EC8652 WIRELESS COMMUNICATION

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OBJECTIVES:

- To study the characteristic of wireless channel
- To understand the design of a cellular system
- To study the various digital signaling techniques and multipath mitigation techniques
- To understand the concepts of multiple antenna techniques

UNIT I WIRELESS CHANNELS 9

Large scale path loss –Path loss models: Free Space and Two-Ray models -Link Budget design –Small scale fading-Parameters of mobile multipath channels –Time dispersion parameters-Coherence bandwidth –Doppler spread & Coherence time, fading due to Multipath time delay spread –flat fading –frequency selective fading –Fading due to Doppler spread –fast fading –slow fading.

UNIT II CELLULAR ARCHITECTURE 9

Multiple Access techniques -FDMA, TDMA, CDMA –Capacity calculations–Cellular concept-Frequency reuse - channel assignment-hand off-interference & system capacity-trunking & grade of service –Coverage and capacity improvement.

UNIT III DIGITAL SIGNALING FOR FADING CHANNELS 9

Structure of a wireless communication link, Principles of Offset-QPSK, p/4-DQPSK, Minimum Shift Keying, Gaussian Minimum Shift Keying, Error performance in fading channels, OFDM principle –Cyclic prefix, Windowing, PAPR.

UNIT IV MULTIPATH MITIGATION TECHNIQUES 9

Equalisation –Adaptive equalization, Linear and Non-Linear equalization, Zero forcing and LMS Algorithms. Diversity –Micro and Macro diversity, Diversity combining techniques, Error probability in fading channels with diversity reception, Rake receiver.

UNIT V MULTIPLE ANTENNA TECHNIQUES 9

MIMO systems –spatial multiplexing -System model -Pre-coding -Beam forming -transmitter diversity, receiver diversity-Channel state information-capacity in fading and non-fading channels.

TOTAL:45 PERIODS

OUTCOMES:

The student should be able to:

1. Characterize a wireless channel and evolve the system design specifications
2. Design a cellular system based on resource availability and traffic demands
3. Identify suitable signaling and multipath mitigation techniques for the wireless channel and system under consideration.

TEXT BOOKS:

1. Rappaport,T.S., ―Wireless communications‖, Pearson Education, Second Edition, 2010.(UNIT I, II, IV)
2. Andreas.F. Molisch, ―Wireless Communications‖, John Wiley –India, 2006. (UNIT III,V)

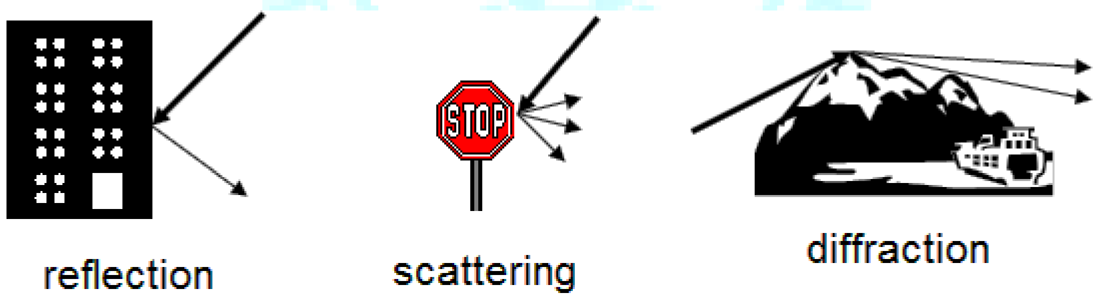
REFERENCES:

1. Wireless Communication –Andrea Goldsmith, Cambridge University Press, 2011
2. Van Nee, R. and Ramji Prasad, ―OFDM for wireless multimedia communications, Artech House, 2000
3. David Tse and Pramod Viswanath, ―Fundamentals of Wireless Communication, Cambridge University Press, 2005.
4. Upena Dalal, ―Wireless Communication, Oxford University Press, 2009.

UNIT I

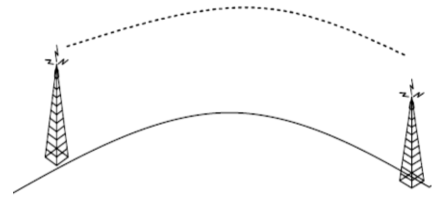
WIRELESS CHANNELS

Large scale path loss – Path loss models: Free Space and Two-Ray models -Link Budget design – Small scale fading- Parameters of mobile multipath channels – Time dispersion parameters-Coherence bandwidth – Doppler spread & Coherence time, Fading due to Multipath time delay spread – flat fading – frequency selective fading – Fading due to Doppler spread – fast fading – slow fading.

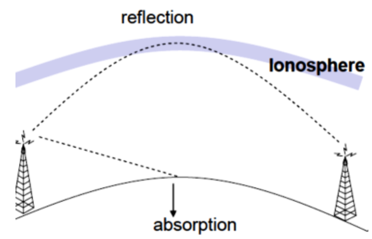


Basics – Propagation

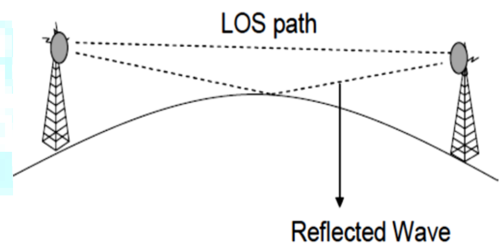
At **VLF, LF, and MF** bands, radio waves follow the ground. AM radio broadcasting uses MF band



At **HF** bands, the ground waves tend to be absorbed by the earth. The waves that reach ionosphere (100-500km above earth surface), are refracted and sent back to earth.



VHF Transmission
Directional antennas are used
Waves follow more direct paths
LOS: Line-of-Sight Communication
Reflected wave interfere with the original signal



Three Basic Radio Propagation Mechanisms

The physical mechanisms that govern radio propagation are complex and diverse, but generally attributed to the following three factors

1. Reflection
2. Diffraction
3. Scattering

Reflection

Occurs when waves impinges upon an obstruction that is much larger in size compared to the wavelength of the signal

Example: reflections from earth and buildings

These reflections may interfere with the original signal constructively or destructively

Diffraction

Occurs when the radio path between sender and receiver is obstructed by an impenetrable body and by a surface with sharp irregularities (edges)

Explains how radio signals can travel urban and rural environments without a line-of-sight path

Scattering

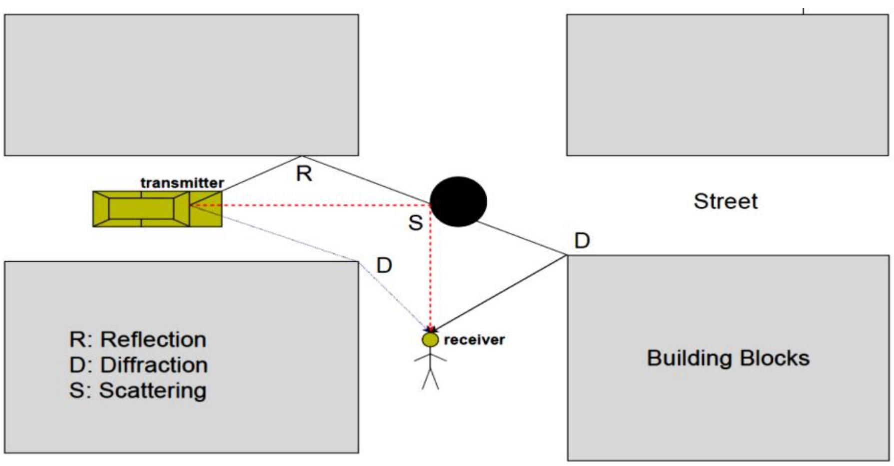
Occurs when the radio channel contains objects whose sizes are on the order of the wavelength or less of the propagating wave and also when the numbers of obstacles are quite large.

They are produced by small objects, rough surfaces and other irregularities on the channel.

Follows same principles with diffraction

Causes the transmitter energy to be radiated in many directions

Lamp posts and street signs may cause scattering



As a mobile move through a coverage area, these 3 mechanisms have an impact on the instantaneous received signal strength.

If a mobile does have a clear line of sight path to the base-station, than diffraction and scattering will not dominate the propagation.

If a mobile is at a street level without LOS, then diffraction and scattering will probably dominate the propagation.

Introduction to Radio Wave Propagation:

The mobile radio channel places fundamental limitations on the performance of wireless communication systems. The transmission path between the transmitter and the receiver can vary from simple line-of-sight to one that is severely obstructed by buildings, mountains, and foliage. Unlike wired channels that are stationary and predictable, radio channels are extremely random and do not offer easy analysis. Even the speed of motion impacts how rapidly the signal level fades as a mobile terminal moves in space. Modeling the radio channel has historically been one of the most difficu1t parts of mobile radio system design, and is typically done in a statistical fashion, based on measurements made specifically for an intended communication system or spectrum allocation.

The mechanisms behind electromagnetic wave propagation are diverse, hut can generally be attributed to reflection, diffraction, and scattering. Most cellular radio systems operate in urban areas where there is no direct line-of-sight path between the transmitter and the receiver, and where the presence of high-rise buildings causes severe diffraction loss. Due to multiple reflections from various objects, the electromagnetic waves travel along different paths of varying lengths. The interaction between these waves causes multipath fading at a specific location, and the strengths of the waves decrease as the distance between the transmitter and receiver increases.

Propagation models have traditionally focused on predicting the average received signal strength at a given distance from the transmitter, as well as the variability of the signal strength in close spatial proximity to a particular location.

Propagation models that predict the mean signal strength for an arbitrary transmitter-receiver (T-R) separation distance are useful in estimating the radio coverage area of a transmitter and are called ***large-scale propagation models,*** since they characterize signal strength over large T-R separation distances (several hundreds or thousands of meters).

On the other hand, propagation models that characterize the rapid fluctuations of the received signal strength over very short travel distances (a few wavelengths) or short time durations (on the order of seconds) are called ***small-scale propagation model*** **or fading models**. As mobile moves over very small distances, the instantaneous received signal strength may fluctuate rapidly giving rise to small-scale fading.

