Satellite system antennas are commonly in the class of aperture antennas. The relationship between the aperture size and spatial characteristics is a most important issue in system sizing. This relationship dictates the antenna's gain levels and beamwidth requirements. Perhaps the most commonly asked question regarding antennas is the size required to meet system requirements. This question is typically followed by a request to explain why the size must be that large. Noise in receiving systems is an important system parameter and is characterized by the antenna noise temperature at the antenna's terminal. The antenna noise temperature added to the receiver noise temperature equals the total system noise temperature, an important factor in the performance of receiving antennas.

1.2.1 Spatial Characteristics

The spatial characteristics describe the spatial variation of the antenna's sensitivity. They also describe the vector nature of the antenna's field distribution in a coordinate system referenced to the antenna's structure. Commonly, satellite systems use aperture antennas that have a distribution of fields in the aperture and a corresponding distribution of fields in space. The coordinate system used for this specification generally places the aperture plane with the XY plane as indicated in Fig. 1-1. At a sufficient distance from the aperture (referred to as the antenna's far field), the variation of the fields becomes invariant with the range from the antenna's aperture. The electric field quantities, E_θ and E_φ, are orthogonal to one another as specified and vary with separation R from the aperture as 1/R. The power density in the far field is proportional to $(|E_\theta|^2 + |E_\varphi|^2)/Z_o$, where Z_o equals 120π and is the free space impedance.

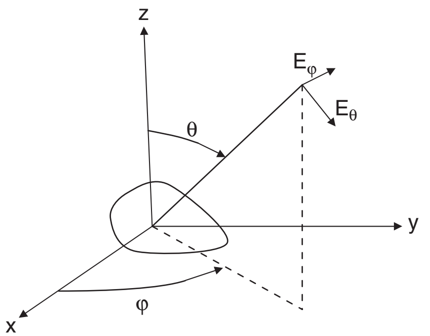

Figure 1-1 Coordinate system for aperture antennas

The relationship between the fields in an antenna's aperture to the spatial distribution is the radiation integral [1]

$$g(k_x, k_y) = \iint F(x, y) \exp(j\ (k_x x + k_y y)\ \mathrm{dx\ dy}$$

where $g(k_x, k_y)$ is the pattern (voltage), $F(x, y)$ is the field distribution in the aperture having coordinates x and y, the integration limits are the physical extent of the aperture, and

$$k_x = k \sin \theta \cos \varphi$$

$$k_y = k \sin \theta \sin \varphi$$

where k is the free space wavenumber equal to $2\pi/\lambda$, λ is the wavelength equal to c/f where c is the speed of light, and f is the RF frequency. The aperture fields are vector functions representing the polarization properties of the aperture fields. The variation of the antenna's sensitivity with direction is referred to as its pattern, and $g(k_x, k_y)$ is proportional to the electric field variation. This relation assumes the spatial fields are sufficiently separated from the aperture that the fields are independent of the range, a condition referred to as the far field. Commonly, the required far field separation for aperture antennas is taken as $2D^2/\lambda$, where D is the aperture width. It should also be noted that antennas generally satisfy reciprocity relations so that at the same frequency, the characteristics are identical independent of whether the antenna is transmitting or receiving. The exception is when the antenna incorporates nonreciprocal devices such as active amplifiers.

The relation between the aperture fields and the far field pattern is a two-dimensional Fourier transform. Similarly, the aperture field is the inverse two-dimensional Fourier transform of the far field pattern. The antenna size is thus related to the beamwidth in the far field, and likewise the beamwidth in the far field is related to the antenna size through the transform. The familiar properties of Fourier transforms are inherent in antenna design. If the aperture fields have an amplitude taper, the far field beamwidth broadens and the sidelobes surrounding the main beam are reduced. If the aperture fields are in phase over the extent of the aperture, the beam maximum is normal to the aperture plane. If the aperture fields have a linear phase gradient over the extent of the aperture, the beam maximum is normal to the phase gradient, a consequence of the familiar shifting theorem of Fourier transforms.

Antenna gain measures the antenna's ability to transfer or receive signals in a particular direction. It is referenced to an idealized lossless antenna having uniform sensitivity in all directions. In a sense, this reference for antenna gain follows the definition of electronics gain that is referenced to the transfer response of an idealized, lossless "straight wire." The maximum value of antenna gain for aperture designs equals

$$G = \eta\,(4\pi A/\lambda^2)$$

where η is the antenna efficiency (≤ 1), A is the physical area of the antenna's aperture, and λ is again the free space wavenumber. Ideally, an antenna having 100% efficiency is lossless and has an aperture distribution uniform in both amplitude and phase. Practically, this ideal antenna efficiency can only be approached, and the antenna efficiency of practical antenna designs falls short of the ideal value because of ohmic and impedance mismatch losses, the aperture amplitude and phase deviations from the ideal, and scattering and blockage from the antenna's structure. In determining the required antenna size or aperture area, an estimated value of the antenna's efficiency is required. The efficiency value depends on the specific antenna design.

Another term defined for receiving antennas is effective aperture, which equals

$$A_e = (\lambda^2/4\pi)\,G$$

The received power equals the product of the incident power density and the effective aperture.

The far field parameters implicitly assume the antenna responds to an incident plane wave or a wave that approximates a plane wave. The far field criteria $2D^2/\lambda$ is derived based on the required range from the point of origin of a spherical wave such that the phase deviation over a planar surface of dimension D has a maximum value of 22.5° relative to an ideal in-phase plane wave.

Directivity or *directive gain* is another term that characterizes an antenna's directional properties. Directivity is a function of the antenna pattern or the variation of the antenna's sensitivity to different signal directions. Directivity differs from antenna gain because ohmic and mismatch losses are not included in directivity. Thus, antenna gain has a lower value than directivity. Directivity is defined by

$$D(\theta, \varphi) = 4\pi P(\theta, \varphi)/\iint (P(\theta, \varphi) \sin\theta \; d\theta d\varphi$$

The integral in the denominator is total power radiated or received from all directions. The fields of an antenna are vector quantities and (as will be discussed) have a principal polarized component with the design polarization state and, unavoidably, a cross-polarized component that is orthogonal to the principal polarization. Directivity is generally computed with the power pattern in the numerator limited to principal polarized fields and the total power in the denominator comprised of both principal and cross-polarized terms. In this way, the directivity is determined relative to the design polarization of the antenna.

