Modern wireless communication

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Chapter 2

Breed and the noise

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2.1 introductions

- Physical model
- Reproduction of the free space
- Reflection
- Refraction
- Count models
- The noise and effect of interfering.

- The electric signal describing
- Through the information wanted of the space, the reproduction of the radiowave
- Estimate from the electric signal that is resumed that sends the receiver of the message.
- The transmission system is
- Display the characteristic of the aerial changed between electric signal and radiowave.
- Assume:
- Media that it may be taking decay or overlapping of different signals as a kind of line of the characteristic that all out of shape.

2.2 Spread of the free space

2.2.1 Isotropic radiation

- Isotropic aerial conveys equally from all directions .
- In fact, isotropic aerial does not exist .
- All aerial let people associate some directivity with them

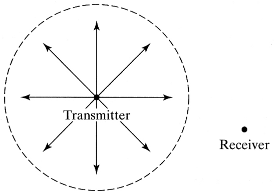

FIGURE 2.2 Illustration of isotropic radiation.

- The power per unit area, or power *flux density* on the surface of a sphere of radius R centered on the source is given by

$$\Phi_R = \frac{P_T}{4\pi R^2} \tag{2.1}$$

- The power P_R received by the antenna depends on the size and orientation of the antenna. The power received by an antenna of *effective area* or *absorption cross section* A_e is given by

$$P_R = \Phi_R A_e = \frac{P_T}{4\pi R^2} A_e \tag{2.2}$$

- An antenna' s Physical area A and its effective area A_e are related by the antenna efficiency

$$\eta = \frac{A_e}{A} \qquad (2.3)$$

- From electromagnetic theory, effective area of an isotropic antenna in any direction is given by

$$A_{ISO} = \frac{\lambda^2}{4\pi} \qquad (2.4)$$

- Relationship between transmitted and received power for isotropic antennas:

$$P_R = \frac{P_T}{(4\pi R/\lambda)^2} = \frac{P_T}{L_P} \qquad (2.5)$$

- Path loss L_P is the free-space path loss between two isotropic antennas, which defined as

$$L_P = \left(\frac{4\pi R^2}{\lambda}\right)^2 \qquad (2.6)$$

2.2 Free-Space Propagation

2.2.2 Directional Radiation

- Most of the antennas have gain or directivity $G(\theta,\phi)$ that is a function of azimuth angle θ the elevation angle ϕ :
- The azimuth angle θ the look angle in the horizontal plane of the antenna relative to a reference horizontal direction
- The elevation angle ϕ is the look angle of the antenna above the horizontal plane

- Transmit gain of an antenna is defined as

$$G_T(\theta,\phi) = \frac{\text{Power flux density in direction } (\theta,\phi)}{\text{Power flux density of an isotropic antenna}} \qquad (2.7)$$

- The corresponding definition for the receive antenna gain is

$$G_R(\theta,\phi) = \frac{\text{Effective area in direction } (\theta,\phi)}{\text{Effective area of an istopic antenna}} \qquad (2.8)$$

- Principle of reciprocity
- Signal transmission over a radio path is reciprocal in the sense that the locations of the transmitter and receiver can be interchanged without changing the transmission characteristics.
- From the principle of reciprocity and definition for the receive antenna gain, the maximum transmit or receive gain of an antenna in any direction is given by

$$G = \frac{4\pi}{\lambda^2} A_e \qquad (2.9)$$

2.2 Free-Space Propagation

2.2.3 The Friis Equation

- When nonisotropic antennas are used, the free-space loss relating the received and transmitted power for general antennas is

$$P_R = \frac{P_T G_T G_R}{L_P} \qquad (2.11)$$

- Rewrite it as decibel equation

$$P_R(dB) = P_T(dB) + G_R(dB) + G_T(dB) - L_P(dB) \qquad (2.12)$$

- Friis equation is the fundamental link budget equation.
- Closing the link refers to the requirement that the right-hand side provide enough power at the receiver to detect the transmitted information reliably.

2.2 Free-Space Propagation

2.2.4 Polarization

- Electromagnetic waves are transmitted in two orthogonal dimensions, referred to as polarizations. Two commonly used orthogonal sets of polarizations are:
- Horizontal and vertical polarization.
- Vertical polarization is used for terrestrial mobile radio communications.
- At frequencies in the VHF band, vertical polarization is better than horizontal polarization.
- Left-hand and right-hand circular polarizations.
- It used in satellite communication.
- For well designed fixed communication link, two orthogonal polarizations can be used to double the transmission capacity in a given frequency ban d.

2.3 Terrestrial Propagation: Physical Models

- In terrestrial communication, buildings, terrain or vegetation may obstruct the line-of-sight path between transmit and receive antennas.
- Communication relies on either reflection or diffraction.
- Multitude of possible paths arise.
- Multipath propagation:
- With multiple waves arriving at the same location and either destructive or constructive interference happened.

- *Ducting occurs when the physical characteristics of the environments create a waveguide like effect.*
- *High frequency radio has carrier frequencies range from 3 to 30MHz.*
- *Differences between the layers of the atmosphere are most pronounced and the layers act like different media to the transmission.*
- *The HF signal can be trapped within a layer of ionosphere.*
- *This allows signal to travel long distances with very little attenuation and is sometimes called skywaves.*

2.3 Terrestrial Propagation: Physical Models

2.3.1 Reflection and the Plane-Earth Model

- The flat-Earth model shows a fixed transmitter with an antenna of height h_T transmitting to fixed receiving antenna of height h_R .

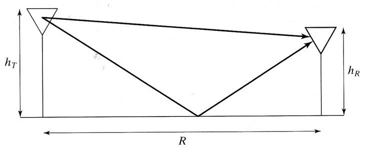

FIGURE 2.5 Plane-Earth reflection model.

