CHAPTER 16.3

FUNDAMENTALS OF WAVE PROPAGATION

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INTRODUCTION: MECHANISMS, MEDIA, AND FREQUENCY BANDS

This chapter deals with radio waves propagated through or along the surface of the earth, through the atmosphere, and by reflection or scattering from the ionosphere or troposphere. The particular propagation mechanism around which a given application is designed depends on the distance to be spanned, the type of information to be transmitted or the service to be provided, and the reliability required. Other propagation mechanisms can affect the performance of the system or lead to interference with and from other systems, depending on frequency and distance.

Over a line-of-sight path within the nonionized atmosphere, transmission is much as through free space, though atmospheric refraction causes bending, reflection, scattering, and possibly fading. At frequencies above about 10 GHz, there may be attenuation because of rainfall and absorption by air and water vapor. The conductivity and permittivity (dielectric constant) of the earth are markedly different from those of the atmosphere. A wave mainly diffracted along the surface of the ground encounters increasing loss with increasing frequency. Very low frequency waves are propagated with little attenuated over thousands of kilometers. At high frequencies, losses along the ground become so great that the usefulness of the ground wave is limited to short distances. At medium and high frequencies, ionospheric reflections permit radio communication to great distances. At frequencies much above 30 MHz, ionospheric reflections are not dependable, and most communications depend on line-of-sight propagation or tropospheric scattering beyond the horizon.

Because of the dependence of propagation characteristics on frequency, much of the discussion of these sections will be in terms of frequency bands, and abbreviations such as VLF for very low frequencies and VHF for very high frequencies will be used.

The International Telecommunication Union (ITU) has defined nine frequency bands designated by integer band numbers; for example, 1 MHz is the approximate midband of band 6, and so on, as listed in Table 16.3.1.

Certain frequency bands, however, are sometimes unofficially designated by letter rather than by the abbreviations given in Table 16.3.1. Since there is no standard correspondence between the letters and the frequency bands concerned, it is advisable that their use be clarified by reference to the approximate limits of the band or to a frequency within the band. For information, letter designations used essentially in the areas of radar and space communications, are given in Table 16.3.2.

Propagation characteristics are discussed according to somewhat different bands, such as 10 to 150 kHz and 150 to 1500 kHz, corresponding to bands of relatively homogeneous propagation characteristics.

References 1 to 6 are basic texts on electromagnetic wave propagation. References 7 to 22 are comprehensive texts on tropospheric and ionospheric radiowave propagation which, in some cases, adopt a systems approach in describing the particular propagation mechanisms of concern. Recent results of theoretical and experimental radio science worldwide are outlined in Ref. 23.

TABLE 16.3.1 Frequency Bands Defined by ITU

ITU band number*	Frequency range (lower limit exclusive, upper limit inclusive)	Corresponding metric subdivision	Abbreviation
4	3–30 kHz	Myriametric waves	VLF
5	30–300 kHz	Kilometric waves	LF
6	300–3000 kHz	Hectometric waves	MF
7	3–30 MHz	Decametric waves	HF
8	30–300 MHz	Metric waves	VHF
9	300–3000 MHz	Decimetric waves	UHF
10	3–30 GHz	Centimetric waves	SHF
11	30–300 GHz	Millimetric waves	EHF
12	300–3000 GHz (3 THz)	Decimillimetric waves	

*Band number N extends from 0.3×10^N to 3×10^N Hz.

The emphasis here is mainly descriptive, key formulas indicating the behavior of parameters and references to significant publications providing material for engineering calculations. Maxwell's uniform plane-wave equations are cited only to show the role of the electrical constants and the vector relationships of electric and magnetic field and power flux.

Wave Propagation in Homogeneous Media

Electromagnetic radiation is composed of two mutually dependent vector fields, electric and magnetic. The electric field is characterized by the vectors **E**, electric field strength in volts per meter, and **D**, dielectric displacement in coulombs per square meter. The magnetic field is characterized by **H**, the magnetic field strength in ampere-turns per meter (or amperes per meter) and **B**, flux density, in webers per square meter. The vector current density **J** is in amperes per square meter.

The relationship between the members of the various pairs of field vectors is characterized by the constitutive parameters, or *electrical constants*, of the medium:

$$\begin{aligned} \epsilon &= \text{permittivity (dielectric constant), F/m} \\ \sigma &= \text{conductivity, S/m} \quad \mu = \text{permeability, H/m} \end{aligned} \tag{1}$$

In some cases these constants are functions of the coordinate. Locally, however, they are always considered to be constant. Nearly always the time factor in these sections is exp $(+i\omega t)$, where ω is the angular frequency $2\pi f$ (f in hertz) and t the time. The *electric field*, then, is the real part of E exp $i\omega t$.

TABLE 16.3.2 Unofficial Letter Designations for Certain Bands (Recomm. ITU-R V.431)

	Radar (GHz)		Space radiocommunications	
Letter symbols	Spectrum regions	Examples	Nominal designations	Examples (GHz)
L	1–2	1.215–1.4	1.5 GHz band	1.525–1.710
		2.3–2.5	2.5 GHz band	
S	2–4	2.7–3.4		
C	4–8	5.25–5.85	4/6 GHz band	3.7–4.2
				5.925–6.425
X	8–12	8.5–10.5	—	
		13.4–14.0	11/14 GHz band	10.7–13.25
Ku	12–18	15.3–17.3	12/14 GHz band	14.0–14.5
K	18–27	24.05–24.25	20 GHz band	
Ka	27–40	33.4–36.0	30 GHz band	20–30 (approx.)

To explain very basic notation,* a short outline of plane electromagnetic waves in a homogeneous medium is given.

Ohm's law in the complex form is

$$\mathbf{J} = (\sigma + i\epsilon\omega)\mathbf{E} \tag{2}$$

where **J** is the current density vector and **E** is the electric field vector.

The analogous relation for magnetic quantities is

$$\mathbf{B} = \mu\mathbf{H} \tag{3}$$

In source-free media the above vector quantities are related by

$$\text{curl } \mathbf{E} = -i\mu\omega\mathbf{H} \tag{4}$$

and

$$\text{curl } \mathbf{H} = (\sigma + i\epsilon\omega\mathbf{E} \tag{5}$$

These are Maxwell's equations.

For a homogenous medium

$$\text{curl curl } \mathbf{E} = \text{grad div } \mathbf{E} - \text{div grad } \mathbf{E} = -i\mu\omega(\sigma + i\epsilon\omega)\mathbf{E} \tag{6}$$

Since div $\mathbf{E} = 0$,

$$(\nabla^2 - \gamma^2)\mathbf{E} = 0 \tag{7}$$

where ∇^2 = div grad = Laplacian operator (which operates on the rectangular components of **E**) and $\gamma^2 = i\mu\omega(\sigma + i\epsilon\omega)$. The quantity γ is called the *propagation constant*.

