Wireless Communication

Chapter 4

Mobile Radio Propagation –

Large-Scale Path Loss

4.1 Introduction to Radio Wave Propagation

- The mechanisms behind electromagnetic wave propagation can generally be attributed to reflection, diffraction, and scattering.
- High-rise buildings $\rightarrow$ diffraction loss

  reflection $\rightarrow$ multipath fading
- Large-scale propagation model : local average signal strength over a measurement track of 5λ to 40λ.

  over distance of 100's ~ 1000's meters.
- Small-scale propagation model : characterize the rapid fluctuation of the received signal strength over short distance $(\lambda's)$ or short time durations (secs).

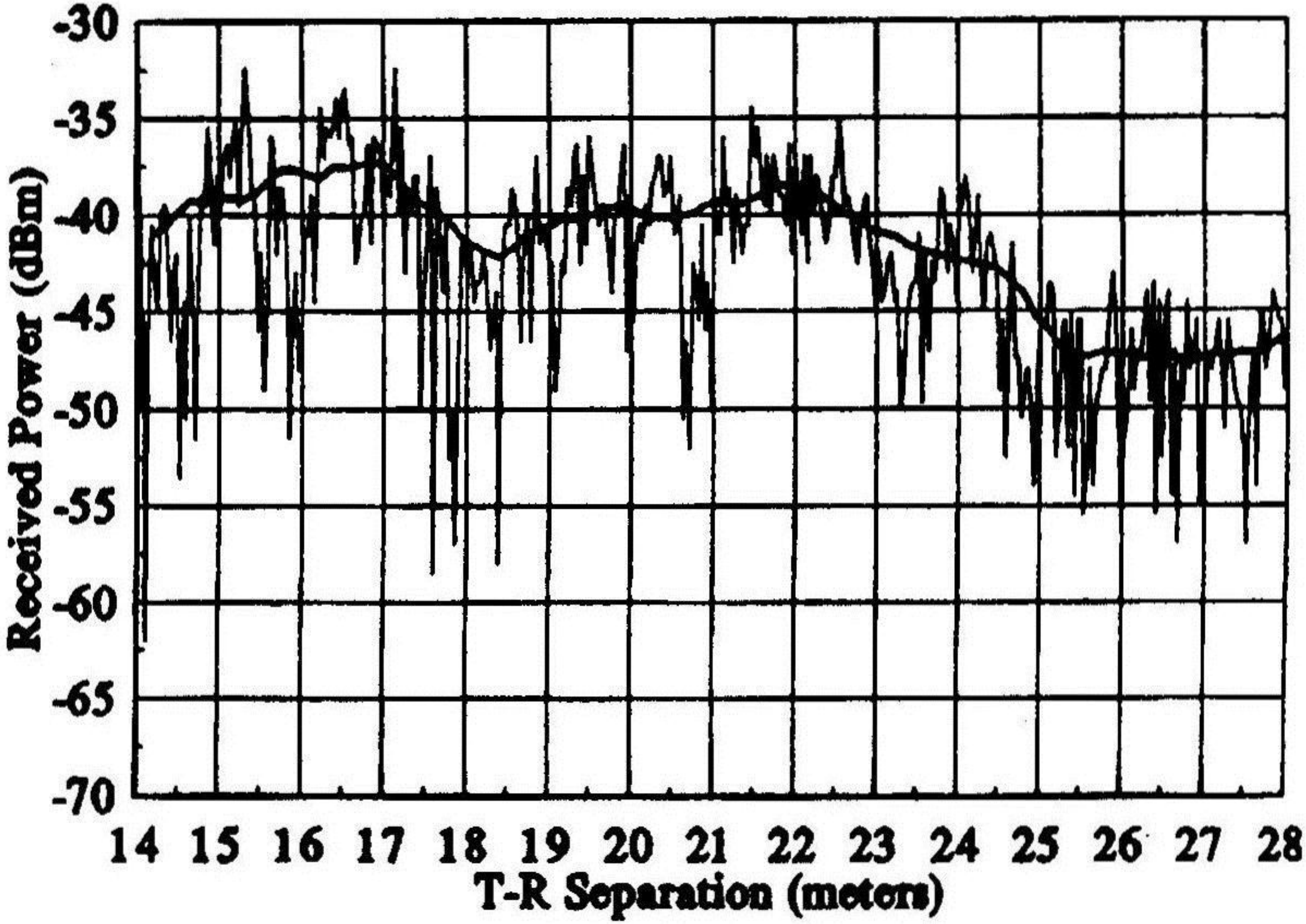


Figure 4.1 Small-scale and large-scale fading.

4.2 Free Space Propagation Model

- The free space propagation model is used to predict received signal strength when the transmitter and receiver have a clear unobstructed line-of-sight path between them, e.g., satellite communication systems and microwave line-of-sight radio links.
- The Friis free space equation

$$\boxed{P_r(d) = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2 L}}$$

d : distance , P_t: the transmitted power, P_r: the received power

G_t: the transmitter antenna gain, G_r : the receiver antenna gain

L : the system loss, λ : the wavelength

- The gain of an antenna is related to its aperture

$$G = \frac{4\pi A_e}{\lambda^2}$$

A_e : the effective aperture, related to antenna size

$$\lambda = \frac{c}{f} = \frac{2\pi c}{\omega_c}$$

$c = 3 \times 10^8$ meters.

- $L = 1$ indicates no loss in the system hardware.
- The effective isotropic radiated power (EIRP) is defined as

$$EIRP = P_t G_t$$

and represents the maximum radiated power available from a transmitter in the direction of maximum antenna gain, as compare to an isotropic radiator.

- The effective radiated power (ERP) is compared to a half-wave dipole antenna (instead of an isotropic antenna).
- Since a dipole antenna has a gain of 1.64 (2.15dB above an isotropic).

$$ERP = EIRP - 2.15(dB)$$

- Antenna gains are given in units of dBi (dB gain with respect to an isotropic antenna) or dBd (w.r.t a half-wave dipole).
- Path loss, when antenna gains are included,

$$PL(dB) = 10\log\frac{P_t}{P_r} = -10\log\left[\frac{G_t G_r \lambda^2}{(4\pi)^2 d^2}\right]$$

when antenna gains are excluded,

$$PL(dB) = 10\log\frac{P_t}{P_r} = -10\log\left[\frac{\lambda^2}{(4\pi)^2 d^2}\right]$$

- The far-field, or *Fraunhofer region*, is defined as the region beyond the far-field distance d_f ,

$$d_f = \frac{2D^2}{\lambda}$$

D :the largest physical linear dimension of the antenna. Additionally

$$d_f >> D,\ d_f >> \lambda$$

- Given $P_r(d_0),\ \ d_0 \geq d_f$

$$P_r(d) = P_r(d_0)(\frac{d_0}{d})^2 \qquad d \geq d_0$$

- This equation may be expressed in units of dBm or dBW. For example:

$$P_r(d)\,\text{dBm} = 10\log[\frac{P_r(d_0)}{0.001\text{W}}] + 20\log(\frac{d_0}{d})$$

when $P_r(d_0)$ is in units of watts.

- d_0 for practical systems using low-gain antennas in the 1-2 GHz region is typically chosen to be

1m	in indoor environments
100m or 1km	in outdoor environments

Example 4.1

Find the far-field distance for an antenna with maximum dimension of 1 m and operating frequency of 900 MHz.

Solution:

Largest dimension of antenna , $D = 1$ m

$f = 900$ MHz, $\lambda = c/f = (3\times10^8$ m/s) / $(900 \times10^6$ Hz)

$d_f = 2(1)^2 / 0.33 = 6$ m

Example 4.2

If a transmitter produces 50 W of power, express the transmit power in units of (a) dBm, and (b) dBW. If 50W is applied to a unity gain antenna with a 900MHz carrier frequency, find the received power in dBm at a free space distance of 100 m from the antenna. What is P_r (10 km)? Assume unity gain for the receiver antenna.

Solution:

$$\text{Transmitter power}, P_t = 50 \text{ W}$$

$$\text{Carrier frequency}, f_c = 900\text{MHz}$$

(a)

$$P_t(\text{dBm}) = 10\log[P_t(\text{mW})/(1\text{mW})] = 10\log[50\times 10^3] = 47.0 \text{ dBm}$$

(b)

$$P_t(\text{dBW}) = 10\log[P_t(\text{W})/(1\text{W})] = 10\log[50] = 17.0 \text{ dBW}$$

The receiver power

$$P_r = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2 L} = \frac{50(1)(1)(1/3)^2}{(4\pi)^2 (100)^2 (1)} = (3.5\times 10^{-6})\text{W} = 3.5\times 10^{-3}\text{mW}$$

$$P_r(\text{dBm}) = 10\log P_r(\text{mW}) = 10\log(3.5\times 10^{-3}\text{mW}) = -24.5\text{dBm}$$

The receiver power at 10km

$$P_r(10\text{km}) = P_r(100) + 20\log[100/1000] = -24.5\text{dBm} - 40\text{dB} = -64.5\text{dBm}$$

4.3 Relating Power to Electric Field

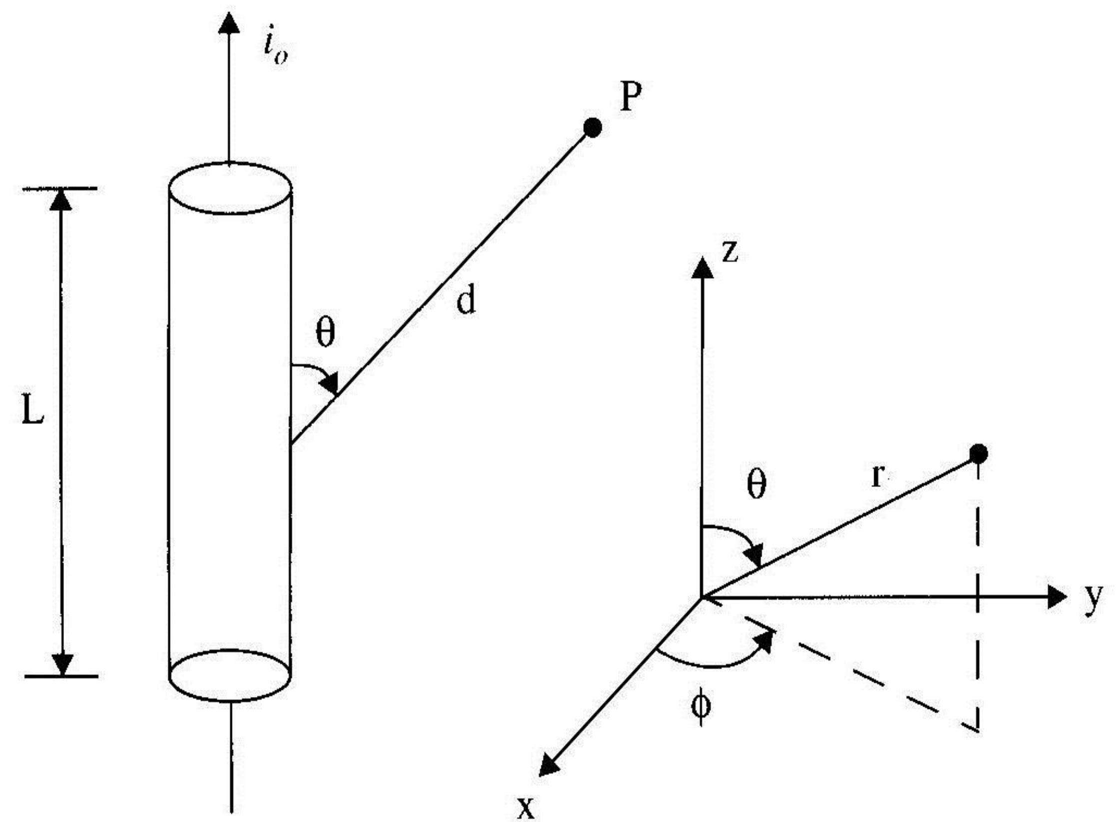


Figure 4.2 Illustration of a linear radiator of length $L (L \ll \lambda)$, carrying a current of amplitude i_0 and making an angle θ with a point, at distance *d*.

- If a current flows through such an antenna, it launches electric and magnetic fields that can be expressed as

$$E_r = \frac{i_0 L \cos\theta}{2\pi\varepsilon_0 c}\{\frac{1}{d^2} + \frac{c}{j\omega_c d^3}\}e^{j\omega_c(t-d/c)}$$

$$E_\theta = \frac{i_0 L \sin\theta}{4\pi\varepsilon_0 c^2}\{\frac{j\omega_c}{d} + \frac{c}{d^2} + \frac{c^2}{j\omega_c d^3}\}e^{-j\omega_c(t-d/c)}$$

$$H_\phi = \frac{i_0 L \sin\theta}{4\pi c}\{\frac{j\omega_c}{d} + \frac{c}{d^2}\}e^{j\omega_c(t-d/c)}$$

$$E_\phi = H_r = H_\theta = 0$$

- In far-field region, only the radiated field components of E_θ and H_ϕ need be considered.
- In free space, the power flux density P_d (W/m^2) is given by

$$P_d = \frac{EIRP}{4\pi d^2} = \frac{P_t G_t}{4\pi d^2} = \frac{E^2}{R_{fs}} = \frac{E^2}{\eta} \text{ W/m}^2$$

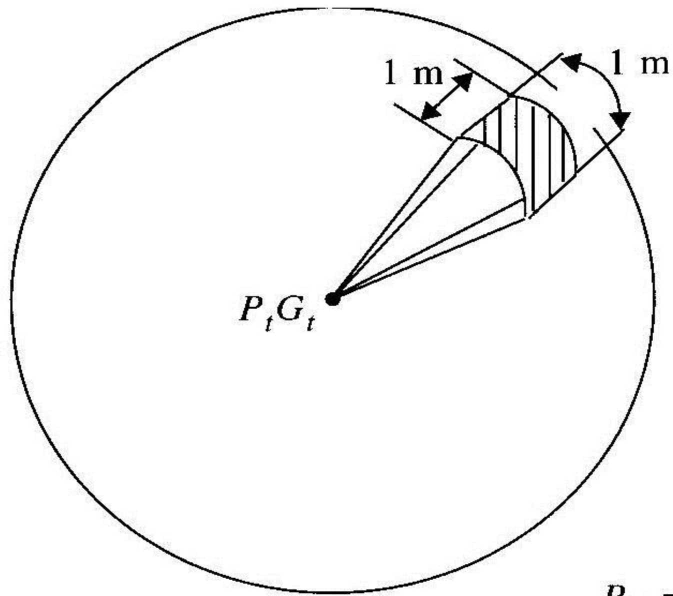
R_{fs} : the intrinsic impedance of free space $= \eta = 120\pi\Omega$ (377 Ω)

$$\Rightarrow P_d = \frac{|E|^2}{377\,\Omega}$$

- The power received at distance *d*

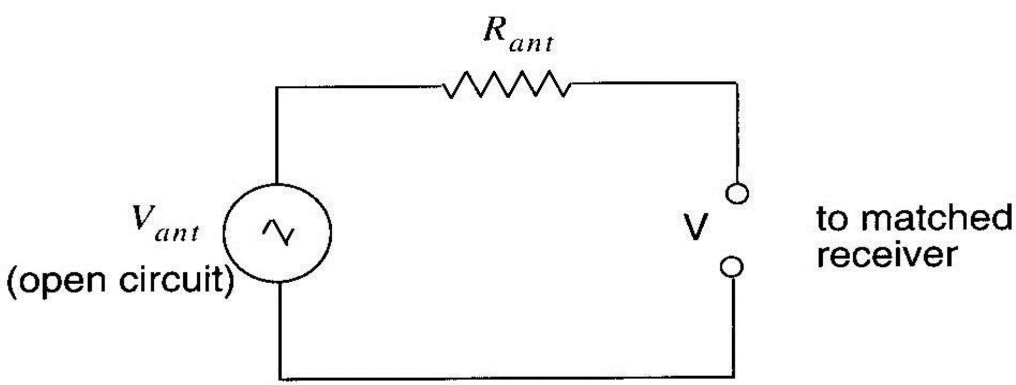
$$P_r(d) = P_d A_e = \frac{|E|^2}{120\pi} A_e = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2} = \frac{|E|^2 G_r \lambda^2}{480\pi^2} \text{ W}$$

A_e : the effective aperture of the receiver antenna



(a)

$$P_d = \frac{P_tG_t}{4\pi d^2} = \frac{EIRP}{4\pi d^2} = \frac{|E|^2}{120\pi} \quad \text{W/m}^2$$



(b)

Figure 4.3 (a) Power flux density at a distance *d* from a point source; (b) model for voltage applied to the input of a receiver.

- If the receiver antenna is modeled as a matched resistive load to the receiver,

$$P_r(d) = \frac{V^2}{R_{ant}} = \frac{[V_{ant}/2]^2}{R_{ant}} = \frac{V_{ant}{}^2}{4R_{ant}}$$

V_{ant} : the open circuit voltage at the antenna.

R_{ant} : the resistance of the matched receiver.

Example 4.3

Assume a receiver is located 10 km from a 50 W transmitter. The carrier frequency is 900MHz, free space propagation is assumed, $G_t = 1$, and $G_r = 2$, find (a) the power at receiver, (b) the magnitude of the E-field at the receiver antenna, (c) the rms voltage applied to the receiver input assuming that the receiver antenna has a purely real impedance of 50 Ω and is matched to the receiver.

Solution:

Transmitter power, P_t = 50W.   Carrier frequency, f_c = 900MHz

Transmitter antenna gain, G_t = 1. Receiver antenna resistance, G_r = 2.

Receiver antenna resistance = 50Ω.