2.3 Terrestrial Propagation: Physical Models

2.3.1 Reflection and the Plane-Earth Model

- Power flux density and the field strength E:

$$\phi = \frac{|E|^2}{\eta_0} \tag{2.13}$$

- The field strength is measured in volts per meter and η_0 is the characteristic wave impedance of free space.
- The antenna acts as impedance transformer. Assume electric field is generated by continuous wave sign al with frequency f, at given point of space,

$$E(t) = \sqrt{2}E_0 \cos(2\pi ft + \theta) = \sqrt{2}\,\mathrm{Re}\left\{E_0 e^{j(2\pi ft+\theta)}\right\} \tag{2.14}$$

- Where field strength E_0 and phase θ depends on the location in space and $\mathrm{Re}\{\}$ denotes the real part of the quantity inside the parentheses.

- For simplification, we define the complex phasor

$$\tilde{E} = E_0 e^{j\theta} \tag{2.15}$$

- In the case of the reflection of a single ray, if $\tilde{E}_d$ is the reflected field, the relationship between the two fields is given by

$$\tilde{E} = \tilde{E}_d \rho e^{j\psi} \tag{2.16}$$

- Where ρ is the attenuation of the electric field and ψ is the phase change caused by reflection

- Differences between the direct and reflected paths depends on path length differences. By Pythagorean theorem, the length of direct paths is

$$R_d = \sqrt{R^2 + (h_T - h_R)^2} \qquad (2.17)$$

- Similarly, the length of the reflected path is

$$R_r = \sqrt{R^2 + (h_T + h_R)^2} \qquad (2.18)$$

- To calculate the field strength at receiving antenna, we assume difference in attenuation between direct and ground-reflected wave is negligible. That is,

$$\left|\tilde{E}_r\right| = \left|\tilde{E}_d\right| \qquad (2.19)$$

- The phase difference between two paths is sensitive to the path length and it cannot be neglected. The path difference between the reflected and incident rays is

$$\Delta R = R_r - R_d \qquad (2.20)$$

- If *R* is large compared with both ht and h_R, the length of the direct path can be approximated by

$$\begin{aligned} R_d &= R\sqrt{1 + \frac{(h_T - h_R)^2}{R^2}} \\ &= R\left(1 + \frac{(h_T - h_R)^2}{2R^2}\right) \end{aligned} \qquad (2.21)$$

- From the above equation, approximation $\sqrt{1+x} \approx (1 + x/2)$ for x <<1. Using the same technique for R_r , we have,

$$\Delta R = \frac{(h_T + h_R)^2 - (h_T - h_R)^2}{2R} = 2\frac{h_T h_R}{R} \qquad (2.22)$$

- The phase difference between two paths is proportional to the transmission wavelength and is given by

$$\Delta\phi = \frac{2\pi}{\lambda}\Delta R = \frac{4\pi h_T h_R}{\lambda R} \qquad (2.23)$$

- The total received field is

$$\begin{aligned} \tilde{E} &= \tilde{E}_d + \tilde{E}_r \\ &= \tilde{E}_d(1 + \rho\exp(j\psi)\exp(-j\Delta\phi)) \end{aligned} \qquad (2.24)$$

- If the earth' s surface is assumed to be smooth and flat, and the reflected ray is at grazing incidence so that the reflection coefficient is $\rho\exp(j\psi) = -1$, then the total received electric field is

$$\tilde{E} = \tilde{E}_d(1 - \exp(-j\Delta\phi)) \qquad (2.25)$$

- The magnitude of the combined electric field is given by

$$\begin{aligned} \left|\tilde{E}\right| &= \left|\tilde{E}_d\right|\left|1 - \exp(j\Delta\phi)\right| \\ &= \left|\tilde{E}_d\right|\left[1 + \cos^2\Delta\phi - 2\cos\Delta\phi + \sin^2\Delta\phi\right]^{1/2} \\ &= \left|\tilde{E}_d\right|\left[2 - 2\cos\Delta\phi\right]^{1/2} \\ &= 2\left|\tilde{E}_d\right|\sin\left(\frac{\Delta\phi}{2}\right) \end{aligned} \qquad (2.26)$$

- Substituting Eq. (2.23)into (2.26) yields

$$\left|\tilde{E}\right| = 2\left|\tilde{E}_d\right|\sin\left(\frac{2\pi h_T h_R}{\lambda R}\right) \qquad (2.27)$$

- From Eqs. (2.2)and (2.13), the received power is given by

$$\begin{aligned} P_R &= \phi A_e \\ &= \frac{\left|\tilde{E}\right|^2}{\eta_0}A_e \\ &= 4\frac{\left|\tilde{E}_d\right|^2}{\eta_0}A_e\sin^2\left(\frac{2\pi_T h_R}{\lambda R}\right) \end{aligned} \qquad (2.28)$$

- The factor $|\tilde{E}_d|^2 A_e/\eta_0$ in Eq. (2.28)is the power received via the direct path. It governed by free-space propagation condition and therefore equivalent to the right-hand side of Eq. (2.11). Substituting Eq. (2.6)into (2.11) and using the factor $|\tilde{E}_d|^2 A_e/\eta_0$ in Eq. (2.28),Friis' equation can be modified as

$$P_R = 4P_T\left(\frac{\lambda}{4\pi R}\right)^2 G_T G_R \sin^2\left(\frac{2\pi h_T h_R}{\lambda R}\right) \qquad (2.29)$$

- If the product $\lambda R >> h_T h_R$, we approximate sin θ by θ and Eq. (2.29) becomes

$$P_R = P_T G_T G_R \left(\frac{h_T {h_R}^2}{R^2}\right)^2 \qquad (2.30)$$

- Equation (2.30)is the plane-Earth propagation equation which differs from the free-space equation in three ways:
- As a consequence of the assumption that $R >> h_T h_R$, the angle $\Delta\phi$ is small, and cancels out of the equation, leaving it to be essentially frequency independent.
- It shows inverse fourth-power law, rather than the inverse-square law of free-space propagation.
- It shows the effect of the transmit and receive antenna heights on propagation losses. The dependence on antenna height makes intuitive sense.