As a simple illustration of the role of the electrical constants of the medium, the field of a wave is assumed to vary only in the z direction in space (time factor exp $i\omega t$ understood), and the electric field is assumed to have only an x component E_x. For this case Eq. (7) reduces to

$$\left(\frac{d^2}{dz^2} - \gamma^2\right) E_x = 0 \tag{8}$$

and the solutions are exp $(+\gamma z)$ and exp $(-\gamma z)$ or, in general,

$$E_x = A \exp \gamma z + B \exp(-\gamma z) \tag{9}$$

where A and B are constants. The magnetic field then has only a y component given by

$$H_y = -\frac{1}{i\mu\omega}\frac{\partial E_x}{\partial z} = -\eta^{-1}\,[A \exp(+\gamma z) - B \exp(-\gamma z)] \tag{10}$$

where

$$\eta = \left(\frac{i\mu\omega}{\sigma + i\epsilon\omega}\right)^{1/2} \tag{11}$$

η is defined as the *characteristic impedance* of the medium for plane-wave propagation. Remembering that the time factor is exp $i\omega t$, we see that the term B exp $(-\gamma z)$ is a wave traveling in the positive z direction with diminishing amplitude and the term A exp γz is a wave traveling in the negative z direction with diminishing amplitude.*† The electric and magnetic fields are both transverse to the direction of propagation and orthogonal to each other.

*From Wait[3] by permission.

†The geometry of subsequent paragraphs uses a different convention for x, y, and z directions, shown in figures.

Such radiation is termed *plane-polarized*. It is by convention designated as *horizontal* or *vertical* according to the orientation of the plane containing the **E** vector.

The quantity η is equal to the complex ratio of the electric and magnetic field components in the x and y directions, respectively, for plane waves in an unbounded homogeneous medium, i.e.,

$$H_y = (i\omega\epsilon/r)E_x = E_x/\eta, \qquad H_x = (-r/i\omega\mu)E_y = -E_y/\eta \tag{12}$$

For a perfect dielectric ($\sigma = 0$)

$$\eta = \sqrt{\mu/\epsilon} \quad \text{and} \quad r = ik \tag{13}$$

where $k = \omega\sqrt{\mu\epsilon} = 2\pi/\lambda$ and λ is the wavelength.

The velocity of this wave is

$$v = 1/\sqrt{\mu\epsilon} \tag{14}$$

v is called the *phase velocity* of the wave. It represents the velocity of propagation of phase and does not necessarily coincide with the velocity with which the energy of a wave of signal is propagated, known as the group velocity. In fact, v may exceed free-space wave velocity without violating relativity in any way.

The wavelength is defined as the distance the wave propagates in one period.

$$\lambda = \frac{2\pi}{\omega\sqrt{\mu\epsilon}} = \frac{v(\text{m/s})}{f(\text{Hz})} \quad \text{m} \tag{15}$$

For free space

$$\epsilon = \epsilon_0 = 8.854 \times 10^{-12}\,\text{F/m}, \qquad \mu = \mu_0 = 4\pi \times 10^{-7}\,\text{H/m}, \qquad \sigma = 0 \tag{16}$$

Then

$$\gamma = ik = i\omega\sqrt{\mu_0\epsilon_0} = i2\pi/\lambda$$

The *velocity of the wave for free space* is

$$v_0 = 1/\sqrt{\mu_0\epsilon_0} \approx 3 \times 10^8 \ \text{m/s} \tag{17}$$

The *characteristic impedance of free space* is

$$\eta_0 = \sqrt{\mu_0/\epsilon_0} \approx 120\pi\ \Omega \approx 4\pi v_0 \times 10^{-7}\,\Omega \tag{18}$$

Energy flow in the electromagnetic field is described by the Poynting vector

$$\mathbf{P} = \mathbf{E} \times \mathbf{H}^* \tag{19}$$

where the complex representation of the time-periodic quantities is implied and the asterisk denotes the complex conjugate. The *real part* of the Poynting vector represents the average power flow over a cycle of the time variation per unit area in the direction of transmission.

$$\mathbf{P}_{\text{av}} = \tfrac{1}{2}\,\text{Re}\,(\mathbf{E} \times \mathbf{H}^*) = E^2/2\eta_0 \quad \text{W/m}^2 \tag{20}$$

$\mathbf{P}_{\text{av}}$ is called the *power flux density* or *field intensity*. Note that **E** and **H** are peak values.

The permeability and permittivity of any medium relative to free space are called the *relative permeability* and *relative permittivity*. These are usually the values given in tables of physical constants; they are dimensionless and designated by μ_r and ϵ_r, respectively.

Polarization of the Wave. *Polarization* is a term characterizing the orientation of the field vector in its travel. In radio, polarization usually refers to the electric vector. In the simplest case E_z and H_z (field components in the direction of propagation) are zero, and **E** and **H** lie in a plane transverse to the direction of propagation and orthogonal to each other. Such a plane wave is *elliptically polarized* when the electric vector **E** describes an ellipse in the plane perpendicular to the direction of propagation over one cycle of the wave.

When the amplitudes of the rectangular components are equal and their phases differ by some odd integral multiple of $\pi/2$, the polarization ellipse becomes a circle and the wave is *circularly polarized*. It is customary to describe as *right-handed* circularly polarized a clockwise rotation of **E** when viewed in the direction of propagation; counterclockwise rotation is *left-handed* polarization.*

An important case for many radio problems is that in which the polarization is a straight line. The wave is then *linearly polarized*. In *horizontal polarization* the electric vector lies in a plane parallel to the earth's surface.

To obtain maximum transfer of power between two antennas the polarization should match. If the transmitting antenna is horizontally polarized, the receiving antenna must likewise be horizontally polarized. If the transmitting antenna is elliptically polarized with a given degree of ellipticity and a specified direction of rotation, the receiving antenna should have the proper direction of rotation and degree of ellipticity in order to maximize the path antenna gain.

It should be noted that in the process of propagation, except through free space, the polarization may be altered. This can be caused by reflections from surfaces and, for frequencies at SHF and above, by hydrometers in the neutral atmosphere such as rain. Passage through the ionosphere in the presence of magnetic field is likely to impart elliptical polarization to a plane-polarized incident wave and rotation of the major axis. For MF or HF ionospheric propagation, the downcoming wave may be randomly polarized.

Reflection

Most problems in wave propagation involve reflection from a boundary between media of different refractive properties, often between air and the ground or between air and the ionosphere. In general, such a boundary may involve dissipative media (finite conductivity), curvature, finite dimensions, roughness, and stratification.

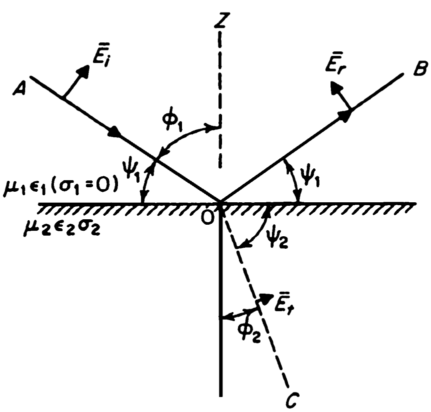

FIGURE 16.3.1 Geometry of reflection and transmission.

The *complex index of refraction* for a conducting medium is

$$n^2 = [i\omega\mu(\sigma + i\omega\epsilon)]/[i\omega\mu_0(i\omega\epsilon_0)] \tag{21}$$

When $\sigma = 0$,

$$n = \sqrt{\mu_r\epsilon_r} \quad \text{where } \mu_r = \mu/\mu_0, \epsilon_r = \epsilon/\epsilon_0$$

For many applications $\mu = 1$ and $\sigma = 0$, and the index of refraction is simply the square root of ϵ_r.