(a)

$$P_r(d) = 10\log\left(\frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2}\right) = 10\log\left(\frac{50\times 1\times 2\times (1/3)^2}{(4\pi)^2 10000^2}\right)$$

$$= -91.5\text{dBW} = -61.5\,\text{dBm}$$

(b)

$$|E| = \sqrt{\frac{P_r(d)120\pi}{A_e}} = \sqrt{\frac{P_r(d)120\pi}{G_r\lambda^2/4\pi}} = \sqrt{\frac{7\times 10^{-10}\times 120\pi}{2\times 0.33^2/(4\pi)}} = 0.0039\ \text{V/m}$$

(c)

$$\text{V} = \sqrt{P_r(d)\times R_{ant}} = \sqrt{7\times 10^{-10}\times 50} = 0.187\text{mV}$$

4.4 The Three Basic Propagation Mechanisms

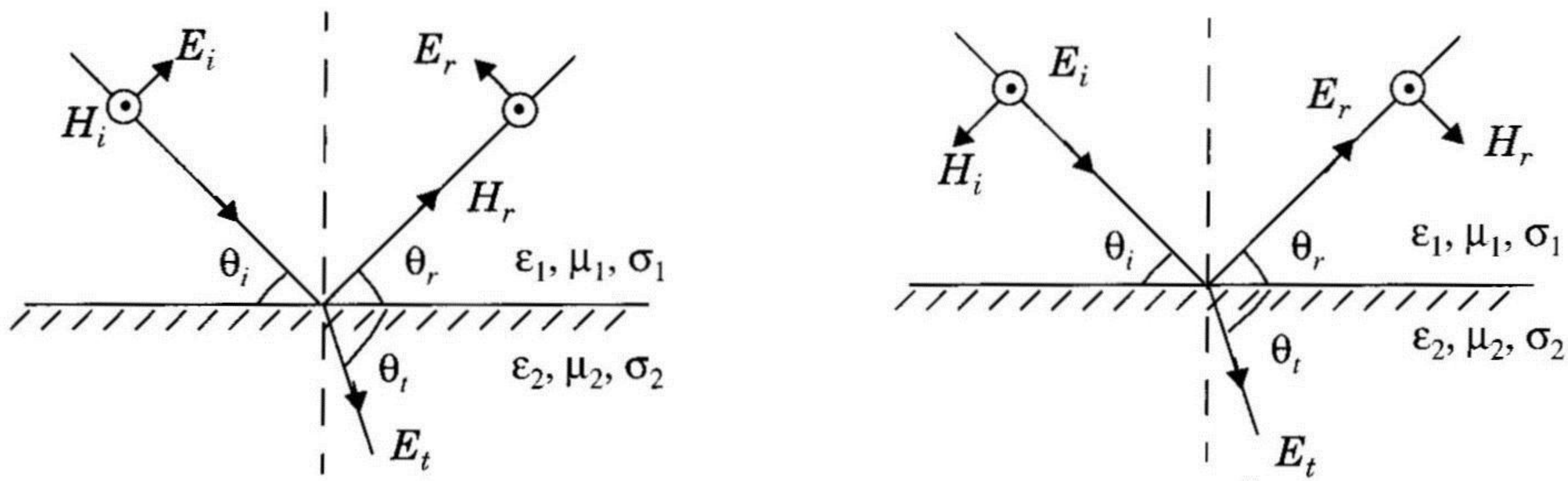
- Reflection – occurs from the surface of the earth or from building and walls.
- Diffraction – occurs when the radio path between the transmitter and receiver is obstructed by a surface that has sharp irregularities (edges).
- Scattering – are produced by rough surface, small objects, or by other irregularities in the channel, e.g., foliage, street signs, and lamp posts.

4.5 Reflection

- The electric field intensity of the reflected and transmitted waves may be related to the incident wave through the *Fresnel reflection coefficient* (Γ).

4.5.1 Reflection from Dielectric

- The behavior for arbitrary directions of polarization can be studied by considering the two distinct cases.



(a) E-field in the plane of incidence

(b) E-field normal to the plane of incidence

Figure 4.4 Geometry for calculating the reflection coefficients between two dielectrics.

- In Figure 4.4a, the E-field polarization is parallel with the plane of incidence. (vertical polarization, normal component, with respect to the reflecting surface).
- In Figure 4.4b, the E-field polarization is perpendicular to the plane of incidence. (parallel to the reflecting surface).
- ε_1, μ_1, σ_1 :the permittivity, permeability, and conductance.
- For perfect (lossless) dielectric, $\varepsilon=\varepsilon_0\varepsilon_r$.

  For lossy dielectric (absorbing power), $\varepsilon=\varepsilon_0\varepsilon_r-j\varepsilon'$

$$\varepsilon' = \frac{\sigma}{2\pi f}$$

Table 4.1 Material Parameters at Various Frequencies

Material	Relative Permittivity ε_r	Conductivity σ (s/m)	Frequency (MHz)
Poor Ground	4	0.001	100
Typical Ground	15	0.005	100
Good Ground	25	0.02	100
Sea Water	81	5.0	100
Fresh Water	81	0.001	100
Brick	4.44	0.001	4000
Limestone	7.51	0.028	4000
Glass, Corning 707	4	0.00000018	1
Glass, Corning 707	4	0.000027	100
Glass, Corning 707	4	0.005	10000

- The reflection coefficients for the two cases are given by

$$\Gamma_{II} = \frac{E_r}{E_i} = \frac{\eta_2 \sin\theta_t - \eta_1 \sin\theta_i}{\eta_2 \sin\theta_t + \eta_1 \sin\theta_i} \quad \text{(E - field in plane of incidence)}$$

$$\Gamma_{\perp} = \frac{E_r}{E_i} = \frac{\eta_2 \sin\theta_i - \eta_1 \sin\theta_t}{\eta_2 \sin\theta_i + \eta_1 \sin\theta_t} \quad \text{(E - field normal to the plane of incidence)}$$

where η_i is the intrinsic impedance of the *i*th medium.

$$\eta_i = \sqrt{\frac{\mu_i}{\varepsilon_i}}$$

The velocity of an electromagnetic wave is

$$v = \frac{1}{\sqrt{\mu\varepsilon}}$$

- Snell's Law

$$\sqrt{\mu_1\varepsilon_1}\sin(90-\theta_i)=\sqrt{\mu_2\varepsilon_2}\sin(90-\theta_t)$$

- The boundary conditions from Maxwell's equations give

$$\theta_i=\theta_r$$

and

$$E_r=\Gamma E_i$$
$$E_t=(1+\Gamma)E_i$$

where Γ is either $\Gamma_{\parallel}$ or $\Gamma_{\perp}$, depending on whether the E-field is in (vertical) or normal (horizontal) to the plane of incidence.

- In the general case of reflection or transmission, the horizontal and vertical axes of the spatial coordinates may not coincide with the perpendicular and parallel axes of the propagation waves.

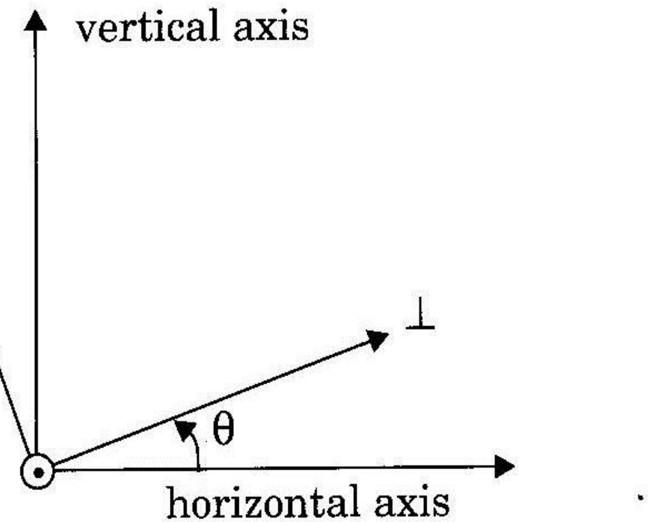


Figure 4.5 Axes for orthogonally polarized components. Parallel and perpendicular components are related to the horizontal and vertical spatial coordinates. Wave is shown propagating out of the page toward the reader.

- The vertical and horizontal field components at a dielectric boundary may be related by

$$\begin{bmatrix} E_H^d \\ E_V^d \end{bmatrix} = R^T D_C R \begin{bmatrix} E_H^i \\ E_V^i \end{bmatrix}$$

$$R = \begin{bmatrix} \cos\theta & \sin\theta \\ -\sin\theta & \cos\theta \end{bmatrix}$$

θ : the angle between the two sets of axes, as shown in Fig 4.5

$$D_C = \begin{bmatrix} D_{\perp\perp} & 0 \\ 0 & D_{\parallel\parallel} \end{bmatrix}$$

where $D_{xx} = \Gamma_x$ for the case of reflection and

$D_{xx} = T_x = 1 + \Gamma_x$ for the case of transmission.

- Figure 4.6 shows a plot of reflection coefficient for both horizontal and vertical polarization.

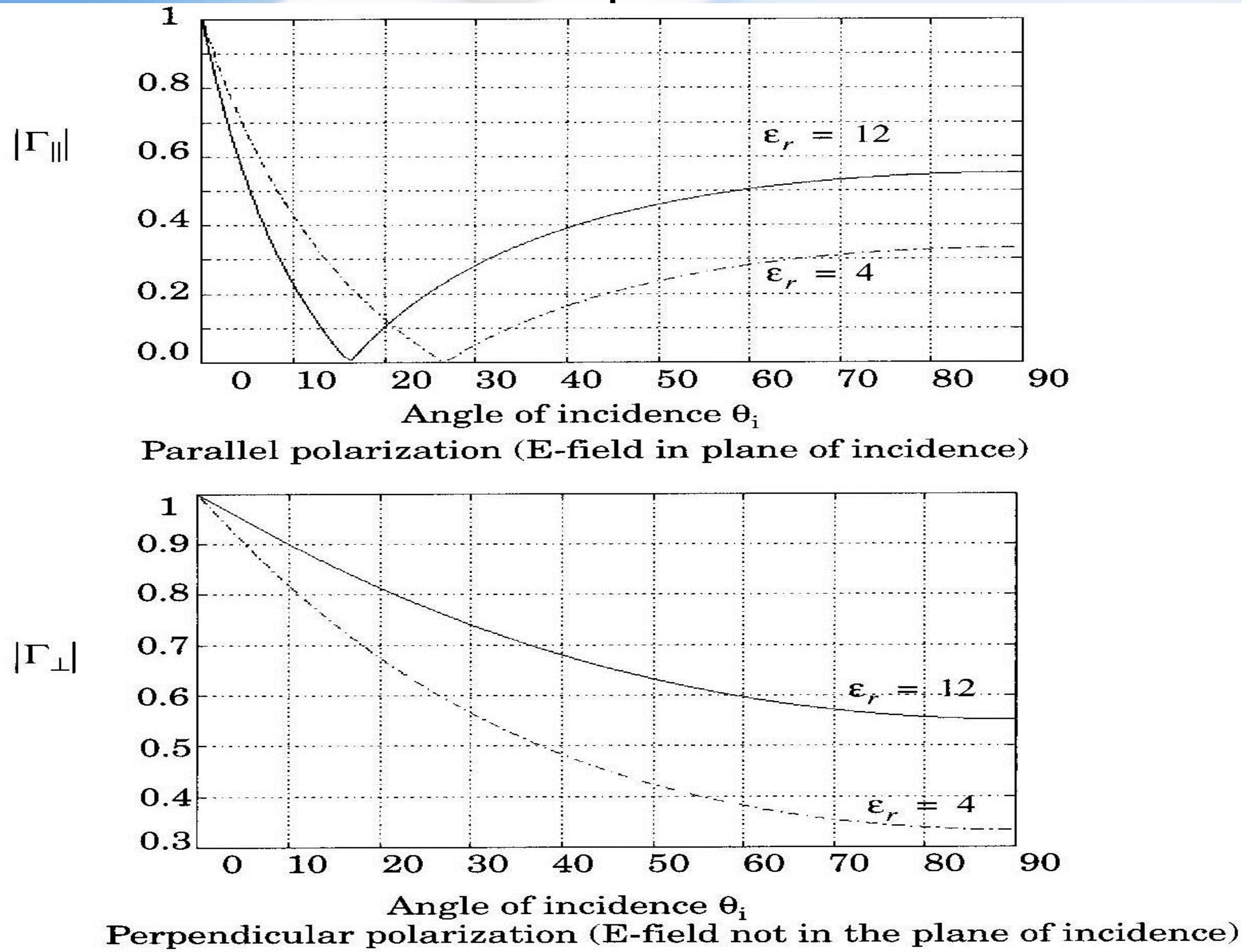


Figure 4.6 Magnitude of reflection coefficients as a function of angle of incidence for $\varepsilon_r = 4$, $\varepsilon_r = 12$, using geometry in Figure 4.4.

Example 4.4

Demonstrate that if medium 1 is free space and medium 2 is a dielectric, both $|\Gamma_{||}|$ and $|\Gamma_{\perp}|$ approach 1 as θ_j approaches 0° regardless of ε_r.

Solution:

$$\theta_i = 0^\circ$$

$$\Gamma_{\parallel} = \frac{-\varepsilon_r \sin 0 + \sqrt{\varepsilon_r - \cos^2 0}}{\varepsilon_r \sin 0 + \sqrt{\varepsilon_r - \cos^2 0}}$$

$$|\Gamma_{\parallel}| = \frac{\sqrt{\varepsilon_r - 1}}{\sqrt{\varepsilon_r - 1}} = 1$$

$$\Gamma_{\perp} = \frac{\sin 0 - \sqrt{\varepsilon_r - \cos^2 0}}{\sin 0 + \sqrt{\varepsilon_r - \cos^2 0}}$$

$$|\Gamma_{\perp}| = \frac{-\sqrt{\varepsilon_r - 1}}{\sqrt{\varepsilon_r - 1}} = -1$$

This example illustrates that ground may be modeled as a perfect reflector regardless of polarization or ground dielectric properties.

4.5.2 Brewster Angle

- The *Brewster angle* is the angle at which no reflection occurs in the medium of origin

$$\Gamma_{\|} = 0$$

$$\sin(\theta_B) = \sqrt{\frac{\varepsilon_1}{\varepsilon_1 + \varepsilon_2}}$$

if $\varepsilon_1 = \varepsilon_0$, $\varepsilon_2 = \varepsilon_0 \varepsilon_r$

$$\sin(\theta_B) = \sqrt{\frac{\varepsilon_r - 1}{\varepsilon_r^2 - 1}}$$

- The *Brewster angle* occurs only for vertical (i.e. parallel) polarization.

Example 4.5

Calculate the Brewster angle for a wave impinging on ground having a permittivity of $\varepsilon_r = 4$.

Solution:

$$\sin(\theta_i) = \frac{\sqrt{(4)-1}}{\sqrt{(4)^2-1}} = \sqrt{\frac{3}{15}} = \sqrt{\frac{1}{5}}$$

$$\theta_i = \sin^{-1}\sqrt{\frac{1}{5}} = 26.56^\circ$$

4.5.3 Reflection from Perfect Conductors

- Since electromagnetic energy cannot pass through a perfect conductor, a plane wave incident on a conductor has all of its energy reflection.
- For E-field in plane of incidence,

$$\theta_i = \theta_r$$

$$E_i = E_r$$

for E-field normal to plane of incidence,

$$\theta_i = \theta_r$$

$$E_i = -E_r$$

4.6 Ground Reflection (Two-Ray) Model

- In a mobile radio channel, a single direct path between the base station and a mobile is seldom the only physical means for propagation.
- The two-ray ground reflection model shown in Figure 4.7 is a useful and more accurate propagation model.

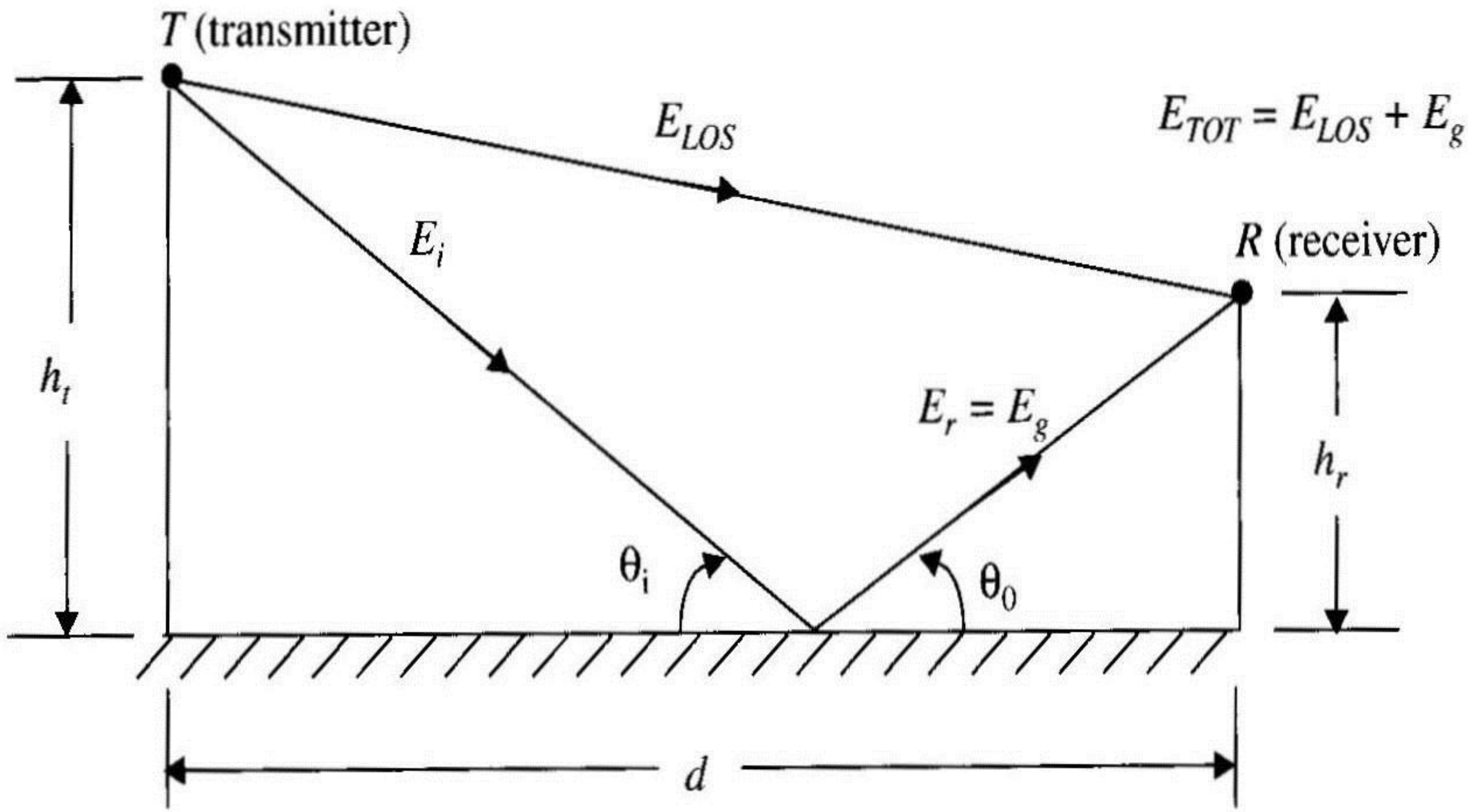


Figure 4.7 Two-ray ground reflection model.

- $E_{TOT} = E_{LOS} + E_g$
- If E_0 is the free space E-field at a reference distance d_0 from the transmitter, then for $d > d_0$,

$$E(d,t) = \frac{E_0 d_0}{d}\cos(\omega_c(t - \frac{d}{c})) \qquad (d > d_0)$$

$$\Rightarrow$$

$$E_{LOS}(d',t) = \frac{E_0 d_0}{d'}\cos(\omega_c(t - \frac{d'}{c}))$$

$$E_g(d'',t) = \Gamma \frac{E_0 d_0}{d''}\cos(\omega_c(t - \frac{d''}{c}))$$

The laws of reflection give

$$\theta_i = \theta_0$$

$$E_g = \Gamma E_i$$

$$E_t = (1+\Gamma) E_i$$

For small values of θ_i (i.e., grazing incidence), as shown in Example 4.4 $\Gamma_\perp = -1$, $E_t = 0$.