2.3 Terrestrial Propagation: Physical Models

2.3.2 Diffraction

- *Diffraction*
- *phenomenon when electromagnetic waves are forced to travel through a small slit and tend to spread ou t on the far end of the slit.*
- *Huygens' s principle*
- *Each point on a wave front acts as a point source for further propagation. However, the point source does not radiate equally in all directions, but favors the forward direction, of the wave front.*

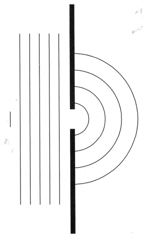

FIGURE 2.6 Behavior of plane wave passing through a slit from left to right.

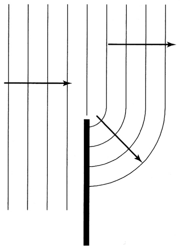

FIGURE 2.7 Illustration of knife-edge diffraction.

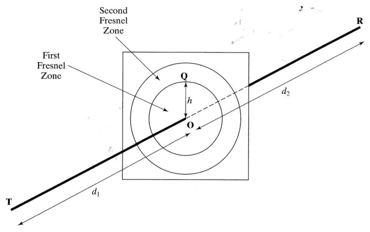

FIGURE 2.8 Fresnel zones of transmitted signal.

- From Fig.2.8, the excess path length between the direct path and the non direct path through the point Q on the circle of radius h is given by

$$\begin{aligned}\Delta R &= |TQR| - |TOR| \\ &= \sqrt{d_1^2 + h^2} + \sqrt{d_2^2 + h^2} - (d_1 + d_2) \\ &= d_1\sqrt{1+\left(\frac{h}{d_1}\right)^2} + d_2\sqrt{1+\left(\frac{h}{d_2}\right)} - (d_1 + d_2) \qquad (2.31)\end{aligned}$$

- Assuming that $h << d_1, d_2$ and using the approximation , $\sqrt{1+x} \approx 1 + x/2 \text{ for } x << 1$,We find that Eq. (2.31)reduces to

$$\Delta R \approx \left(\frac{h^2}{2}\frac{d_1 + d_2}{d_1 d_2}\right) \qquad (2.32)$$

- The phase difference, corresponding to a transmission wavelength λ given by

$$\begin{aligned}\Delta\phi &= \frac{2\pi}{\lambda}\Delta R \\ &= \frac{2\pi}{\lambda}\frac{h^2}{2}\left(\frac{d_1 + d_2}{d_1 d_2}\right) \qquad (2.33) \\ &= \frac{\pi}{2}h^2\left(\frac{2(d_1 + d_2)}{\lambda d_1 d_2}\right)\end{aligned}$$

- The Fresnel-Kirchhoff diffraction parameter

$$v = h\sqrt{\frac{2(d_1 + d_2)}{\lambda d_1 d_2}} \qquad (2.34)$$

- The Fresnel-Kirchhoff diffraction parameter is a dimensionless quantity that characterize diffraction losses in general condition.

- The phase difference of Eq. (2.33)can be written as $\Delta\phi \approx \frac{\pi}{2}v^2$
- According to Fig. 2.8, a family of hypothetical circles are constructed with total path length from **T** to **R** via each circle is $\Delta R = q\lambda/2$
- The radii depend on the location of **TR**-axis.
- The set of points for which excess path length is an integer number of half wavelengths defines a family of ellipsoids with **TOR** being the axis revolution.
- If we slice a specific ellipsoid perpendicular to **TOR**, the radius of resulting circle is

$$h = r_q = \sqrt{\frac{q\lambda d_1 d_2}{d_1 + d_2}} \qquad (2.36)$$

- The corresponding diffraction parameter is $v_q = \sqrt{2q}$

- From Eq. (2.32),this approximation is valid for $d_1, d_2 >> r_q$.
- In words, a particular Fresnel-Kirchhoff parameter v_q defines a corresponding ellipsoid of constant excess path.
- The volume enclosed by the q th ellipsoid is known as the q th Fresnel zone.
- Differential path length $\frac{(q-1)\lambda}{2} \le \Delta R \le \frac{q\lambda}{2}$ Phase difference $(q-1)\pi \le \Delta\phi \le \frac{q\lambda}{2}$
- As a general rule of thumb, we must keep the " first Fresnel zone" free of obstructions in order to obtain transmission under free-space condition.

2.3 Terrestrial Propagation: Physical Models

2.3.3 Diffraction Losses

- In fig.2.8, when an obstruction is present, the Fresnel-Kirchhoff parameter is still given by

$$v = h\sqrt{\frac{2(d_1+d_2)}{\lambda d_1 d_2}} \qquad (2.39)$$

- The height h and thus also v, is considered positive if the obstruction extends above the line of sight and negative if it does not.

- From advanced diffraction theory, the field strength at the point **R** can be expressed as

$$E = E_d \frac{1+j}{2}\int_{v}^{\infty} \exp\left(-j\frac{\pi}{2}t^2\right)dt \qquad (2.40)$$

- where v is the Fresnel-Kirchhoff parameter and $j=\sqrt{-1}$ The Eq. (2.40)is known as the complex Fresnel integral
- For a fixed obstruction at a distance d1 from the transmitter and a transmission wavelength λ , the diffraction parameter

$$v = h\sqrt{\frac{2(d_1+d_2)}{\lambda d_1 d_2}} \qquad (2.41)$$

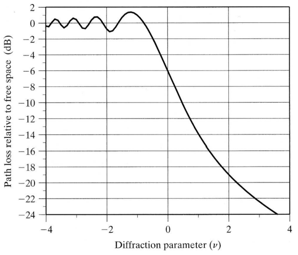

FIGURE 2.10 Diffraction loss over a single knife edge as a function of the Fresnel–Kirchhoff parameter (v).

