Velammal College of Engineering & Technology
Department of Information Technology

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UNIT 1: Wireless Channels

Wireless channels places fundamental limitations on the performance of wireless communication systems. The transmission path between the transmitter and the receiver would be either simple Line Of Sight (LOS) or severely obstructed by buildings, mountains and foliage. Modeling these kind of channels has historically been of the difficult in system design and is typically done in a statistical fashion based on the measurements made specifically for an intended communication system or spectrum allocation.

Wave Propagation and propagation channel models:

Usually, propagation of wave in a wireless channel is generally attributed to reflection, diffraction and scattering. Most cellular radio systems operate in urban areas where there is no direct line of sight path between transmitter and receiver, but there is a presence of high rise buildings which 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 strength of the waves decrease as the distance between the transmitter and receiver increases.

Propagation models is statistically analysed to predict the average received signal strength at a given distance from the transmitter as well as the variability of the signal strength to a particular strength. Propagation models that predict the mean signal strength for a transmitter – receiver separation distance are useful in estimating the coverage area of a transmitter and are called as **large scale propagation models**. On the other hand, propagation models that characterize the rapid fluctuations of the received signal strength over very short travel distances or short time durations are called as **small scale or fading models.**

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 LOS path between them. Satellite communication systems and microwave LOS radio links typically undergo free space propagation. This model predicts that received power decays as a function of T-R separation distance raised to some power. The free space power received by a receiver antenna which is separated from a radiating transmitter antenna by a distance d, is given by Friss free space equation

$$P_r(d) = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2 L} \quad \text{---------------------------- (1)}$$

Where
P_t is the transmitted power,
$P_r(d)$ is the received power which is a function of T-R separation,
G_t is the transmitter antenna gain,
G_r is the receiver antenna gain,
d is the T-R separation distance in meters,
L is the system loss factor not related to propagation
λ is the wavelength in meters

The gain of the antenna is given as

$$G = \frac{4\pi A_e}{\lambda^2} \quad \text{-- (2)}$$

Where
A_e is effective aperture of an antenna and is related to physical size of the antenna,
λ is wavelength and is given as $\lambda = c/f = (2\pi c)/(\omega_c)$

where
f is carrier frequency in Hertz,
ω_c is carrier frequency in radians per second
c is speed of light in meters/s
P_t and P_r must be expressed in the same units
G_t and G_r are dimensionless quantities
L is a loss due to transmission line attenuation, filter losses, and antenna losses in the communication system. If L=1, it indicates no loss in the system hardware.

Eqn (1) shows that the received power falls off as the square of the T-R separation distance.

An isotropic radiator is an ideal antenna which radiates power with unit gain uniformly in all directions, and is often used to reference antenna gains in wireless systems. The Effective Isotropic Radiated Power (EIRP) is defined as maximum radiated power available from a transmitter in direction of maximum antenna gain and it is given as,

$$EIRP = P_t G_t \quad \text{-------------------------------- (3)}$$

In practice, antenna gains is expressed in dBi or dBd.

The path loss is defined as the difference between the effective transmitted power and the received power, and may or may not include the effect of the antenna gains. The path loss for the free space model when antenna gains are included is given by,

$$PL(dB) = 10 \log \frac{P_t}{P_r} \text{ ------------------------ } (4)$$

$$= -10 \log \frac{G_t G_r \lambda^2}{(4\pi)^2 \, d^2} \text{ ----------------- } (5)$$

When antenna gains are excluded, the antennas are assumed to have unity gain and path loss is given by,

$$PL(dB) = -10 \log \frac{\lambda^2}{(4\pi)^2 \, d^2} \text{ ------------------ } (6)$$

This free space model is applicable for P_r for values of d which are in the far-field of the transmitting antenna. The far field or fraunhofer region, of a transmitting antenna is defined as the region beyond the far-field distance d_f, which is related to the largest linear dimension of the transmitter antenna aperture and the carrier wavelength. The fraunhofer distance is given by,

$$d_f = \frac{2D^2}{\lambda} \text{ ------------------------------------ } (7)$$

where
D is the largest physical dimension of the antenna

Additionally, to be in the far field region, d_f must satisfy

$$d_f \gg D \text{ -------------------------------------- } (8)$$

and

$$d_f \gg \lambda \text{ -------------------------------------- } (9)$$

furthermore, it is clear that Eqn (1) does not hold for d=0 . For this reason, large scale propagation models use a close in distance, d_0, as a known received power reference point. The received power, $P_r(d)$, at any distance $d>d_0$ may be related to P_r at d_o. The value $P_r(d_0)$ may be predicted from eqn (1) or may be measured in the radio environment by taking the average received power at many points located at a close in radial distance d_0 from the transmitter. The reference distance must be chosen such that it lies in the far field region, that is $d_0 \geq d_f$, and d_0 is chosen to be smaller than any practical distance used in the mobile communication system. Thus, using eqn (1), the received power in free space at a distance greater than d_0 is given by,

$$P_r(d) = P_r(d_0)\left(\frac{d_0}{d}\right)^2 \quad , \quad d \geq d_0 \geq d_f \text{ --------------------- } (10)$$

In mobile radio systems, it is uncommon to find that P_r may change by many orders of magnitude over a typical coverage area of several square kilometres. Because of the large dynamic range of received power levels, often dBm or dBW units are used to express received power levels. Eqn (10) may be expressed in units of dBm or dBW units by simply taking the logarithm of both sides and multiplying by 10. For example, if P_r is in units of dBm , the received power is given by,

$$P_r(d) \text{ dBm} = 10 \log\left[\frac{P_r(d_0)}{0.001\,W}\right] + 20 \log\left(\frac{d_0}{d}\right) , \quad d \geq d_0 \geq d_f \text{ ------}(11)$$

Where
$P_r(d_0)$ is in units of watts

The reference distance d_0 for practical systems using low gain antenna in the 1-2 GHz region is typically chosen to be 1 m in indoor environments and 100m or 1km in outdoor environments, so that the numerator in eqn (10) and (11) is a multiple of 10. This makes path loss computations easy in dB units.