Antenna gain defines the signal power transfer and varies with angular coordinates. The antenna's beamwidth describes the angular width of the antenna's maximum response and is defined by the angular extent of the pattern within 3 dB of the peak antenna gain value or the HPBW, half-power beamwidth. The beamwidth of practical antennas can vary depending on which plane of the antenna pattern is used. Commonly, principal plane patterns display the patterns in the XZ and YZ planes in Fig. 1-1 when the beam maximum is coincident with the Z-axis. These patterns are great circle cuts through the sphere surrounding the antenna. When the beam is not coincident with the Z-axis, great circle cuts that intersect the peak gain level of the antenna are used. Depending on requirements, the patterns in other planes, also, great circle cuts are taken and sometimes referred to as φ cuts. When φ equals 45° or 135°, the patterns are referred to as diagonal cuts. Generally, multiple pattern cuts are used to judge the symmetry of the antenna's pattern. For aperture antennas, the beamwidth, θ_{hp}, equals

$$\theta_{hp} = K\lambda/D$$

where K is a constant that depends on the aperture distribution, λ is the wavelength, and D is the aperture width.

The parameters and their variation are illustrated by a simple analytic model. A circular aperture is assumed to have a uniform phase distribution and a rotationally symmetric amplitude having a $(1 - r^2)^p$ variation, where r is the aperture's radius. Example characteristics of this family of distributions are given in Table 1-1 where $J_{P+1}(x)$ is the Bessel function of order $p + 1$, and X equals $(\pi \mathrm{D}\lambda) \sin \theta$ with D equal to the aperture's diameter. When p equals 0, the amplitude distribution, like the phase distribution, is uniform over the aperture. The uniform aperture distribution has the maximum efficiency, a beamwidth factor of 58 in degrees, and a first sidelobe level that is 17.6 dB lower than the peak gain level. As the value of p increases, the efficiency decreases, the beamwidth broadens, and the sidelobe level decreases, all very familiar consequences of the Fourier transform relation between the

TABLE 1-1 Amplitude Taper Effects for Circular Apertures

P	Efficiency Loss, dB	Beamwidth Factor, K, degrees	First Sidelobe Level, dB	Pattern Variation
0	0	58	17.6	$J_1(X)/X$
1	1.2	73	24.6	$J_2(X)/X^2$
2	2.5	84	30.6	$J_3(X)/X^3$

aperture and the far field patterns. The pattern characteristics of reflector antennas are sometimes represented for p having a value of 1. These simple analytic forms lend themselves to simulation activities, and as the simulation is refined, characteristics of the actual antenna can be used to increase the simulation fidelity.

The antenna gain and the antenna beamwidth depend on the electrical size of the antenna, that is, the size in wavelengths. The antenna gain increases with the square of the electrical size while the beamwidth is inversely related to the electrical size. Both values clearly depend on the specifics of the antenna's design. For preliminary system sizing, an efficiency of 55% and a beamwidth factor of 70^{o} are often used. As the design evolves, such values are updated. Using these parameter values, the gain and beamwidth are plotted in Fig. 1-2 for various aperture sizes in wavelengths. Values of antenna gain and beamwidth for specific cases as a function of frequency are given in Figs. 1-3 and 1-4, respectively.

In practice, detailed computer codes are available to accurately project the performance of a wide variety of antenna technology used in communication satellite systems. Such analyses provide the means of refining the values of the nominal parameters used in preliminary system sizings, as indicated here. These nominal values can also be useful for "mental estimates" of antenna performance. Notice that the speed of light is approximately 1 ft/nsec and therefore the number of wavelengths per foot equals the frequency in GHz. For example, a 10-ft antenna at 10 GHz has a diameter of 100 wavelengths. Using a beamwidth factor of 70, the beamwidth equals about 0.7^{o}. For a circular aperture, the antenna gain

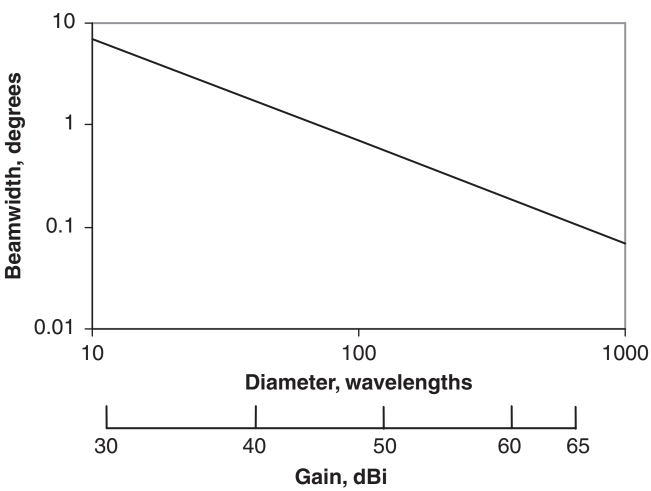

Figure 1-2 Nominal antenna gain and beamwidth values

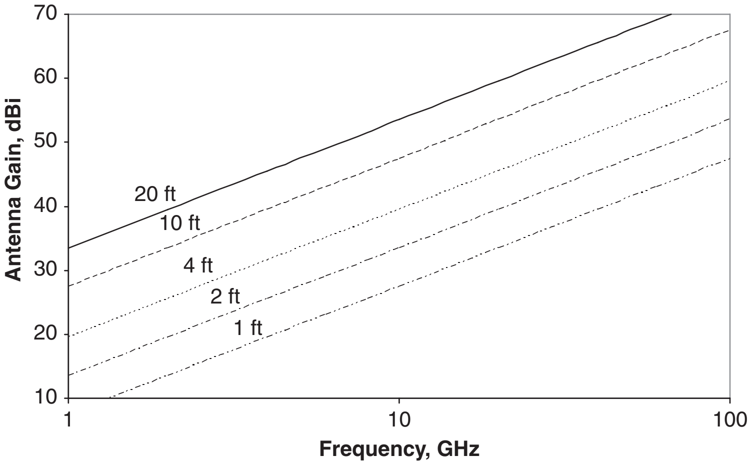

Figure 1-3 Antenna gain variation with frequency

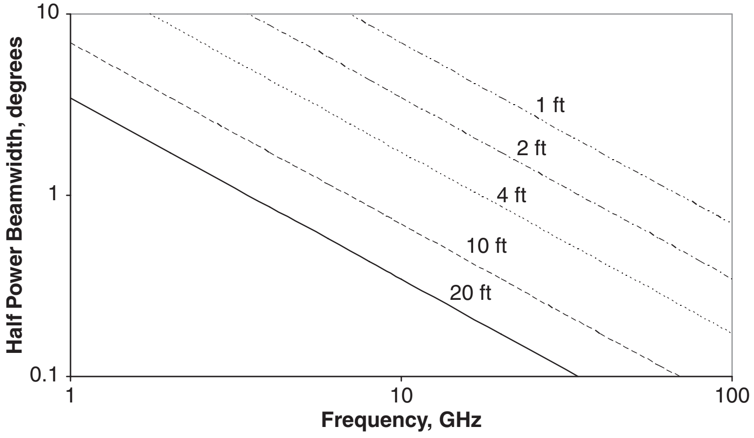

Figure 1-4 Antenna beamwidth variation with frequency

equals $\eta(\pi D/\lambda)^2$. The product $\eta\pi^2$ corresponds to about 7.3 dB, for a 55% antenna efficiency. The aperture diameter equals 100 wavelengths, and 20 times the log of 100 equals 40 dB. The antenna gain thus equals about 47.3 dBi (this indicates antenna gain relative to an isotropic gain level). This process may prove useful when rough estimates of antenna parameters are required and detailed calculation is unavailable.