2.4 Terrestrial propagation: Statistical models

- Statistical approach is used in propagation characteristics are empirically approximated on the basis of measurements in certain general types of environments.
- Two components of statistical approach
- Median path loss
- Local variations

2.4 Terrestrial propagation: Statistical models

2.4.1 Median Path Loss

- If $\tilde{E}$ is the total received electrical field and $\tilde{E}_d$ is the electrical field of an equivalent direct path, assuming there are *N* different paths between the transmitter and receiver, the total electric field is

$$\tilde{E} = \tilde{E}_d \sum_{k=1}^{N} L_k e^{j\phi_k} \qquad (2.42)$$

- The $\{L_k\}$ represent the relative losses for the different paths and the $\{\phi_k\}$ represent the relative phase rotations. If a direct path exists, it would be characterized by $L_0 = 1$ and $\phi_0 = 0$

- General propagation model for median path loss having the form

$$\frac{P_R}{P_T} = \frac{\beta}{r^n} \qquad (2.43)$$

- where the path-loss exponent n typically ranges from 2 to 5, depending on the propagation environment. The parameter β represents a loss that is related to frequency.
- The right-hand size of Eq. (2.43)can be rewritten as

$$L_p = \beta(dB) - 10n\log_{10}(r/r_0) \qquad (2.44)$$

- β_0 represents the measured path loss at the reference distance r_0 . In the absence of other information, β_0 is often taken to be the free-space path loss at a distance of 1 meter

TABLE 2.1 Sample path-loss exponents.

Environment	n
Free space	2
Flat rural	3
Rolling rural	3.5
Suburban, low rise	4
Dense urban, skyscrapers	4.5

2.4 Terrestrial propagation: Statistical models

2.4.2 Local Propagation Loss

- In particular, if μ_{50} is the median value of the path loss at a specified distance r from the transmitter, the distribution x_{dB} of observed path losses at this distance has the probability density function which called lognormal model for local shadowing.

$$f(x_{dB}) = \frac{1}{\sqrt{2\pi\sigma_{dB}}} e^{-(x_{dB} - \mu_{50}/2\sigma_{dB}^2)} \qquad (2.45)$$

- It is the Gaussian distribution in which all quantities are measured in decibels

- By plotting the lognormal distribution $prob(x_{dB} > x) = \int_x^{\infty} f(x_{dB})dx_{dB}$

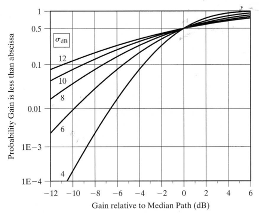

FIGURE 2.11 The lognormal distribution.

2.5 Indoor Propagation

- Understanding the physics of indoor applications
- determining how the locations and numbers of transmitters and receivers will affect the quality of service.
- Simple model for indoor path loss in statistical approach is given by

$$L_p(dB) = \beta(dB) + 10\log_{10}\left(\frac{r}{r_0}\right)^n + \sum_{p=1}^{P} WAF(p) + \sum_{q=1}^{Q} FAF(q) \qquad (2.47)$$

- where r is distance between transmitter and receiver
- r_0 is the nominal reference distance
- n is the path-loss exponent
- $\mathrm{WAF}(p)$ is the wall attenuation factor
- $\mathrm{FAF}(q)$ is the floor attenuation factor and P and Q are the number of walls and floors.

- Shortcomings of statistical model
- Dependence of path-loss exponent.
- Dependence of the WAF on the angle of incidence.
- If the transmitter is outside the building, there is an additional building penetration loss
- the size depends on the frequency of operation and what floor of the building the receiver is located on.

2.6 Local Propagation Effects with Mobile Radio

- The major difficulties:
- mobile antenna is below the surrounding buildings.
- Most communication is via scattering of electromagnetic waves from surfaces or diffraction over and a round building.
- These multipaths have both slow and fast aspects.

- Slow fading
- Arises from large reflectors and diffracting objects with distant path from the small terminal.
- With slow propagation changes, these factors contribute to the median path losses between a fixed transmitter and receiver.
- The statistical variation of these mean losses due to
- variation was modeled as lognormal distribution for terrestrial application.

- Fast fading is the
- rapid variation of signal levels when user terminal moves short distances.
- It is due to reflections of local objects and motion of terminal. That is, the received signal is the sum of a number of signals reflected from local surfaces and signals sum i n the constructive or destructive manner.
- The resulting phase relationships are dependent on relative path lengths to the local object, speed of motion and frequency of transmission

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2.6 Local Propagation Effects with Mobile Radio

2.6.1 Rayleigh Fading

- *Portable terminals:*
- *easily moved but communications occur when the terminal is stationary.*

- *Mobile terminals:*
- *easily moved and communication can occur while the terminal is moving*

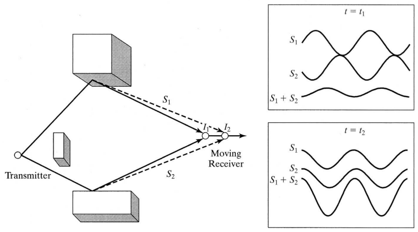


FIGURE 2.13 Illustration of constructive and destructive interference.