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. Satellite communication systems and microwave line-of-sight radio links undergo free space propagation.

Consider an isotropic source that radiates power P_T (w) equally in all directions. The flux density, on the surface of a sphere of radius *d* centered on the source is given by

$$\Phi_R = \frac{P_T}{4\pi d^2}$$

The power received by an antenna of effective area is given by Friis free space equation

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

The factor $\left(\frac{\lambda}{4\pi d}\right)^2$ is also known as the **free space loss factor**.

P_t : transmitted power

P_d (d) : received power

L : system loss

d : T-R separation distance (m)

G_t: transmitter antenna gain

λ : wave length (m)

G_r: receiver antenna gain

The gain of the antenna

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

A_e: Effective aperture is related to the physical size of the antenna

Antenna Efficiency $\eta = \frac{A_e}{A}$

A = antenna's physical area (cross sectional)

The wave length λ is related to the carrier frequency by

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

f : carrier frequency in Hertz c : speed of light (meters/s)

ω_c: carrier frequency in radians

The losses $(L \geq 1)$ are usually due to transmission line attenuation, filter losses, and antenna losses in the communication system. A value of $L=1$ indicates no loss in the system hardware.

Isotropic radiator is an ideal antenna which radiates power with unit gain.

Effective isotropic radiated power (EIRP) is defined as $EIRP = P_t G_t$ and represents the maximum radiated power available from transmitter in the direction of maximum antenna gain as compared to an isotropic radiator.

Path loss for the free space model with antenna gains $PL(dB) = 10\log\frac{P_t}{P_r} = -10\log\left(\frac{G_t G_r \lambda^2}{(4\pi)^2 d^2}\right)$

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 **path loss**, which represents signal attenuation as a positive quantity measured in dB, is defined as the difference (in dB) between the effective transmitted power and the received power.

The Friis free space model is only a valid predictor for P_r for values of d which is in the far-field (**Fraunhofer region**) of the transmission antenna.

The far-field region of a transmitting antenna is defined as the region beyond the far-field distance

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

where D is the largest physical linear dimension of the antenna.

To be in the far-filed region the following equations must be satisfied

$$d_f >> D \text{ and } \quad d_f >> \lambda$$

Furthermore the following equation does not hold for $d=0$.

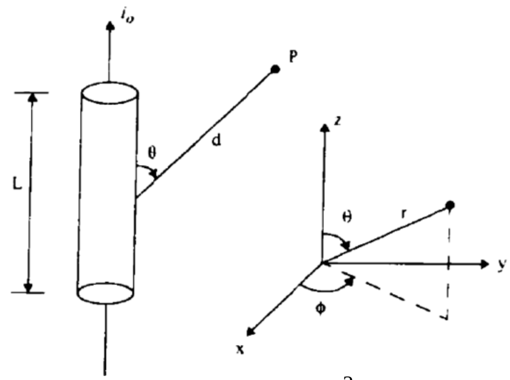
Use close-in distance d_0 and a known received power $P_r(d_0)$ at that point

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

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

$$P_r(d) \text{ dBm} = 10\log\left(\frac{P_r(d_0)}{0.001\,\text{W}}\right) + 20\log\left(\frac{d_0}{d}\right) \qquad d \geq d_0 \geq d_f$$

RELATING POWER TO ELECTRIC FIELD



In free space, the power flux density (expressed in W/m^2) is given by

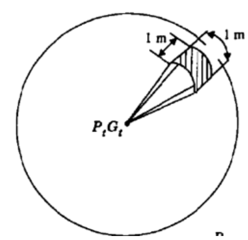
$$\Phi_R = \frac{EIRP}{4\pi d^2} = \frac{P_T G_T}{4\pi d^2} = \frac{E^2}{\eta}\ \mathrm{W/m^2}$$

where η is the intrinsic impedance of free space given by $\eta = 120\pi\ \Omega$. Thus, the power flux density is

$$\Phi_R = \frac{|E|^2}{120\pi}\ \mathrm{W/m^2}$$

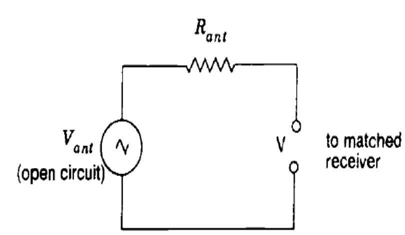
where $|E|$represents the magnitude of the radiating portion of the electric field in the far field.

The power received at distance *d* is given by the power flux density times the effective aperture of the receiver antenna



$$P_R(d) = \Phi_R A_R = \frac{|E|^2}{120\pi} A_R = P_T G_T G_R \left(\frac{\lambda}{4\pi d}\right)^2\ \mathrm{W}$$

Often it is useful to relate the received power level to a receiver input voltage, as well as to an induced E-fleld at the receiver antenna. If the receiver antenna is modeled as a matched resistive load to the receiver, then the receiver antenna will induce an RMS voltage into the receiver which is half of the open circuit voltage at the antenna. Thus, if V is the rms voltage at the input of a receiver (measured by a high impedance voltmeter), and Rant is the resistance of the matched receiver, the received power is given by



$$P_r(d) = \frac{v^2}{R_{ant}} = \left[\frac{v_{ant}/2}{R_{ant}}\right]^2 = \frac{{v_{ant}}^2}{4R_{ant}}$$

$$v_{ant} = v \text{ when there is no load}$$

EXAMPLE

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

Solution

Largest dimension of antenna, D = 1 m

Operating frequency f= 900 MHz, $\lambda = \frac{c}{f} = \frac{3\times10^8}{900\times10^6} = 0.33\text{m}$

Far-field distance is given by

$$d_f = \frac{2D^2}{\lambda} \qquad d_f = \frac{2*1^2}{0.33} = 6m$$

EXAMPLE

If a transmitter produces 50 watts of power, express the transmit power in units of (a) dBm, and (b) dBW. If 50 watts is applied to a unity gain antenna with a 900 MHz 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

Transmit power $P_t = 50$ W

Carrier frequency $f_c = 900$ MHz, $\lambda = \frac{c}{f} = \frac{3\times10^8}{900\times10^6} = 0.33\text{m}$

a) Transmit power $P_t\ (\text{dBm}) = 10\log\left(\frac{P_t}{1\times10^{-3}}\right) = 10\log\left(\frac{50}{1\times10^{-3}}\right) = 47\ \text{dBm}$

b) Transmit power $P_t\ (\text{dB}) = 10\log(P_t) = 10\log(50) = 47\ \text{dBW}$

The received power at a free space distance 100 m from the antenna is given by

$$P_R(d) = P_T G_T G_R \left(\frac{\lambda}{4\pi d}\right)^2$$

$$= 50\times1\times1\left(\frac{0.33}{4\pi\times100}\right)^2$$

$$= 3.5\times10^{-6}\ \text{W}$$

$$P_R(d)\ \text{dBm} = 10\log\left(\frac{P_R(d)}{1\times10^{-3}}\right) = 10\log\left(\frac{3.5\times10^{-6}}{1\times10^{-3}}\right) = -24.5\ \text{dBm}$$

The received power at 10 km is given by

Now d=10km, d_0=100m

$$P_r(d)\ \text{dBm} = 10\log\left(\frac{P_r(d_0)}{0.001\ \text{W}}\right) + 20\log\left(\frac{d_0}{d}\right) \qquad d \ge d_0 \ge d_f$$

$$P_r(10km) = p_r(100) + 20\ log\left(\frac{100}{10000}\right) = -24.5\ dBm - 40\ dBm = -64.5\ dbm$$

EXAMPLE

Assume a receiver is located 10 km from a 50 W transmitter. The carrier frequency is 900 MHz, free space propagation is assumed, $G_T = 1$, and $G_R = 2$, find (a) the power at the 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

Transmit power $P_t = 50$ W

Carrier frequency $f_c = 900$ MHz, $\lambda = \frac{c}{f} = \frac{3\times10^8}{900\times10^6} = 0.33\text{m}$

Transmitter antenna gain $G_T = 1$
Receiver antenna gain $G_R = 2$

a) The power received at a dstance d=10 km is

$$P_R(d) = P_T G_T G_R \left(\frac{\lambda}{4\pi d}\right)^2$$

$$= 50\times1\times2\left(\frac{0.33}{4\pi\times10000}\right)^2 = 7\times10^{-10}\ \text{W}$$

$$P_R(d)\ \text{dBm} = 10\log\left(\frac{P_R(d)}{1\times10^{-3}}\right) = -61.5\ \text{dBm}$$

b) The magnitude of the E-field at the receiver antenna.

$$P_R(d) = \frac{|E|^2}{120\pi}A_R \Rightarrow |E|^2 = \frac{120\pi P_R(d)}{A_R} \Rightarrow |E| = \sqrt{\frac{120\pi P_R(d)}{A_R}} \quad \text{where} \quad A_R = \frac{\lambda^2}{4\pi}G_R$$

The magnitude of the E-field at the receiver antenna is

$$|E| = \sqrt{\frac{120\pi P_R(d)}{\lambda^2 G_R / 4\pi}} = \sqrt{\frac{120\pi\times7\times10^{-10}}{0.33^2\times2/4\pi}} = 0.0039\ \text{V/m}$$

c) The open circuit RMS voltage at the receiver input is

$$P_r(d) = \frac{v^2}{R_{ant}} = \left[\frac{v_{ant}/2}{R_{ant}}\right]^2 = \frac{{v_{ant}}^2}{4R_{ant}}$$

$$v_{ant} = \sqrt{P_r(d) * 4R_{ant}} = \sqrt{7 * 10^{-10} * 4 * 50} = 0.374\ mV$$

BASIC PROPAGATION MECHANISMS

Basic propagation mechanisms

Reflection occurs when a propagating electromagnetic wave impinges upon an object which has very large dimensions when compared to the wavelength of the propagating wave. Reflections occur from the surface of the earth and from buildings and walls.

Diffraction occurs when the radio path between the transmitter and receiver is obstructed by a surface that has sharp irregularities (edges).