1.2.2 Polarization

The vector nature of electromagnetic waves is specified by the polarization produced by an antenna, and propagating in free space. Polarization specifies the orientation of the electric field during one RF cycle. The most general polarization state is elliptical where the electric field traces out an ellipse. For every polarization, a unique orthogonal polarization exists, where orthogonal denotes ideal isolation between a receiving antenna and an incident field having orthogonally polarized states. Polarization is characterized by three parameters. One is an axial ratio equal to the ratio of the major and minor axes of the polarization ellipse. The second is the tilt angle specified by the alignment of the major axis of the ellipse in a reference frame. The third is the polarization sense specified by the familiar right- or left-hand rotation, as viewed in the direction of propagation.

The nominal orthogonal polarizations are linear and circular. Linear polarizations are typically indicated as vertical and horizontal, and ideal linear polarization confines the electric field to a plane. Linear vertically polarized antennas do not respond to horizontally polarized fields and thus the linear polarizations must be spatially aligned in use. Circular polarization is comprised of two orthogonal linear components having a 90° phase difference. Over one RF period, the electric field traces out a circle. Circular polarization does not require the polarization alignment that linear polarization does, and for that reason circular polarization is widely used in satellite communication systems. Circular polarization components are orthogonal when their sense differs. Right-hand circular polarization sense is orthogonal to left-hand circular polarization sense. These polarization senses follow the familiar right- and left-hand rules when viewed in the direction of propagation.

Practical antennas are not ideally polarized and are mixtures of the two orthogonal components. The cross-polarized antenna response quantifies the degree to which the antenna deviates from the ideal polarization. At a system level, two issues result from the finite cross-polarized components:

1. What signal loss results from the cross-polarized components, a parameter referred to as polarization mismatch loss?
2. When orthogonally polarized signals are used to communicate independent data streams in polarization reuse designs, what is the isolation between orthogonal pairs?

The axial ratio, r, can be expressed [2] in terms of the circularly polarized components as

$$r = (E_R + E_L)/(E_R - E_L)$$

where E_R and E_L are the amplitudes of the right- and left-hand polarization components, respectively. Notice that the numeric value of axial ratio is positive for right-hand components and negative for left-hand components. Normally, axial ratio is given in a logarithmic value that involves the magnitude of the axial ratio. The level of the cross-polarized component relative to the principally polarized component can be calculated as presented in Fig. 1-5.

Normally, the axial ratio of incident fields and antenna systems are known, but the relative orientation of the tilt angles of their respective polarization ellipses are unknown. Both the polarization mismatch loss and polarization isolation depend on the relative orientation of the two polarization ellipses of the incident field and receiving antenna. A statistical approach [3] is presented as a means of understanding the variations resulting from unknown polarization ellipse alignment.

The polarization efficiency has been defined in terms of the axial ratios and orientation of the polarization ellipses of the incident field and receiving antenna. Polarization mismatch loss is determined from polarization efficiency when the incident field and receiving antenna have the same polarization sense. Polarization isolation is determined from polarization efficiency when the incident field and receiving antenna have opposite polarization senses. Polarization efficiency [2] equals

$$\eta_p = \tfrac{1}{2} + A + B\cos\Delta$$

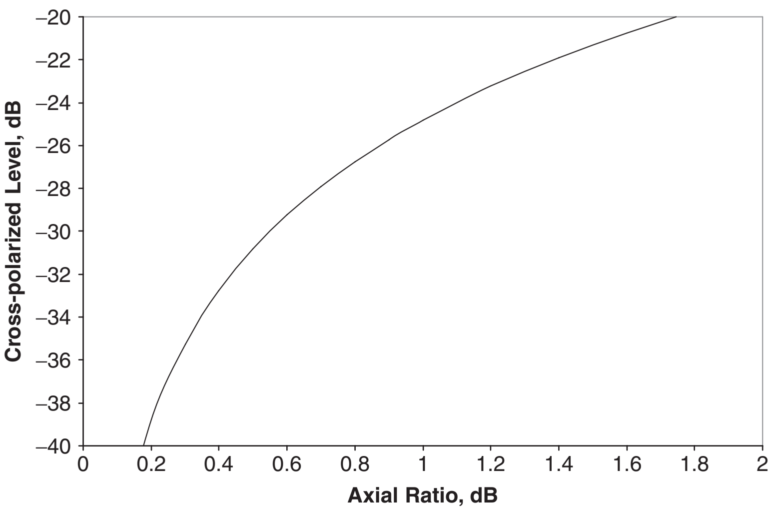

Figure 1-5 Cross-polarized level versus axial ratio

where

$$A = 2r_w r_r \Big/ \left[\left(1 + r_w^2\right)\left(1 + r_r^2\right)\right]$$

$$B = \left(1 - r_w^2\right)\left(1 - r_r^2\right) \Big/ \left[2\left(1 + r_w^2\right)\left(1 + r_r^2\right)\right]$$

where the subscripts "*w*" and "*r*" refer to the axial ratios of the incident wave and the receiving antenna, respectively. When the sense of incident field and the receiving antenna have the same polarization sense, *A* is positive because the product of the numeric value of axial ratios is positive when both senses are the same. When the senses of the incident field and receiving antenna have opposite polarization senses, *A* is negative. The angle, Δ, is the phase difference between the polarization components and equals twice the difference in the tilt angle orientations of the ellipses.

The commonly used bounds on polarization efficiency are $\frac{1}{2} + A \pm B$. The statistical variation of the polarization efficiency is derived by assuming the relative orientations of the tilt angle of the incident field and receiving antenna are equally likely and uniformly distributed over 0 to π, corresponding to Δ being equally likely and uniformly distributed over 0 to 2π. The first order (mean) statistics are determined from

$$E_p = (\tfrac{1}{2}\pi) \int \eta \, d\Delta$$

$$= \tfrac{1}{2} + A$$

where the integration extends over 0 to 2π. When both the incident field and receiving antenna have the same polarization sense, the mean efficiency is $\geq\frac{1}{2}$, and when their polarization senses are opposite, the efficiency is $\leq\frac{1}{2}$. Additionally, if either the incident field or the receiving antenna or both is ideally linear, the mean polarization efficiency is ½, or the familiar 3 dB loss.