- The complex phasor of the N signal rays is given by

$$\tilde{E} = \sum_{n=1}^{N} E_n e^{j\theta_n} \qquad (2.48)$$

- Where E_n is the electric field strength of the nth path and θ_n is relative phase. $\tilde{E}$

$$\sum_{n=1}^{N} E_n e^{j\theta_n} \to Z_r + jZ_i \text{ as } N \to \infty \qquad (2.49)$$

- Considering one of the components of the sum, the expectation of each component is

$$\begin{aligned} E\left[E_n e^{j\phi_n}\right] &= E\left[E_n\right] E\left[e^{j\theta_n}\right] \\ &= E\left[E_n\right] \frac{1}{2\pi} \int_0^{2\pi} e^{j\theta} d\theta \qquad (2.50) \\ &= 0 \end{aligned}$$

- E denotes the statistical expectation operator

- Mean of the complex envelope is given by

$$E\left[\tilde{E}\right]=E\left[\sum_{n=1}^{N}E_n e^{j\theta_n}\right] = \sum_{n=1}^{N}E_n E\left[e^{j\theta_n}\right] = 0 \qquad (2.51)$$

- The variance (power) in the complex envelope is given by mean-square value

$$E\left[\left|\tilde{E}\right|^2\right]=E\left[\sum_{n=1}^{N}E_n e^{j\theta_n}\sum_{m=1}^{N}E_m e^{-j\theta_m}\right] = \sum_{n=1}^{N}\sum_{m=1}^{N}E_n E_m E\left[e^{j(\theta_n-\theta_m)}\right] = \sum_{n=1}^{N}E_n^2 = P_0 \qquad (2.52)$$

- Difference of two random phases is a random phase. By symmetry, the power is equally distributed between the real and imaginary parts of complex envelope.

- Since the complex envelope has zero mean, for $\sigma^2 = P_0/2$,the probability density function of Z_r in Eq. (2.49)is given by the Gaussian density function

$$f_{Z_r}(z_r)=\frac{1}{\sqrt{2\pi}\sigma}e^{-z_r^2/2\sigma^2} \qquad (2.53)$$

- Define the amplitude of complex envelope as $R=\sqrt{Z_i^2+Z_r^2}$
- Rayleigh probability density function

$$f_R(r)=\frac{r}{\sigma^2}e^{-r^2/2\sigma^2} \qquad (2.55)$$

- Integrating the Rayleigh probability density function yields the corresponding cumulative probability distribution function:

$$prob(r<R)=\int_0^R f_R(r)dr = 1-e^{-R^2/2\sigma^2} \qquad (2.56)$$

- The mean value of Rayleigh distribution is given by

$$E[R]=\int_0^\infty r f_R(r)dr = \sigma\sqrt{\frac{\pi}{2}} \qquad (2.57)$$

- Mean-square value is given by

$$E[R]=\int_0^\infty r^2 f_R(r)dr = 2\sigma^2 = R_{rms}^2 \qquad (2.58)$$

- The root-mean-square (rms) amplitude is

$$R_{rms}=\sqrt{2}\sigma$$

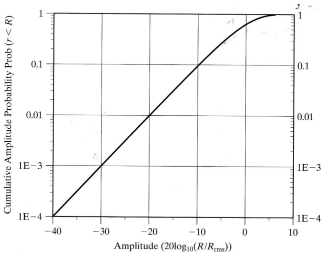


FIGURE 2.14 Rayleigh amplitude distribution (R/R_{rms}).

2.6 Local Propagation Effects with Mobile Radio

2.6.2 Rician Fading

- For the direct ling-of-sight path in mobile radio channels and indoor wireless, the reflected paths tend to be weaker than the direct path and the complex envelope is

$$\tilde{E} = E_0 + \sum_{n=1}^{N} E_n e^{j\theta_n} \qquad (2.59)$$

- For $s^2 = |E_0|^2$, Rician factor defines as

$$K = \frac{s^2}{\sum_{n=1}^{N} |En|^2} \qquad (2.60)$$

- Amplitude density function

$$f_R(r) = \frac{r}{\sigma^2} e^{-\left(r^2+s^2\right)/2\sigma^2} I_0\left(\frac{rs}{\sigma^2}\right) \qquad r \geq 0 \qquad (2.61)$$

- where $I_0(.)$ is the modified Bessel Function of zeroth order.

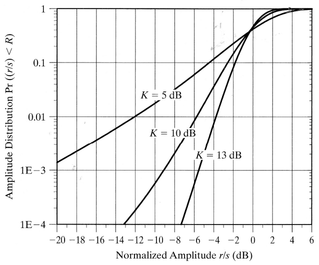


FIGURE 2.15 Amplitude distribution for the Rician channel.

2.6 Local Propagation Effects with Mobile Radio

2.6.3 Doppler

Doppler shift: A receiver is moving toward the source. Zero crossings of the signal appear faster therefore the received frequency is higher The opposite effect occurs if the receiver is moving away from the source.

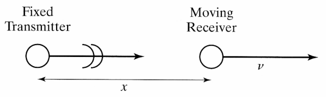


FIGURE 2.16 Illustration of Doppler effect.

- For complex envelope emitted by transmitter is $Ae^{j2\pi f_0}$,with the $A(x)$ is the amplitude and c is the speed of light, then the signal at a point along the x-axis is given by

$$\tilde{r}(t, x) = A(x) e^{j2\pi f_0 (t - x/c)} \qquad (2.62)$$

- x_0 is the receiver' s initial position and v is its velocity, we have

$$x = x_0 + vt \qquad (2.63)$$

- Substituting Eq. (2.63)into (2.62) the signal at the receiver is

$$\begin{aligned} \tilde{r}(t) &= A(x_0 + vt) e^{j2\pi f_0 \left(t - \frac{x_0 + vt}{c}\right)} \\ &= A(x_0 + vt) e^{-j2\pi f_0 x_0 / c} e^{j2\pi f_0 (1 - v/c) t} \end{aligned} \qquad (2.64)$$

- Given f_0 is the carrier transmission frequency, the received frequency
$$f_r = f_0\left(1 - \frac{v}{c}\right) \qquad (2.65)$$
- Doppler shift
$$f_D = f_r - f_0 = -f_0\frac{v}{c} \qquad (2.66)$$
- Relationship between Doppler frequency and velocity
$$\frac{v}{c} = -\frac{f_D}{f_0} \qquad (2.67)$$
- If the terminal motion and the direction of radiation are at an angle ψ, shift can be expressed as
$$f_D = -\frac{f_0}{c}v\cos\psi \qquad (2.68)$$

2.6 Local Propagation Effects with Mobile Radio

2.6.4 Fast Fading

- Rapid variation in the signal strength about a median value.
- Clarke model: all rays are arriving from a horizontal direction

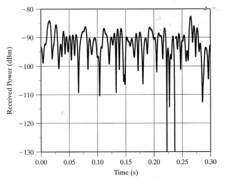


FIGURE 2.17 Illustration of variations in received signal power due to Rayleigh fading.