- Assuming perfect horizontal E-field polarization

$$E_{TOT}(d,t) = \frac{E_0 d_0}{d^{'}}\cos(\omega_c(t - \frac{d^{'}}{c})) + (-1)\frac{E_0 d_0}{d^{''}}\cos(\omega_c(t - \frac{d^{''}}{c}))$$

using the method of images in next page,

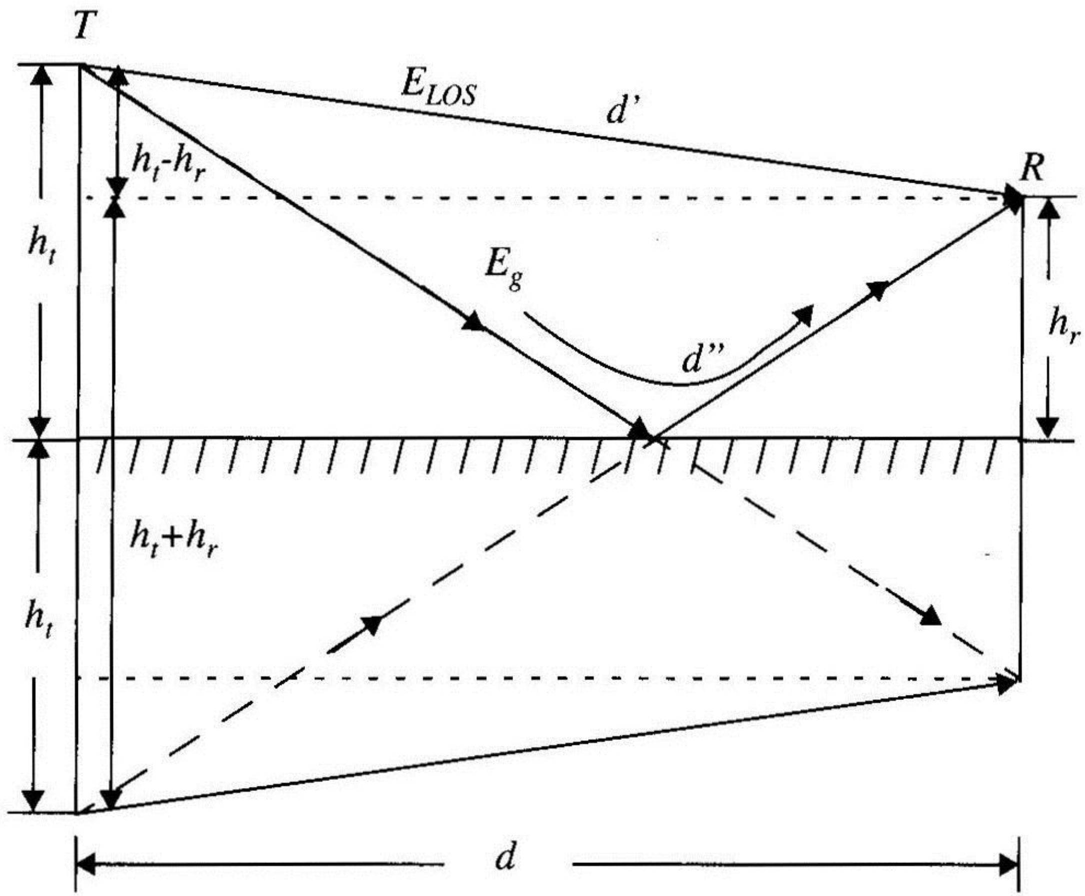


Figure 4.8 The method of images is used to find the path difference between the line-of-sight and the ground reflected paths.

$$\Delta = d'' - d' = \sqrt{(h_t + h_r)^2 + d^2} - \sqrt{(h_t - h_r)^2 + d^2}$$

$$\approx \frac{2h_t h_r}{d} \quad (\because d >> (h_t + h_r))$$

$$\theta_\Delta = \frac{2\pi\Delta}{\lambda} = \frac{\Delta\omega_c}{c}$$

$$\tau_{\mathrm{d}} = \frac{\Delta}{c} = \frac{\theta_\Delta}{2\pi f_c}$$

- As d becomes large, the difference between the distances d' and d'' becomes very small

$$\left|\frac{E_0 d_0}{d}\right| \approx \left|\frac{E_0 d_0}{d'}\right| \approx \left|\frac{E_0 d_0}{d''}\right|$$

E_{LOS} and E_g differ only in phase.

- At $t = d''/c$

$$E_{TOT}\left(d, t = \frac{d''}{c}\right) = \frac{E_0 d_0}{d} \cos\left(\omega_c \left(\frac{d'' - d'}{c}\right)\right) - \frac{E_0 d_0}{d''} \cos 0^\circ$$

$$\approx \frac{E_0 d_0}{d} [\angle \theta_\Delta - 1]$$

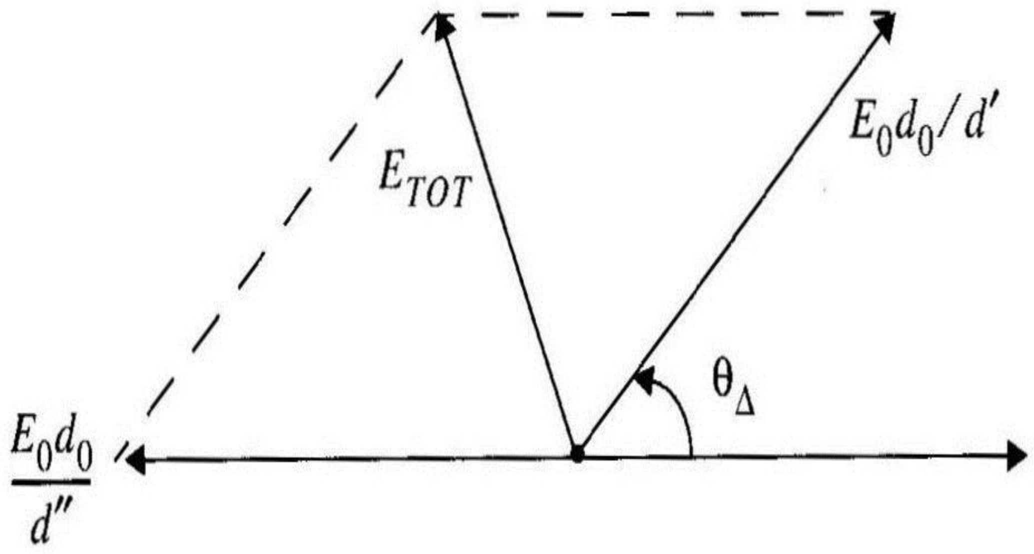


Figure 4.9 Phasor diagram showing the electric field components of the line-of-sight, ground reflected, and total received E-fields, derived from Equation (4.45).

$$|E_{TOT}(d)| = \left(\frac{E_0 d_0}{d}\right)\sqrt{(\cos\theta_\Delta - 1)^2 + \sin\theta_\Delta}$$

$$= \frac{E_0 d_0}{d}\sqrt{2 - 2\cos\theta_\Delta}$$

$$= 2\frac{E_0 d_0}{d}\sin\left(\frac{\theta_\Delta}{2}\right)$$

$E_{TOT}(d)$ decays in an oscillatory fashion, with local maxima being 6dB greater than the free space value and the local minima plummeting to –∞ dB (although in reality this never happens).

- As d is large

$$\Rightarrow \frac{\theta_\Delta}{2} < 0.3$$

$$\sin\frac{\theta_\Delta}{2} \approx \frac{\theta_\Delta}{2} \approx \frac{2\pi h_t h_r}{\lambda d} < 0.3 \quad \text{which implies}$$

$$E_{TOT}(d) \approx \frac{2E_0 d_0}{d}\frac{2\pi h_t h_r}{\lambda d} \approx \frac{k}{d^2} \quad \text{V/m}$$

$$\text{as} \quad d > \frac{20\pi h_t h_r}{3\lambda} \approx \frac{20 h_t h_r}{\lambda}$$

where k is a constant related to E_0 .

- Combining Equations (4.2) ,(4.15) and (4.51), the received power at distance d can be expressed as

$$P_r = P_t G_t G_r \frac{h_t^{\,2} h_r^{\,2}}{d^4}$$

$$\left(P_r = \frac{|E|^2}{120\pi} A_e = \frac{|E|^2}{120\pi} \times \frac{G_r \lambda^2}{4\pi} = \frac{4\pi |E_0|^2 d_0^2}{120\pi} G_r \frac{h_t^{\,2} h_r^{\,2}}{d^4}\right)$$

$$\left(\frac{|E_0|^2}{120\pi} = \frac{P_t G_t}{4\pi d_0^2}\right)$$

- The receiver power falls off with distance raised to the fourth power, or at a rate of 40 dB/decade. Path loss is more rapid than in free space and independent of frequency.

$$PL(dB) = 40 \log d - (10 \log G_t + 10 \log G_r + 20 \log h_t + 20 \log h_r)$$

Example 4.6

A mobile is located 5 km away from a base station and uses a vertical $\lambda/4$ monopole antenna with a gain of 2.55 dB to receive cellular radio signals. The E-field at 1 km from the transmitter is measured to be 10^{-3} V/m. The carrier frequency used for this system is 900 MHz.

(a) Find the length and the effective aperture of the receiving antenna.

(b) Find the received power at the mobile using the two-ray ground reflection model assuming the height of the transmitting antenna is 50 m and the receiving antenna is 1.5 m above ground.

Solution:

T-R separation distance = 5 km ,

E-field at a distance of 1km=10^{-3} V/m

Frequency of operation, f = 900MHz

$$\lambda = \frac{c}{f} = \frac{3\times10^8}{900\times10^6} = 0.333\text{m}$$

(a) $L = \lambda/4 = 0.333/4 = 0.0833\text{m} = 8.33\text{cm}$

$$\text{Effective aperature of antenna} = 0.016\text{m}^2$$

(b) since $d >> \sqrt{h_t h_r}$

$$E_R(d) \approx \frac{2E_0 d_0}{d}\frac{2\pi h_t h_r}{\lambda d} \approx \frac{k}{d^2} \text{ V/m}$$

$$= \frac{2\times10^{-3}\times1\times10^3}{5\times10^3}\left[\frac{2\pi(50)(1.5)}{0.333(5\times10^3)}\right] = 113.1\times10^{-6}\text{ V/m}$$

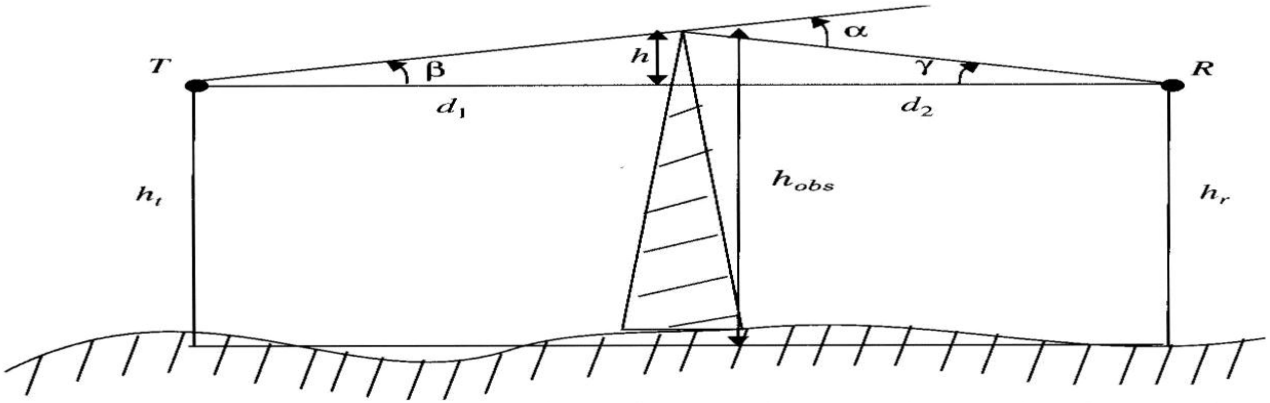
$$P_r(d) = \frac{(113.1\times10^{-6})^2}{377}\left[\frac{1.8(0.333)^2}{4\pi}\right]$$

$$P_r(d = 5\text{km}) = 5.4\times10^{-13}\text{W} = -122.68\text{ dBW or } -92.68\text{ dBm}$$

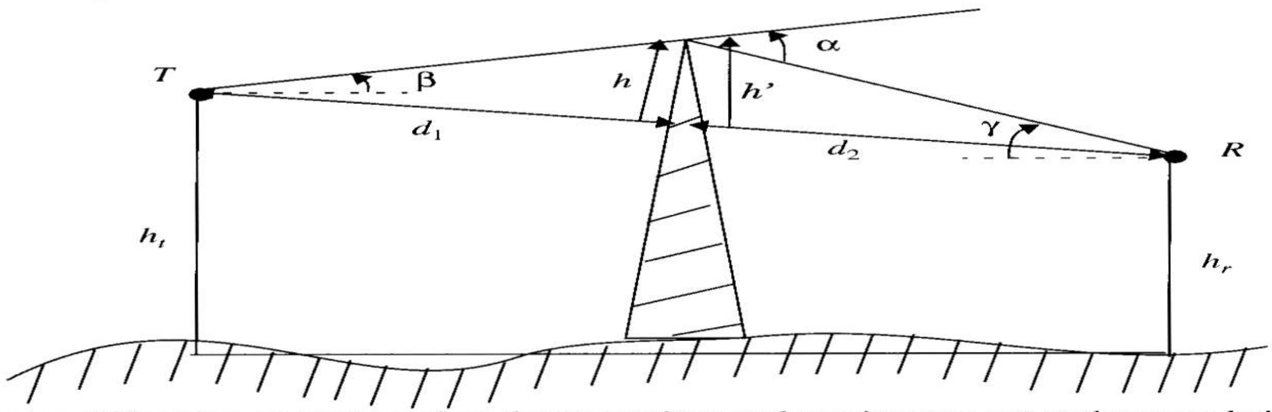
4.7 Diffraction

- Diffraction allows radio signals to propagation behind obstructions.
- Huygen's principle: all points on a wavefront can be considered as point sources of the production of secondary wavelets, and that these wavelets combine to produce a new wavefront in the direction of propagation.
- Diffraction is caused by the propagation of secondary wavelets into a shadowed region.

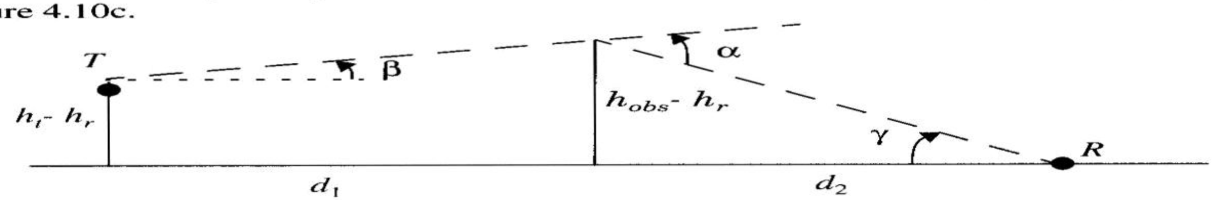
4.7.1 Fresnel Zone Geometry



(a) Knife-edge diffraction geometry. The point T denotes the transmitter and R denotes the receiver, with an infinite knife-edge obstruction blocking the line-of-sight path.



(b) Knife-edge diffraction geometry when the transmitter and receiver are not at the same height. Note that if α and β are small and $h << d_1$ and d_2, then h and h' are virtually identical and the geometry may be redrawn as shown in Figure 4.10c.



(c) Equivalent knife-edge geometry where the smallest height (in this case h_r) is subtracted from all other heights.

Figure 4.10 Diagrams of knife-edge geometry.

- Assuming h <<d_1,d_2 and h >> λ, the excess path length Δ (the difference between the direct path and the diffracted path) is

$$\Delta \approx \frac{h^2}{2}\frac{(d_1 + d_{2)}}{d_1 d_2}$$

- The corresponding phase difference is given by

$$\phi = \frac{2\pi\Delta}{\lambda} \approx \frac{2\pi}{\lambda}\frac{h^2}{2}\frac{(d_1 + d_{2)}}{d_1 d_2}$$

- Figure 4.10c

  when tan$x \approx x$, α = β+γ $\approx h\frac{(d_1 + d_{2)}}{d_1 d_2}$

- Define *Fresnel-Kirchoff* diffraction parameter v as

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

$$\Rightarrow \quad \phi = \frac{\pi}{2}v^2 \qquad \text{(a convenient form)}$$

- It is clear that ϕ is a function of h and the position of the obstruction.
- *Fresnel-Kirchoff* represent successive regions where $\Delta = n\frac{\lambda}{2}$

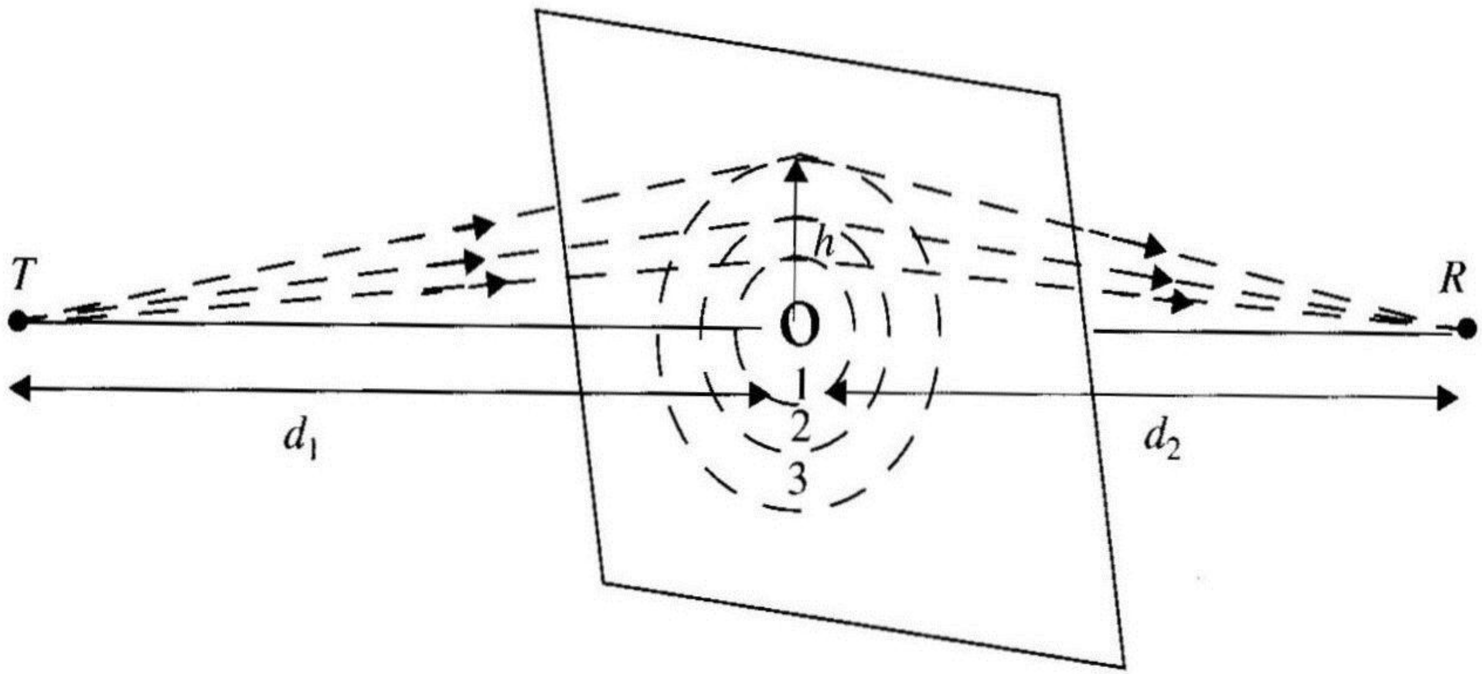


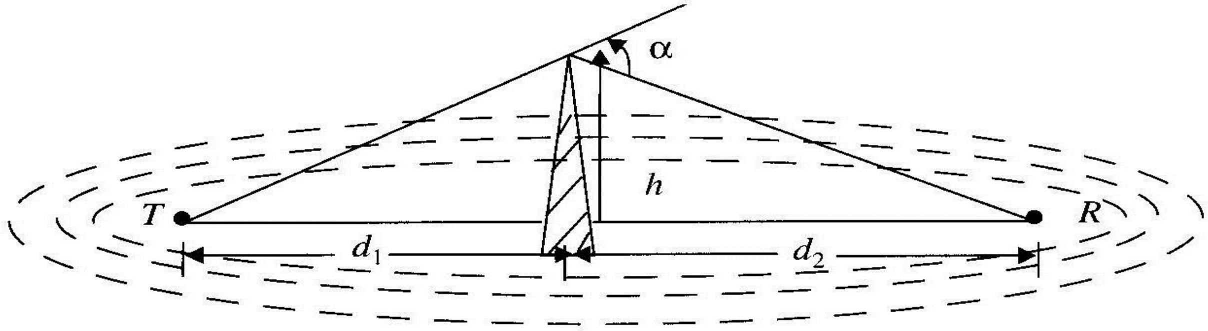
Figure 4.11 Concentric circles which define the boundaries of successive Fresnel zones.