Similarly, the second order (variance) statistics are determined from

$$V_p = 1/2\pi \int (\eta - E_p)^2 \, d\Delta$$

$$= B^2/2$$

where again the integration extends over 0 to 2π. The standard deviation of the polarization efficiency, σ, about its mean value equals $B/2^{1/2}$. The polarization efficiency statistics have a non-zero mean value, so the second-order statistics are generally expressed as the $\pm 1\ \sigma$ spread about the mean value. Further, the polarization efficiency statistics are not Gaussian. The peak-to-peak bounds are $\pm B$ for these statistics, while the rms spread is ±0.707 B. The bounds equal 1.41 times the standard deviation. By contrast, peak-to-peak variations for Gaussian statistics are often taken as $\pm 3\sigma$, well in excess of the possible peak-to-peak excursion for the polarization efficiency.

Example values illustrate the statistical variations and two different cases, one for an incident field having a 0.5 dB axial ratio and the second with a 2 dB axial ratio. The polarization mismatch loss levels in Fig. 1-6

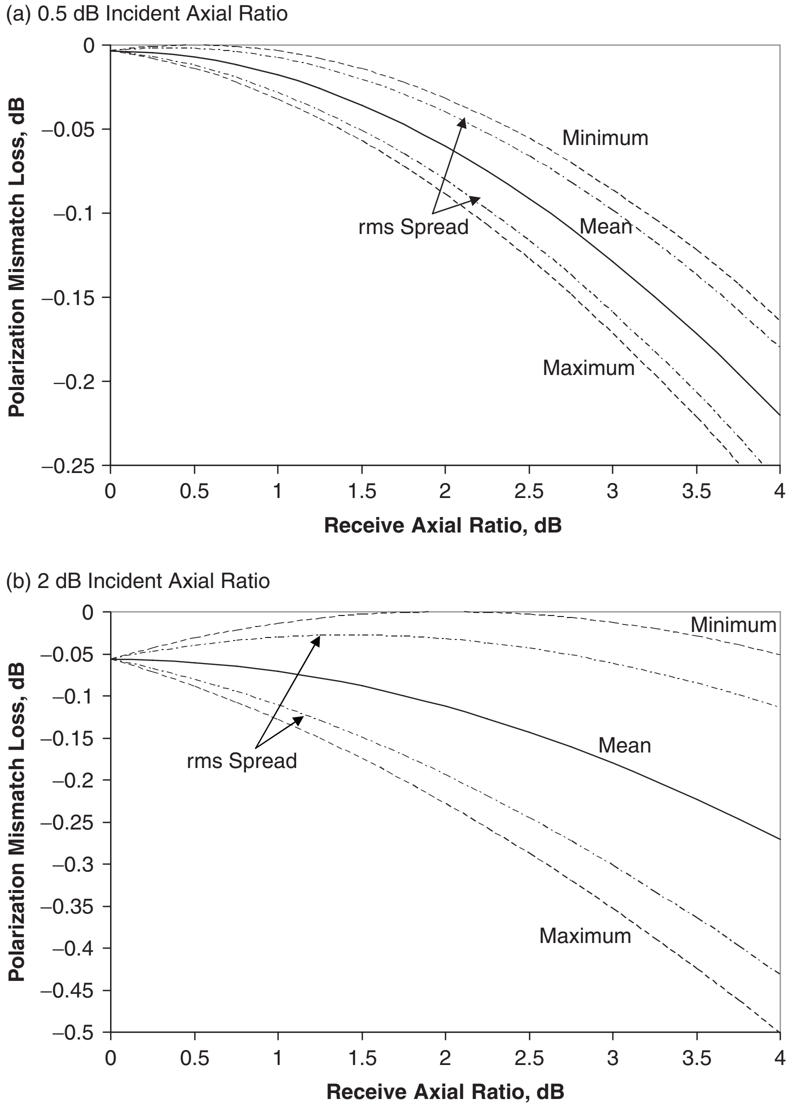

Figure 1-6 Polarization mismatch loss statistics

illustrate the mean, the minimum, the maximum, and the mean ± 1 σ values. The two examples illustrate the variation in the statistical values increase as the axial ratio of the incident field increases and as the axial ratio of the receiving antenna increases. Further, notice that the matched polarization condition when the incident field and receiving antenna have the same axial ratio value and their polarization tilt angles are coincident has a finite probability of having no polarization mismatch loss. System applications for polarization reuse require high polarization purity and incident fields having axial ratio values on the order of 0.5 dB. Other applications that seek reasonable polarization mismatch loss generally limit axial ratio values to about 2 dB. For example, if both the axial ratios of the incident field and receiving antenna are limited to 2 dB, the maximum possible polarization mismatch loss is less than 0.25 dB, and on average the polarization mismatch loss is about 0.1 dB, corresponding to the mean value.

The polarization isolation values in Fig. 1-7 similarly illustrate the statistical variations when the incident field has a 0.5 and 2 dB axial ratio. When orthogonally polarized components are used in frequency reuse designs to increase system capability, high levels of polarization purity are required of both the incident field and the receiving antenna. For example, if the incident field and the receiving antenna both have 0.5 dB axial ratios, the minimum polarization isolation is about 25 dB. When design attention is not paid to polarization purity, the isolation significantly degrades. Ideal polarization isolation requires the incident field and receiving antenna to have the same axial ratio and orthogonal polarization tilt angle orientations. The results indicate a finite probability of that condition being satisfied.

Often, circular polarization is produced by combining orthogonal linear components with a quadrature hybrid to produce circular polarization. The axial ratio resulting from amplitude and phase imbalance [4] in combining two orthogonal linearly polarized components is illustrated in Fig. 1-8. As the authors point out, the inherent cross polarization in the linear components is not included in this analysis and must be considered in practical designs.

1.2.3 Impedance

The interface between the antenna and the system electronics must also be specified. Maximum power transfer requires matched impedance characteristics, and deviations from ideal matched impedance result in reduced power transfer referred to as mismatch loss. Measured terminal parameters are generally performed using network analyzer instrumentation, and such measurements are expressed in scattering matrix parameters. The reflected components are expressed in the voltage