Figure 16.3.1 illustrates Snell's law for refraction of plane waves at an infinite plane interface. The angle ϕ between the direction of propagation and the normal to the boundary is called the *angle of incidence*. The angle ψ between the direction of propagation and the boundary, called the *grazing angle*

*Right- and left-hand polarization are sometimes misinterpreted with regard to the direction of viewing; the definition given here corresponds to the International Radio Regulations, Geneva, 1990.

or *Elevation angle*, is often more convenient. If the medium containing the incident wave is lossy, the angle of incidence is complex and can be defined in various ways.[3] At the boundary, the tangential components of **E** and **H** must be continuous; the phase of the reflected wave is in step with the phase of the incident wave to satisfy this requirement.

Snell's law of refraction for the direction of the transmitted wave toward C is

$$n_1 \sin\phi_1 = n_2 \sin\phi_2 \quad \text{or} \quad n_1 \cos\ \psi_1 = n_2 \cos\ \psi_2 \tag{22}$$

The *penetration depth* δ, or the depth at which the transmitted wave E_t has attenuated to $1/e$ of its incident value (for a conducting medium where $\sigma >> \omega\epsilon$), is

$$\delta = \frac{1}{\sqrt{\omega\mu\sigma/2}} \quad \text{m} \tag{23}$$

Ground Reflection, Reflection Coefficients, Fresnel Zones

A wave incident on a plane surface can be resolved into two components, one polarized normal and the other parallel to the plane of incidence. The reflection coefficients for the two components differ, and consequently the polarization of the reflected wave depends on the angle of incidence. Consider an air-earth boundary, taking the media to be nonmagnetic; for the case where the **H** vector is parallel to the ground surface the complex reflection coefficient[1] is

$$\mathbf{R}_v = \frac{(\epsilon_r - i60\sigma\lambda)\sin\psi - (\epsilon_r - \cos^2\psi - i60\sigma\lambda)^{1/2}}{(\epsilon_r - i60\sigma\lambda)\sin\psi + (\epsilon_r - \cos^2\psi - i60\sigma\lambda)^{1/2}} \tag{24}$$

where σ = conductivity (S/m)
ψ = grazing angle (Fig. 16.3.1)
ϵ_r = relative permittivity of earth to air or free space
$\epsilon_r - i60\sigma\lambda$ is referred to as the *complex permittivity*.

If E is parallel to the ground surface and H is in the plane of incidence,

$$\mathbf{R}_h = \frac{\sin\psi - (\epsilon_r - \cos^2\psi - i60\sigma\lambda)^{1/2}}{\sin\psi + (\epsilon_r - \cos^2\psi - i60\sigma\lambda)^{1/2}} \tag{25}$$

These are reflection coefficients for vertical and horizontal polarization, respectively. Curves of values for a range of ϵ and σ are given in Ref. 1.

An important property for vertical polarization is that there exists an angle of incidence for which the reflection coefficient approaches zero (for purely dielectric media it equals zero). This is the *Brewster angle*, also called the *polarizing angle*, given by

$$\phi_0 = \tan^{-1}\sqrt{\epsilon_1/\epsilon_2} \tag{26}$$

It is equal to the angle of incidence for which the reflected and refracted (transmitted) rays are at right angles. If the incidence occurs at the Brewster angle, the reflected wave is polarized entirely in the direction normal to the plane of incidence.

Wave tilt is a property frequently used to determine the electrical constants of the earth. For waves traveling at nearly grazing incidence along the surface of the earth, wave tilt may be interpreted geometrically as the angle between the normal to the wavefront and the tangent to the earth's surface. Wave tilt is defined to be the ratio of the horizontal to the vertical component of the electric field in the air just above the ground:

$$W = \mathbf{E}_h/\mathbf{E}_v \tag{27}$$

Wave tilt is related to the electrical constants of a homogeneous earth by

$$W = \frac{\sqrt{\mu_1/\epsilon_{1c}}}{\sqrt{\mu/\epsilon_0}} \sqrt{1 - \frac{\mu_0\epsilon_0}{\mu_1\epsilon_{1c}}} \tag{28}$$

The subscript 1 refers to earth constants and 0 refers to free space; ϵ_{1c} = complex permittivity = $\epsilon_1 - i\sigma_1/\omega$.

This procedure assumes that $\mu_0 = \mu_1$, generally a valid assumption; if it is not, μ_1 must be determined by some other procedure.

An important consideration in many propagation problems is the interference pattern generated by vector addition of the fields corresponding to the direct ray from an antenna to a point within line of sight plus the ground-reflected ray. In fact, the ground acts as a partial reflector and as a partial absorber, and the resultant field strength at a receiving point is given by

$$E_R = E_1[\underbrace{1}_{a} + \underbrace{Re^{i\Delta}}_{b} + \underbrace{(1 - R)Ae^{i\Delta}}_{c} + \cdots] \tag{29}$$

Term a corresponds to the direct wave while term b represents the reflected wave with R as the reflection coefficient and Δ the phase difference (in radians) corresponding to the path difference between the direct and reflected rays.

Term c corresponds to the surface wave, with the attenuation factor A a function of frequency, polarization, and electrical constants of the ground. The surface wave is particularly important at medium and low frequencies.

Using the notation in Fig. 16.3.2a the phase difference Δ may be written as

$$\Delta = 4\pi h_1 h_2/\lambda d \quad \text{rad} \tag{30}$$

for a wavelength λ and for grazing angles less than about 0.5 rad, Eq. (29) may be written:

$$E_R = E_1 \frac{4\pi h_1 h_2}{\lambda d} \tag{31}$$

omitting any contribution from the surface wave.

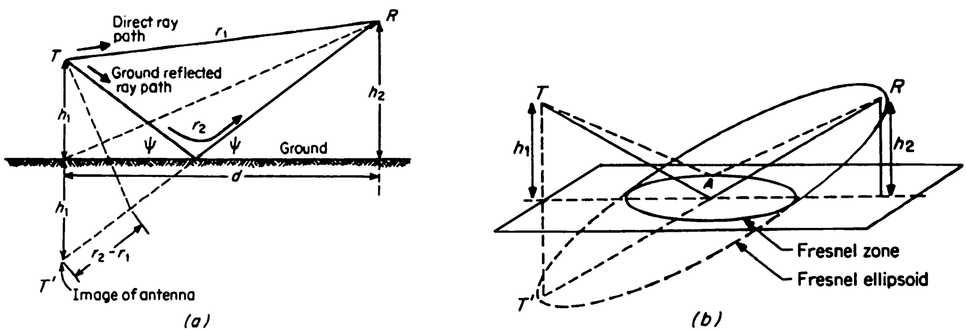

FIGURE 16.3.2 Geometry of ground reflection, image antennas, and Fresnel zones for plane earth: (a) ray paths; (b) Fresnel zone and ellipsoid.