Scattering occurs when the medium through which the wave travels consists of objects with dimensions that are small compared to the wavelength, and where the number of obstacles per unit volume is large. Scattered waves are produced by rough surfaces, small objects, or by other irregularities in the channel.

REFLECTION

When a radio wave propagating in one medium impinges upon another medium having different electrical properties, the wave is partially reflected and partially transmitted.
The electric field intensity of the reflected and transmitted waves may be related to the incident wave in the medium of origin through the **Fresnel reflection coefficient** (Γ).
The reflection coefficient is a function of' the material properties, and generally depends on the wave polarization, angle of incidence, and the frequency of the propagating wave.

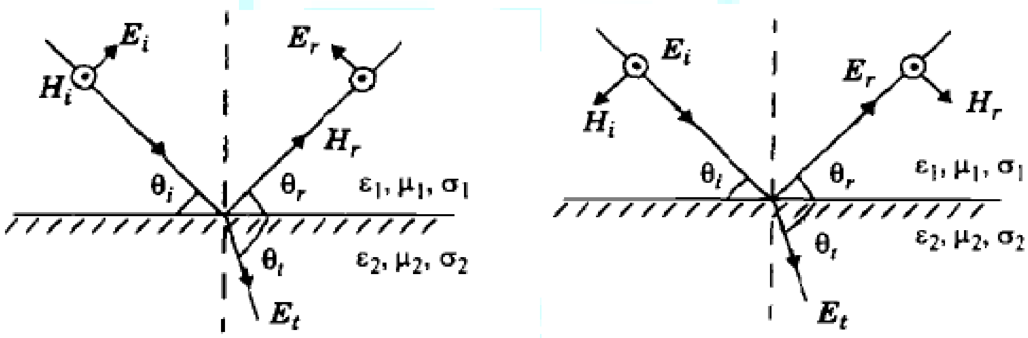
Reflection from Dielectrics

If radio wave is incident on a perfect dielectric

Part of energy is reflected back

Part of energy is transmitted

In addition to the change of direction, the interaction between the wave and boundary causes the energy to be split between reflected and transmitted waves
Figure shows an electromagnetic wave incident at an angle with the plane of the boundary between two dielectric media.



E-field in the plane of incidence E-field normal to the plane of incidence

Part of the energy is reflected back to the first media at an angle θ_r and part of the energy is transmitted (refracted) into the second media at an angle θ_t.

Snell's law

The angle of incidence is the same as the reflected angle:

$$\theta_i = \theta_r$$

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

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

For lossless dielectric material

$$\varepsilon = \varepsilon_0\varepsilon_r$$ `

If a dielectric material is lossy, it will absorb power and may be described by a complex dielectric constant given by

$$\varepsilon = \varepsilon_0\varepsilon_r - \varepsilon'$$

where $\varepsilon' = \dfrac{\sigma}{2\pi f}$

The reflection coefficients for parallel and perpendicular E-field polarization at the boundary of two dielectrics are given by

$$\Gamma_{\parallel} = \frac{E_r}{E_i} = \frac{\eta_2 \sin_{\theta_t} - \eta_1 \sin_{\theta_i}}{\eta_2 \sin_{\theta_t} + \eta_1 \sin_{\theta_i}}$$

$$\Gamma_{\perp} = \frac{E_r}{E_i} = \frac{\eta_2 \sin_{\theta_i} - \eta_1 \sin_{\theta_t}}{\eta_2 \sin_{\theta_i} + \eta_1 \sin_{\theta_t}}$$

Where η_i is the intrinsic impedance of the *i* th medium (i = 1,2), and is given by

$$\eta_i = \sqrt{\frac{\mu_i}{\varepsilon_i}} \qquad \mu\text{=permeability,}\varepsilon\text{=permittivity}$$

The velocity of an electromagnetic wave is given by $\frac{1}{\sqrt{\mu_i \varepsilon_i}}$

When the first medium is free space second medium dielectric then, $\mu_1 = \mu_2$,

The reflection coefficients for the two cases of vertical and horizontal polarization can be simplified to

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

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

Brewster Angle

The **Brewster angle** is the angle at which no reflection occurs in the medium of origin. It occurs when the incident angle θ_i is such that the reflection coefficient is equal to zero. The Brewster angle is given by the value of θ_B which satisfies

$$\sin(\theta_B) = \sqrt{\frac{\varepsilon_1}{\varepsilon_1 + \varepsilon_2}} \qquad \theta_B = \sin^{-1}\sqrt{\frac{\varepsilon_1}{\varepsilon_1 + \varepsilon_2}}$$

When the first medium is free space and the second medium has a relative permittivity ε_r, then

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

Reflection from Perfect Conductors

Electromagnetic energy cannot pass through a perfect conductor. So a plane wave incident on a conductor has all of its energy reflected.

When E-field polarization is in the plane of incidence, the boundary conditions require that

$$\theta_i = \theta_r$$

and $E_r = E_i$ (E-field in the plane of incidence)

When the E-field is horizontally polarized, the boundary conditions require that

$$\theta_i = \theta_r$$

and $E_r = -E_i$ (E-field normal to the plane of incidence)

EXAMPLE

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

Solution

The Brewster angle can be found by using the formula

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

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

Thus the Brewster angle for $\varepsilon_r = 4$ is 26.56°

EXAMPLE

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

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

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

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

Subtitling $\Theta_i = 0^{\Theta}$ $= 1$ $= -1.$

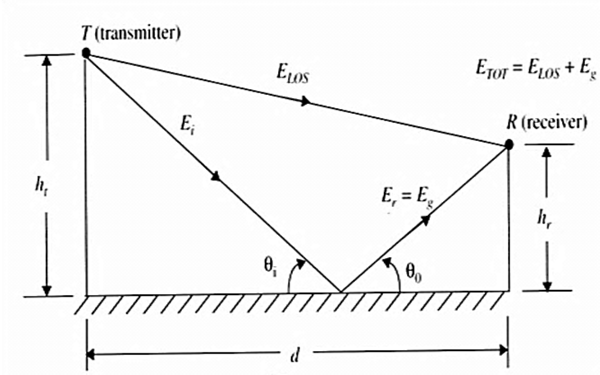
GROUND REFLECTION (TWO-RAY) MODEL

In mobile radio channel, single direct path between base station and mobile and is seldom only physical means for propagation. Free space model as a standalone is inaccurate.

The 2-ray ground reflection model consists of both the direct path and a ground reflected propagation path between transmitter and receiver.

Two ray ground reflection model is useful

- Based on geometric optics
- Considers both direct and ground reflected path
- Reasonably accurate for predicting large scale signal strength over several kms that use tall tower height
- Assumption: The height of Transmitter >50 meters



We assume that the electric field is generated by a continuous wave signal with transmission frequency *f*. at a given point in space,

$$E(t) = \sqrt{2}E_0 \cos(2\pi ft + \theta)$$

ETOT is the electric field that results from a **combination** of a direct line-of-sight path and a ground reflected path

$$\vec{E}_{TOT} = \vec{E}_{LOS} + \vec{E}_g$$

Let E_0 be $|\vec{E}\,|$ at reference point d_0 then

$$\vec{E}(d,t) = \left(\frac{E_0 d_0}{d}\right)\cos\left(w_c(\,t - \frac{d}{c})\right) \qquad d > d_0$$

- $\frac{E_0 d_0}{d}$ is the amplitude of the electric field at distance *d*
- $\omega_c = 2\pi f_c$ where f_c is the carrier frequency of the signal
- Notice at different distances *d* the wave is at a different phase because of the form similar to $E(t) = \sqrt{2}E_0 \cos(2\pi ft + \theta)$

For the direct path let $d = d'$; for the reflected path $d = d''$ then

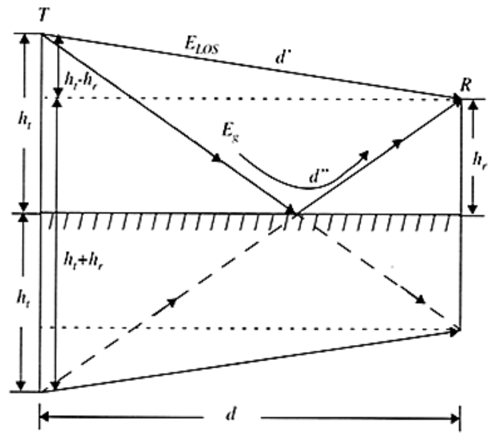
$$E_{LOS}(d',t) = \left(\frac{E_0 d_0}{d}\right)\cos\left(w_c(\,t - \frac{d'}{c})\right)$$

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

$$\vec{E}_{TOT} = \left(\frac{E_0 d_0}{d}\right)\cos\left(w_c(\,t - \frac{d'}{c})\right) + \Gamma\left(\frac{E_0 d_0}{d}\right)\cos\left(w_c(\,t - \frac{d''}{c})\right)$$

for large T−R separation : θ_i goes to 0 (angle of incidence to the ground of the reflected wave) and $\Gamma = -1$

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



The method of images is used to find the path difference between the line of sight and ground reflected paths.
Phase difference can occur depending on the phase difference between direct and reflected E fields
The **phase difference** is θ_Δ due to **Path difference**,
$\Delta = d'' - d'$, between E_{LOS} *and* E_g

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

Using Taylor series the expression can be simplified as:-

$$\Delta = d'' - d' \approx \frac{2h_t h_r}{d}$$

Now as Path Difference is known, Phase difference and time delay can be evaluated as:-

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

$$\tau_d = \frac{\Delta}{c} = \frac{\theta\Delta}{2\pi f_c}$$

If d *is* large than path difference become negligible and amplitude E_{LOS} & E_g are virtually identical and differ only in phase,.i.e

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

For large distances $d = \sqrt{h_t h_r}$ it can be shown that

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

Path loss model for two ray model has d^4 instead of d^2 for free space

$$P_r \propto \frac{1}{d^4} P_r$$

$$P_r\,(d) = P_r\,(d_0)\left(\frac{d_0}{d}\right)^4$$

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

Received power will be calculated in two ray model by

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

P_r is the **plane-Earth propagation equation**. It differs from the free space propagation equation in three ways:

It is frequency independent

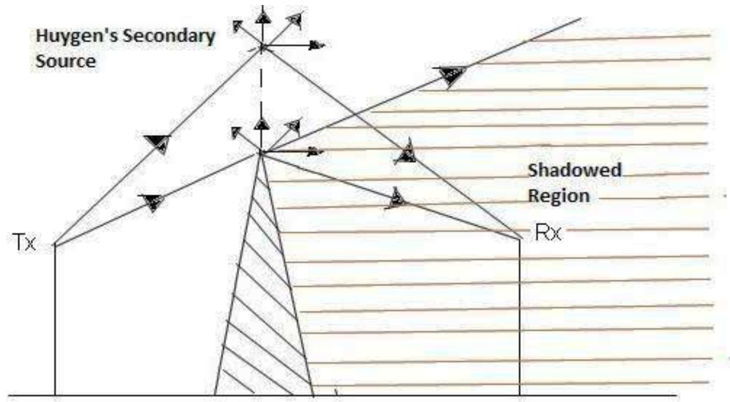
It shows a inverse fourth-power law rather than inverse-square law of the free space propagation

It is dependent on the antenna height

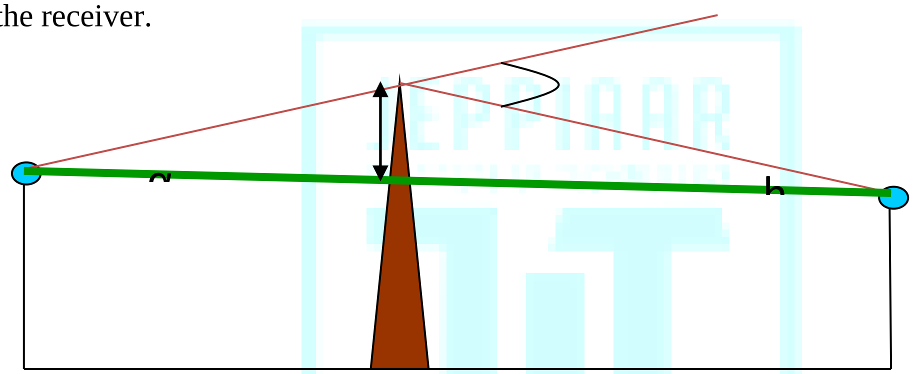
DIFFRACTION

Diffraction allows radio signals to propagate around the curved surface of the earth, beyond the horizon, and to propagate behind obstructions. Diffraction is caused by the propagation of secondary wavelets into a shadowed region.

- $P_r(d) = P_r(d_0)\left(\frac{d_0}{d}\right)^4$



Consider a transmitter and receiver separated in free space as shown in Figure. Let an obstructing screen of effective height h be placed between them at a distance d_1 from the transmitter and d_2 from the receiver.



Diffraction occurs when waves hit the edge of an obstacle "Secondary" wave propagated into the shadowed region Excess path length results in a phase shift

The difference between the direct path and the diffracted path, called the excess path length is

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

$$\alpha = \beta + \gamma$$

The corresponding phase difference is given by

$$\Delta\phi = \frac{2\pi}{\lambda}\Delta R$$

$$\approx \frac{2\pi}{\lambda}\frac{h^2}{2}\left(\frac{d_1+d_2}{d_1 d_2}\right)$$

$$= \frac{\pi}{2}h^2\left(\frac{2(d_1+d_2)}{\lambda d_1 d_2}\right)$$

We define the **Fresnel-Kirchhoff diffraction parameter** as

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

The phase difference can then be written as

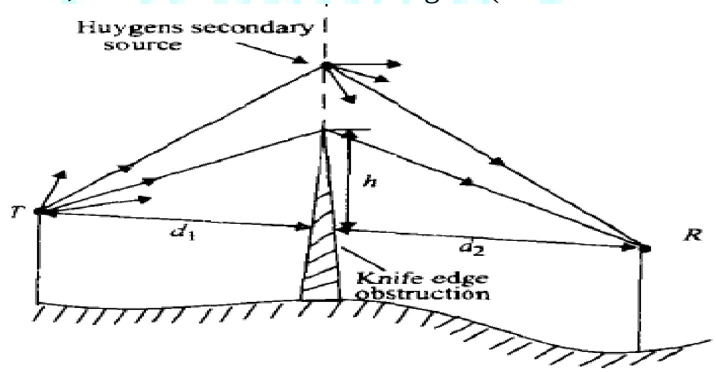
$$\Delta\phi \approx \frac{\pi}{2}v^2$$

Knife-edge Diffraction Model

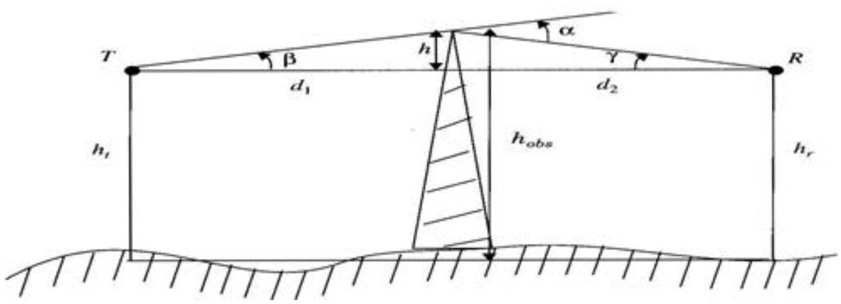
Estimating the signal attenuation caused by diffraction of radio waves over hills and buildings is essential in predicting the field strength in a given service area. Generally, it is impossible to make very precise estimates of the diffraction losses, and in practice prediction is a process of theoretical approximation modified by necessary empirical corrections. Though the calculation of diffraction losses over complex and irregular terrain is a mathematically difficult problem, expressions for diffraction losses for many simple cases have been derived. As a starting point, the limiting case of propagation over a knife-edge gives good insight into the order of magnitude of diffraction loss.

When shadowing is caused by a single object such as a hill or mountain, the attenuation caused by diffraction can be estimated by treating the obstruction as a diffracting knife edge. This is the simplest of diffraction models, and the diffraction loss in this case can be readily estimated using the classical Fresnel solution for the field behind a knife edge

Consider a receiver at point R, located in the shadowed region (also called the diffraction zone).

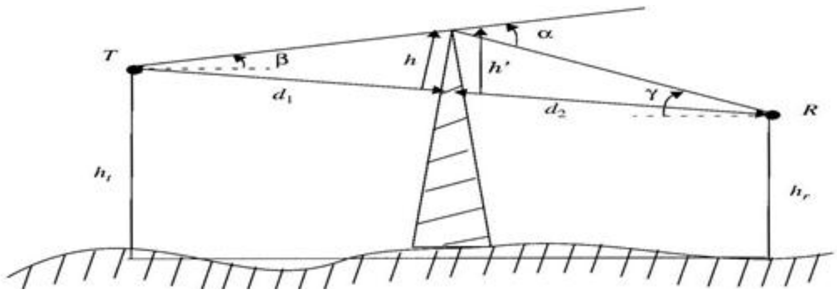


The field strength at point R is a vector sum of the fields due to all of the secondary Huygen's sources in the plane above the knife edge.

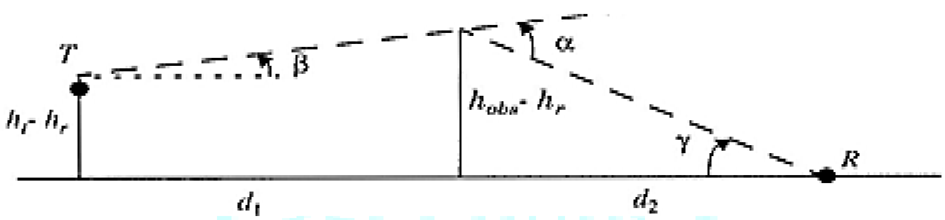


a) Knife Edge Diffraction Geometry.

The point T denotes the transmitter and R denotes the receiver with an infinite knife edge obstruction blocking the line of sight.



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 << d1 and d2, then h and h' are virtually identical and the geometry may be redrawn as



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

From the above diffraction equations it is clear that the phase difference between a direct line of sight path and diffracted path is a function of height and position of the obstruction, as well as the transmitter and receiver location. In practical diffraction problems, it is advantageous to reduce all heights by a constant, so that the geometry is simplified without changing the values of the angles.

The concept of diffraction loss as a function of the path difference around an obstruction is explained by Fresnel zones. Fresnel zones represent successive regions where secondary waves have a path length from the transmitter to receiver which are nλ/2 greater than the total path length of a line of sight path.

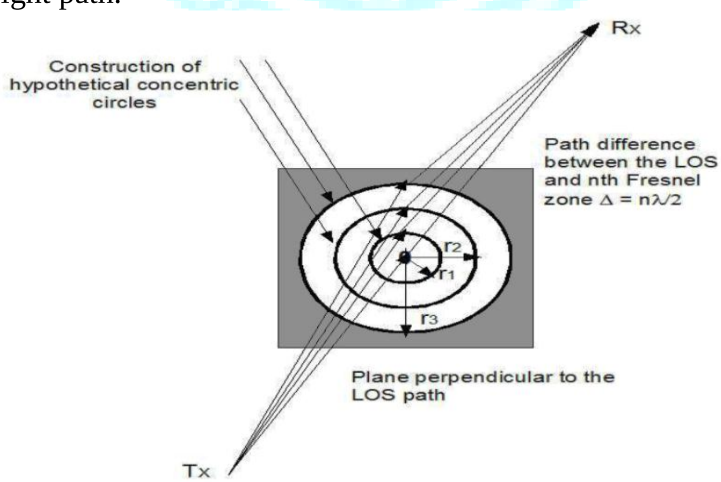


Figure demonstrates a transparent plane located between a transmitter and receiver. The concentric circles on the plane represent the loci of the origins of secondary wavelets which propagate to the receiver such that the total path length increases by for successive circles. These circles are called **Fresnel zones.** The successive Fresnel zones have the effect of alternately providing constructive and destructive interference to the total received signal.