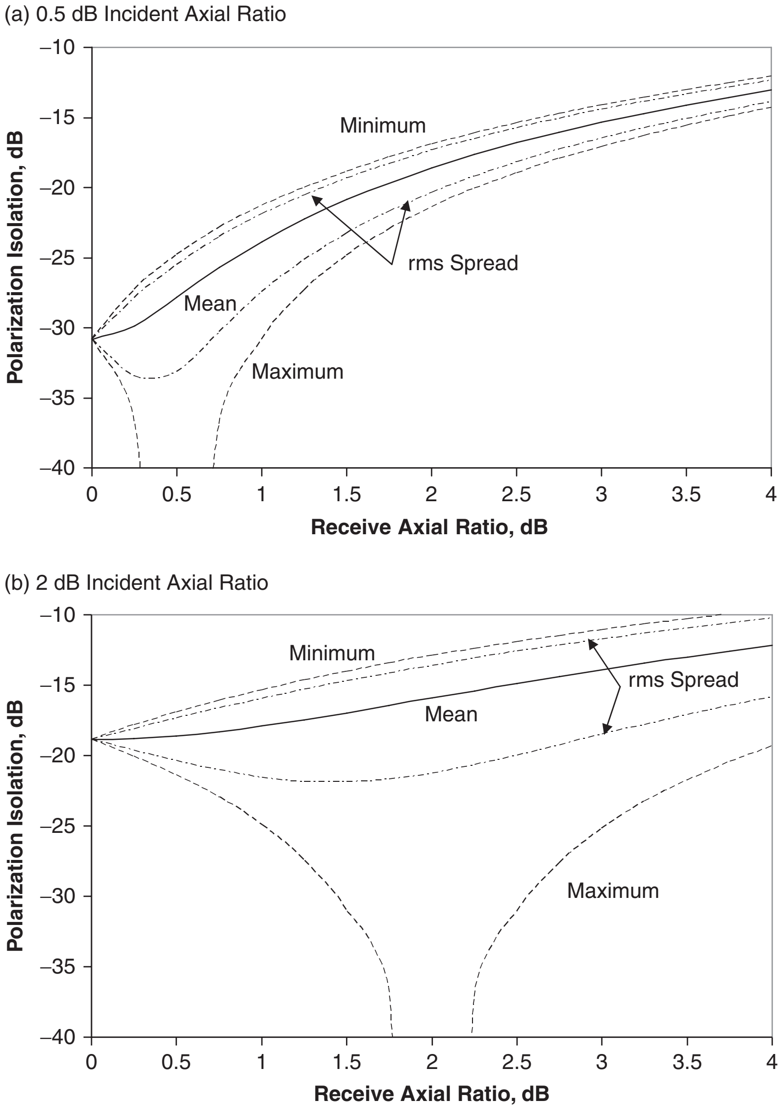

Figure 1-7 Polarization isolation statistics

reflection coefficient denoted by S_{11}, while the voltage transmission properties are denoted by S_{21}. The insertion loss for passive components equals $|S_{12}|^2$ between terminals 1 and 2 and for active electronics, $|S_{12}|^2$ is the insertion gain. The mismatch loss is $1 - |S_{11}|^2$. Antenna impedance

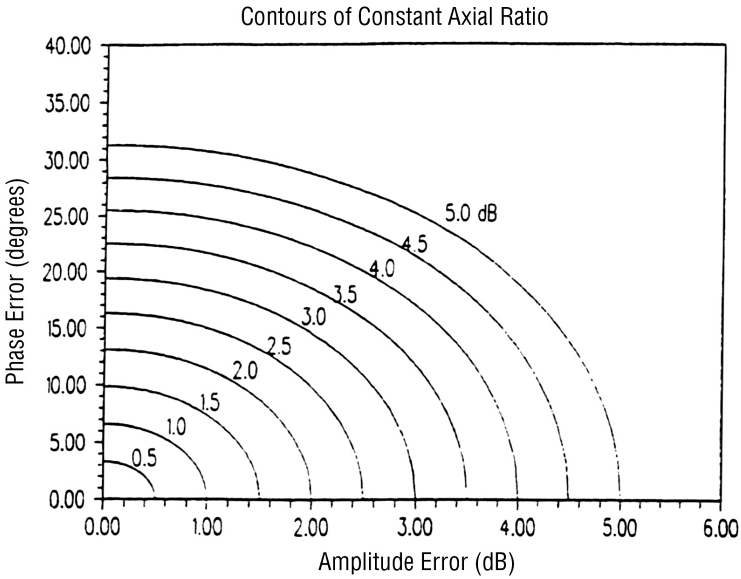

Figure 1-8 Hybrid imbalance impacts on axial ratio [4] (©1990 IEEE)

values are commonly specified by the voltage reflection coefficient S_{11}, the return loss RL, or the VSWR (voltage standing wave ratio), and are expressed as

$$\text{RL} = 20 \log (|S_{11}|)$$

and

$$\text{VSWR} = (1 + |S_{11}|)/(1 - |S_{11}|)$$

where physically VSWR is the ratio of the maximum and minimum values of the voltages along a transmission line. The return loss that equals $20 \log |S_{11}|$ is commonly used when network analyzer measurements are used. The mismatch loss is $1 - |S_{11}|^2$. Example values of impedance mismatch loss for the three common ways of expressing mismatch are given in Fig. 1-9. In specifications, VSWR and return loss are most commonly used, and the relationship between these parameters is given in Fig. 1-10.

In practical system designs, neither the antenna nor the interface electronics are ideally matched. As a result, multiple reflections between the antenna and electronics occur giving rise to amplitude and phase ripples over the system's operating bandwidth. Typically, the amplitudes of the reflection coefficients are known, but the phase values and their