These circles are called Fresnel zone. The radius of the *n*th Fresnel zone circle

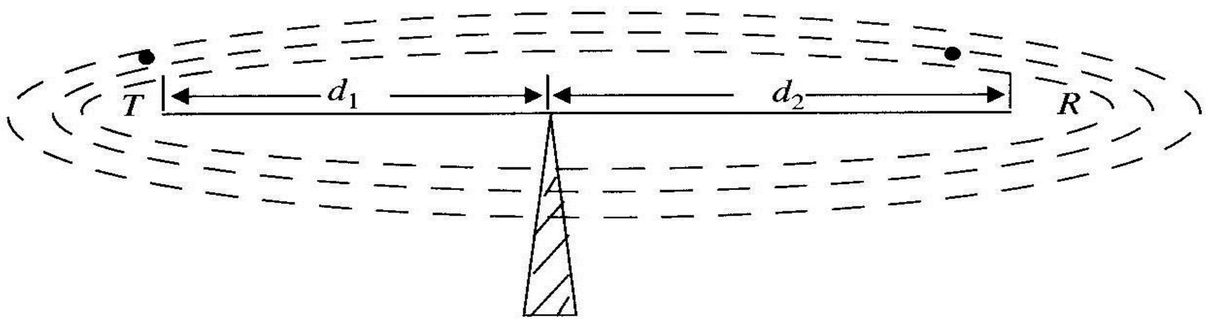
$$r_n = \sqrt{\frac{n\lambda d_1 d_2}{d_1 + d_2}}$$

For $d_1, d_2 >> r_n$.

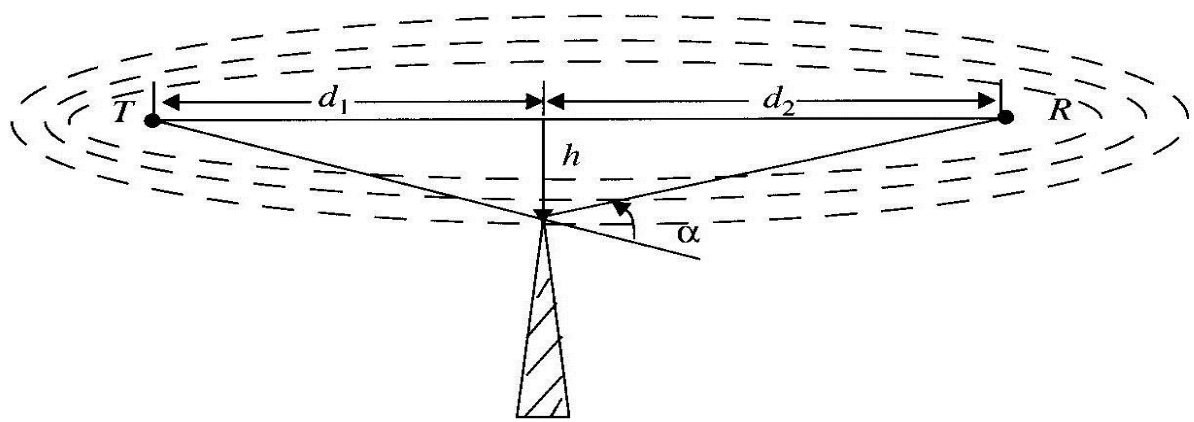
- r_n have maximum radius if $d_1 = d_2$, i.e., the plane is midway between transmitter and receiver. Hence , this illustrates how shadowing is sensitive to the frequency as well as the location of obstructions with relation to the transmitter and receiver.
- Diffraction loss occurs from the blockage of secondary waves such that only a portion of the energy is diffracted around an obstacle.



(a) α and ν are positive, since h is positive



(b) α and ν are equal to zero, since h is equal to zero



(c) α and ν are negative, since h is negative

Figure 4.12 Illustration of Fresnel zones for different knife-edge diffraction scenarios.

- As long as 55% of first Fresnel zone is kept clear, then further Fresnel zone clearance does not significantly alter the diffraction loss.

4.7.2 Knife-edge Diffraction Model

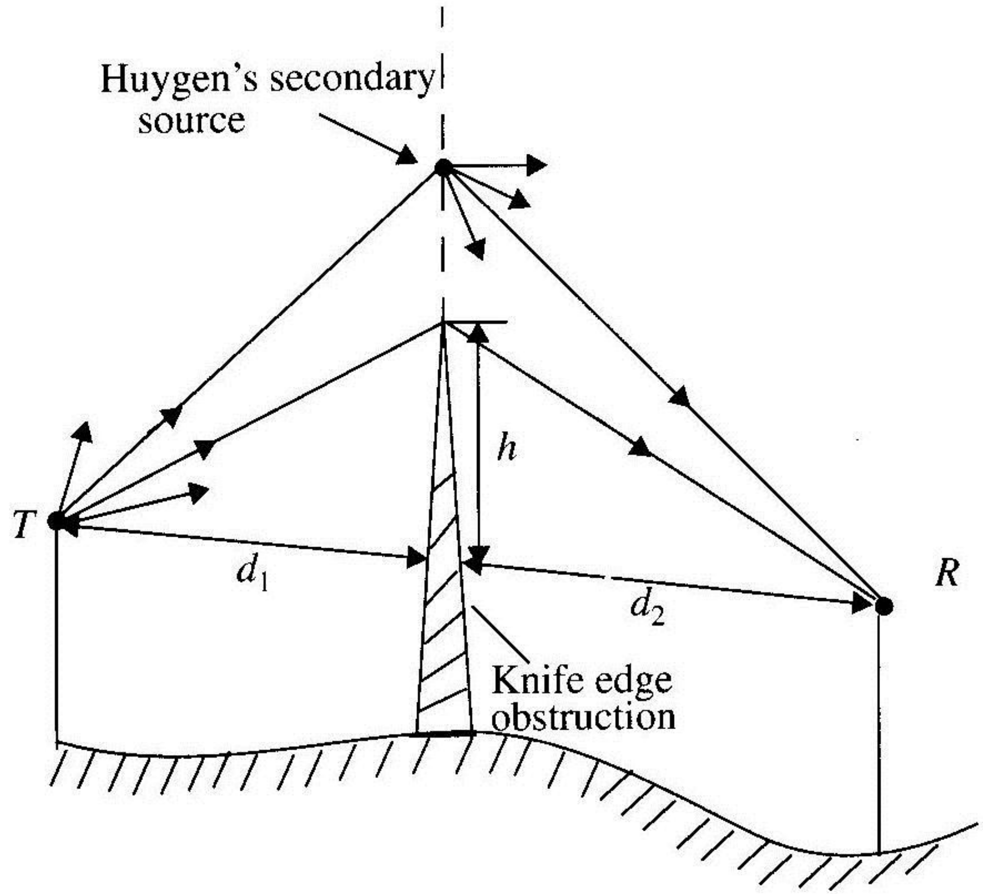


Figure 4.13 Illustration of knife-edge diffraction geometry. The receiver R is located in the shadow region.

- The field strength at point *R*, located in the shadowed region (diffraction zone) is a vector sum of the fields due to all of the secondary Huygen's sources in the plane above the knife edge. The electric field strength, E_d, is given by

$$\frac{E_d}{E_0} = F(v) = \frac{(1+j)}{2}\int_v^{\infty} \exp((-j\pi t^2)\ /2)\ dt$$

where E_0 is the free space field strength in the absence of both the ground and the knife edge and $F(v)$ is the complex Fresnel integral.

- The diffraction gain due to the presence of a knife edge, as compared to the free space E-field, is given by

$$G_d(dB) = 20\log|F(v)|$$

- An approximate solution is (provided by Lee)

$$G_d(dB) = 0 \qquad v \le -1$$

$$G_d(dB) = 20\log(0.5 - 0.62v) \qquad -1 \le v \le 0$$

$$G_d(dB) = 20\log(0.5\exp(-0.95v)) \qquad 0 \le v \le 1$$

$$G_d(dB) = 20\log\left(0.4 - \sqrt{0.1184 - (0.38 - 0.1v)^2}\right) \qquad 1 \le v \le 2.4$$

$$G_d(dB) = 20\log\left(\frac{0.225}{v}\right) \qquad v > 2.4$$

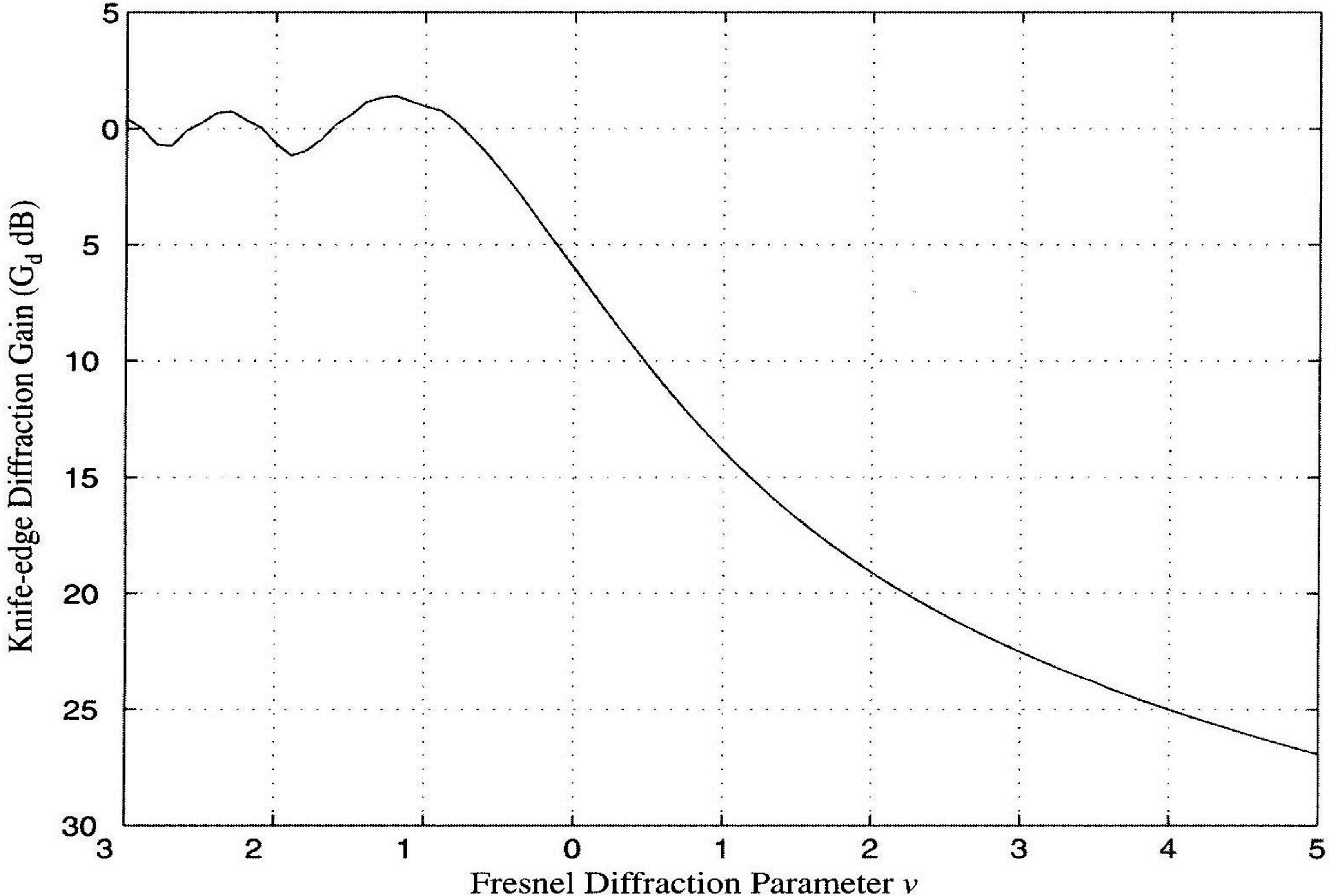


Figure 4.14 Knife-edge diffraction gain as a function of Fresnel diffraction parameter v.

Example 4.7

Compute the diffraction loss for the three cases shown in Figure 4.12. Assume λ=1/3 m, d_1 =1 km, d_2 = 1 km ,and (a) h = 25 m, (b) h = 0, (c) h = –25 m. Compare your answers using values from Figure4.14, as well as the approximate solution given by Equation (4.61.a)-(4.61.e). For each of these cases, identify the Fresnel zone within which the tip of the obstruction lies.

Solution:

λ = 1/3, d_1 = 1 km, d_2 = 1 km.

$(a)\ \ h = 25m,$

$$v = h\sqrt{\frac{2(d_1 + d_2)}{\lambda d_1 d_2}} = 25\sqrt{\frac{2(1000+1000)}{(1/3)\times 1000\times 1000}} = 2.74$$

The diffraction loss is 22 dB.

The path length difference between the direct and diffraction rays is

$$\Delta \approx \frac{h^2}{2}\frac{(d_1+d_2)}{d_1 d_2} = \frac{25^2}{2}\frac{(1000+1000)}{1000\times 1000} = 0.625m$$

$$\Delta = n\lambda/2 \ \ \Rightarrow \ \ n = \frac{2\Delta}{\lambda} = \frac{2\times 0.625}{0.3333} = 3.75$$

Solution:

(b) $h = 0$ m, the Fresnel diffraction parameter $v = 0$.

From Figure 4.14 the diffraction loss is 6 dB.

For this case $h = 0$ and $\Delta = 0$, the tip of the obstruction lies in the middle of the first Fresnel zone.

(c) $h = -25$ m , the Fresnel diffraction parameter $v = -2.74$.

From Figure 4.14 the diffraction loss is 1 dB.

Using the numerical approximation in Equation(4.61a), the diffraction loss is equal to 0 dB.

Example 4.8

Given the following geometry, determine (a) the loss due to knife-edge diffraction ,and (b) the height of the obstacle required to induce 6 dB diffraction loss. Assume f = 900 MHz.

Solution:

(a) The wavelength $\lambda = \dfrac{c}{f} = \dfrac{3\times10^{8}}{900\times10^{6}} = \dfrac{1}{3}$ m

$$\beta = \tan^{-1}\left(\frac{75-25}{10000}\right) = 0.2865^{\circ}$$

$$\gamma = \tan^{-1}\left(\frac{75}{2000}\right) = 2.15^{\circ}$$

$$\alpha = \beta + \gamma = 2.434^{\circ} = 0.0424 \text{ rad}$$

$$v = 0.0424\sqrt{\frac{2\times10000\times2000}{(1/3)\times(10000+2000)}} = 4.24$$

The diffraction loss is 25.5 dB.

(b) For 6 dB loss, $v = 0$.

$$\frac{h}{2000} = \frac{25}{12000}, \quad h = 4.16\text{m}.$$

4.7.3 Multiple Knife-edge Diffraction

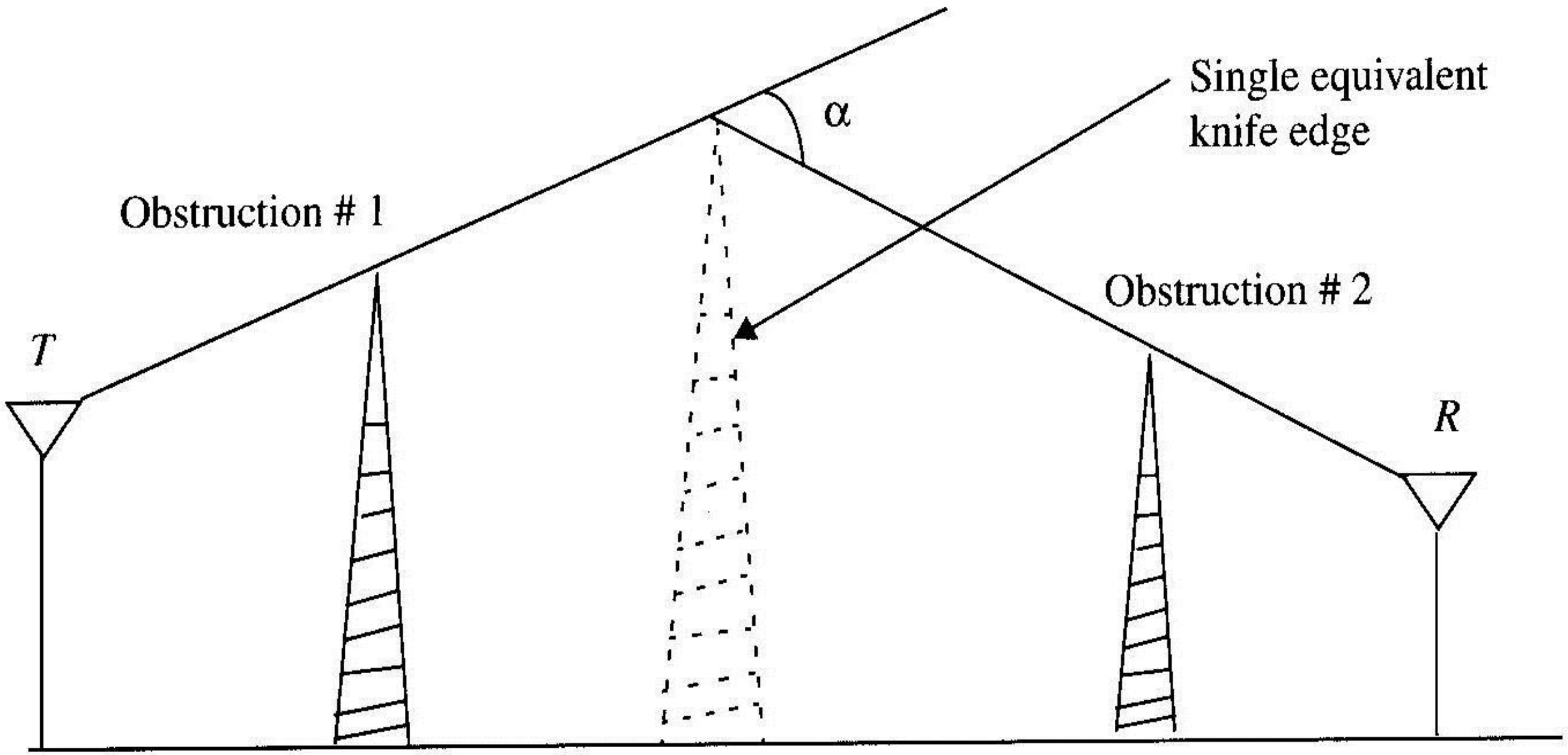


Figure 4.15 Bullington's construction of an equivalent knife edge [from [Bul47] © IEEE].