Fresnel zones represent successive regions where secondary waves have a path length from the transmitter to receiver which are $n\lambda/2$ greater than the total path length of a line-of-sight path.

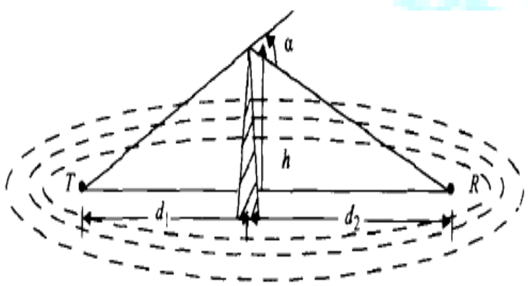
The radius of the n th Fresnel zone circle is denoted by r_n and can be expressed in terms of n, λ, d1, and d2 by

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

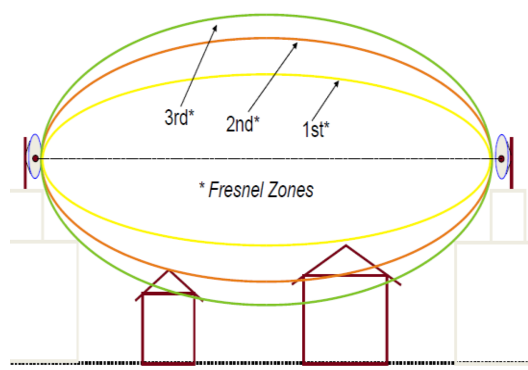
This approximation is valid for d1,d2 >> r_n
The excess total path length traversed by a ray passing through each circle is nλ/2, where n is an integer. Thus, the path traveling through the smallest circle corresponding to n = 1 in Figure will have an excess path lengths of λ/2 as compared to a line-of-sight path, and circles corresponding to n = 2, 3, etc. will have an excess path length of λ, 3λ/2, etc. The radii of the concentric circles depend on the location of the plane.

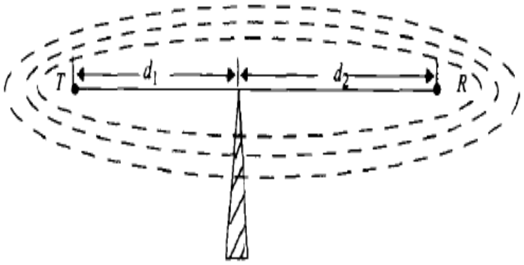
The Fresnel zones of Figure will have maximum radii if the plane is midway between the transmitter and receiver, and the radii become smaller when the plane is moved towards either the transmitter or the receiver. This effect illustrates how shadowing is sensitive to the frequency as well as the location of obstructions with relation to the transmitter or receiver.

In mobile communication systems, diffraction loss occurs from the blockage of secondary waves such that only a portion of the energy is diffracted around an obstacle. That is, an obstruction causes a blockage of energy from some of the Fresnel zones, thus allowing only some of the transmitted energy to reach the receiver. Depending on the geometry of the obstruction, the received energy will be a vector sum of the energy contributions from all unobstructed Fresnel zones.

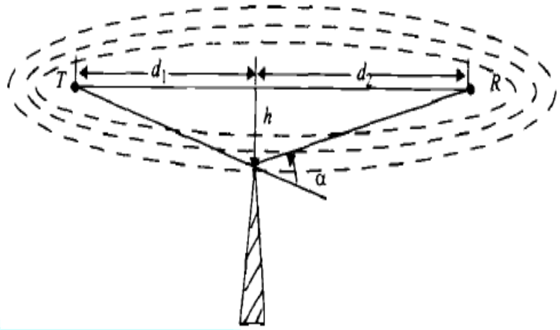


α and v are positive, since h is positive





α and v are equal to zero, since h is equal to zero



α and v are negative, since h is negative

As shown in Figure, an obstacle may block the transmission path, and a family of ellipsoids can be constructed between a transmitter and receiver by joining all the points for which the excess path delay is an integer multiple of half wavelengths. The ellipsoids represent Fresnel zones. Note that the Fresnel zones are elliptical in shape with the transmitter and receiver antenna at their foci.

In general, if an obstruction does not block the volume contained within the first Fresnel zone, then the diffraction loss will be minimal, and diffraction effects may be neglected.

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

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

Approximated by

$$G_d\,(dB) = 0 \qquad for\ v \leq\ -1$$

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

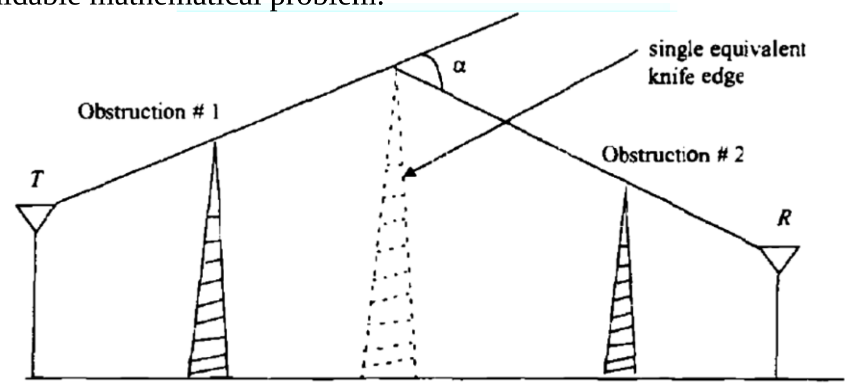
$$G_d\,(dB) = 20\log(0.5\exp(-0.95v)) \qquad for\ 0\ \leq v \leq\ 1$$

Multiple Knife Edge Diffraction

In many practical situations, especially in hilly terrain, the propagation path may consist of more than one obstruction, in which case the total diffraction loss due to all of the obstacles must be computed.

Bullington suggested that the series of obstacles be replaced by a single equivalent obstacle so that the path loss can be obtained using single knife-edge diffraction models. This method, illustrated in Figure, oversimplifies the calculations and often provides very optimistic estimates of the received signal strength.

In a more rigorous treatment, Millington et. al., gave a wave-theory solution for the field behind two knife edges in series. This solution is very useful and can be applied easily for predicting diffraction losses due to two knife edges. However, extending this to more than two knife edges becomes a formidable mathematical problem.



SCATTERING

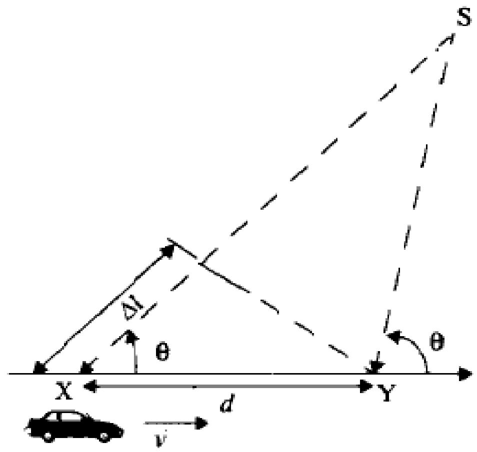
When a radio wave impinges on a rough surface, the reflected energy is spread out (diffused) in all directions due to scattering.
Flat surfaces that have much larger dimension than a wavelength may be modeled as reflective surfaces. However, the roughness of such surfaces often induces propagation effects different from the specular reflection.
Surface roughness is often tested using the Rayleigh criterion which defines a critical height (h_c) of surface protuberances for a given angle of incidence, given by

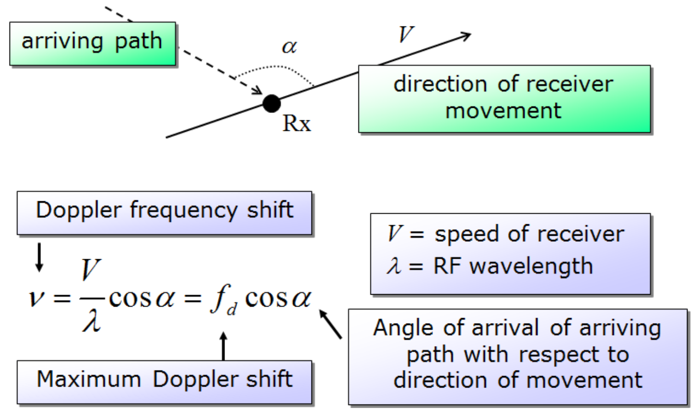
$$h_c = \frac{\lambda}{8 \sin \theta_i}$$

A surface is considered smooth if its minimum to maximum protuberance h is less than h_c, and is considered rough if the protuberance is greater than h_c

DOPPLER SHIFT

Consider a mobile moving at a constant velocity v, along a path segment having length d between points X and Y, while it receives signals from a remote source S, as shown in the Figure
The difference in path lengths traveled by the wave from source S to the mobile at points X and Y is





$$\Delta l = d\cos\theta = v\Delta t\cos\theta$$

The phase change in the received signal due to the difference in path lengths is

$$\Delta\phi = \frac{2\pi\Delta l}{\lambda} = \frac{2\pi v\Delta t\cos\theta}{\lambda} \qquad \text{---------- (10)}$$

The apparent change in frequency, or **Doppler shift** is given by

$$f_d = \frac{1}{2\pi}\frac{\Delta\phi}{\Delta t} = \frac{v}{\lambda}\cos\theta \qquad \text{---------- (11)}$$

If the mobile is moving toward the direction of arrival of the wave, the Doppler shift is positive and if the mobile is moving away from the direction of arrival of the wave, the Doppler shift is negative.

Example

An aircraft is headed towards an airport control tower with a speed of 500 km/h at an elevation of 20°. safety communications between the aircraft tower and the plane occurs at a frequency of approximately 128 MHz. What is the expected Doppler shift of the received signal?

Solution

Carrier frequency $f_c = 128$ MHz

Therefore, wavelength $\lambda = \frac{c}{f_c} = \frac{3\times10^8}{128\times10^6} = 2.34\text{m}$

Aircraft speed v = 500 × 1000/3600 m/s = 138.89 m/s

The Doppler shift of the received signal is

$$f_d = \frac{v}{\lambda}\cos\theta = \frac{138.89}{2.34}\cos 20^\circ = 55.775$$

Example

Consider a transmitter which radiates a sinusoidal carrier frequency of 900 MHz. For a vehicle moving 70 km/h, compute the received carrier frequency if the mobile is moving (a) directly towards the transmitter, (b) directly away from the transmitter, (c) in a direction which is perpendicular to the direction of arrival of the transmitted signal.