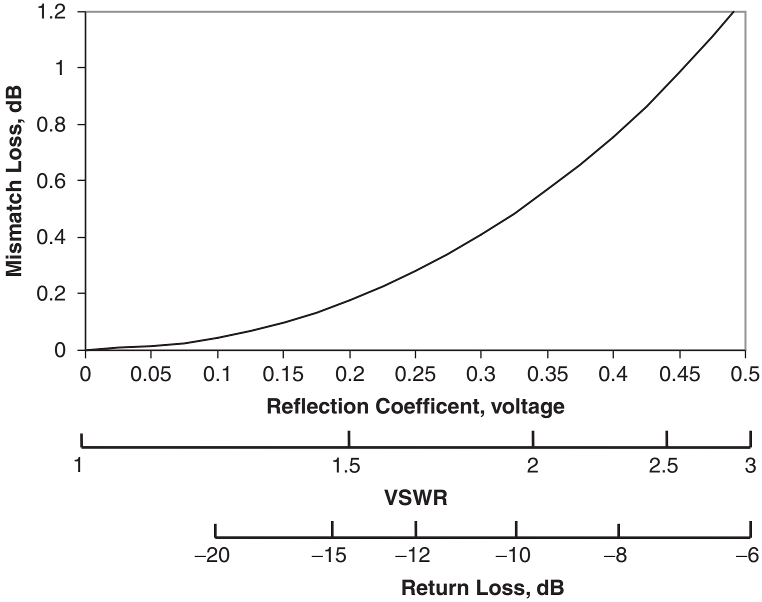

Figure 1-9 Impedance mismatch loss versus common impedance parameters

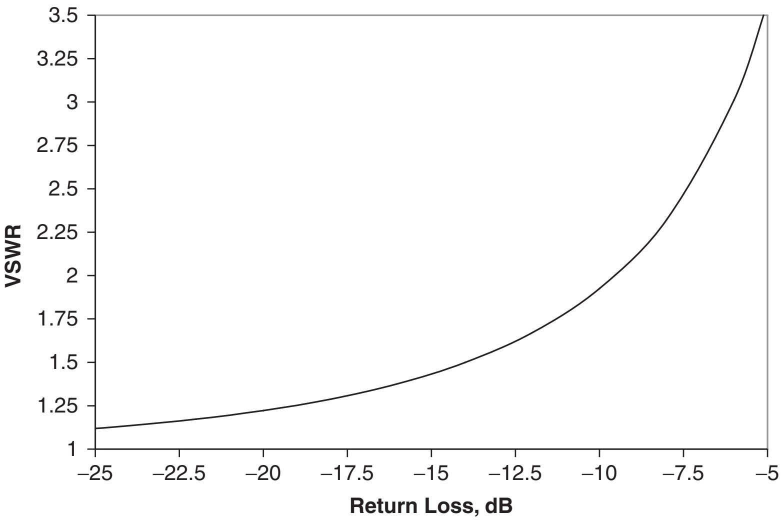

Figure 1-10 Relationship between return loss and VSWR

variation with frequency are not. Amplitude and phase ripple degrades signal detection performance and generally a specification is placed on the tolerable ripple. One approach to addressing the ripple uses coherent error statistics [5]. Other applications of these statistics, besides assessing VSWR interactions, include assessing the effects of multipath facility reflection errors and antenna cross-polarization errors.

The coherent error statistics can be visualized by the phasor diagram in Fig. 1-11. The true value assumes a unit level, and the coherent error is represented by a phasor having an amplitude a and a phase α. Since the phase is assumed to be unknown, the error statistics are derived by assuming any phase value is equally likely and uniformly distributed between 0 and 2π. The resulting first- and second-order statistical values for power, voltage, and phase are given in Table 1-2. The statistics for power can be exactly integrated, while series expressions are derived for both voltage and phase statistics. The maximum errors for a < 0.5 (VSWR = 3, RL = –6 dB) are indicated and apply to most practical cases. The power and voltage statistics are non-zero mean and the peak-to-peak error bounds have finite values. The rms spread about the mean error and the peak-to-peak errors are presented in Fig. 1-12. The coherent error statistics are clearly not a Gaussian distribution and have finite bounds whose values are much less than that that would be anticipated from 3 σ confidence values for Gaussian statistics. In the case of limiting the amount of amplitude ripple resulting from VSWR interactions, the product of the return loss values for both impedance interfaces must be less than values indicated in Fig. 1-13. For example, if the ripple is to be less than 1 dB, the product of the return loss values must be less than about –25 dB. In some cases (e.g., a cable run from an LNA output to a downconverter input), the addition of loss can be advantageous in reducing the ripple and the product of the return loss

TABLE 1-2 Coherent Error Statistics

	Power	Voltage	Phase
Mean	$1 + a^2$	$\approx 1 + a^2/4$	0
Error a < 0.5	N/A	1.6%	N/A
Standard Deviation	$2^{1/2}a$	$\approx a/(2^{1/2})(1 - 3a^2/16)^{1/2}$	$\approx (a^2/2 + a^4/8)^{1/2}$
Error a < 0.5	N/A	0.1%	0.5%

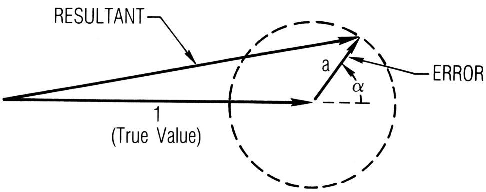

Figure 1-11 Phasor diagram for coherent errors [5] (©1989 IEEE)

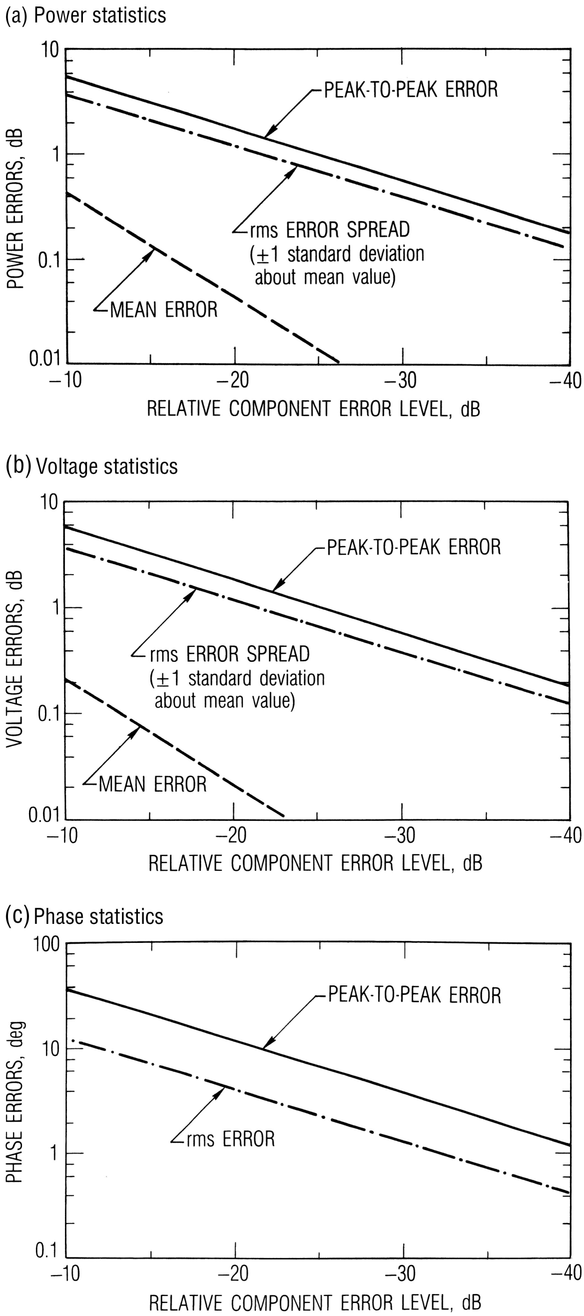

Figure 1-12 Coherent error statistical values [5] (©1989 IEEE)

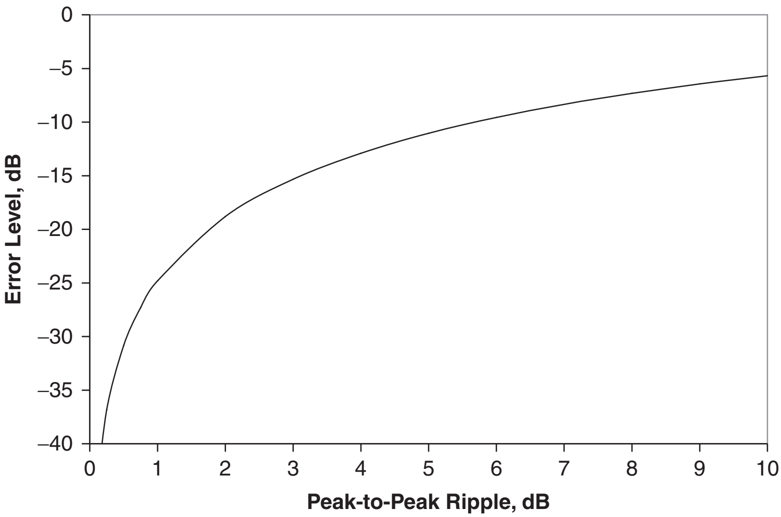

Figure 1-13 Error component level versus peak-to-peak ripple

values is increased by twice the attenuation value because the VSWR interaction component incurs a two-way path through the attenuation. The insertion of attenuators when measuring antennas that have significant mismatch is commonly done to reduce errors resulting from VSWR interactions. To be effective, the attenuators must have a good impedance match.