4.8 Scattering

- Objects such as lamp posts and trees tend to scatter energy in all directions, thereby providing additional radio energy at a receiver.
- Surface roughness is often tested using the Rayleigh criterion which defines a critical height h_c given by

$$h_c = \frac{\lambda}{8 \sin \theta_i}$$

- The surface protuberance $h < h_c \longrightarrow$ the surface is smooth .

  $h > h_c \longrightarrow$ the surface is rough.

- For rough surfaces, $\Gamma_{rough} = \rho_s \Gamma$ where the scattering loss factor ρ_s is to account for the diminished reflected wave.
- Ament found that

$$\rho_s = \exp[-8(\frac{\pi \sigma_h \sin \theta_i}{\lambda})^2]$$

Boithias modified to

$$\rho_s = \exp[-8(\frac{\pi \sigma_h \sin \theta_i}{\lambda})^2] I_0[8(\frac{\pi \sigma_h \sin \theta_i}{\lambda})^2]$$

4.8.1 Radar Cross Section Model

- Knowledge of the physical location and size of the scattering object is helpful to predict the received signal strength.
- The radar cross section (RCS) of a scattering object is defined as the ratio of the power density of the signal scattered in the direction of the receiver to the power density of the radio wave incident upon the scattering object, and has units of squared meters.
- For urban mobile radio systems, models based on the bistatic radar equation may be used to compute the received power due to scattering in the far field,

$$P_R(\text{dBm}) = P_T(\text{dBm}) + G_T(\text{dBi}) + 20\log(\lambda) + RCS[\text{dB m}^2]$$
$$-30\log(4\pi) - 20\log d_T - 20\log d_R$$

- For medium and large size buildings located 5-10 km away, RCS values were found to be in the range of 14.1 to 55.7 dB m^2.

4.9 Practical Link Budget Design Using Path Loss Models

4.9.1 Log-distance Path Loss Model

- The average large-scale path loss is expressed as

$$\overline{PL}(d) \propto \left(\frac{d}{d_0}\right)^n$$

or

$$\overline{PL}(\mathrm{dB}) = \overline{PL}(d_0) + 10n\log\left(\frac{d}{d_0}\right)$$

where n is the path loss exponent.

- In free space, $n = 2$ when obstructions are present, $n \geq 2$

Table 4.2 Path Loss Exponents for Different Environments

Environment	Path Loss Exponent, *n*
Free space	2
Urban area cellular radio	2.7 to 3.5
Shadowed urban cellular radio	3 to 5
In building line-of-sight	1.6 to 1.8
Obstructed in building	4 to 6
Obstructed in factories	2 to 3

- d_0 = 1 km, for large coverage cellular systems.
  d_0 = 100 m or 1 m, for microcellular systems.

4.9.2 Log-normal Shadowing

- The path loss $PL(d)$ at a particular location is random and distributed log-normally (normal in dB) about the mean distance-dependent value

$$PL(d)[dB] = \overline{PL}(d) + X_\sigma = \overline{PL}(d_0) + 10n\log\left(\frac{d}{d_0}\right) + X_\sigma$$

and

$$P_r(d)[dBm] = P_t[dBm] - PL(d)[dB]$$

where X_σ is a zero-mean Gaussian distributed random variable (in dB) with standard deviation σ (also in dB).

- The log-normal distribution describes the random shadowing effects, or is refered to as log-normal shadowing.

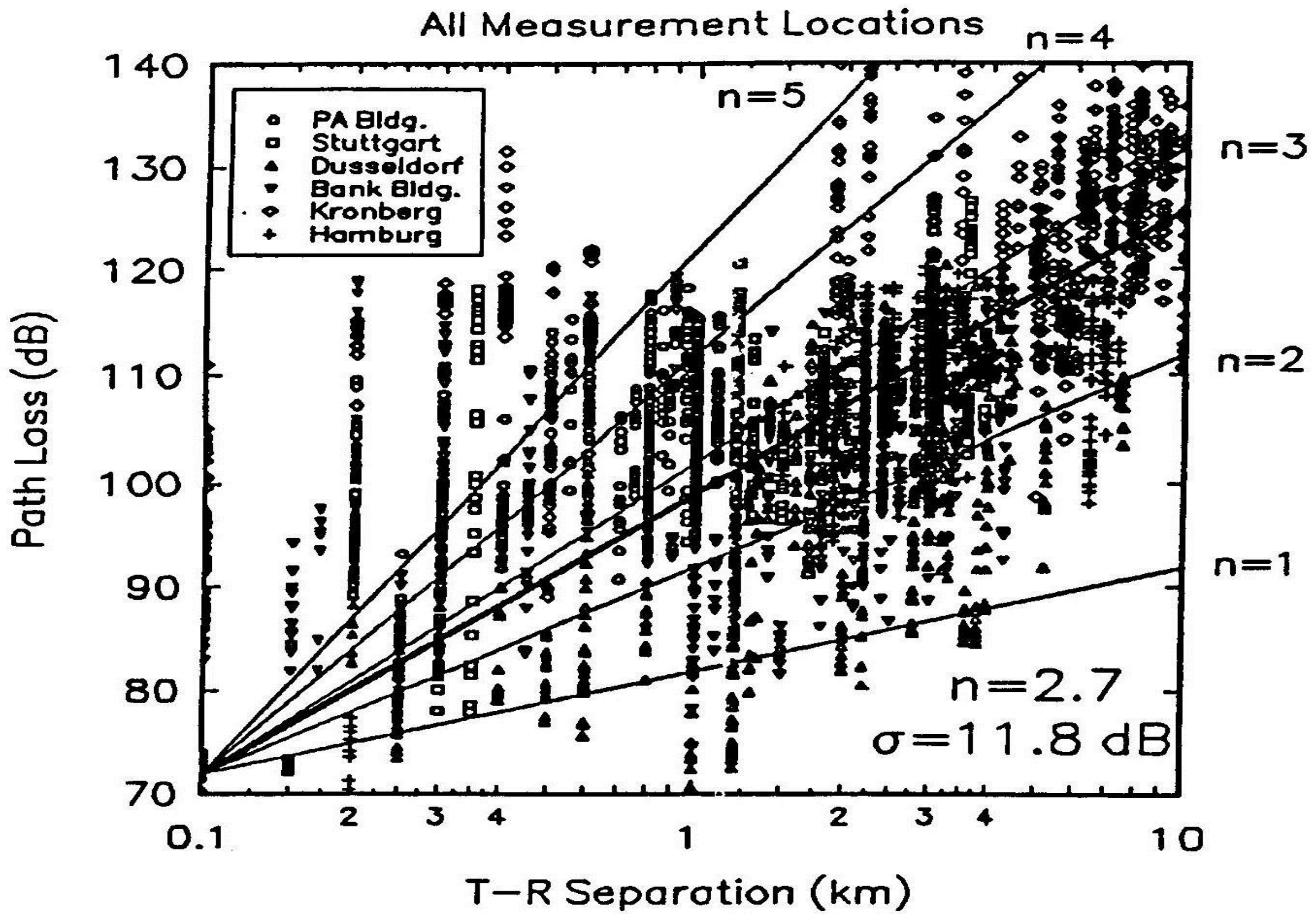


Figure 4.17 Scatter plot of measured data and corresponding MMSE path loss model for many cities in Germany. For this data, $n = 2.7$ and $\sigma = 11.8$ dB [from [Sei91] © IEEE].

- Since $PL(d)$ is normal in dB, so is $P_r(d)$. The probability that is $P_r(d)$ (in dB) will exceed a certain value γ can be calculated as

$$\Pr[P_r(d) > \gamma] = Q\left(\frac{\gamma - \overline{P_r(d)}}{\sigma}\right)$$

where Q-function or error function are defined as

$$Q(z) = \frac{1}{\sqrt{2\pi}} \int_z^{\infty} \exp\left(-\frac{x^2}{2}\right)$$

$$= \frac{1}{2}\left[1 - erf\left(\frac{z}{\sqrt{2}}\right)\right]$$

and

$$Q(z) = 1 - Q(-z)$$

- The probability that the received signal level will be below γ is given by

$$\Pr[P_r(d) < \gamma] = Q\left(\frac{\overline{P_r(d)} - \gamma}{\sigma}\right)$$

4.9.3 Determination of Percentage of Coverage Area

- Since $Pr[P_r(r) > \gamma]$ is the probability that the random received signal at $d = r$ exceeds the threshold γ within an incremental area dA, the percentage of useful service area $U(\gamma)$ (i.e., the percentage of area with a received signal that is equal or greater than γ)

$$U(\gamma) = \frac{1}{\pi R^2}\int \Pr[P_r(r) > \gamma]dA$$

$$= \frac{1}{\pi R^2}\int_0^{2\pi}\int_0^{R} \Pr[P_r(r) > \gamma]r\,dr d\theta$$

- Using (4.71) and substituting $t = a + b\log(r/R)$, it can be shown that

$$U(\gamma) = \frac{1}{2}(1 - erf(a) + \exp(\frac{1-2ab}{b^2})[1 - erf(\frac{1-ab}{b})])$$

If $\overline{P_r}(R) = \gamma$, i.e., $a = 0$, (4.78) becomes

$$U(\gamma) = \frac{1}{2}[1 + \exp(\frac{1}{b^2})(1 - erf(\frac{1}{b}))]$$

- Equation (4.78) may be evaluated for a large number of values of σ and n, as shown in

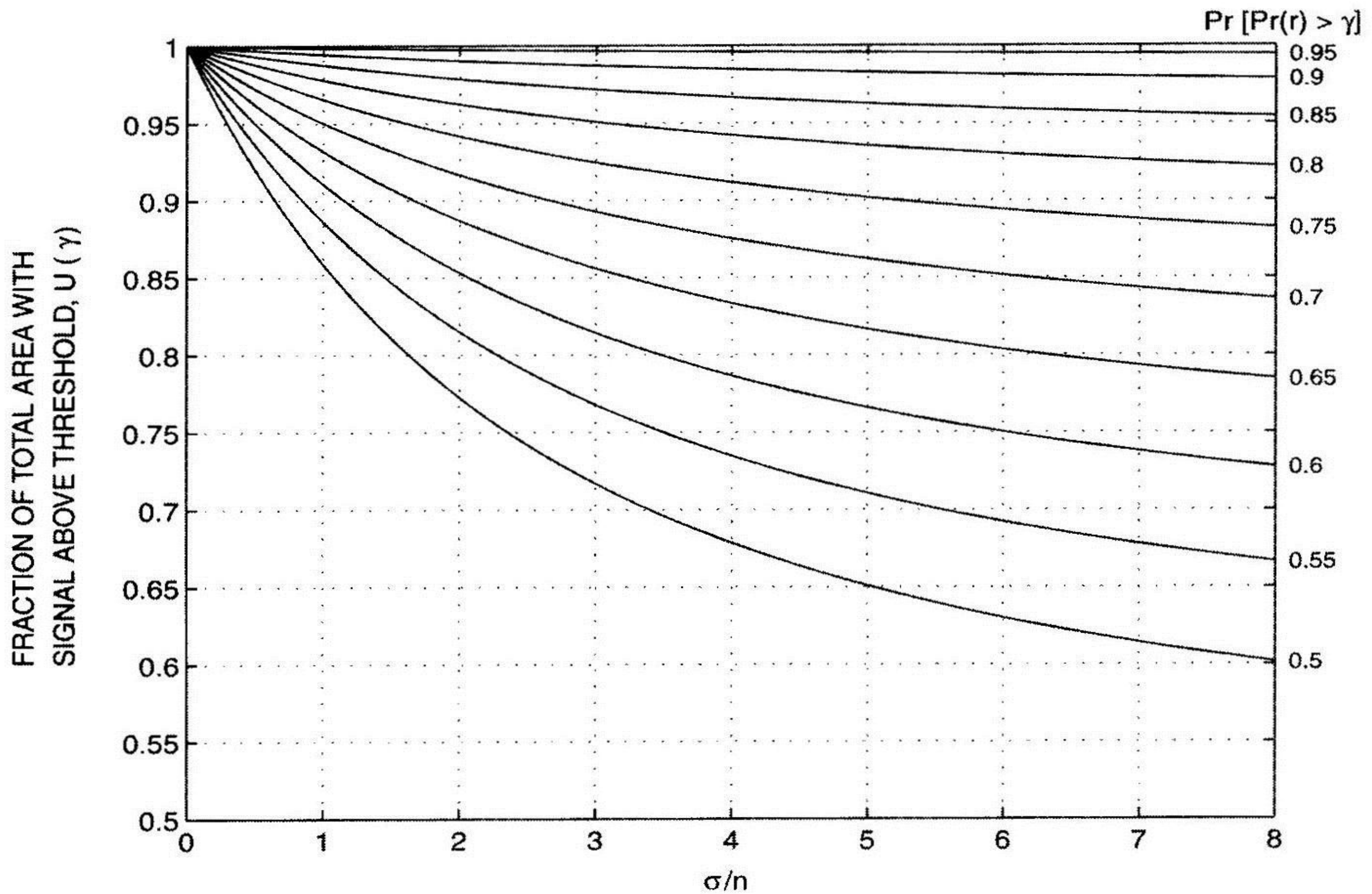


Figure 4.18 Family of curves relating fraction of total area with signal above threshold, $U(\gamma)$ as a function of probability of signal above threshold on the cell boundary.

- 75% boundary coverage means 75% of the time the signal is to exceed the threshold at the boundary.
- If $n = 4$, $\sigma = 8$ dB, 75% boundary coverage $\Rightarrow$ then the area coverage = 90%.

  If $n = 2$, $\sigma = 8$ dB, 75% boundary coverage $\Rightarrow$ area coverage = 86%.
- It is often useful to compute how the boundary coverage relates to the percent of area covered within the boundary.

Example 4.9

Four received power measurements were taken at distance of 100 m, 200 m, 1 km, and 3 km from a transmitter. These measured values are given in the following table. It is assumed that the path loss for these measurements follows the model in Equation(4.69a), where d_0 = 100 m:

(a) Find the minimum mean square error (MMSE) estimate for the path loss exponent, n ;

(b) Calculate the standard deviation about the mean value;

(c) Estimate the received power at d = 2 km using the resulting model;

(d) Predict the likelihood that the received signal level at 2 km will be greater than -60 dBm;

(e) Predict the percentage of area within a 2 km radius cell that receives signals greater than -60 dBm.

Solution:

The sum of squared errors between the measured and estimated value is

$$J(n) = \sum_{i}^{k} (p_i - \hat{p}_i)^2$$

(a) We find $\hat{p}_i = p_i(d_0) - 10n\log(d_i/100\mathrm{m})$.

$$\hat{p}_1 = 0,\ \hat{p}_2 = -3n,\ \hat{p}_3 = -10n,\ \hat{p}_4 = -14.77n.$$

$$J(n) = (0-0)^2 + (-20-(-3n))^2 + (-35-(-10n))^2 + (-70-(-14.77n))^2$$

$$= 6525 - 2887.8n + 327.153n^2$$

$$\frac{\mathrm{d}J(n)}{\mathrm{d}n} = 654.306n - 2887.8.$$

$$n = 4.$$

Solution:

$(b)\ \sigma^2 = J(n)/4$ at $n = 4.$

$$J(n) = (0+0) + (-20+13.2)^2 + (-35+44)^2 + (-70+64.988)^2$$
$$= 152.36.$$
$$\sigma^2 = 152.36/4 = 38.09\,\text{dB}^2,\ \ \sigma = 6.17\text{dB}.$$

$(c)\ d = 2$ km,

$$\hat{p}(d = 2\text{km}) = 0 - 10(4.4)\log(2000/100) = -57.24\text{dBm}.$$

(d) $P_r[\,P_r(d) > -60\text{dBm}] = Q(\frac{\gamma - \overline{P_r(d)}}{\sigma}) = Q(\frac{-60+57.24}{6.17}) = 67.4\%$

(e) Equation (4.78) or Figure 4.18 may be used to determine that 88% of the cell area receives coverage above -60 dB.

4.10 Outdoor Propagation Models

4.10.1 Longley-Rice Model

- The Longley-Rice model is applicable in the frequency range from 40 MHz to 100 GHz using two-ray ground reflection model, Fresnel-kirchoff knife edge model, forward scatter theory, and modified van der Pol-Bremmer method (predict far field diffraction).
- Also called the ITS irregular terrain model.
- It is also available as a computer program for frequencies between 20MHz and 10GHz.
- This model operates in two modes: point-to-point mode and area mode (depending on whether the terrain path profile is available or not).

- The urban factor (UF) deals with the additional attenuation due to urban clutter.
- Shortcomings: No correction factors to account for the effects of buildings and foliage, multipath is not considered.

4.10.2 Durkin's Model – A Case Study

- A computer simulator for predicting field strength contours over irregular terrain, that was adopted by the Joint Radio Committee (JRC) in the U.K..
- The execution of the Durkin path loss simulator consists of two parts:
  - Access a topographic data base and reconstruct the ground profile information along the radial joining of the transmitter to the receiver.
  - Calculates the expected path loss along that radial.
- The assumption is that the receiving antenna receives all of its energy along the radial (LOS and diffraction models) and experiences no multipath propagation (somewhat pessimistic).