The received field strength at a given distance will therefore display maxima and minima around the free-space value (corresponding to the direct wave alone) as the height of either antenna is altered. The angle at which maxima and minima occur are given by

$$\sin\psi = n\lambda/4h_1 \quad \text{maxima for } n \text{ odd, minima for } n \text{ even} \tag{32}$$

For the first maximum to occur at a specified elevation ψ_1,

$$h_1 = \lambda/(4\ \sin\psi_1) \tag{33}$$

In Fig. 16.3.2*a* the ray reflections are shown as though they occurred at a point. Actually, the surface of the earth is illuminated over a wide region corresponding to the radiation patterns of the two antennas and, in accordance with Huygens' principle, reradiates elementary wavelets in all directions. In any particular direction, as toward *R*, these elementary wavelets arrive with strength and phase such that the waves from an elliptical zone in the neighborhood of the ray reflection add nearly in phase. From successive ring areas, similarly bounded by larger ellipses, the waves alternately cancel and add.

These zones of physical reflection are called *Fresnel zones*, since they are closely related to the Fresnel zones of diffraction theory.[5,10] Most of the energy can be thought of as being reflected from the first Fresnel zone; it is defined, with the aid of Fig. 16.3.2*b*, for reflection paths between points such as *T* and *R*, as the area from which all the reradiated elementary wavelets arrive, according to geometric optics, within half wavelength of the phase of the direct ray. Thus the length of the geometric ray path at the edge of the *n*th Fresnel zone is *n* half wavelengths greater than the geometric ray path.

From Fig. 16.3.2*b*, the boundary of the first Fresnel zone is defined as the locus of points *A* such that *TA* + *AR* differs by half a wavelength from *TR*. This locus is an ellipsoid of revolution with foci at *T′* and *R*. The minor axis of the first Fresnel zone has the same dimension as the diameter of the first Fresnel ellipsoid, which can be easily shown to be $\sqrt{\lambda d}$, where *d* is the length of the propagation path. The length of the major axis will depend on the antenna heights h_1 and h_2 and is given by

$$L = \frac{d\sqrt{1 + 4h_1h_2/\lambda d}}{1 + (h_1 + h_2)^2/\lambda d} \tag{34}$$

For a well-developed ground reflection, the ground should be flat over an area that includes at least the first Fresnel zone. The degree of flatness depends on the wavelength and angle of incidence; assuming that phase-path changes less than $\lambda/16$ are unimportant, Rayleigh's criterion limits height deviations in terrain from a smooth surface to a magnitude less than $\Delta h = \lambda/(16\ \sin\ \psi)$ over the area of the first Fresnel zone for waves incident at angle ψ. Methods allowing for surface roughness, finite conductivity, and divergence owing to the spherical shape of the earth are outlined on the accompanying CD-ROM under "Propagation Over the Earth Through the Nonionized Atmosphere."

The general topic of reflection from the surface of the earth and a description of the influence-reflected signals have on the performance of telecommunication systems is given in Refs. 7 and 8.

Diffraction and Scattering

The spherical shape of the earth and irregularities of terrain give rise to *diffraction* as an important part of the ground wave and as a mechanism for propagation beyond the optical horizon. Diffraction is included in the methods for calculation given on the accompanying CD-ROM under "Propagation Over the Earth Through the Nonionized Atmosphere."

Scattering takes place from the rough surface of the earth and from small-scale irregularities in the index of refraction of the atmosphere or the ionosphere. It is analogous to scattering of light, although the radio problem is complicated by the wide range of relationships of radio wavelength to size of irregularity.

Scattering is discussed on the accompanying CD-ROM (Tropospheric forward scattering under "Propagation Over the Earth Through the Nonionized Atmosphere," and ionospheric scattering under "Propagation Via the Ionosphere").

Reciprocity

Reciprocity in wave propagation means that the source and receiver can be interchanged, with the transmission loss and phase unaffected by direction of propagation. In most radio-wave-propagation problems, with the notable exception of those involving an ionized medium with magnetic field, such reciprocity obtains. The refraction index of the ionosphere depends on magnetic field effects; the direction of propagation of the wave affects attenuation, phase, and bending, and the medium is called *anisotropic*. Thus, especially at very low and medium frequencies propagated via the ionosphere, reciprocity does not obtain. Reciprocity does not in any case imply the same signal-to-noise ratio in both directions. The noise environment may be very different at the transmitting and receiving locations.

Transmission Loss; Free-Space Attenuation; Field Strength; Power Flux Density

The concept of transmission loss for radio links consisting of a transmitting antenna, a receiving antenna, and the intervening propagation medium is addressed and defined in *Recomm. ITU-R P.341*.[69]

In general terms, the transmission loss on a radio link between a transmitter and a receiver is defined by the ratio between the power supplied by the transmitter and the power available at the receiver input. The transmission loss depends on several factors such as the losses in the antennas or in the transmission feed lines, the attenuation in the propagation medium, the losses because of faulty adjustment of the impedance or polarization, and so forth. As a consequence there are several definitions employed to characterize transmission loss and its components.

The *system loss* (L_s) on a radio link is the ratio, usually expressed in decibels, of the radio-frequency power input to the terminals of the transmitting antenna p_t, and the resultant radio-frequency signal power available at the terminals of the receiving antenna p_a. The available power is the maximum real power that a source can deliver to a load, i.e., the power that would be delivered to the load if the impedances were conjugately matched. The system loss may be expressed by*

$$L_s = 10\ \log(p_t/p_a) = P_t - P_a \quad \text{dB} \tag{35}$$

and as defined here excludes losses in feeder lines but includes all losses in radio-frequency circuits associated with the antennas, such as ground losses, dielectric losses, antenna loading coil losses, and terminating resistor losses.

The *transmission loss* (L) on a radio line is the ratio, usually expressed in decibels, between the power radiated by the transmitting antenna and the power that would be available at the receiving antenna output if there were no loss in the radio-frequency circuits, assuming that the antenna radiation diagrams are retained. The transmission loss may be expressed by

$$L = L_s - L_{\text{tc}} - L_{\text{rc}} \quad \text{dB} \tag{36}$$

where L_{tc} and L_{rc} are the losses, expressed in decibels, in the transmitting and receiving antenna circuits, respectively, excluding the dissipation associated with antenna radiation, i.e., L_{tc} and L_{rc} may be expressed by $10\ \log(r'/r)$, where r' is the resistive component of the antenna circuit and r is the radiation resistance.

The *basic transmission loss* (L_b) on a radio link is the transmission loss that would occur if the antennas were replaced by isotropic antennas with the same polarization as the real antennas, the propagation path being retained, but the effects of obstacles close to the antennas being disregarded. An isotropic antenna is one that radiates (or receives radio energy equally in (or from) all directions. In determining the basic transmission loss, any local features, such as the ground or nearby structures, which affect the power gain and directivity of the antenna, but which do not affect the overall propagation path, are assumed to be removed. The effect of the local ground is included in computing the antenna gain, but not in L_b. For instance, in the case of ionospheric propagation using an antenna near the ground which has a strong influence on the effective gain for the sky-wave path,

*Capital letters are used to denote quantities in decibels.

the ground is removed for the calculation of L_b so as to maintain the gain in the desired direction. In the case of a tropospheric propagation path involving diffraction over a distance obstacle, that obstacle is not removed in estimating L_b.