1.2.4 System Noise Temperature

The system figure of merit for receiving antennas is G/T, the antenna gain divided by the total system noise temperature T. The system noise temperature has two components. The antenna noise temperature includes noise components from the environment surrounding the antenna and the noise generated by losses within the antenna. A common reference terminal must be specified where antenna gain and system noise figures are both established. G/T is independent of the location of the reference terminal plane, but both antenna gain and system noise temperature values vary with the location of the terminal plane used for G/T determination. For example, the input terminal of an LNA is often convenient in measuring the receiver noise temperature. Generally, cabling and RF filtering follow the antenna terminals where the antenna gain values have been established. The losses in such components must be used to adjust the antenna gain value and antenna noise temperature so that the antenna gain level is referenced to the LNA's input terminal. Alternatively, the receiver noise temperature can be measured, including the filter and cabling loss, so that the G/T is determined at the antenna terminal.

The antenna noise temperature can be measured as described in Chapter 8 or calculated in the following way. When calculating the antenna noise temperature, the antenna is initially assumed to be lossless, and the noise temperature is calculated from

$$T_{ant}' = \iint P(\theta - \theta_o, \varphi - \varphi_o)\, T_e(\theta, \varphi) \sin\theta\, d\theta\, d\varphi$$

where $P(\theta - \theta_o, \varphi - \varphi_o)$ is the power pattern of the antenna pointed with its beam maxima in the direction θ_o, φ_o and normalized to a unity value at the beam peak, $T_e(\theta, \varphi)$ is the emission background temperature that is described in further detail in Chapter 3. The power pattern includes both principal and cross-polarized components. An example of such a calculation for a reflector antenna is given in Fig. 1-14. The antenna is an 8-ft reflector fed with a low-loss conical horn illuminating a Cassegrain subreflector at a frequency of 11 GHz. The analysis was performed using NEC REF [7], a geometrical theory of diffraction code. The measurements and analytic results agree well.

The example in Fig. 1-14 has a low-loss feed that contributes little to the antenna noise temperature. However, practical antennas generally have losses in the antenna feed, the interconnecting cabling, and the filters are

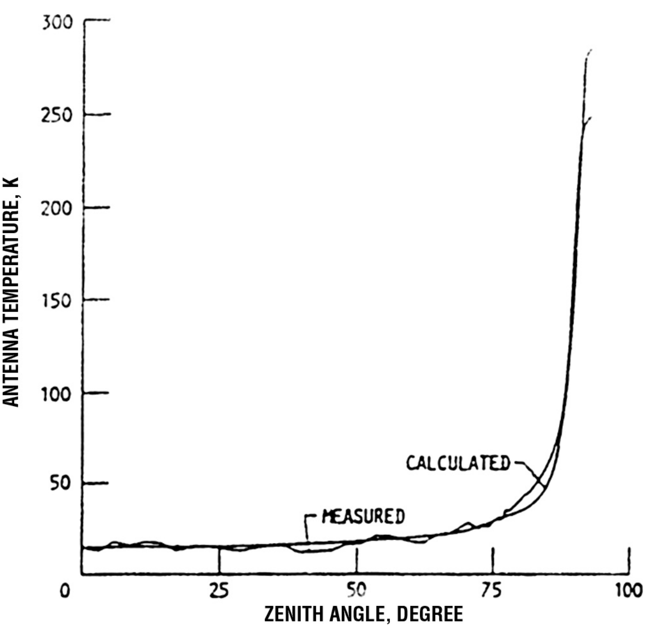

Figure 1-14 Example antenna temperature values [6]

required to limit the signal spectra to the operating bandwidth. The lossless antenna temperature [8] is then corrected for the noise contributed by these losses. The antenna noise temperature at the output of the filter and input to the receiver's LNA is given by

$$T_{\text{ant}} = (1 - \Gamma^2)\ [T_{\text{ant}}'L + 290(1 - L)]$$

where Γ is the magnitude of the reflection coefficient [for well-matched systems, $(1 - \Gamma^2)$ is very close to 1], T_{ant}' is the antenna noise temperature at the antenna terminals, and L is the ohmic loss. Physically, the noise power received by a lossless antenna is reduced by ohmic loss, but additional noise is generated by the loss. When impedance mismatch loss is significant, the antenna noise temperature is reduced by the impedance mismatch loss $(1 - \Gamma^2)$.

The receiver noise temperature at the LNA input terminal includes contributions from the LNA and other receiver components following the LNA. The receiver noise temperature at the LNA input, taking into account the LNA and receiver components following the LNA, is the cascade noise temperature, which equals

$$T_{\text{rec}} = T_{\text{lna}} + \sum T_i/G_{i-1}$$

where T_i is the noise temperature of the i^{th} component in the receiver and G_{i-1} is the insertion gain at the input to that component. System designs generally strive to make the LNA noise temperature dominate the receiver noise temperature, but contributions from the other receiver components can become significant when the receiver is required to have a wide linear dynamic range. The noise performance of receiver components is generally stated in terms of noise figure, NF, which is related to the noise temperature, T_n, by

$$T_n = 290(NF - 1)$$

Noise figure is defined as the input SNR (signal-to-noise ratio) divided by the output SNR. The total input noise can have many different values depending on how the component is used. Specifying the component's noise figure to a standard 290 K reference temperature allows the noise temperature to be calculated and used in applications where the input noise temperature differs from 290 K. In this application, the antenna noise temperature generally differs from 290 K, for example. Numerical values of the conversion between noise figure and noise temperature are given in Fig. 1-15.