Solution

Carrier frequency $f_c = 900$ MHz

Therefore, wavelength $\lambda = \frac{c}{f_c} = \frac{3\times10^8}{900\times10^6} = 0.33\text{m}$

Vehicle speed v = 70 × 1000/3600 = 19.44 m/s

(a) The vehicle is moving directly towards the transmitter. So the Doppler shift is positive and is given by

$$f_d = \frac{v}{\lambda}\cos\theta$$
$$= \frac{19.44}{0.33}\cos 0 = 58.9091$$

The received frequency is given by

$$f = f_c + f_d = 900\times10^6 + 58.9091 = 900.0000589 \text{ MHz}$$

(b) The vehicle is moving directly away from the transmitter. So the Doppler shift in this case is negative and hence the received frequency is given by

$$f = f_c - f_d = 900\times10^6 - 58.9091 = 899.9999411 \text{ MHz}$$

(c) The vehicle is moving perpendicular to the angle of arrival of the transmitted signal. In this case $\theta = 90°$, $\cos 90° = 0$,, and there is no Doppler shift. The received signal frequency is the same as the transmitted frequency of 900 MHz.

Link Budget Design using Path Loss Models

Most radio propagation models are derived using a combination of analytical and empirical methods.

The empirical approach is based on fitting curves or analytical expressions that recreate a set of measured data. This has the advantage of implicitly taking into account all propagation factors, both known and unknown, through actual field measurements.

However, the validity of an empirical model at transmission frequencies or environments other than those used to derive tile model can only be established by additional measured data in the new environment at the required transmission frequency. Over time, some classical propagation models have emerged, which are now used to predict large-scale coverage for mobile communication systems design.

By using path loss models to estimate the received signal level as a function of distance, it becomes possible to predict the SNR for a mobile communication system

Log Distance Path Loss Model

Both theoretical and measurement-based propagation models indicate that average received signal power decreases logarithmically with distance, whether in outdoor or indoor radio channels. Such models have been used extensively in the literature.

The average large-scale path loss for an arbitrary T-R separation is expressed as a function of distance by using a path loss exponent, n.

$$\mathrm{PL}(\mathrm{d}) \propto \left(\frac{\mathrm{d}}{d0}\right)^{n}$$

In dB format: $(\mathrm{PL})\mathrm{dB} = \mathrm{PL}(do) + 10n\log(d/do)$

The 'PL' includes all possible average path losses. Where n is the path loss exponent, which indicates the rate at which the path loss increases with distance, d0 is the close in reference distance which is determined from measurements close to the transmitter, and d is the T-R separation distance.

The bars in equations denote the ensemble average of all possible path loss values for a given value of d. When plotted on a log-log scale, the modeled path loss is a straight line with a slope equal to 10n dB per decade. The value of n depends on the specific propagation environment. For example, in free space, n is equal to 2, and when obstructions are present, n will have a larger value.

It is important to select a free space reference distance that is appropriate for the propagation environment. In large coverage cellular systems, 1 km reference distances are commonly used

Whereas in microcellular systems, much smaller distances (such as 100 m or 1 m) are used.

The reference distance should always be in the far field of the antenna so that near-field effects do not alter the reference path loss. Table lists typical path loss exponents obtained in various mobile radio environments.

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

Log-Normal Shadowing

The log distance model equation does not consider the fact that the surrounding environmental clutter may be vastly different at two different locations having the same T-R separation. This leads to measured signals which are vastly different than the average value predicted by equation.

Measurements have shown that at any value of d, the path loss PL(d) at a particular location is random and distributed log-normally (nonnal in dB) about the mean distance dependent value.

That is

$$[PL(d)]\ dB = PL(d) + X\sigma = PL(do) + 10n\log(d/do) + X\sigma$$

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

where Xσ is Gaussian distributed random variable with zero mean (in dB) and standard deviation σ (dB).

The log-normal distribution describes the random shadowing effects which occur over a large number of measurement locations which have the same T-R separation, but have different levels of clutter on the propagation path. This phenomenon is referred to as log-normal shadowing.

Simply put, log-normal shadowing implies that measured signal levels at a specific T-R separation have a Gaussian (normal) distribution about the distance-dependent mean of log distance equation, where the measured signal levels have values in dB units. The standard deviation of the Gaussian distribution that describes the shadowing also has units in dB.

The close-in reference distance d0, the path loss exponent n, and the standard deviation σ , statistically describe the path loss model for an arbitrary location having a specific T-R separation, and this model may be used in computer simulation to provide received power levels for random locations in communication system design and analysis.

In practice, the values of n and σ are computed from measured data, using linear regression such that the difference between the measured and estimated path losses is minimized in a mean square error sense over a wide range of measurement locations and T-R separations. The value of Pt (d0) in Log Normal Model is based on either close-in measurements or on a free space assumption from the transmitter to d0.

Indoor Propagation Models

The indoor radio channel differs from the traditional radio channel in two aspects:

1. The distances covered are much smaller.
2. The variability of the environment is much greater for a much smaller range of T - R separation distances.

Propagation within building is strongly affected by:

1. The layout of the building.
2. The construction materials.
3. The building type.

Small-Scale Fading and Multipath

- The term **fading** is used to describe rapid fluctuation of the amplitude of a radio signal over a short period of time or travel distance
- **Fading** is caused by destructive interference between two or more versions of the transmitted signal being slightly out of phase due to the different propagation time.
- This is also called multipath propagation
- The different components are due to reflection and scattering form trees buildings and hills etc.
- At a receiver the radio waves generated by same transmitted signal may come
    - From Different direction
    - With Different propagation delays,
    - With Different amplitudes
    - With Different phases
- Each of the factor given above is random
- The multipath components combine vectorially at the receiver and produce a fade or distortion.

Effects of Fading/Multipath

Multipath propagation creates small-scale fading effects. The three most important effects are:

- Rapid changes in signal strength over a small travel distance or time interval;
- Random frequency modulation due to varying Doppler shifts on different multipath signals; and
- Time dispersion (echoes) caused by multipath propagation delays.

Even when a mobile receiver is stationary, the received signal may fade due to a non-stationary nature of the channel (reflecting objects can be moving)

Factors Influencing Small-Scale Fading

Multipath Propagation

- **The presence of reflecting** objects and scatterers in the space between transmitter and receiver creates a constantly changing channel environment
- Causes the signal at receiver to fade or distort

Speed of mobile receiver

- **The relative motion** between the transmitter and receiver results in a random frequency modulation due to different Doppler shifts on each of the multipath signals. Doppler shift may be positive or negative depending on direction of movement of mobile

Speed of Surrounding Objects:

- If the speed of surrounding objects is greater than mobile, the fading is dominated by those objects
- If the surrounding objects are slower than the mobile, then their effect can be ignored

The Transmission Bandwidth:

- Depending on the relation between the signal bandwidth and the coherence bandwidth of the channel, the signal is either distorted or faded
- If the signal bandwidth is greater than coherence bandwidth it creates distortion
- If the signal bandwidth is smaller than coherence bandwidth it create small scale fading
- The coherence bandwidth of a wireless channel is the range of frequencies that are allowed to pass through the channel without distortion. **This is the bandwidth over which the channel transfer function remains virtually constant.**

Impulse Response of Multipath Channel

- The small scale variations of a mobile radio signal can be directly related to the impulse response of mobile radio channel.
- Impulse response contains information to Simulate and Analyze the channel
- The mobile radio channel can be modeled as Linear filter with time varying impulse response
- In case of mobile reception, the length and attenuation of various paths will change with time i.e. Channel is time varying.
- The time variation is strictly due to receiver movement (t=d/v).
- At any distance d=vt, the received signal is the combination of different signals coming with different propagation delays depending on the distance between transmitter and receiver.
- So the impulse response is a function of d, which is the separation between the transmitter and receiver.

PARAMETERS OF MOBILE MULTIPATH CHANNELS

TIME DISPERSION PARAMETERS

Power delay profile: Integrating the scattering function over the Doppler shift gives the **multipath intensity profile**, or **power delay profile** (PDP). PDP gives the average power at the channel output as a function of the time delay.

The PDP can be obtained from the complex impulse responses $h(t, \tau)$ as

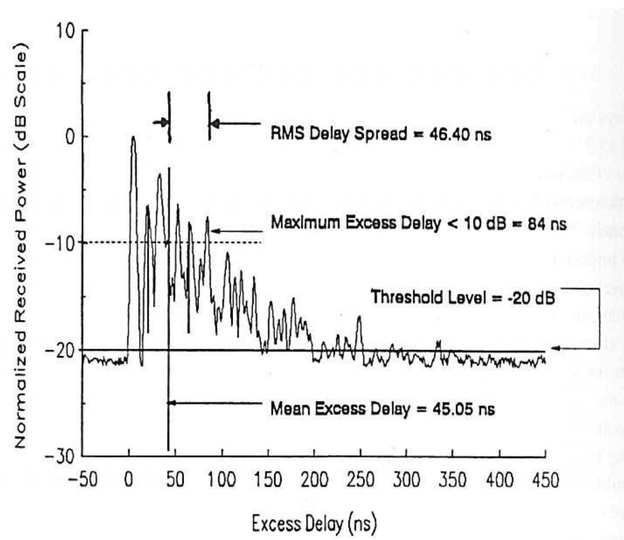
$$P_h(\tau) = \lim_{T\to\infty} \frac{1}{2T} \int_{-T}^{T} |h(t,\tau)|^2 \, dt$$

The multipath channel parameters that can be determined from a power delay profile are

- Mean Delay
- RMS delay spread

The **mean delay** or **mean excess delay** μ_τ is the first moment of the power delay profile and is defined to be

$$\bar{\tau} = \frac{\sum_k a_k^2 \tau_k}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k)(\tau_k)}{\sum_k P(\tau_k)}$$



The **rms delay spread** σ_τ is the square root of the second central moment of the power delay profile and is defined to be

$$\sigma_\tau = \sqrt{\overline{\tau^2} - (\bar{\tau})^2}$$

$$\overline{\tau^2} = \frac{\sum_k a_k^2 \tau_k^2}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k)(\tau_k^2)}{\sum_k P(\tau_k)}$$

Maximum Excess Delay (X dB):
Defined as the time delay value after which the multipath energy falls to X dB below the maximum multipath energy (not necessarily belonging to the first arriving component). It is also called ***excess delay spread*.**

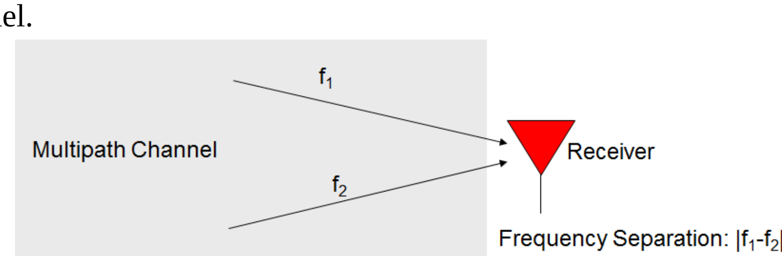
Noise Threshold

- The values of time dispersion parameters also depend on the noise threshold (the level of power below which the signal is considered as noise).