The *free-space basic transmission loss* (L_{bf}) is the transmission loss that would occur if the antennas were replaced by isotropic antennas located in a perfectly dielectric, homogeneous, isotropic, and unlimited environments, the distance between the antennas being retained. At a distance d very much greater than the wavelength λ, the power flux density (field intensity), expressed in watts per square meter, is simply $p_t'/4\pi d^2$ since the power p_t' is radiated uniformly in all directions. The effective absorbing area of the isotropic receiving antenna is $\lambda^2/4\pi$, and the available power at the terminals of the loss-free isotropic receiving antenna is given by

$$p_a' = \frac{\lambda^2}{4\pi}\frac{p_t'}{4\pi d^2} \tag{37}$$

Consequently, the free-space *basic transmission loss* can be expressed by

$$L_b = 20\log\frac{4\pi d}{\lambda} \quad \text{dB} \tag{38}$$

d and λ are expressed in the same units.

A practical form of this equation is

$$L_b = 32.45 + 20\log f_{\text{MHz}} + 20\log d_{\text{km}} \quad \text{dB} \tag{39}$$

In many cases, it is important to know the *loss relative to free space* (L_m) on a radio link which is the difference between the basic transmission loss and the free-space basic transmission loss, expressed in decibels, and may be expressed by

$$L_m = L_b - L_{\text{bf}} \quad \text{dB} \tag{40}$$

The loss relative to free space may be divided into losses of different types, such as

- Absorption loss (ionospheric, atmospheric gases or precipitation)
- Diffraction loss as for ground waves
- Effective reflection or scattering loss, as for ionospheric propagation and including the results of any focusing or defocusing because of curvature of a reflection layer
- Polarization coupling loss, which can arise from any polarization mismatch between the antennas for the particular ray path considered
- Aperture to medium coupling loss or antenna gain degradation, which may be a result of the presence of substantial scatter phenomena on the path
- Effect of wave interference between the direct ray and rays reflected from the ground, other obstacles, or atmospheric layers

For broadcasting and mobile services, where characteristics and locations of receiving installations vary, it is convenient to calculate the electric *field strength* at some distance form the transmitter. The rms field strength e (V/m) of a plane wave of wavelength λ (m) is related to the power p_a' (W) available from an ideal loss-free isotropic receiving antenna by

$$e = (480\,\pi^2 p_a'/\lambda^2)^{1/2} \tag{41}$$

It then follows from Eq. (37) that the field strength produced by a transmitter having an *equivalent isotopically radiated power* (EIRP) of p_t (W), at the distance d (m) sufficiently large for the wavefront to be considered plane, is given by

$$e = (30 p_t'/d^2)^{1/2} \tag{42}$$

or in more practical units,

$$e(\text{mV/m}) = 173[p_t' \text{ (kW)}]^{1/2}/d \text{ (km)} \tag{43}$$

In decibels, the equivalent expression is

$$E\,[\text{dB}(\mu\text{V/m})] = 105 - 20 \log d \text{ (km)} + P_t' \tag{44}$$

where p_t is the EIRP in dB(kW).

In the above expressions, the EIRP includes the gain of the transmitting antenna with respect to an isotropic antenna. Often it is more convenient to express antenna gain with respect to a standard antenna other than an isotope, and an appropriate conversion must be made in order to determine the EIRP. For example, if the gain is given with respect to a half-wave dipole, the gain must be multiplied by 1.64 or, in decibels, be increased by 2.1 dB to obtain the EIRP. Alternatively, the following formula may be used

$$E \approx 7(p_t'')^{1/2}/d \tag{45}$$

where E is the field strength in V/m, d is the distance in m, and p_t'' is the effective radiated power (ERP), which includes the transmitter gain with respect to a half-wave dipole.

Similarly, the antenna gain may be expressed with respect to a short vertical monopole above a perfectly conducting ground plane. In this case, the gain must be multiplied by 3 or, in decibels, be increased by 4.8 dB to obtain the EIRP.

For microwave and satellite services, power flux density is a convenient term. The *power flux density* is given by the characteristic relations of a plane wave

$$p = e^2/\eta \quad \text{W/m}^2 \tag{46}$$

where η is the characteristic impedance of the medium in which the measurement is made ($\eta_0 = 120\pi$ in free space). In free space

$$p = e^2/120\pi \quad \text{and} \quad p = 4\pi p_a/\lambda^2 \tag{47}$$

where p_a is available power received by an isotropic antenna in the field.

The relations between field strength, power flux density, and available power in the receiving antenna are outlined below.

The absorbing area of a receiving antenna with gain g_r relative to an isotropic antenna can be written

$$a_e = \lambda^2 g_r r_f / 4\pi r \tag{48}$$

where λ = wavelength in medium
r = radiation resistance of antenna
r_f = radiation resistance of antenna in free space

Combining the above two equations, we find the following formula for the available power p_a' from a lossless receiving antenna:

$$p_a' = e^2\lambda^2 g_r r_f / 4\pi\eta r = v^2/4r \tag{49}$$

The v in this equation denotes the open-circuit voltage induced in the receiving antenna. The field strength is related to the open-circuit voltage by

$$v = e\sqrt{\lambda^2 g_r r_f/\pi\eta_0} = el \tag{50}$$

Field-strength meters usually are calibrated in terms of the effective length l of the antenna.

The relation between the available power p_a from the receiving antenna (neglecting losses) and the field strength e can be expressed in decibels as

$$E = 10 \log[4\pi\eta_0 p_a \times 10^{12})/\lambda^2 g_r] = P_a + 20 \log f - G_r + 107.22 \text{ dB } (\mu\text{V/m}) \tag{51}$$

where f is in megahertz.

For field strength E in decibels referred to 1 μV/m, referred to 1 kW radiated from a half-wave dipole over perfectly conducting earth, propagation loss is

$$L_p = 139.4 - G_r + 20 \log f - E \quad \text{dB} \tag{52}$$

If the reference radiation is a short electric dipole, the constant becomes 136.0.

Power Fading and Time-Variant Multipath Fading

Random variations appear in the signal received via various transmission media, especially at frequencies above about 100 kHz when propagation is by the troposphere or ionosphere. Such variation is usually of two types: one is *attenuation*, or *power fading*, which may be quite slow (minute to minute, hour to hour, and so forth) and is associated with comparatively large-scale changes in the medium, such as absorption; the other is *variable-multipath* or *phase-interference fading*.

Power fading is usually allowed for in the power margin designed into the system. Phase-interference fading, on the other hand, affects not only the amplitude but also the variable phase-vs.-frequency characteristic of the channel, limiting coherence bandwidth and introducing extraneous fluctuation in received-signal parameters. Alleviation of the effects of variable multipath is possible by diversity techniques, signal design, and signal receiving, processing, and detection operations.

The amplitude probability distribution of the fading envelope is usually determined from samples of duration much shorter than the shortest fade duration; observation intervals over which statistical averages are taken are about 1000 times the reciprocal of the nominal fading rate. The fit of experimental distributions of envelope fading to the Rayleigh distribution is often excellent for ionospheric and tropospheric scatter propagation, and similarly to a Nakagami-Rice[8] distribution for situations where a specular component is mixed with scattered components. (See also Recomm. ITU-R PN. P.1057).

It is often necessary, however, to consider the combination of long-term (power) fading, represented by a log-normal distribution, with short-term variations, represented by a Rayleigh distribution. The distribution of instantaneous values over a long period can be obtained from the Rayleigh law, whose mean is itself a random variable having a log-normal distribution.

A description of the fundamental properties of the most common probability distribution used in the statistical study of radio-wave propagation is given in Recomm. ITU-R PN. P.1057.