The system noise temperature is the sum of the antenna and receiver noise temperature and equals

$$T = T_{\text{ant}} + T_{\text{rec}}$$

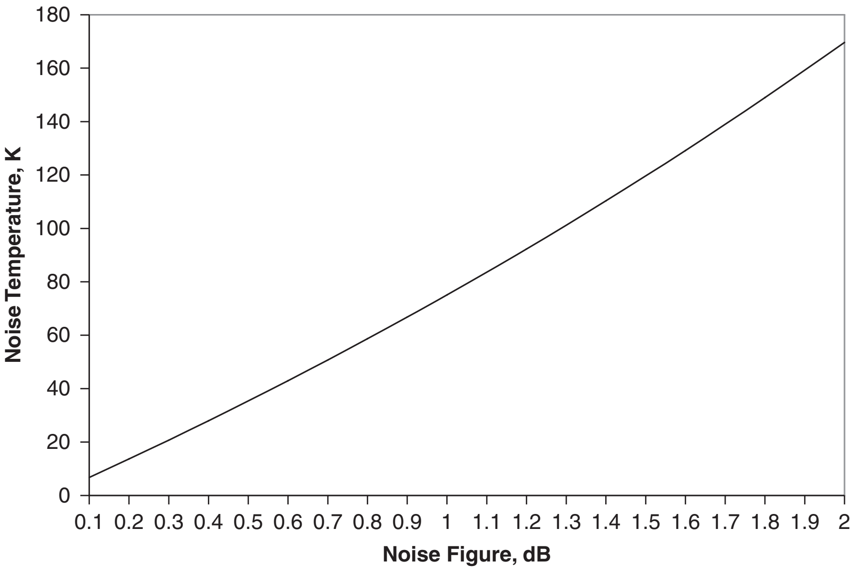

Figure 1-15 Noise temperature versus noise figure

The system noise temperature depends on the terminals in the system used for the specification and both the antenna and receiver noise temperature values must be specified for the same terminal.

In communication satellite applications, the space segment antennas for uplink receiving service view the earth's surface and generally have an ambient 290 K antenna temperature. User segment antennas look out towards space but incur atmospheric loss as discussed in Chapter 4, and as illustrated in Fig. 1-14, their antenna noise temperature varies with elevation angle and frequency of operation. Crosslink antennas that provide connectivity between satellites view a 3 K cosmic background temperature. Receiving system performance, however, depends on the system noise temperature. Today's LNA technology offers very low noise performance. However, the system noise performance can be dominated by the antenna noise temperature. This is illustrated in Fig. 1-16 where an ideal noiseless receiver is used as a reference value. The loss in receiving sensitivity or system temperature for various antenna noise temperature values is parametrically plotted as a function of receiver noise temperature. As the antenna noise temperature increases, the loss in sensitivity is less sensitive to the receiver noise temperature. Thus, low noise receiver temperatures can be effectively used if a low antenna noise temperature exists, whereas higher antenna noise temperature values derive less benefit from low noise receiver technology.

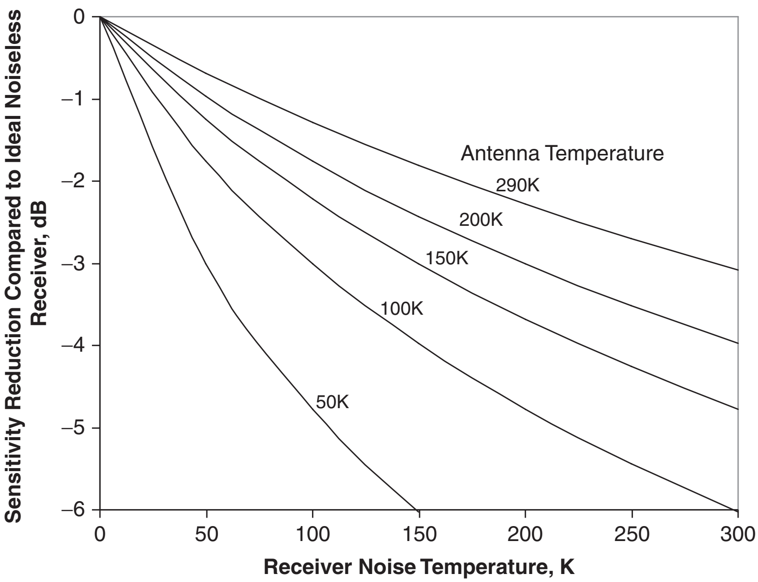

Figure 1-16 Sensitivity reduction from receiver noise temperature

1.2.5 System Parameters

The system parameters are *G/T* for receiving systems and ERP (Effective Radiated Power) for transmitting systems. The values of these parameters do not vary with the system terminals at which they are determined but their specification must use a consistent set of terminals. Examples will be used to illustrate the determination of these system parameters.

The *G/T* of an example system to operate at 10 GHz assumes a 4-ft antenna is used with a 0.5 dB noise figure receiver and a 0.5 dB loss exists between the antenna and receiver because of a bandpass filter and interconnection loss. The antenna is assumed to have a 55% efficiency, so that at 10 GHz the antenna has a 40.6 wavelength diameter, a gain value of 39.5 dBi, and a beamwidth of 1.7° if a beamwidth factor value of 70 is assumed. The input terminals for *G/T* determination are the LNA input. The antenna gain level after the filter loss therefore equals 39.0 dBi. The antenna system specification references the *G/T* value to a 20° elevation angle where the antenna noise temperature at the antenna terminals is assumed to be 30 K. The antenna noise temperature at the LNA input terminal after the 0.5 dB loss equals 58.2 K. The receiver noise temperature corresponding to the 0.5 dB noise figure equals 35.4 K so that the system noise temperature equals 93.6 K or 19.7 dBK (dB referenced to 1 K). The *G/T* value is obtained from the antenna gain and system noise temperature referenced to the LNA input terminal. The G/T at the specified 20° elevation angle thus equals 19.3 dBi/K.

In the preceding example, the loss not only reduces the antenna gain value but also increases the system noise temperature. In the design of user receiving systems where the antenna noise temperature values are generally significantly lower than 290 K, attention to reducing system loss is important in achieving good G/T performance. For space segment antennas where the antenna noise temperature is 290 K, the antenna noise temperature remains 290 K after loss and the G/T of uplink receiving antennas is reduced only by the loss. At VHF and UHF frequencies, the situation for user receiving antennas differs. As discussed in Chapter 4, the antenna noise temperature greatly exceeds 290 K and dominates the system noise temperature. The expression for G/T with loss can be written as

$$G/T = GL/(T_{\text{ant}}L + 290(1 - L) + T_{\text{rec}})$$

$$\approx G/T_{\text{ant}}$$

The G/T for high antenna noise temperature values becomes independent of the loss. Physically, because the antenna noise temperature greatly dominates the system noise temperature, the loss reduces signal and noise equally so that the system's G/T remains unaffected. When operating in the VHF and UHF frequencies where the antenna noise temperature can be several thousand K, the loss in filters needed at these frequencies does not reduce system sensitivity.

The determination of ERP is also illustrated by an example. Suppose the same 4-ft antenna used in the previous example is connected to a 10 W transmitter. The loss between the transmitter and antenna is assumed to be 1 dB. The 10 W transmitter power level is its saturated power output (discussed in Chapter 3) and must be backed off to satisfy linearity requirements. In this case, a 3 dB backoff is assumed to be adequate and so the transmitted power output at this design operating point is 5 W or 7 dBW (dB relative to a 1 W level). The ERP in this case equals 45.5 dBW.

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