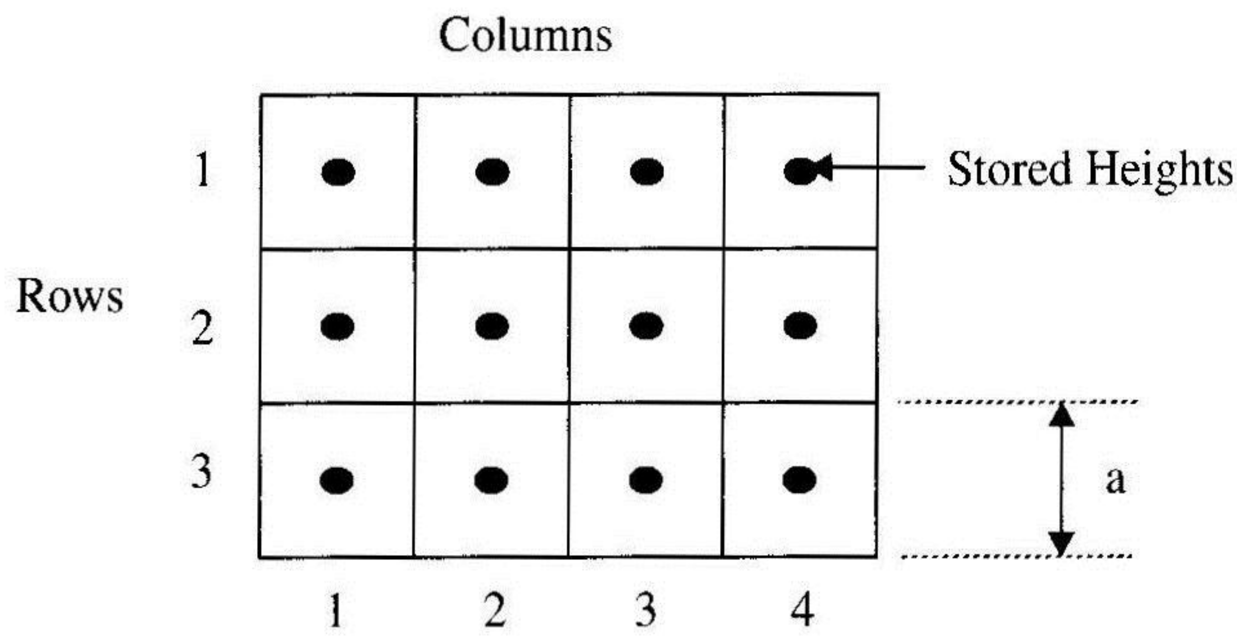
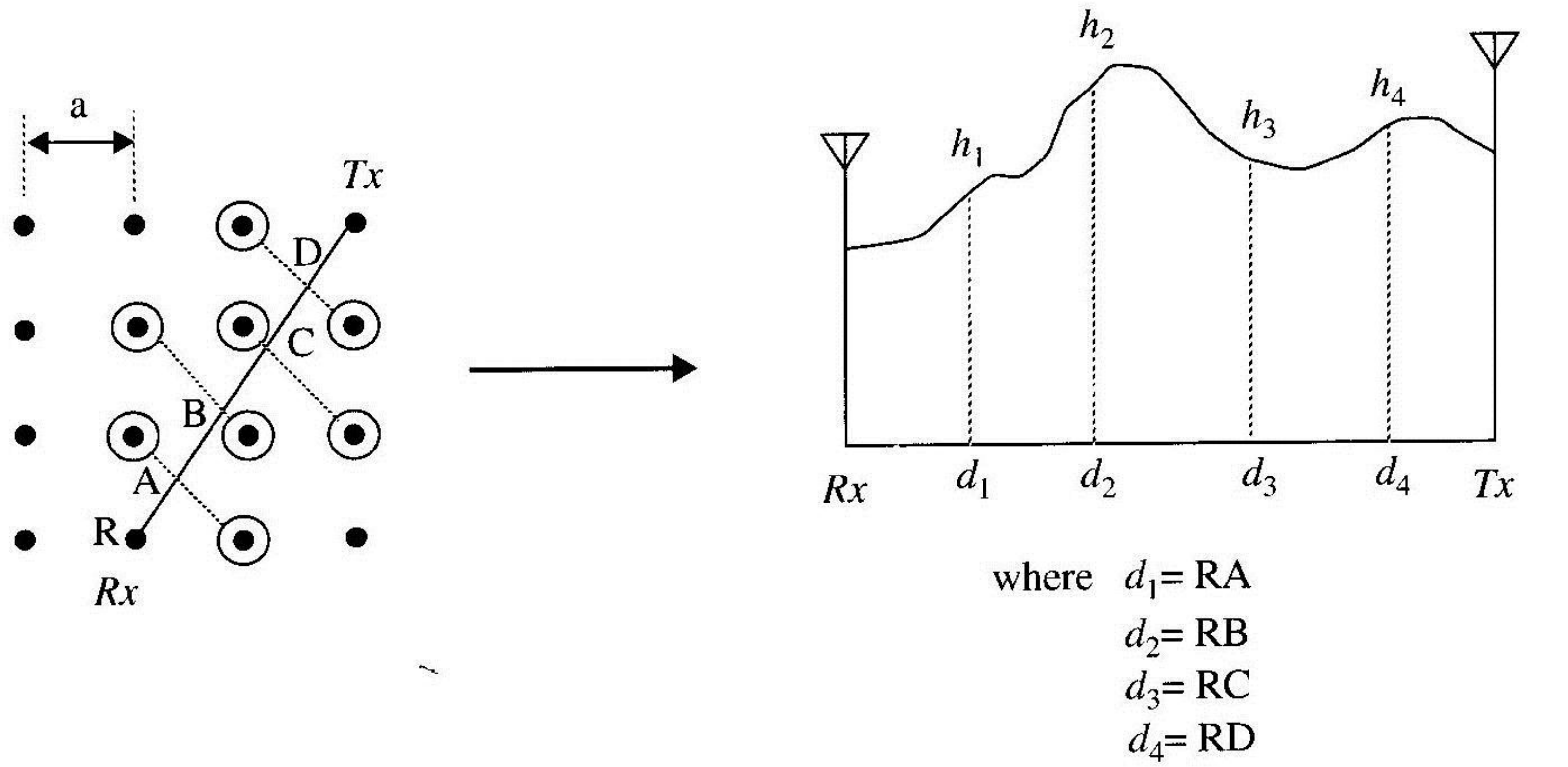


Figure 4.19 Illustration of a two-dimensional array of elevation information.



(a) Top view of interpolated map and line between Tx and Rx

(b) Side view showing reconstructed terrain profile between Tx and Rx

Figure 4.20 Illustration of terrain profile reconstruction using diagonal interpolation.

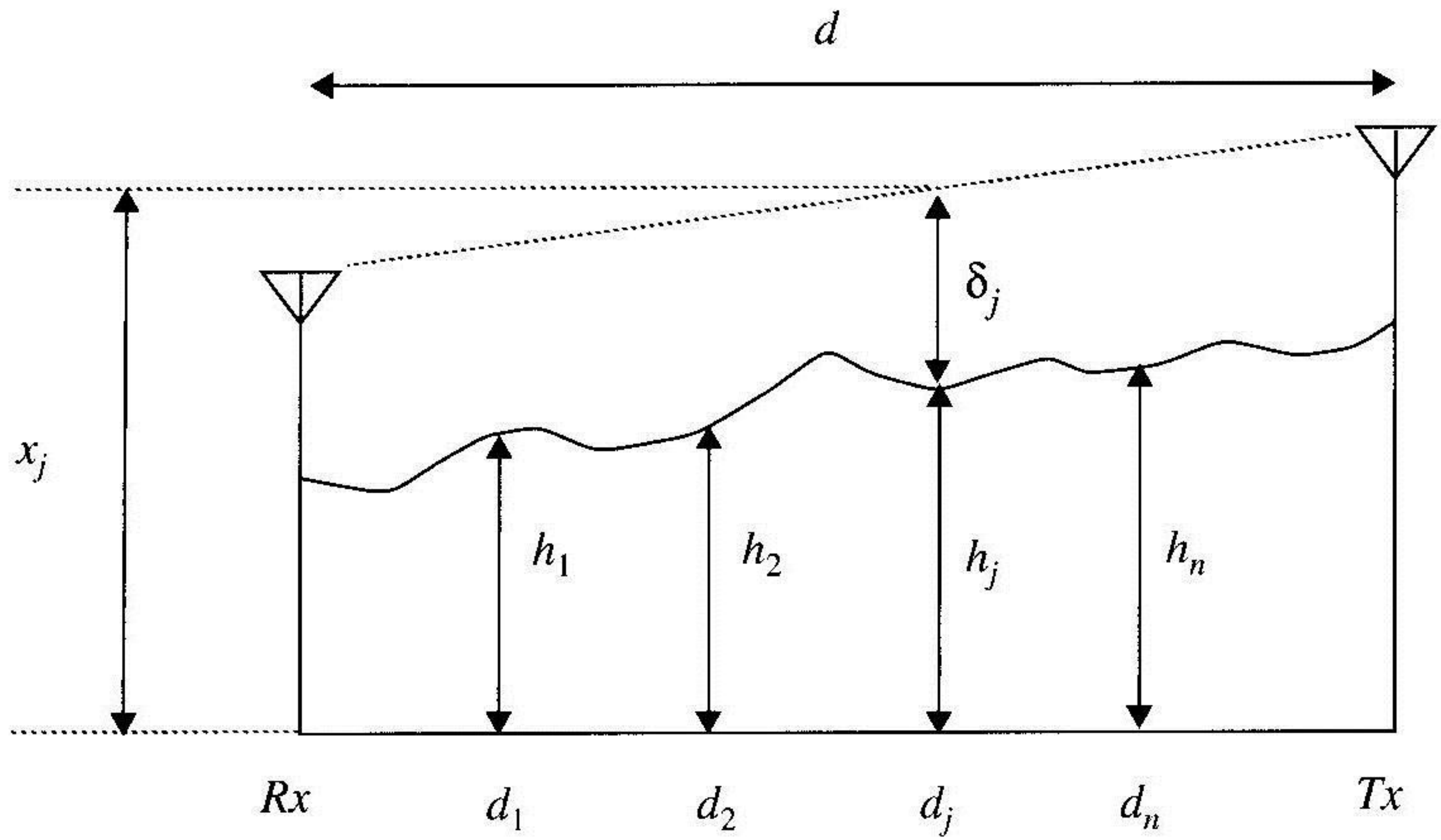


Figure 4.21 Illustration of line-of-sight (LOS) decision making process.

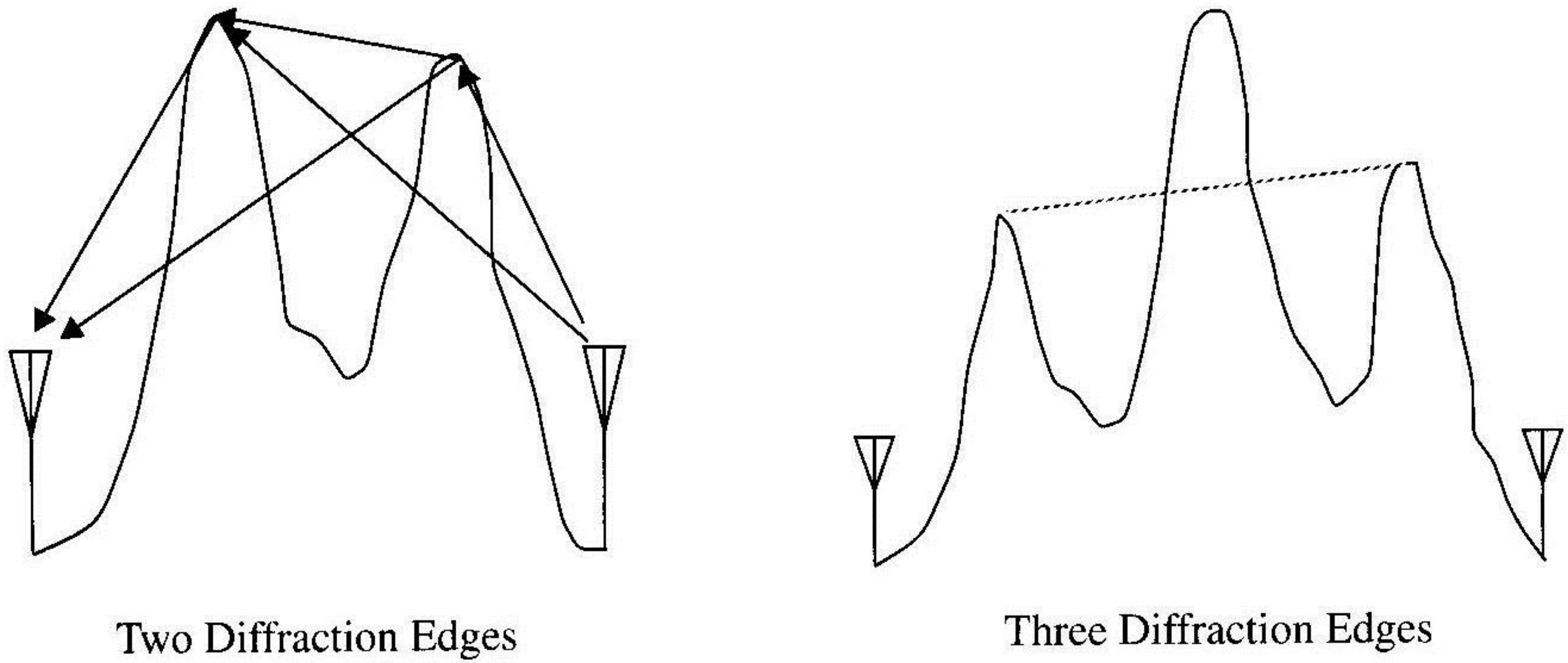


Figure 4.22 Illustration of multiple diffraction edges.

- Disadvantages :

  Do not predict propagation effects due to foliage, building, other man-made structures.

  Do not account for multipath propagation other than ground reflection, so additional loss factors are often included.

4.10.3 Okumura Model

- One of the most widely used models in urban areas.
- Applicable for frequencies in the range 150 MHz to 1920 MHz, distances of 1 km to 100 km, base station antenna heights ranging from 30 m to 1000m.
- Extensive measurements giving the median attenuation relative to free space.

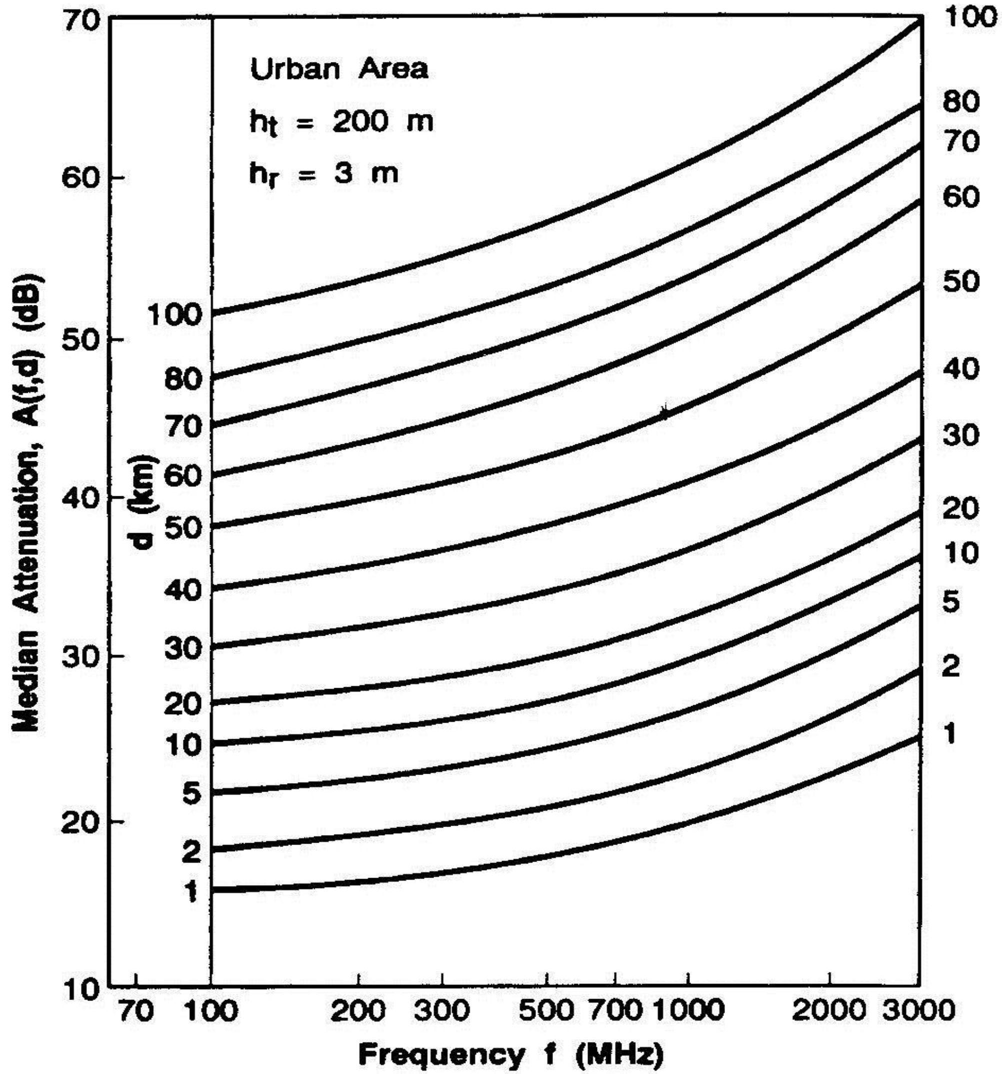


Figure 4.23 Median attenuation relative to free space $(A_{mu}(f,d))$, over a quasi-smooth terrain [from [Oku68] © IEEE].

- The model can be expressed as

$$L_{50}(dB) = L_F + A_{mu}(f,d) - G(h_{te}) - G(h_{re}) - G_{AREA}$$

G_{AREA} : the gain due to the type of environment.