- Autocorrelation of complex envelope
$$\begin{aligned} R_E(\tau) &= P_0E\left[e^{-j2\pi f_D\tau\cos\psi}\right] \\ &= P_0\left(\frac{1}{2\pi}\int_{-\pi}^{\pi} e^{-j2\pi f_D\tau\cos\psi}\,d\psi\right) \qquad (2.75) \\ &= P_0J_0(2\pi f_D\tau) \end{aligned}$$
- where $J_0(x)$ is the zeroth-order Bessel function of the first kind
$$J_0(x) = \frac{1}{\pi}\int_0^{\pi} e^{-jx\cos\theta}\,d\theta \qquad (2.76)$$
- The power spectrum of the fading process is determined by Fourier transform of autocorrelation function of envelope and is given by
$$\begin{aligned} S_E(f) &= F[R_E(\tau)] \\ &= F[P_0J_0(2\pi f_D\tau)] \qquad (2.77) \\ &= \begin{cases} \dfrac{P_0}{\sqrt{1-(f/f_D)^2}} & f < f_D \\ 0 & f > f_D \end{cases} \end{aligned}$$

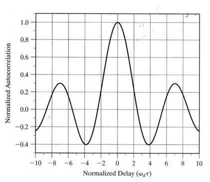


FIGURE 2.19 Autocorrelation of the complex envelope of the received signal according to Clarke's model.

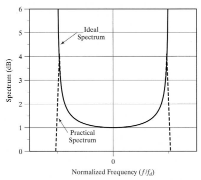


FIGURE 2.20 Power spectrum of the fading process for Clarke's model.

2.7 Channel Classification

- *Large-scale effects*
- *due to terrain, density and height of the building.*
- *characterized statistically by median path loss and lognormal shadowing*

- *Small-scale effects*
- *due to local environment and the movement of radio terminal.*
- *They are characterized statistically as fast Rayleigh fading.*

- *Channels are classified on the basis of the properties of time-varying impulse response*

2.7 Channel Classification

2.7.1 Tim-Selective Channels

- For time selective channel, the channel impulse response is
- $$\tilde{h}(t,\tau)=\tilde{\alpha}(t)\delta(\tau) \qquad (2.84)$$
- Where $\delta(t)$ is the Dirac delta function or unit-impulse function
- Frequency-flat channel:
- the frequency response of the channel is approximately constant and does not change the spectrum of t he transmitted signal

2.7 Channel Classification

2.7.2 Frequency-Selective Channels

- With large-scale effect, the complex phasor is

$$\tilde{x}(t)=\sum_{l=1}^{L}\tilde{\alpha}_l\tilde{s}(t-\tau_l) \qquad (2.86)$$

- Channel impulse response

$$\tilde{h}(t,\tau)=\sum_{l=1}^{L}\tilde{\alpha}_l\delta(\tau-\tau_l) \qquad (2.87)$$

- This channel is time invariant, but shows a frequency-dependent response.

2.7 Channel Classification

2.7.3 General Channel

- Time-varying impulse response

$$\tilde{h}(t,\tau)=\sum_{l=1}^{L}\tilde{\alpha}_l(t)\delta(\tau-\tau_l(t)) \qquad (2.90)$$

- Received signal

$$\tilde{x}(t)=\int_{-\infty}^{\infty}\tilde{h}(t,\tau)\tilde{s}(t-\tau)d\tau \qquad (2.91)$$

- Time-varying frequency

$$H(t,f)=F[h(t,\tau)] \qquad (2.92)$$

2.7 Channel Classification

2.7.4 WSSUS Channels

- A random process is
- wide-sense stationary if it has a mean that is time independent and a correlation function $R(t_1,t_2)=R(t_1-t_2)$
- In multipath channels, the gain and phase shift at one delay are uncorrelated with the gain and phase shift at another delay. This refers to as uncorrelated scattering

$$R_h(t_1-t_2;\tau_1,\tau_2)=R_h^W(t_1-t_2;\tau_1-\tau_2)$$

- The combination of wide-sense stationary signal and uncorrelated scattering is called WSSUS.

2.7 Channel Classification

2.7.5 Coherence Time

- Coherence time
- the period over which there is a strong correlation of the channel time response.
- Doppler power spectrum

$$S_D(f)=F[R_h^W(\Delta t;0)] \qquad (2.100)$$

- Coherence time

$$T_{coherence}\approx\frac{1}{2f_D} \qquad (2.101)$$

2.7 Channel Classification

2.7.6 Power-Delay Profile

- Power-delay profile

$$P_h(\tau) = R_h^W(0;\tau) \qquad (2.103)$$

- It provides an estimate of the average multipath power as a function of the relative delay τ

2.7 Channel Classification

2.7.7 Coherence Bandwidth

- The coherence bandwidth of the channel is the bandwidth over which the frequency response is strongly correlated.
- Relationship between time and frequency domains

$$BW_{coh} \approx \frac{1}{T_M} \qquad (2.116)$$

- If the coherence bandwidth is small with respect to bandwidth of transmitted signal
- frequency selective.
- If the coherence bandwidth is large with respect to bandwidth of transmitted signal
- frequency nonselective or frequency-flat.