Most theoretical treatments of communication performance in the presence of variable multipath fading resulting from several signal components have been carried out for channels characterized by Rayleigh envelope distribution. The probability density function is given by

$$p(V) = (2V/v^2)\exp[-(V/v)^2] \tag{53}$$

where V is the fluctuating envelope and v^2 is the mean square value of V over distribution. For the Rayleigh fading channel, the probability that the received signal envelope will fall at or below some specified value of V is given by the cumulative distribution

$$p(V \le V') = \int_0^{V'} \frac{2V}{v^2} \exp\left[-\left(\frac{V}{v^2}\right)^2\right] dV = 1 - \exp\left[-\left(\frac{V'}{v}\right)^2\right] \tag{54}$$

The Rayleigh probability distribution function Eq. (54) is often used in the form

$$p(V \le V') = 1 - \exp[-0.693(V'/V_M)^2] \tag{55}$$

where V_M is the median value, about 1.6 dB below the rms value. *Fading rate* is important to certain systems. One measure of fading rate is the number of times per second (hertz) the carrier envelope crosses its median with a positive or negative slope. Another measure is the width of the received carrier-envelope spectral density.

It is also important to obtain information on fade durations; the effect of a large number of very short fades is quite different from that of a few long-duration fades, the former being serious for digital links, the latter for analog links. The rate of change of fade is important in the design of switching systems for diversity reception.

For some time, design concern (besides its emphasis on the amplitude variation) has centered on the dispersion and multipath characteristics of the medium, in terms of linear time-variant amplitude and phase- and frequency-distortion parameters, often referred to as *multiplicative noise*, which cannot be overcome by power increase.

A more comprehensive characterization is in terms of the *system function* or *impulse function*. This approach relates the response and excitation of a channel at its input and output terminals.[12,24,25]

The expressions for output are formulated in terms of operations on the input time function $x(t)$ or the spectral function (Fourier transform) $X(\omega)$, to produce the output function $y(t)$ or $Y(\omega)$. Each path is characterized by a system function $h(t, \tau)$ that operates on the replica of $x(t)$ traversing it; t is the time at which the observation is made, and τ is the delay or transit time for the path. The spread of delays between the input and output is determined by the system function $h(t, \tau)$, which can be called the *delay-spread system function*. For any particular elemental path x in a distribution of paths that covers some range of delays, the output for a range of delay $\Delta\tau$ centered on τ is given by $h(t, \tau)x(t - \tau)\,\Delta\tau$, and the total output of the channel is the sum of all such weighted and delayed contributing paths, namely.

$$y(t) = \int_{-\infty}^{\infty} x(t-\tau)h(t, \tau)\, d\tau \tag{56}$$

For a Fourier transformable input $x(t)$, the output can be expressed in terms of the input spectral function $X(i\omega)$:

$$x(t) = \int_{-\infty}^{\infty} X(i\omega)\exp(i\omega t)d(\omega/2\pi) \tag{57}$$

Here we characterize the channel by stating that it modifies the contribution to the structure of $x(t)$ from spectral components in the range $\Delta\omega$ centered at ω by multiplying it by the transfer function $H(i\omega, t)$. The total channel response to $x(t)$ is then

$$x(t) = \int_{-\infty}^{\infty} H(i\omega, t)X(i\omega)\exp(i\omega t)d(\omega/2\pi) \tag{58}$$

which is the limit of the sum of the channel responses to the components of $x(t)$ from various infinitesimally wide spectral elements. The time-variant, *frequency-dependent transfer function* $H(i\omega, t)$ can be shown to be the *Fourier transform over the delay-spread variable* τ of the *delay-spread function* $h(t, \tau)$.

The randomness of the channel with time is reflected in the treatment of the system functions $h(t, \tau)$ and $H(i\omega, t)$ as sample functions of processes that are random over the space of the time variable t. The autocorrelation function of the channel response process is given by an inverse Fourier transform operation on the product of the spectral density function of the input process and the autocorrelation function of the time-variant, frequency-dependent transfer function $H(i\omega, t)$ of the channel. Further transformations produce a combined time-shift and frequency-shift correlation function and a so-called *scattering function* $S(\tau_0, f_0)$, which has the physical significance of a function that determines the weighting of the signal power as a function of the time delay τ_0 and Doppler shift f_0 incurred in transmission.

On the basis of the above functions, a set of transmission parameters for random time-variant linear filters is defined.[25]

1. The *multipath spread* or *delay spread* is determined by the relative delays of the component paths, or the "duration" of $h(t, \tau)$ over the delay variable τ.
2. The *coherence bandwidth*, the bandwidth over which correlation of amplitude fading (or coherence of phase for some applications) remains to a desired degree is usually defined in terms of specified degradation of error rate, distortion, or other parameter.

3. *Diversity bandwidth* is the frequency separation between two sinusoidal inputs which results in a specified decorrelation of the fluctuating responses, usually taken to be a correlation coefficient of $1/e$.
4. The *fading rate, fading bandwidth,** *frequency smear,* or *Doppler spread* is a measure of the bandwidth of the received signal when the input to the channel is a stable single-frequency signal.
5. The *diversity time* (or decorrelation time) is a measure of the time separation between input signals to yield correlation of less than $1/e$ between the envelopes of the responses.

These parameters are not all independent; the coherence bandwidth and delay spread are inversely proportional to each other, as are fading bandwidth and decorrelation time.

Most channel characteristics can be measured. The delay-spread response $h(t, \tau)$ can be measured directly by transmitting very short, widely spaced pulses; each received replica will correspond to one path, which can be resolved to examine the relative amplitudes and delays. The amplitude characteristic of the frequency-dependent transfer function $|H(\omega, t)|$ can be measured for short intervals by transmitting a constant-amplitude test signal with repetitive linear sweep covering the desired frequency range. The envelope of the received signal will give a very close approximation to $|H(\omega, t)|$.

Diversity techniques for counteracting short-term fading are used extensively for HF ionospheric communication, forward scatter systems, microwave line-of-sight systems, and Earth-space systems, where high reliability is required. The most common mechanism is to use *spaced antennas*, taking advantage of the fact that fading at one antenna tends to be independent of the signal fluctuation received on another antenna; provision is made to switch between signals or to combine two or more of them. Depending on the propagation mechanism, other kinds of useful diversity include *frequency, angle of arrival, polarization*, and *time*. Line-of-sight links frequently employ vertical space diversity where vertical separation of the antennas is found to be more effective than horizontal spacing. A simple, generally effective design procedure for this so-called height diversity gives the required vertical spacing between the centers of the antennas as

$$\Delta h = 0.3\sqrt{\lambda d} \tag{59}$$

where the path length d, wavelength λ, and the spacing Δh are all in the same units. A more thorough treatment is found in Recomm. ITU-R P.530.[69]

At frequencies affected by rain, i.e., above about 10 GHz, route diversity is used in which the spacing between the terminals is sufficiently large so that rain affecting one route is unlikely to affect the other.

The performance of a diversity system may be quantified by two parameters, *diversity gain* or *diversity advantage*. The diversity gain of a system is the ratio of the power output at a given time percentage to the corresponding power that would be obtained from a single channel. The diversity advantage is the ratio of the time percentage for which a given fade level is exceeded for a single channel to that for a diversity system. Both parameters may be obtained from cumulative distributions of fade statistics.