- If noise threshold is set too low, then the noise will be processed as multipath and thus causing the parameters to be higher.

COHERENCE BANDWIDTH

- **Coherence bandwidth** $B_c$is the range of frequencies over which the channel can be considered flat.
- Coherence bandwidth is used to characterize the channel in the frequency domain.
- It is a statistical measure of the range of frequencies over which the channel can be considered flat.
- Two sinusoids with frequency separation greater than B_c are affected quite differently by the channel.



- Frequency correlation between two sinusoids: 0 <=Cr1, r2<=1.
- Coherence bandwidth is the range of frequencies over which two frequency components have a strong potential for amplitude correlation.
  - σ is rms delay spread
  - If correlation is above 0.9, then $B_C = \dfrac{1}{50\sigma}$
  - If correlation is above 0.5, then $B_C = \dfrac{1}{5\sigma}$
    - This is called 50% coherence bandwidth.

DOPPLER SPREAD AND COHERENCE TIME

Doppler spread B_D is a measure of the spectral broadening caused by the time rate of change of the mobile channel. Doppler spread is defined as the range of frequencies over which the received Doppler spectrum is essentially non-zero.

- Delay spread and Coherence bandwidth describe the time dispersive nature of the channel in a local area.
- They don't offer information about the time varying nature of the channel caused by relative motion of transmitter and receiver.
- It is measure of spectral broadening caused by motion, the time rate of change of the mobile radio channel, and is defined as the range of frequencies over which the received Doppler spectrum is essentially non-zero
- We know how to compute Doppler shift: f_d
- Doppler spectrum have components in the range of f_c-f_d to f_c+f_d
- Doppler spread, B_D, is defined as the maximum Doppler shift: f_m = v/l
- If the baseband signal bandwidth is much less than B_D then effect of Doppler spread is negligible at the receiver. This is slow fading channel characteristics.

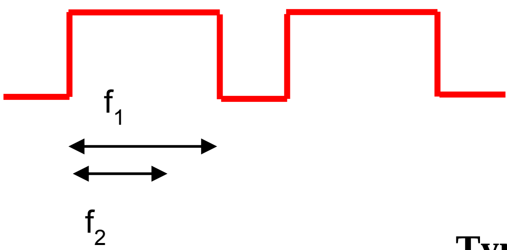
Coherence time:

Coherence time T_c is the time duration over which the channel impulse response is essentially invariant. Coherence time is the time duration over which two received signals have a strong potential for amplitude correlation.

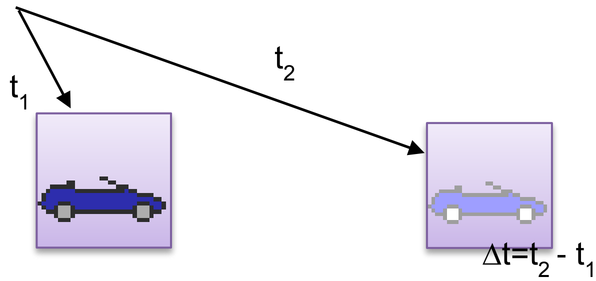
- Coherence time is also defined as: $T_C \approx \sqrt{\dfrac{9}{16\pi f_m^2}} = \dfrac{0.423}{f_m}$

  The time domain dual of B_D
- Coherence time definition implies that two signals arriving with a time separation greater than Tc are affected differently by the channel.
- If the coherence time is defined as the time over which the time correlation function is above 0.5, then the coherence time is approximately, $T_c \approx \dfrac{9}{16\pi f_m}$ where $f_m = \dfrac{v}{\lambda}$

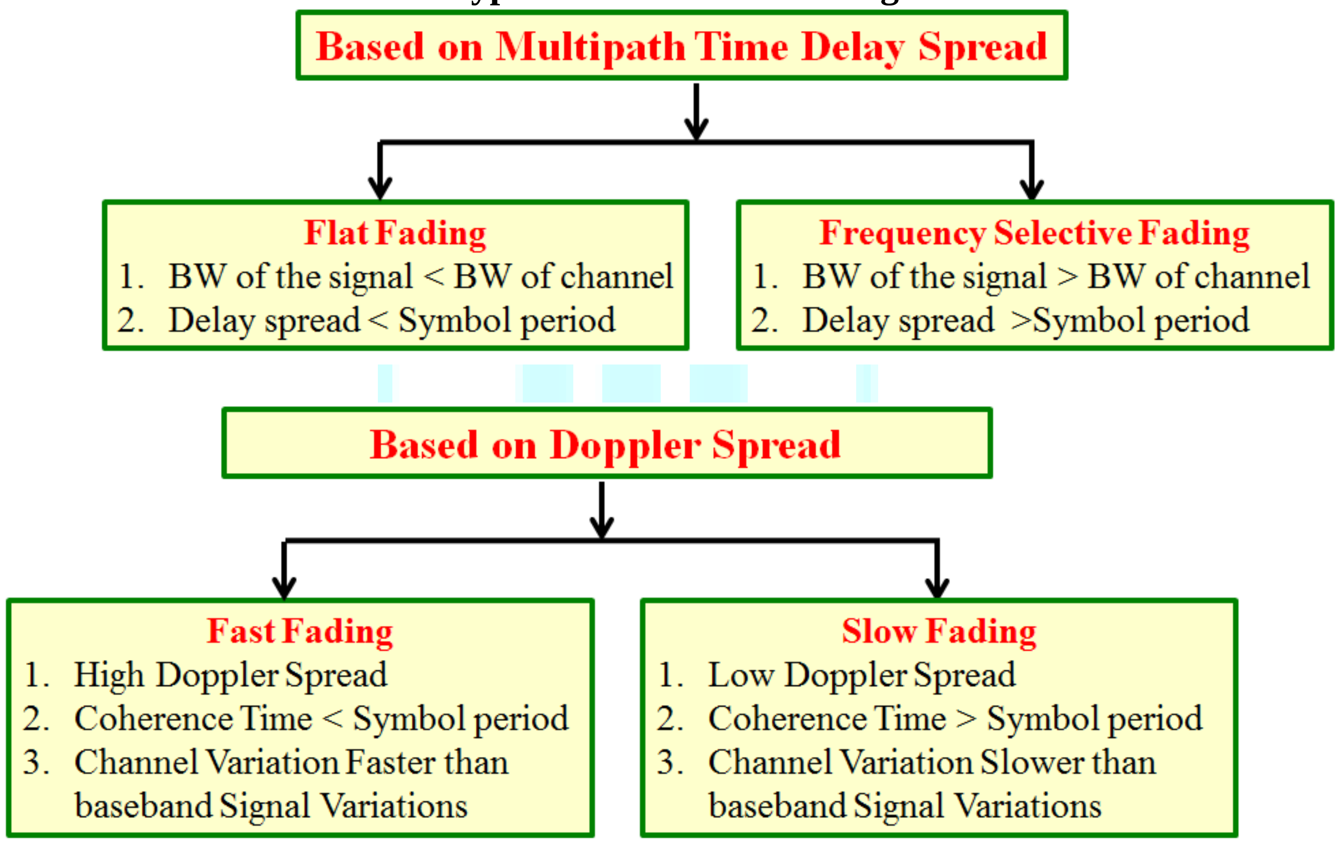
- If the symbol period of the baseband signal (reciprocal of the baseband signal bandwidth) is greater the coherence time, than the signal will distort, since channel will change during the transmission of the signal .
- Coherence time (T_C) is defined as:



$$T_C \approx \frac{1}{f_m}$$



Types of Small-Scale Fading



Depending on the relation between signal parameters (bandwidth and symbol period) and channel parameters (delay spread and Doppler spread) different signals undergo different types of fading

Based on delay spread the types of small scale fading are

Flat fading

Frequency selective fading

Based on Doppler spread the types of small scale fading are

Fast fading

Slow fading

FLAT FADING CHANNEL

- Occurs when the amplitude of the received signal changes with time

- Occurs when symbol period of the transmitted signal is much larger than the Delay Spread of the channel
- Bandwidth of the applied signal is narrow.
- The channel has a flat transfer function with almost linear phase, thus affecting all spectral components of the signal in the same way
- May cause deep fades.
- Increase the transmit power to combat this situation.