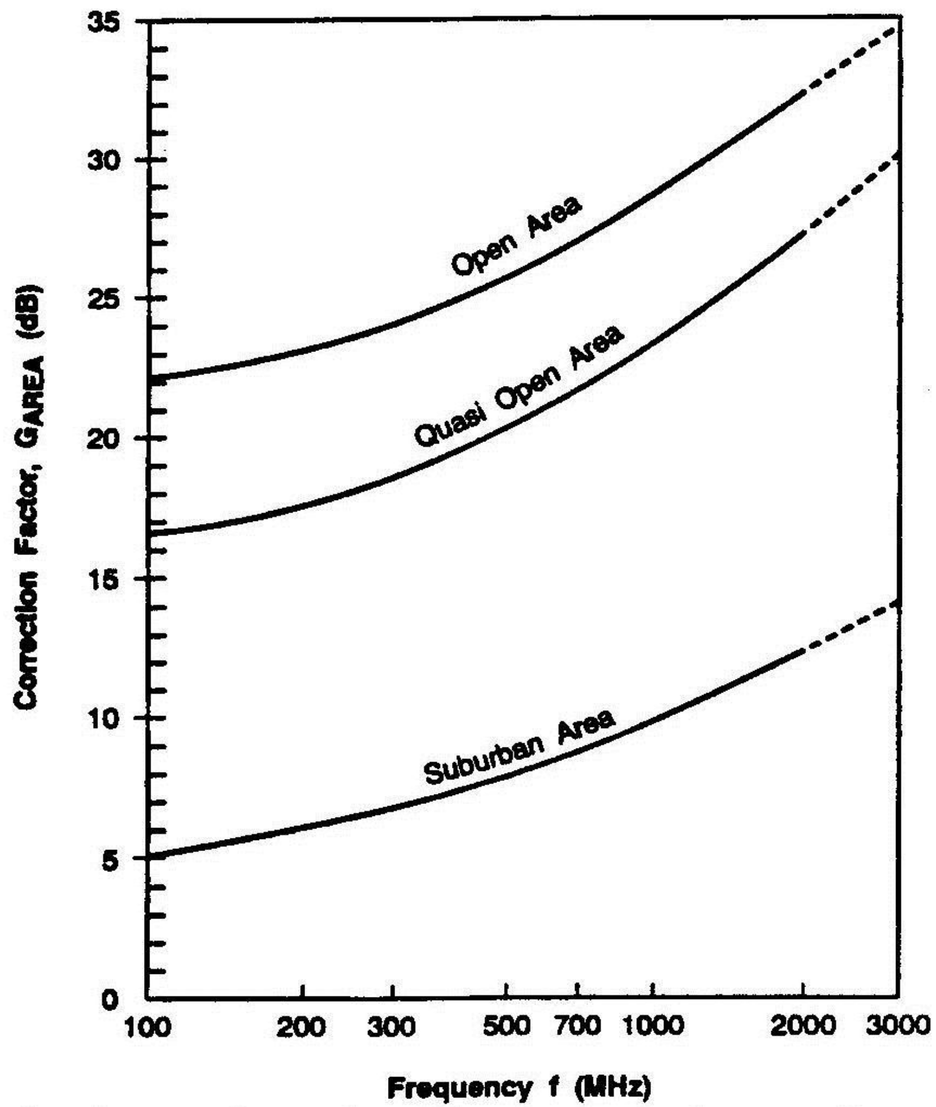


Figure 4.24 Correction factor, G_{AREA}, for different types of terrain [from [Oku68] © IEEE].

- $G(h_{te})$ varies at a rate of 20dB/decade and $G(h_{re})$ varies at a rate of 10dB/decade for $h_{re} \le 3$m.

$$G(h_{te}) = 20\log\left(\frac{h_{te}}{200}\right) \qquad 1000\text{m} > h_{te} > 30m$$

$$G(h_{re}) = 10\log\left(\frac{h_{re}}{3}\right) \qquad h_{re} \le 3m$$

$$G(h_{re}) = 20\log\left(\frac{h_{re}}{3}\right) \qquad 10\text{m} > h_{re} > 3\text{m}$$

- Other corrections may also be applied to Okumura's model.
- A standard for system planning in modern land mobile radio system in Japan.
- Fairly good in urban and suburban areas, bit not as good in rural areas.

<u>Example 4.10</u>

Find the median path los using Okumura's model for d = 50 km, h_{te} = 100 m, h_{re} = 10 m in a suburban environment. If the base station transmitter radiates as EIRP of 1 kW at a carrier frequency of 900 MHz, find the power at the receiver (assume a unity gain receiving antenna).

Solution:

The free space path loss L_F can be calculated as

$$L_F = 10\log[\frac{\lambda^2}{(4\pi)^2 d^2}] = 10\log[\frac{(3\times10^8/900\times10^6)^2}{(4\pi)^2\times(50\times10^3)^2}] = 125.5\ \text{dB}$$

From the Okumura curves

$$A_{mu}(900\ \text{MHz}(50\ \text{km})) = 43\ \text{dB}$$

and

$$G_{AREA} = 9\text{dB}$$

$$G(h_{te}) = 20\log(\frac{h_{te}}{200}) = 20\log(\frac{100}{200}) = -6\text{dB}$$

$$G(h_{re}) = 20\log(\frac{h_{re}}{3}) = 20\log(\frac{10}{3}) = 10.46\text{dB}$$

$$L_{50}(\text{dB}) = L_F + A_{mu}(f,d) - G(h_{te}) - G(h_{re}) - G_{AREA}$$

$$= 125.5 + 43\text{-}(\text{-}6)\text{-}10.46\text{-}9$$

$$= 155.04\ \text{dB}$$

Solution:

The median received power is

$$P_r(d) = EIRP(\text{dBm}) - L_{50}(\text{dB}) + G_r(\text{dB})$$
$$= 60\text{dBm} - 155.04\text{dB} + 0\text{dB} = -95.04\text{dBm}.$$

4.10.4 Hate Model

- The Hate model is an empirical formulation of the graphical path loss data provided by Okumura.

$$L_{50}(urban)(\mathrm{dB}) = 69.55 + 26.16\log f_c - 13.82\log h_{te} - a(h_{re}) + (44.9 - 6.55\log h_{te})\log d$$

$a(h_{re})$ is the correction factor for h_{re} which is a function of the size of the coverage area.

- For small to medium city

$$a(h_{re}) = (1.1\log f_c - 0.7)h_{re} - (1.56\log f_c - 0.8)\ \mathrm{dB}$$

- For large city

$$a(h_{re}) = 8.29(\log 1.54 h_{re})^2 - 1.1\text{dB} \quad \text{for } f_c \le 300\,\text{MHz}$$
$$a(h_{re}) = 3.2(\log 11.75 h_{re})^2 - 4.97\text{dB} \quad \text{for } f_c \ge 300\,\text{MHz}$$

- For suburban areas, the Hata model is

$$L_{50}(\text{dB}) = L_{50}(urban) - 2[\log(f_c/28)]^2 - 5.4$$

- For rural areas

$$L_{50}(\text{dB}) = L_{50}(urban) - 4.78(\log f_c)^2 + 18.33 \log f_c - 40.94$$

- The Hata model predictions are very closely to that of the Okumura's models when $d > 1$ km.
- This model is well suited for large cell mobile systems, but not personal communication system (PCS) which have cells on the order of 1 km radius.

4.10.5 PCS Extension to Hata Model

- The EURO-COST formed COST-231 working committee to develop an extended version of the Hata model to 2 GHz.

$$L_{20}(urban) = 46.3 + 33.9 \log f_c - 13.82 \log h_{te} - a(h_{re})$$
$$+ (44.9 - 6.55 \log h_{te}) \log d + C_M$$

4.10.6 Walfisch and Bertoni Model

- A model considers the impact of rooftops and building height by using diffraction to predict average signal strength at street level.
- The model considers the path loss, S, to be

$$S = P_0 Q^2 P_1$$

P_0 : free space path loss

Q^2 : the reduction in the rooftop signal due to the row of buildings which immediately shadow the receiver at street level.

P_1 : diffraction loss from the rooftop to the street.

- In dB

$$S(\text{dB}) = L_0 + L_{rts} + L_{ms}$$

L_0 : free space loss.

L_{rts} : rooftop-to-street diffraction and scatter loss.

L_{ms} : multiscreen diffraction loss due to the rows of buildings.

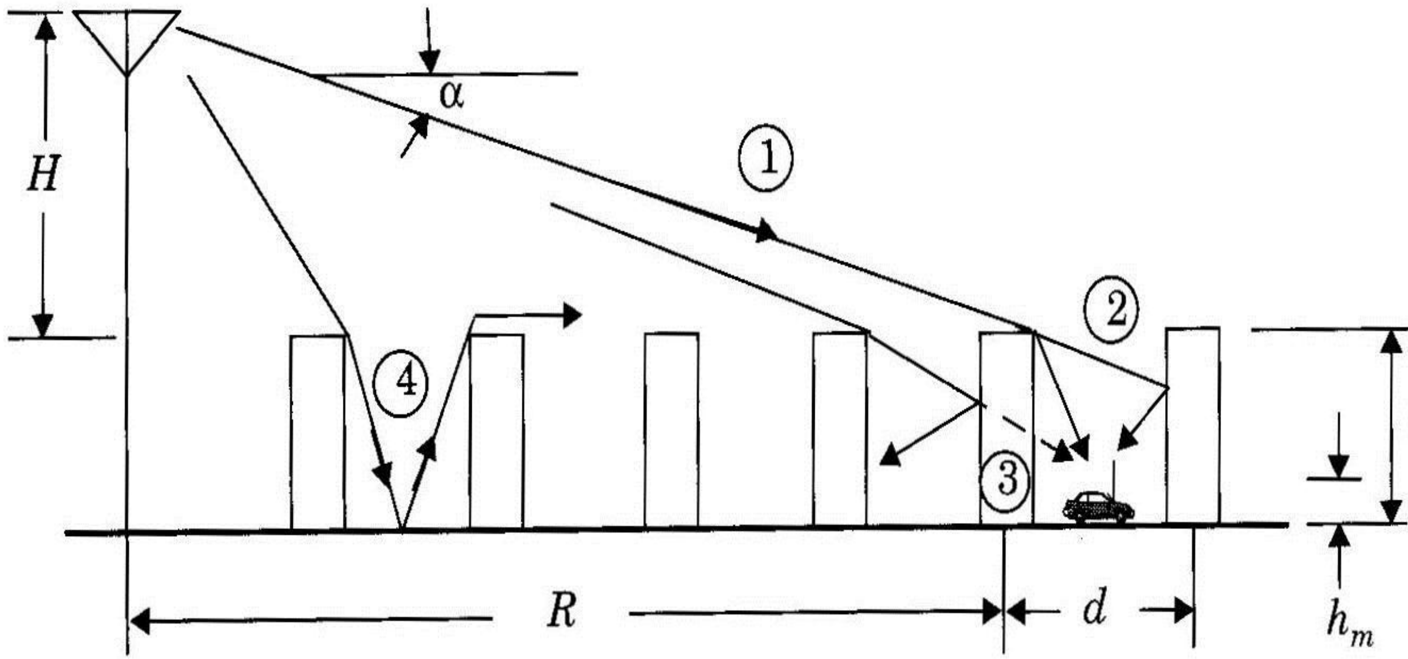


Figure 4.25 Propagation geometry for model proposed by Walfisch and Bertoni [from [Wal88] © IEEE].

4.10.7 Wideband PCS Microcell Model

- Work by Feuerstein et al. in 1991 used a 20MHz pulse transmitter at 1900 MHz to measure path loss, outage, and delay spread in typical microcellular systems in San Francisco and Oakland.
- This work revealed that a two-ray ground reflection model is a good estimate for oath loss in LOS microcells and a simple log-distance path loss model holds well for OBS (obstructed) microcell environments.

- For LOS cases,

$$d_f = \frac{1}{\lambda}\sqrt{(\Sigma^2 - \Delta^2)^2 - 2(\Sigma^2 + \Delta^2)(\frac{\lambda}{2})^2 + (\frac{\lambda}{2})^4}$$

$$= \frac{1}{\lambda}\sqrt{16 h_t^{\ 2} h_r^{\ 2} - \lambda^2 (h_t^{\ 2} + h_r^{\ 2}) + \frac{\lambda^4}{16}}$$

$$\overline{PL}(d) = \begin{cases} 10 n_1 \log(d) + p_1 & \text{for } 1 < d < d_f \\ 10 n_2 \log(d / d_f) + 10 n_1 \log d_f + p_1 & \text{for } \ d > d_f \end{cases}$$

- For the OBS case,

$$\overline{PL}(d)[\mathrm{dB}] = 10n\log(d) + p_1$$

Transmitter Antenna Height	1900 MHz LOS			1900 MHz OBS	
	n_1	n_2	σ(dB)	n	σ(dB)
Low (3.7 m)	2.18	3.29	8.76	2.58	9.31
Medium (8.5 m)	2.17	3.36	7.88	2.56	7.67
High (13.3 m)	2.07	4.16	8.77	2.69	7.94

Figure 4.26 Parameters for the wideband microcell model at 1900 MHz [from [Feu94] © IEEE].

4.11 Indoor Propagation Models

- The indoor radio channel differs from the traditional mobile radio channel in two aspects –

  (1) the distances covered are much smaller,

  (2) the variability of environment is much greater.
- The propagation within buildings is strongly influenced by the layout of the building, the construction materials, and the building type.

- Indoor radio propagation is dominated by the same mechanisms as outdoor : reflection, diffraction, and scattering.

  However, conditions are much more variable. For example, the interior doors are open or closed; antennas positions (desk or ceiling).
- In general, indoor channels may be classified either as LOS or obstructed (OBS) with varying degrees of clutter.

4.11.1 Partition Losses (same floor)

- Houses v.s. office buildings.
- Hard partitions v.s. Soft partitions.
- Data bases of losses for a great number of partitions.

Table 4.3 Average Signal Loss Measurements Reported by Various Researchers for Radio Paths Obstructed by Common Building Material

Material Type	Loss (dB)	Frequency	Reference
All metal	26	815 MHz	[Cox83b]
Aluminum siding	20.4	815 MHz	[Cox83b]
Foil insulation	3.9	815 MHz	[Cox83b]
Concrete block wall	13	1300 MHz	[Rap91c]
Loss from one floor	20-30	1300 MHz	[Rap91c]
Loss from one floor and one wall	40-50	1300 MHz	[Rap91c]
Fade observed when transmitter turned a right angle corner in a corridor	10-15	1300 MHz	[Rap91c]
Light textile inventory	3-5	1300 MHz	[Rap91c]
Chain-like fenced in area 20 ft high containing tools, inventory, and people	5-12	1300 MHz	[Rap91c]
Metal blanket — 12 sq ft	4-7	1300 MHz	[Rap91c]
Metallic hoppers which hold scrap metal for recycling — 10 sq ft	3-6	1300 MHz	[Rap91c]
Small metal pole — 6" diameter	3	1300 MHz	[Rap91c]
Metal pulley system used to hoist metal inventory — 4 sq ft	6	1300 MHz	[Rap91c]
Light machinery < 10 sq ft	1-4	1300 MHz	[Rap91c]
General machinery — 10 - 20 sq ft	5-10	1300 MHz	[Rap91c]
Heavy machinery > 20 sq ft	10-12	1300 MHz	[Rap91c]
Metal catwalk/stairs	5	1300 MHz	[Rap91c]
Light textile	3-5	1300 MHz	[Rap91c]
Heavy textile inventory	8-11	1300 MHz	[Rap91c]
Area where workers inspect metal finished products for defects	3-12	1300 MHz	[Rap91c]
Metallic inventory	4-7	1300 MHz	[Rap91c]
Large 1-beam — 16 - 20"	8-10	1300 MHz	[Rap91c]
Metallic inventory racks — 8 sq ft	4-9	1300 MHz	[Rap91c]
Empty cardboard inventory boxes	3-6	1300 MHz	[Rap91c]
Concrete block wall	13-20	1300 MHz	[Rap91c]
Ceiling duct	1-8	1300 MHz	[Rap91c]
2.5 m storage rack with small metal parts (loosely packed)	4-6	1300 MHz	[Rap91c]
4 m metal box storage	10-12	1300 MHz	[Rap91c]

Table 4.3 Average Signal Loss Measurements Reported by Various Researchers for Radio Paths Obstructed by Common Building Material (Continued)

Material Type	Loss (dB)	Frequency	Reference
5 m storage rack with paper products (loosely packed)	2-4	1300 MHz	[Rap91c]
5 m storage rack with large paper products (tightly packed)	6	1300 MHz	[Rap91c]
5 m storage rack with large metal parts (tightly packed)	20	1300 MHz	[Rap91c]
Typical N/C machine	8-10	1300 MHz	[Rap91c]
Semi-automated assembly line	5-7	1300 MHz	[Rap91c]
0.6 m square reinforced concrete pillar	12-14	1300 MHz	[Rap91c]
Stainless steel piping for cook-cool process	15	1300 MHz	[Rap91c]
Concrete wall	8-15	1300 MHz	[Rap91c]
Concrete floor	10	1300 MHz	[Rap91c]
Commercial absorber	38	9.6 GHz	[Vio88]
Commercial absorber	51	28.8 GHz	[Vio88]
Commercial absorber	59	57.6 GHz	[Vio88]
Sheetrock (3/8 in) — 2 sheets	2	9.6 GHz	[Vio88]
Sheetrock (3/8 in) — 2 sheets	2	28.8 GHz	[Vio88]
Sheetrock (3/8 in) — 2 sheets	5	57.6 GHz	[Vio88]
Dry plywood (3/4 in) — 1 sheet	1	9.6 GHz	[Vio88]
Dry plywood (3/4 in) — 1 sheet	4	28.8 GHz	[Vio88]
Dry plywood (3/4 in) — 1 sheet	8	57.6 GHz	[Vio88]
Dry plywood (3/4 in) — 2 sheets	4	9.6 GHz	[Vio88]
Dry plywood (3/4 in) — 2 sheets	6	28.8 GHz	[Vio88]
Dry plywood (3/4 in) — 2 sheets	14	57.6 GHz	[Vio88]
Wet plywood (3/4 in) — 1 sheet	19	9.6 GHz	[Vio88]
Wet plywood (3/4 in) — 1 sheet	32	28.8 GHz	[Vio88]
Wet plywood (3/4 in) — 1 sheet	59	57.6 GHz	[Vio88]
Wet plywood (3/4 in) — 2 sheets	39	9.6 GHz	[Vio88]
Wet plywood (3/4 in) — 2 sheets	46	28.8 GHz	[Vio88]
Wet plywood (3/4 in) — 2 sheets	57	57.6 GHz	[Vio88]
Aluminum (1/8 in) — 1 sheet	47	9.6 GHz	[Vio88]
Aluminum (1/8 in) — 1 sheet	46	28.8 GHz	[Vio88]
Aluminum (1/8 in) — 1 sheet	53	57.6 GHz	[Vio88]

17]

4.11.2 Partition Losses between Floors

- The losses between floors of a building are determined by the external dimensions, materials of the building, the type of the construction of the floor, and external surroundings.
- Even the number of windows and the presence of tinting.
- FAF – *floor attenuation factors*.