2.7 Channel Classification

2.7.8 Stationary and Nonstationary Channels

- Stationary models for channel
- characteristics are convenient for analysis but often except for short time intervals
- not accurate description of reality.
- Terrestrial mobile channels are highly non-stationary because:
- Propagation path often consists of several discontinuities.
- Environment itself is physically nonstationary.
- The interference caused by other users sharing the same frequency channel will vary dynamically.

2.8 Noise and Interference

- Noise is defined as
- unwanted electrical signal interfering with the desired signal.
- Sources of noise
- Natural
- Artificial

2.8 Noise and Interference

2.8.1 Thermal Noise

- Fundamental property of matter above the absolute zero temperature.
- The available thermal noise spectral density is

$$S_n(f) = kT/2$$
$$\equiv N_0/2 \qquad -\infty < f < \infty$$

- White noise contains all frequencies at an equal level.

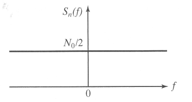


FIGURE 2.26 Spectral model for white noise.

- Noise power in bandwidth B is

$$P = (N_0/2)(2B) = N_0 B \qquad (2.120)$$

- Noise-equivalent bandwidth of the filter as

$$B_{eq} = \frac{\int_{-\infty}^{\infty} |H(f)|^2 (N_0/2)\,df}{N_0/2} \qquad (2.121)$$

2.8 Noise and Interference

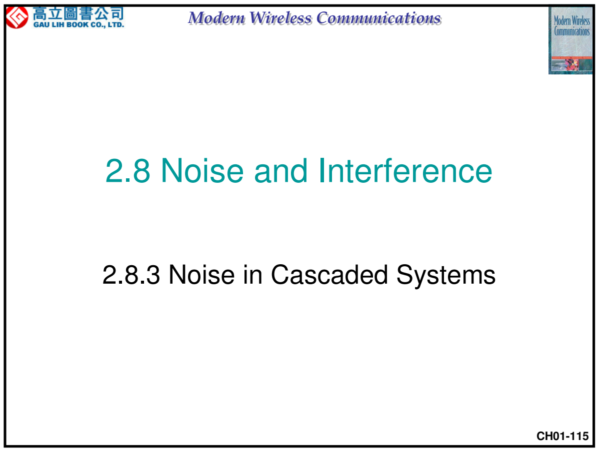
2.8.2 Equivalent Noise Temperature and Noise Figure

- The noise figure represents the increase in noise at the output of the amplifier, referenced to the output.
- $G(f)$ be the available power gain of the device as a function of frequency.
- $S_{no}(f)$ is the spectrum of output noise power.
- $S_{ni}(f)$ is the spectrum of input noise power.

$$F = \frac{S_{no}(f)}{G(f)S_{ni}(f)} \qquad (2.122)$$

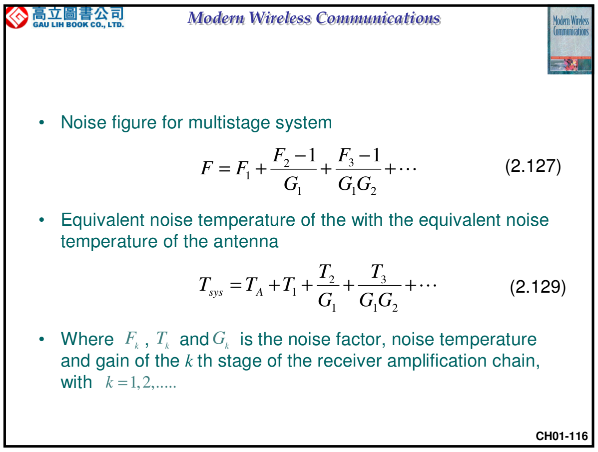
$$T_e = (F-1)T_0 \qquad (2.124)$$

- Relationship between equivalent noise temperature and noise figure:



2.8 Noise and Interference

2.8.3 Noise in Cascaded Systems



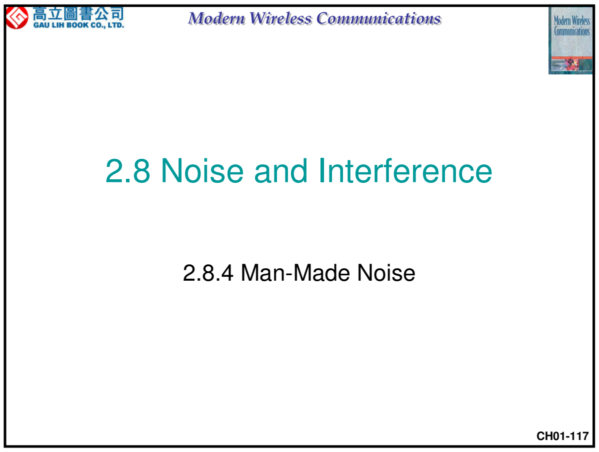
- Noise figure for multistage system

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots \qquad (2.127)$$

- Equivalent noise temperature of the with the equivalent noise temperature of the antenna

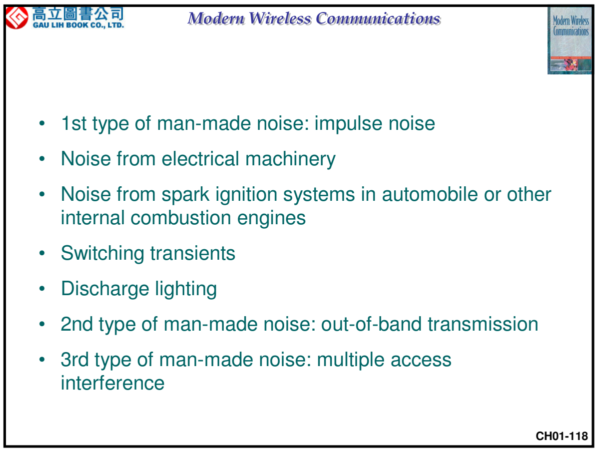
$$T_{sys} = T_A + T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1 G_2} + \cdots \qquad (2.129)$$