Recommendations ITU-R P.530 and 618[69] give expressions for diversity improvement under different propagation conditions for line-of-sight and Earth-space links, respectively, with further discussion given elsewhere.[7,8,14]

The outline of fading and diversity improvement given here has assumed *flat* (nonfrequency-selective) fading and Gaussian (white) noise. References 11, 14, and 25 discuss frequency-selective fading for the various types of system. Non-Gaussian noise effects are considered in the next paragraph.

Noise: Signal-to-Noise Ratio

Several types of radio noise must be considered in any design, though, in general, one type will be the dominant factor. In broad categories, the noise can be divided into two types: noise internal to the receiving system and noise external to the receiving antenna.

The *noise of the receiving system* is often the controlling noise in systems operating above about 300 MHz. This type of noise is a result of antenna losses, transmission-line losses, and the circuit noise of the receiver itself and has the characteristics of thermal (i.e. white) noise and thus is Gaussian in nature.[10]

**Fading bandwidth* is also often used in the same sense as diversity bandwidth, or coherence bandwidth, above.

The second of the broad categories, *external radio noise*, can be subdivided further into natural and artificial sources. Natural sources of radio noise are: (1) atmospheric, (2) galactic, (3) solar noise from antennas pointing at the sun, (4) precipitation (blowing snow or dust), (5) corona, and (6) noise reradiating from any absorbing medium through which the wanted radio signal passes. Very low noise systems used in space communications can be limited by such absorption by clouds, water vapor, and oxygen. Such *sky noise* has the Gaussian characteristic of receiver noise. Examples of artificial noise sources are: (1) power lines or generating equipment, (2) automotive ignition systems, (3) fluorescent lights, (4) switching transients, and (5) electrical equipment in general. Unlike internal noise, external noise is generally non-Gaussian, being impulsive in nature.

Noise power is generally the most significant single parameter in relating the interference value of the noise to system performance. This parameter, however, is seldom sufficient in the case of impulsive noise, and a more detailed statistical description of the received-noise waveform is generally required.

A useful parameter for expressing the noise power external to the antenna is the effective antenna noise factor f_a which is defined by

$$f_a = p_n/Kt_0b = t_a/t_0 \tag{60}$$

where p_n = noise power available from an equivalent loss-free antenna (W)
K = Boltzmann's constant, 1.38×10^{-23} J/K
t_0 = reference temperature, taken as 288 K
b = effective receiver noise bandwidth, Hz
t_a = effective antenna temperature in the presence of external noise

The noise factor f_a is commonly given as the corresponding noise figure F_a, which is given by $F_a = 10 \log f_a$(dB).

Figure 16.3.3 shows the median value of the *available noise-power*, F_a (dB above Kt_0), from various sources. While the solar noise, galactic noise, and sky noise are Gaussian, the atmospheric and artificial noises are very impulsive.

Based on measurements from a worldwide network of stations, Recomm. ITU-R P.372[69] gives detailed estimates of F_a for atmospheric noise as a function of geographic location, season, frequency, and time. Additional data are also provided on noise variability and character which include the standard deviation of F_a, upper and lower deciles of F_a with standard deviations, and estimates of V_d, the ratio of the rms envelope voltage to the average noise envelope voltage. The values of F_a are for a lossless short vertical antenna over a perfectly conducting ground plane; they are related to the vertical rms field strength by

$$E_n = F_a - 95.5 + 20 \log f_{\mathrm{MHz}} + 10 \log b \tag{61}$$

where E_n is the rms noise field strength [dB(μV/m)] in bandwidth b (Hz), and F_a is the noise figure for the center frequency, f_{MHz}.

Estimates of the noise-power spectral density from artificial noise expected in business, residential, rural, and quiet rural areas have been developed from measurements.[27–29] These expected values are the means of a number of location medians, with variation from location to location in each type of area and temporal variation indicated. Generally the noise below 20 MHz is associated with power lines. At 20 MHz and above, automotive electrical systems, especially ignition systems, are the dominant sources in all but rural locations. Median values of artificial noise power expressed in terms of F_a are given in Recomm. ITU-R P.372[69] for different categories of environment, of which curve 3 and 4 in Fig. 16.3.3 are examples.

Impulsive atmospheric or *artificial noise* disturbs communications in a way quite different from Gaussian noise. Figure 16.3.4 shows the amplitude probability distribution of Gaussian noise and a sample of atmospheric noise. The parameter V_d is the ratio in decibels of the rms voltage to the average voltage and is commonly used as an *impulsive index*. The two distributions are plotted relative to their rms level; i.e., both noises shown have the same energy or power. The probability distribution of the noise envelope determines the performance of most basic digital receivers, as indicated by the two error-rate curves for a binary coherent phase-shift-keying system.

Digital receivers are frequently designed for optimum performance in white Gaussian noise. Their performance in impulsive artificial or atmospheric noise can be summarized as follows, with comparisons made on the basis of equal noise power:

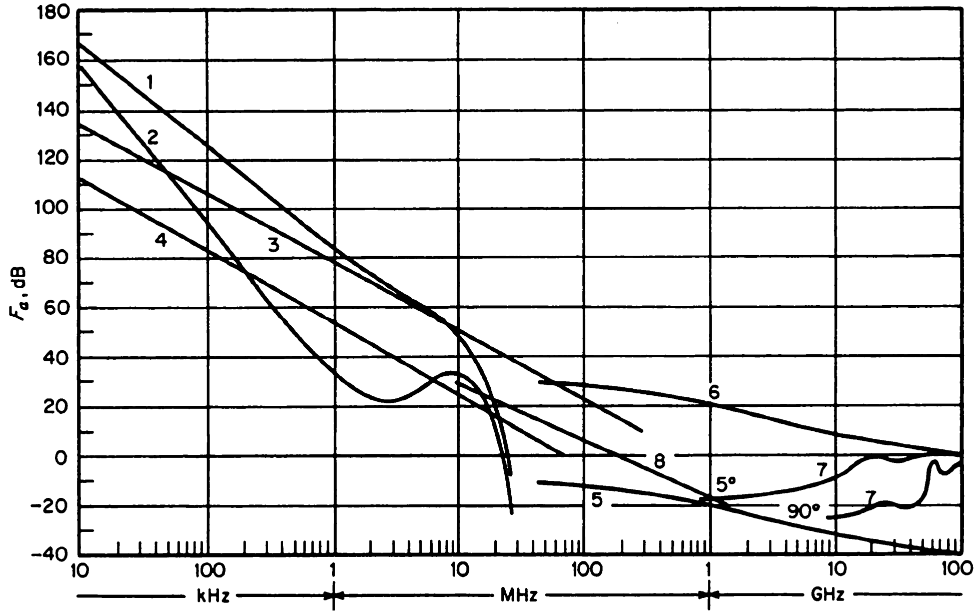

FIGURE 16.3.3 Median radio noise-power spectral density from various sources. Curve 1 = atmospheric noise, summer, 2000 to 2400 h. Washington, D.C., omnidirectional antenna near ground; curve 2 = atmospheric noise, winter 0800 to 1200 h. Washington, D.C., omnidirectional antenna near ground; curve 3 = artificial noise, business area, omnidirectional antenna near ground; curve 4 = artificial noise, quiet rural areas, omnidirectional antenna near ground; curve 5 = quiet sun, isotropic (0 dB gain) antenna; curve 6 = disturbed sun, isotropic (0 dB gain) antenna; curve 7 = sky noise, narrow-beam antenna (degrees from vertical); curve 8 = galactic noise, omnidirectional antenna near ground.