Table 4.4 Total Floor Attenuation Factor and Standard Deviation σ (dB) for Three Buildings. Each Point Represents the Average Path Loss Over a 20λ Measurement Track [Sei92a]

Building	915 MHz FAF (dB)	σ (dB)	Number of locations	1900 MHz FAF (dB)	σ (dB)	Number of locations
Walnut Creek						
One Floor	33.6	3.2	25	31.3	4.6	110
Two Floors	44.0	4.8	39	38.5	4.0	29
SF PacBell						
One Floor	13.2	9.2	16	26.2	10.5	21
Two Floors	18.1	8.0	10	33.4	9.9	21
Three Floors	24.0	5.6	10	35.2	5.9	20
Four Floors	27.0	6.8	10	38.4	3.4	20
Five Floors	27.1	6.3	10	46.4	3.9	17
San Ramon						
One Floor	29.1	5.8	93	35.4	6.4	74
Two Floors	36.6	6.0	81	35.6	5.9	41
Three Floors	39.6	6.0	70	35.2	3.9	27

- The attenuation between one floor of the building is greater than the incremental attenuation caused by each additional floor.
- After about five or six floor separations, very little additional path loss is experienced.

Table 4.5 Average Floor Attenuation Factor in dB for One, Two, Three, and Four Floors in Two Office Buildings [Sei92b]

Building	FAF (dB)	σ (dB)	Number of locations
Office Building 1:			
Through One Floor	12.9	7.0	52
Through Two Floors	18.7	2.8	9
Through Three Floors	24.4	1.7	9
Through Four Floors	27.0	1.5	9
Office Building 2:			
Through One Floor	16.2	2.9	21
Through Two Floors	27.5	5.4	21
Through Three Floors	31.6	7.2	21

4.11.3 Log-distance Path Loss Model

- $$PL(\mathrm{dB}) = PL(d_0) + 10n\log(\frac{d}{d_0}) + X_\sigma$$

  where the value of n depends on the surroundings and building type, and $X_\sigma \sim N(0, \sigma)$.
- This equation is identical in form to the log-mormal shadowing model.

Table 4.6 Path Loss Exponent and Standard Deviation Measured in Different Buildings [And94]

Building	Frequency (MHz)	n	σ (dB)
Retail Stores	914	2.2	8.7
Grocery Store	914	1.8	5.2
Office, hard partition	1500	3.0	7.0
Office, soft partition	900	2.4	9.6
Office, soft partition	1900	2.6	14.1
Factory LOS			
Textile/Chemical	1300	2.0	3.0
Textile/Chemical	4000	2.1	7.0
Paper/Cereals	1300	1.8	6.0
Metalworking	1300	1.6	5.8
Suburban Home			
Indoor Street	900	3.0	7.0
Factory OBS			
Textile/Chemical	4000	2.1	9.7
Metalworking	1300	3.3	6.8

4.11.4 Ericsson Multiple Breakpoint Model

- The Ericsson model provides a deterministic limit on the range of path loss at a particular distance.
- This model was obtained by measurements in a multiple floor office building.

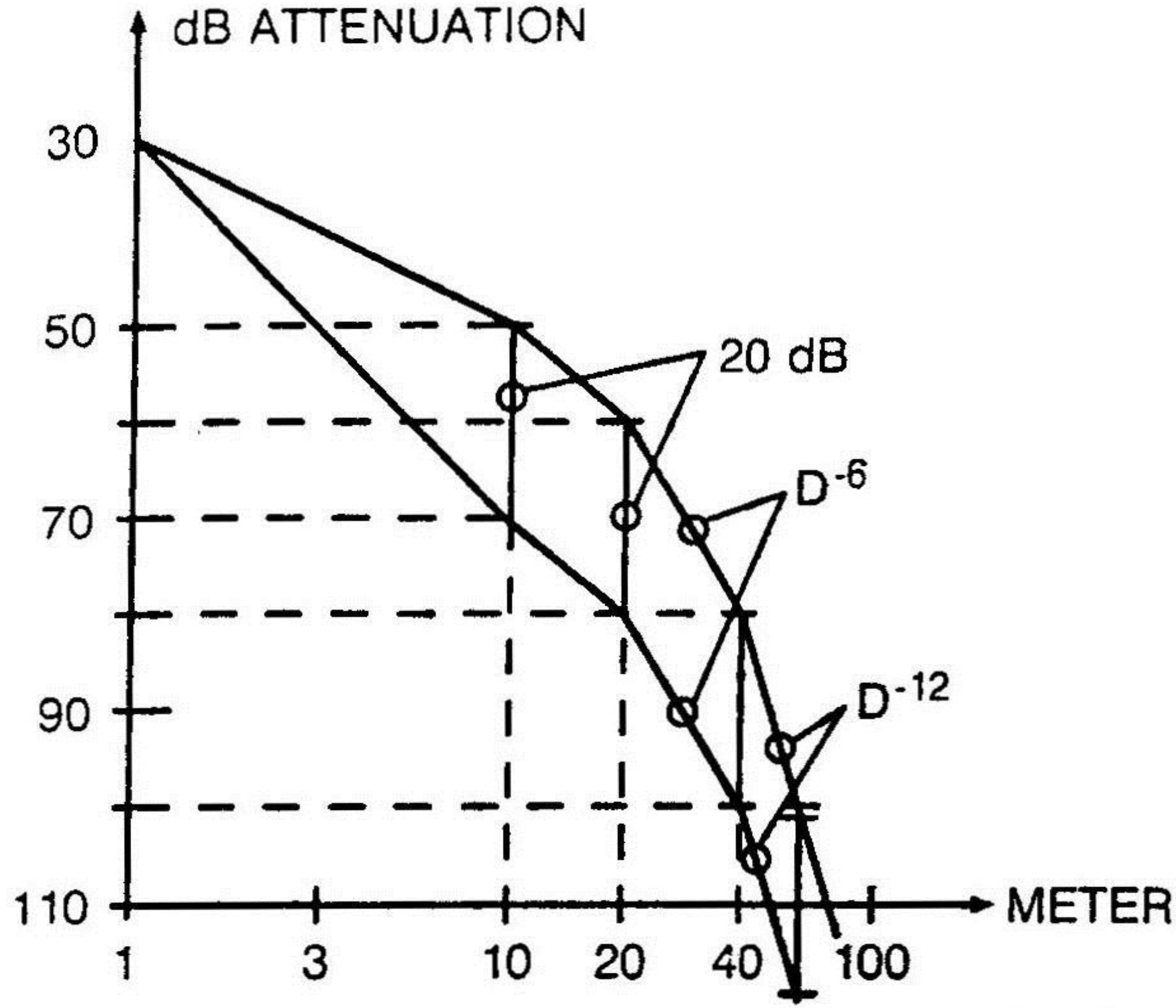


Figure 4.27 Ericsson in-building path loss model [from [Ake88] © IEEE].

4.11.5 Attenuation Factor Model

- This model provides flexibility and was show to reduce X_σ to around 4 dB, as compared to 13 dB when only log-distance model was used.
- The attenuation factor model is given by

$$\overline{PL}(d)[\mathrm{dB}] = \overline{PL}(d_0)[\mathrm{dB}] + 10 n_{SF} \log(\frac{d}{d_0}) + FAF[\mathrm{dB}] + \sum PAF[\mathrm{dB}]$$

n_{SF} : the exponent value for the same floor.

FAF : a floor attenuation.

PAF : the partition attenuation factor in a ray joining t_x and R_x in 3D (the primary ray tracing technique).

- Alternatively, FAF may be replaced by an exponent which already considers the effects of multiple floor separation

$$\overline{PL}(d)[\mathrm{dB}] = \overline{PL}(d_0) + 10 n_{MF} \log(\frac{d}{d_0}) + \sum PAF[\mathrm{dB}]$$

n_{MF} : the multiple floor path loss exponent.

- Typical values of *n*

Table 4.7 Path Loss Exponent and Standard Deviation for Various Types of Buildings [Sei92b]

	n	σ (dB)	Number of locations
All Buildings:			
All locations	3.14	16.3	634
Same Floor	2.76	12.9	501
Through One Floor	4.19	5.1	73
Through Two Floors	5.04	6.5	30
Through Three Floors	5.22	6.7	30
Grocery Store	1.81	5.2	89
Retail Store	2.18	8.7	137
Office Building 1:			
Entire Building	3.54	12.8	320
Same Floor	3.27	11.2	238
West Wing 5th Floor	2.68	8.1	104
Central Wing 5th Floor	4.01	4.3	118
West Wing 4th Floor	3.18	4.4	120
Office Building 2:			
Entire Building	4.33	13.3	100
Same Floor	3.25	5.2	37

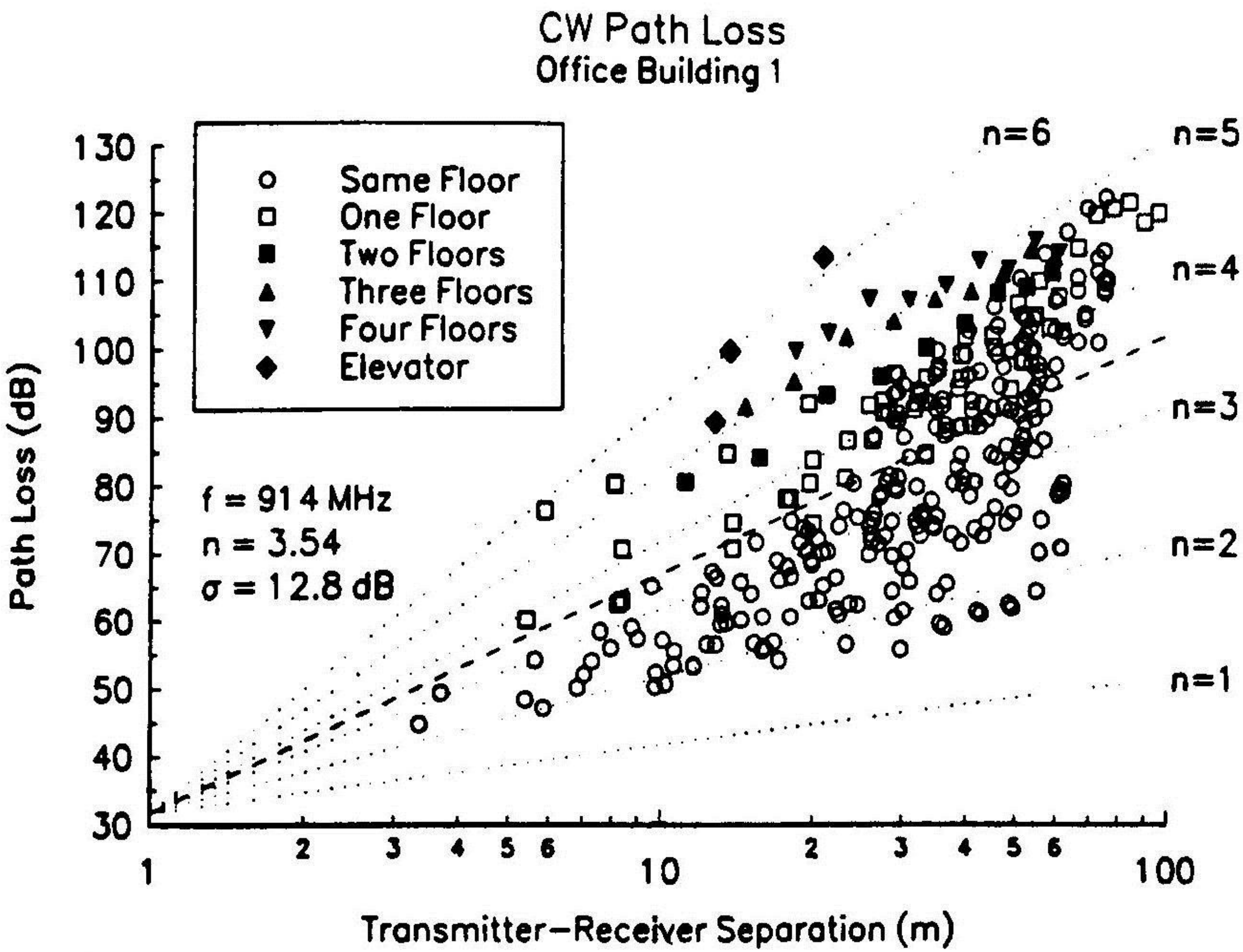


Figure 4.28 Scatter plot of path loss as a function of distance in Office Building 1 [from [Sei92b] © IEEE].

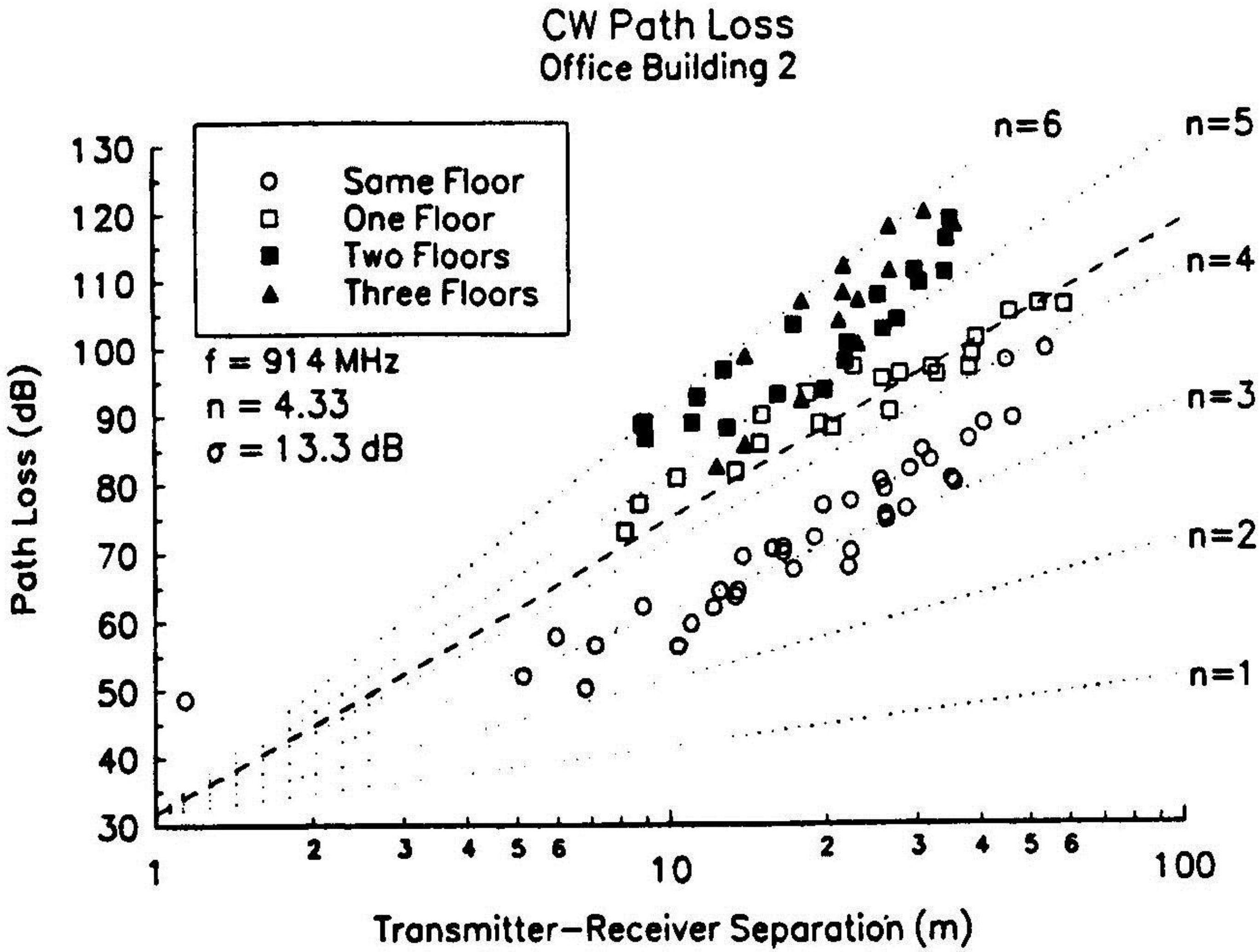


Figure 4.29 Scatter plot of path loss as a function of distance in Office Building 2 [from [Sei92b] © IEEE].

- A modification of Equation (4.94)

$$\overline{PL}(d)[\mathrm{dB}] = \overline{PL}(d_0)[\mathrm{dB}] + 20\log(\frac{d}{d_0}) + \alpha d + FAF[\mathrm{dB}] + \sum PAF[\mathrm{dB}]$$

α : a function of frequency.

Table 4.8 Free Space Plus Linear Path Attenuation Model [Dev90b]

Location	Frequency	α–Attenuation (dB/m)
Building 1: 4 story	850 MHz	0.62
	1.7 GHz	0.57
	4.0 GHz	0.47
Building 2: 2 story	850 MHz	0.48
	1.7 GHz	0.35
	4.0 GHz	0.23

Example 4.11

This example demonstrates how to use Equations (4.94) and (4.95) to predict the mean path loss 30 m from the transmitter, through three floors of Office Building 1 (see Table 4.5). Assume that two concrete block walls are between the transmitter and receiver on the intermediate floors. From Table 4.5 and 4.7, the mean path loss exponent for same-floor measurements in a building is $n = 3.27$, the mean path loss exponent for three-floor measurements is $n = 5.22$, and the average floor attenuation factor is FAF = 24.4 dB for three floors between the transmitter and receiver. Table 4.3 shows a concrete block wall has about 13dB of attenuation. Let $d_0 = 1$m.

Solution:

The mean path loss using Equation(4.94)

$$\overline{PL}(30\text{m})[\text{dB}] = \overline{PL}(1\text{m})[\text{dB}] + 10\times 3.27\times \log(30) + 24.4 + 2\times 13$$

$$= 130.2\text{dB}$$

The mean path loss using Equation(4.95)

$$\overline{PL}(30\text{m})[\text{dB}] = \overline{PL}(1\text{m})[\text{dB}] + 10\times 5.22\times \log(30) + 2\times 13$$

$$= 108.6\text{dB}$$

4.12 Signal Penetration into Buildings

- The signal strength received inside of a building due to an external transmitter is important for wireless systems that share frequencies with neighboring buildings or with outdoor systems.
- Signal strength received inside a building increases with height.

  At lower floors, clutter induces greater attenuation.

  At higher floors, a LOS path may exist.

- RF penetration is also a function of frequency as well as height.

  Measurements in Liverpool : 16.4 dB 11.6dB 7.6dB
  (penetration lose) 441MHz 896.5MHz 1400MHz

  Measurements by Turkmani : 14.2dB 13.4dB 12.8 dB
  900MHz 1800MHz 2300MHz

- Windows v.s. without windows : 6dB less penetration loss.
- Height : decrease at a rate 1.9 dB per floor from the ground level up to 15th floor.

  decrease at a rate 2 dB per floor from the ground level up to 9th floor.

- The percentage of windows, Metallic tints (3dB ~ 30dB attenuation) the angle of incidence of the transmitted wave.

4.13 Ray Tracing and Site Specific Modeling

- Using computer simulation.
- Involve the use of Site Specific (SISP) propagation models and graphical information system (GIS) databases.
- Ray tracing techniques are used with aerial photographs or satellite photographs which convert into usable 3-D databases for reflection, diffraction, and scattering models.
- For example, the SitePlanner® computer-aided design (CAD).
- Some day, these modeling techniques may be downloadable into wireless phones and used to determine instantaneous air interface parameters.