In a flat fading channel,

1. The multipath time delay spread of the channel is smaller than the signal duration of the transmitted signal

$$T_S >> \sigma_\tau$$

T_S : Symbol period
σ_τ : Delay Spread

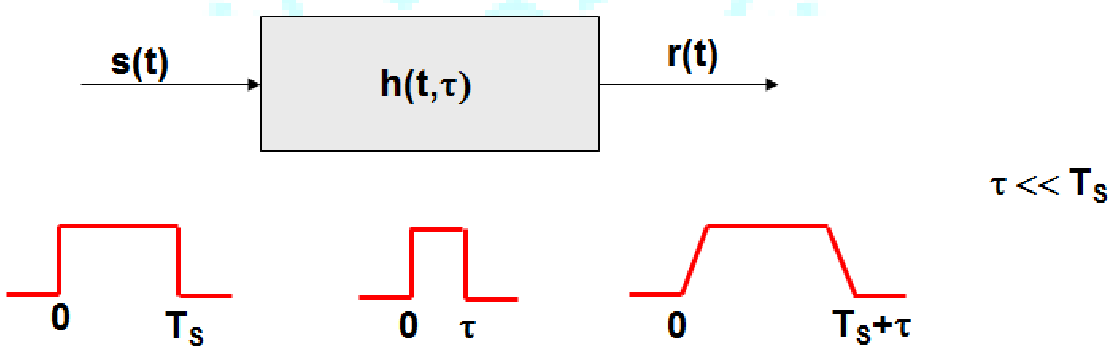
2. Bandwidth of the applied signal is smaller than the coherence bandwidth of the channel

$$B_S << B_C$$

B_C: Coherence bandwidth
B_S: Signal bandwidth

Flat fading channels are also known as **amplitude varying channels** or **narrowband channels**.

The characteristics of a flat fading channel are illustrated below.



If the mobile radio channel has a constant gain and linear phase response over a bandwidth which is greater than the bandwidth of the transmitted signal, then the received signal will undergo **flat fading**.

FREQUENCY SELECTIVE FADING CHANNEL

- A channel that is not a flat fading channel is called *frequency selective fading* because different frequencies within a signal are attenuated differently by the MRC.
- Occurs when channel multipath delay spread is greater than the symbol period.

  Symbols face time dispersion

Channel induces Intersymbol Interference (ISI)

- Bandwidth of the signal s(t) is wider than the channel impulse response

In a Frequency selective fading channel,

1. The multipath time delay spread of the channel is greater than the signal duration of the transmitted signal

$$T_S < \sigma_\tau$$

T_S : Symbol period & σ_τ : Delay Spread

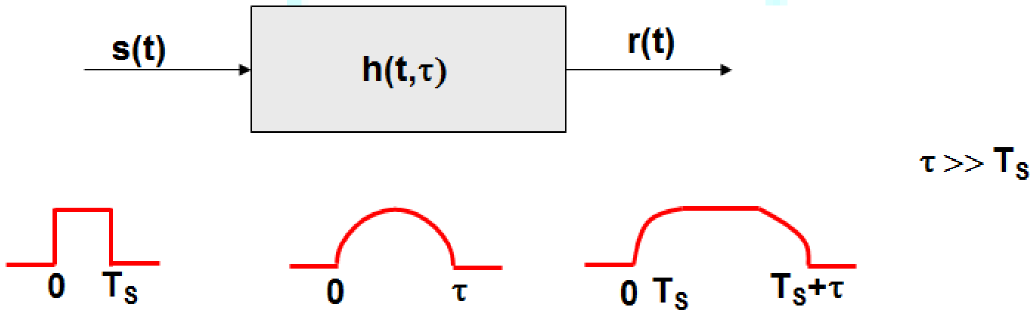
2. Bandwidth of the applied signal is smaller than the coherence bandwidth of the channel

$$B_S > B_C$$

B_C: Coherence bandwidth & B_S: Signal bandwidth

Frequency selective fading channels are also known as **wideband channels**.

The characteristics of a frequency selective fading channel are illustrated below.



If the channel possesses a constant-gain and linear phase response over a bandwidth that is smaller than the bandwidth of transmitted signal, then the channel creates **frequency selective fading** on the received signal.

Frequency selective fading is due to time dispersion of the transmitted symbols within the channel. Thus the channel induces **Intersymbol Interference** (ISI).

FAST FADING CHANNEL

- Fast fading is the rapid variation of the signal levels when the user device moves a short distances.
- Fast fading is due to reflections of local objects and the motion of the user device relative to those objects.
- The received signal is the sum of number of signals reflected from the local surfaces. These signals sum in a constructive or destructive manner, depending on their relative phase relationships.

In a fast fading channel,

1. The channel impulse response changes rapidly within the symbol duration. That is, the coherence time of the channel is smaller than the symbol period of the transmitted

signal. This causes frequency dispersion due to Doppler spreading, which leads to signal distortion.

$$\mathbf{T_S > T_C}$$

TS: Symbol Period & TC: Coherence Bandwidth

2. Doppler spread is greater than the bandwidth of the transmitted signal

$$\mathbf{B_S < B_D}$$

BS: Bandwidth of the signal & BD: Doppler Spread

SLOW FADING CHANNEL

- Most of the large reflectors and diffracting objects along the transmission path are distant from the user device.
- The motion of the user devices relative to these distant objects is small. So the propagation changes are slow. These factors contribute to the mean path losses between a fixed transmitter and a fixed receiver.
- The variation of these mean losses was modeled as lognormal distribution. The slow-fading process is also referred to as **shadowing** or **lognormal fading**.

In a slow fading channel,

1. The channel impulse response changes at a rate much slower than the transmitted baseband signal

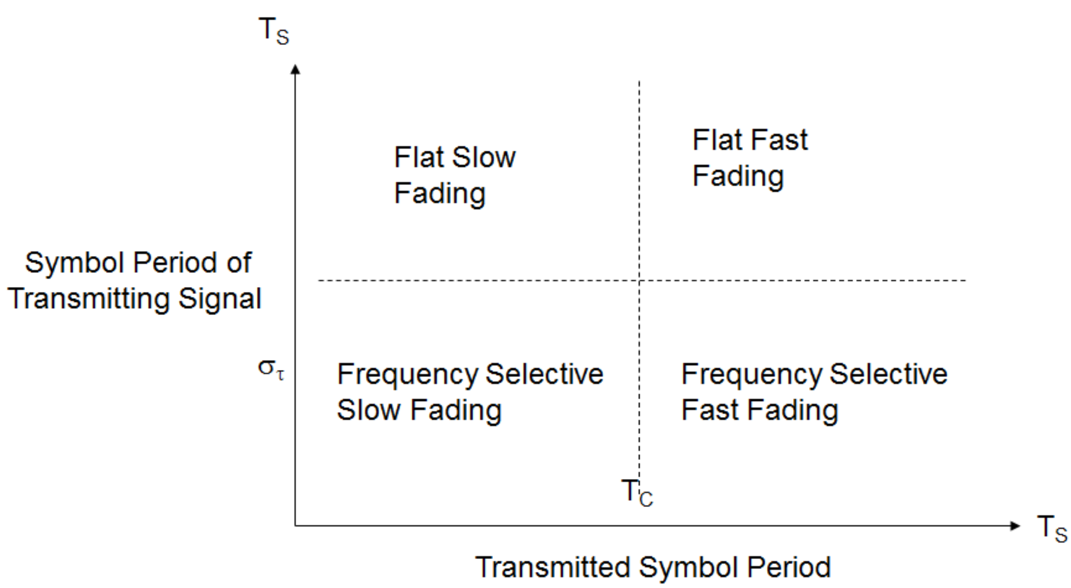
$$\mathbf{T_S \ll T_C}$$

2. The Doppler spread of the channel is much less than the bandwidth of the baseband signal

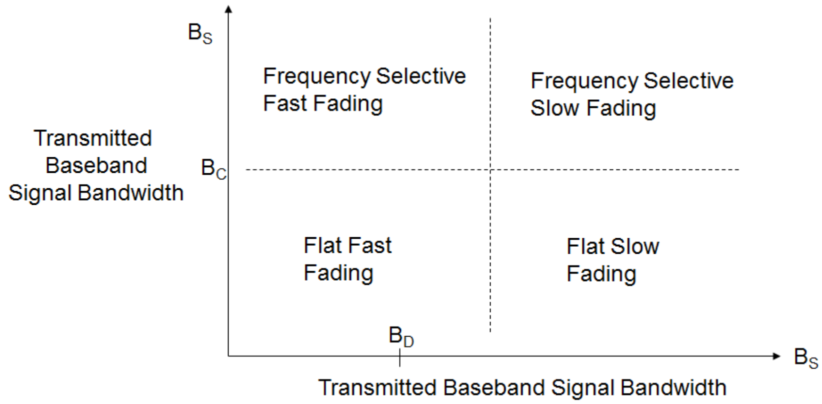
$$\mathbf{B_S \gg B_D}$$

Different Types of Fading

With Respect To SYMBOL PERIOD



With Respect To BASEBAND SIGNAL BANDWIDTH



RAYLEIGH FADING

When there is no line-of-sight component, the distribution of the amplitude fluctuations of a radio signal caused by small-scale fading has a Rayleigh pdf and hence it is known as **Rayleigh fading**.

The Rayleigh fading model assumes that all paths are relatively equal – that is, there is no dominant path. Consider that the transmitted signal reaches a stationary receiver via multiple paths, but the path differences are only due to local reflections.

RICIAN FADING

When a radio wave signal is made of multiple reflective rays and a non-faded line-of-sight component, then the fluctuations of the amplitude of the signal due to small-scale fading follows Rician probability density function. Hence it is known as **Rician fading**.
Occasionally there is a direct line-of-sight path in mobile radio channels and in indoor wireless channels.

Summary of channel classification

Time-flat channels are time-invariant channels. An example is a transmitter and receiver that are both physically stationary, with the propagation environment unchanging.
Frequency-flat fading channels have a frequency response that is approximately flat over a bandwidth greater than the bandwidth of the transmitted signal.
Time-selective channels are time-varying channels. An example is a wireless device moving through the environment and undergoing Rayleigh fading.
Frequency-selective channels have a frequency response that can not be assumed flat over a bandwidth of the transmitted signal. Frequency selective is due to multipath that has delay spread greater than the symbol period.
Wide-sense stationary uncorrelated scattering channel *is* combination of wide-sense stationary and uncorrelated scattering. WSSUS channels display uncorrelated scattering in both the time-delay and Doppler shift.