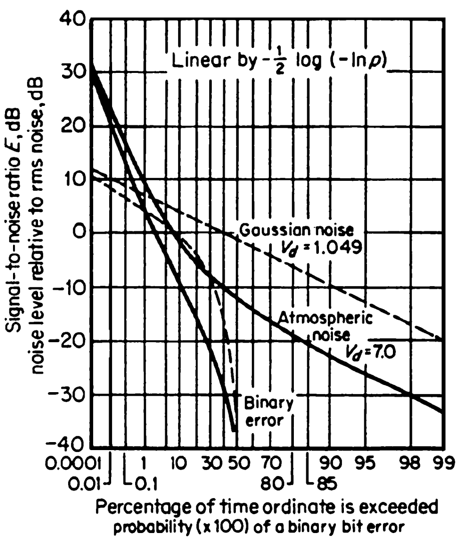

FIGURE 16.3.4 Comparison of noise distribution and error probabilities for Gaussian and non-Gaussian noise (same noise power, coherent phase-shift keying).

1. For constant signal, at high signal-to-noise (S/N) ratio, impulsive noise causes more errors than Gaussian noise; at lower S/N ratio, Gaussian noise causes more errors.
2. For Rayleigh-fading signals, Gaussian noise causes more errors at all S/N ratios; flat-fading cases do arise, for which impulsive noise will cause more errors than Gaussian nose; for diversity reception, impulsive noise is more harmful.
3. While pairing of errors in differentially coherent phase-shift keying (DCPSK) becomes more unlikely as the S/N ratio increases, pairing of errors increases as the noise becomes more impulsive.
4. For systems with time-bandwidth products in the order of unity, the standard matched filter-receiver is also optimum for impulsive noise.
5. Noise-suppression schemes, such as wide-band limiting and smear-desmear, are not particularly effective at high S/N ratios.
6. Receivers especially designed to reject a particular type of impulsive noise perform substantially better then receivers using the "standard" noise-suppression techniques.

TABLE 16.3.3 ITU-R Recommendations

	Title	Year
Recommendation ITU-R		
P.313	Exchange of information for short-term forecasts and transmission of ionospheric disturbance warnings	1995
F.339	Bandwidths, signal-to-noise ratios and fading allowances in complete systems	1990
P.368	Ground-wave propagation curves for frequencies between 10 kHz and 30 MHz	1992
P.369	Reference atmosphere for refraction	1994
P.370	VHF and UHF propagation curves for the frequency range from 30 to 1000 MHz. *Broadcasting services*	1995
P.371	Choice of indices for long-term ionospheric predictions	1995
P.372	Radio noise	1994
V.431	Nomenclature of the frequency and wavelength bands used in telecommunications	1994
P.434	ITU-R reference ionospheric characteristics and methods of basic MUF, operational MUF and ray-path prediction	1995
P.435	Sky-wave field-strength prediction method for the broadcasting service in the frequency range 150 to 1600 kHz	1992
P.452	Prediction procedure for the evaluation of microwave interference between stations on the surface of the earth at frequencies above about 0.7 GHz	1995
P.453	The radio refractive index: its formula and refractivity data	1995
P.526	Propagation by diffraction	1995
P.527	Electrical characteristics of the surface of the earth	1992
P.529	Prediction methods for the terrestrial land mobile service in the VHF and UHF bands	1995
P.530	Propagation data and prediction methods required for the design of terrestrial line-of-sight systems	1995
P.531	Ionospheric effects influencing radio systems involving spacecraft	1994
P.532	Ionospheric effects and operational considerations associated with artificial modification of the ionosphere and the radio-wave channel	1992
P.533	HF propagation prediction method	1995
P.534	Method for calculating sporadic-E field strength	1990
P.617	Propagation prediction techniques for data required for the design of trans-horizon radio-relay systems	1992
P.618	Propagation data and prediction methods required for the design of Earth-space telecommunications systems	1995
P.676	Attenuation by atmospheric gases	1995
P.684	Prediction of field strength at frequencies below about 500 kHz	1994
P.832	World atlas of ground conductivities	1992
P.834	Effects of tropospheric refraction on radio-wave propagation	1994
P.837	Characteristics of precipitation for propagation modeling	1994
P.838	Specific attenuation model for rain for use in prediction methods	1992
P.840	Attenuation due to clouds and fog	1994
P.842	Computation of reliability of HF radio systems	1994
P.843	Communication by meteor-burst propagation	1992
P.844	Ionospheric factors affecting frequency sharing in the VHF (30–300 MHz) band	1994
P.1057	Probability distributions relevant to radio-wave propagation modeling	1994
Other texts		
CCIR Report 340-4 (Geneva, 1983) CCIR Atlas of ionospheric characteristics (deleted)		
CCIR: Handbook of Curves for Radio Wave Propagation over the Surface of the Earth (Geneva, 1991)		

The performance of analog voice systems in impulsive noise can be summarized as follows:

1. For a given articulation index, a much lower S/N ratio is required for impulsive noise than for Gaussian noise.
2. Various forms of limiting in AM systems (pre-i.f. limiting, i.f. limiting, postdetection limiting) are quite effective in further reducing the required S/N ratio when the noise is impulsive.

As mentioned above, noise power is also generated from atmospheric gases, clouds, and rain. Recomm. ITU-R P.372[69] discusses radio emission at frequencies above about 50 MHz from natural sources, not only in the atmosphere but also from extraterrestrial sources and from the surface of the earth. This topic is discussed on the accompanying CD-ROM under "Propagation Over the Earth Through the Nonionized Atmosphere."

Minimum external noise levels to be expected at terrestrial receiving sites owing to natural and artificial noise sources are also specified in Recomm. ITU-R P.372.[69] With such data, the appropriate minimum receiver noise figure can be determined so that a terrestrial receiving system can be designed to be almost limited by external noise. Finally, Recomm. ITU-R F.339[69] gives ratios of required signal energy to noise power spectral density for various systems operating in the presence of atmospheric noise.

Ionospheric Modification by High-Power Radio Transmissions

High-power radio waves can modify the ionospheric plasma by classical ohmic heating, which changes the density and distribution of the ionization, and by generating parametric instabilities which in turn leads to field-aligned ionospheric irregularities (see Recomm. ITU-R P.532, Table 16.3.3).[69] At HF, ionospheric modification experiments generally use purpose-built transmitters, operating close to the *F*-region critical frequency, to modify the upper ionosphere (150–400 km). The ionosphere can also be appreciably modified by oblique high-power transmission at frequencies considerably in excess of the critical frequency. However, transmitters operating over the range from VLF to UHF can give rise to modifications in all regions of the ionosphere, with the resulting modified region having an effect on other radio signals passing through it. In particular, the scattering properties of ionospheric irregularities, artificially produced in the *F*-region, have been used to establish communications up to UHF between two points on the earth's surface. Such experiments demonstrate the interference potential to existing services from such mechanisms.

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69. ITU-R See Table 16.3.3 for a list of Recommendations and other texts of the ITU Radiocommunication Sector.

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