- Where F_k , T_k and G_k is the noise factor, noise temperature and gain of the k th stage of the receiver amplification chain, with $k = 1, 2, ...$

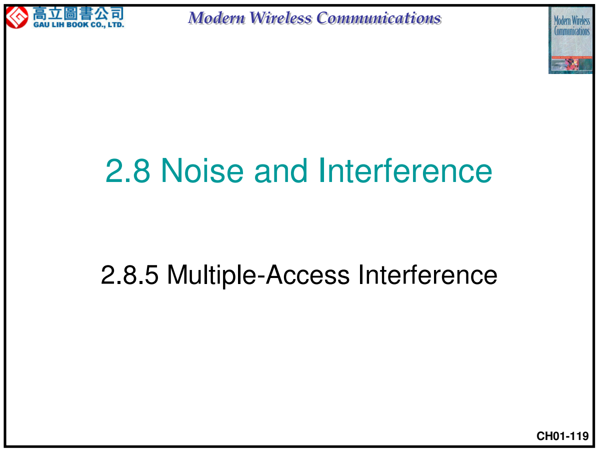


2.8 Noise and Interference

2.8.4 Man-Made Noise

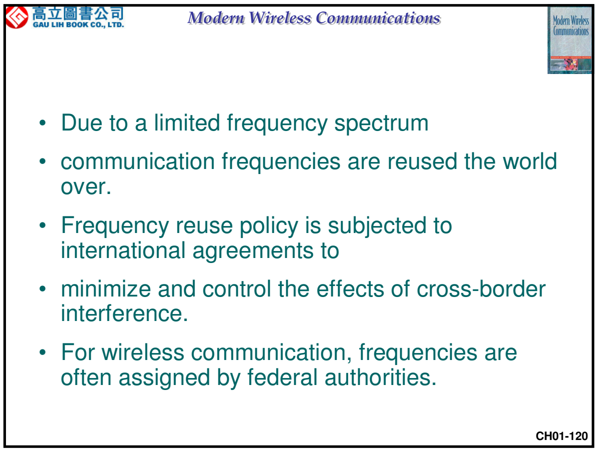


- 1st type of man-made noise: impulse noise
- Noise from electrical machinery
- Noise from spark ignition systems in automobile or other internal combustion engines
- Switching transients
- Discharge lighting
- 2nd type of man-made noise: out-of-band transmission
- 3rd type of man-made noise: multiple access interference



2.8 Noise and Interference

2.8.5 Multiple-Access Interference



- Due to a limited frequency spectrum
- communication frequencies are reused the world over.
- Frequency reuse policy is subjected to international agreements to
- minimize and control the effects of cross-border interference.
- For wireless communication, frequencies are often assigned by federal authorities.

- Cellular system:
- a single service provider is given a band of frequencies for a given service area.
- The reuse pattern can be defined relative to a given reference:
- Move i cells along any chain of hexagons and turn clockwise 60 degrees, then move j cells along the chain that lies on this new heading.
- With hexagonal geometry, the cells form natural clusters around the reference cell and each of its cochannel cells.

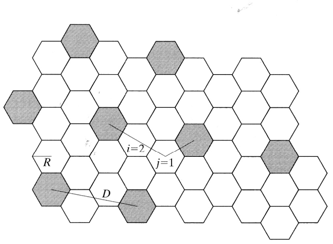


FIGURE 2.31 Hexagonal pattern of cells used in cellular telephone system.

- Number of cells per cluster

$$N = i^2 + ij + j^2 \qquad (2.130)$$

- Ratio between centers of nearest-neighboring cochannel cells to Normalized reuse distance

$$\frac{D}{R} = \sqrt{3N} \qquad (2.131)$$

- The mean carrier-to-interference ratio

$$\begin{aligned}\frac{C}{I} &= \frac{\text{Received power of desired user}}{\sum \text{Received powers of interfering users}} \\ &= \frac{\beta P_T / (r_d / r_0)^{n_d}}{\sum_{k \neq d} \beta P_T / (r_d / r_0)^{n_k}} \qquad (2.133) \\ &= \frac{r_d^{-n_d}}{\sum_{k \neq d} r_k^{-n_k}}\end{aligned}$$

2.9.Link Calculation

2.9.1 Free-Space Link Budget

- Ensure that sufficient power is available at the receiver to close the link and meet SNR requirement.
- For free-space propagation, the basic budget equation is given by

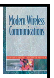
$$\frac{P_R}{N_0} = \frac{P_T G_R G_T}{L_p k T_e} \qquad (2.136)$$

- where $N_0 = kT_e$ and Te is the equivalent noise temperature of the system

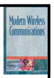
- For satellite applications, the basic link budget equation is

$$\frac{C}{N_0} = EIRP - L_p + (G/T) - k \qquad (2.137)$$

$C/N_0 = P_R/N_0$	is the received carrier - to - noise density ratio (dB - Hz);
$EIRP = G_T P_T$	is the equivalent isotropic radiated power of the transmitter (dBW);
L_p	is the path loss (dB)
$G/T = G_R/T_e$	is the ratio of receiver antenna gain to noise temperature (dB - K^{-1})
k	is Boltzmann's constant (-228.6dBW - sK^{-1})



2.9.2 Terrestrial Link Budget



- For a terrestrial link budget,
- Many of the line items are similar to those in the free-space link budget examples.
- But their calculations are different.