Two-Ray or Ground reflection model:

The two ray ground reflection model is a useful propagation model that is based on geometric optics, and considers both direct path and ground reflected path between transmitter and receiver. This model is found to be reasonably accurate for predicting the large scale signal strength over distances of several kilometres for mobile radio systems that use tall tower systems as well as LOS microcell channels in urban environments.

In most mobile communication systems, the maximum T-R separation distance is atmost only a few tons of kilometres, and the earth may be assumed to be flat. The total received E-field , E_{tot} is then a result of the direct line of sight component, E_{LOS} and the ground reflected component E_g .

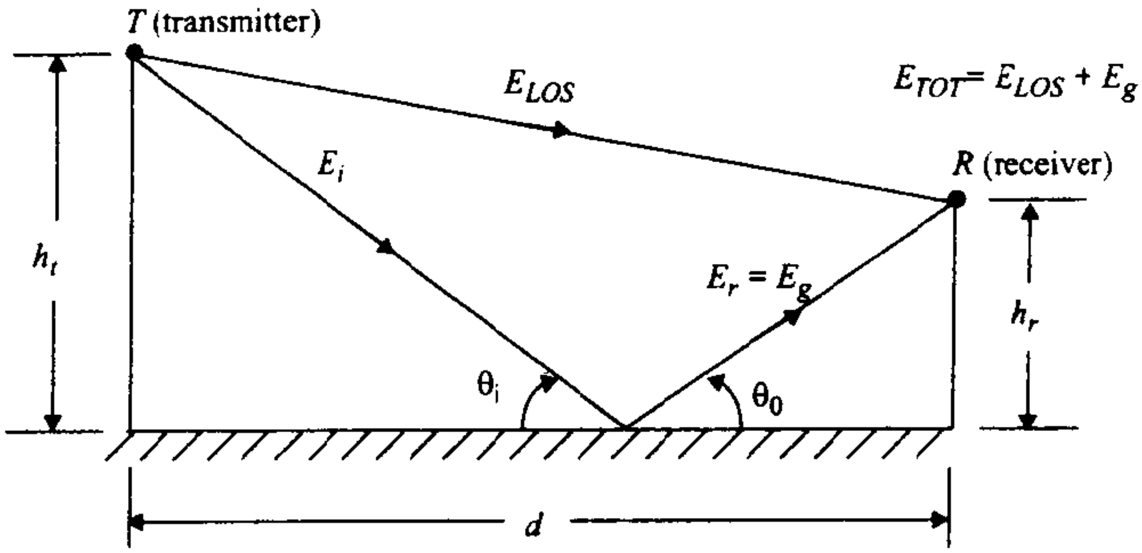


Figure (1) : Two ray Ground reflection model

Referring to the above figure (1), h_t is the height of the transmitter , h_r is the height of the receiver.

If E_0 is the free space E-field at a reference distance d_0 from the transmitter, then for $d>d_0$, the free space propagating E-field is given by,

$$E(d,t) = \frac{E_0 d_0}{d} \cos\left(\omega_c\left(t - \frac{d}{c}\right)\right) \text{ ------------------------------ (12)}$$

Where
$|E(d,t)| = E_0 d_0 / d$ represents the envelope of the E-field at d meters from the transmitter.

Two propagating waves arrives at the receiver:
(1) the direct wave that travels a distance d'.
(2) the reflected wave that travels a distance d''.

The E-field due to the line of sight component at the receiver can be expressed as,

$$E_{LOS}(d',t) = \frac{E_0 d_0}{d'} \cos\left(\omega_c\left(t - \frac{d'}{c}\right)\right) \text{ --------------------- (13)}$$

The E-field due to the line of sight component at the receiver can be expressed as,

$$E_g(d'',t) = \frac{E_0 d_0}{d''} \cos\left(\omega_c\left(t - \frac{d''}{c}\right)\right) \text{ --------------------- (14)}$$

According to the laws of reflection in dielectrics,

$$\theta_i = \theta_0 \text{ --- (15)}$$

and

$$E_g = \Gamma E_i \text{ --- (16)}$$
$$E_t = (1+\Gamma)E_i \text{ --- (17)}$$

Where
Γ is the reflection coefficient for ground.

For small values of θ_i, the reflected wave is equal in magnitude and 180° out of phase with the incident wave. The resultant E-field, assuming perfect horizontal E-field polarization and ground reflection, is the vector sum of E_{LOS} and E_g, and the resultant total E-field envelope is given by,

$$|E_{tot}| = |E_{LOS} + E_g| \text{ ------------------------------------ (18)}$$

The electric field $E_{tot}(d,t)$ can be expressed as the sum of eqn (13) & (14)

$$E_{tot}(d,t) = \frac{E_0 d_0}{d'} \cos\left(\omega_c\left(t - \frac{d'}{c}\right)\right) + (-1)\frac{E_0 d_0}{d''} \cos\left(\omega_c\left(t - \frac{d''}{c}\right)\right) \text{ ---- (19)}$$

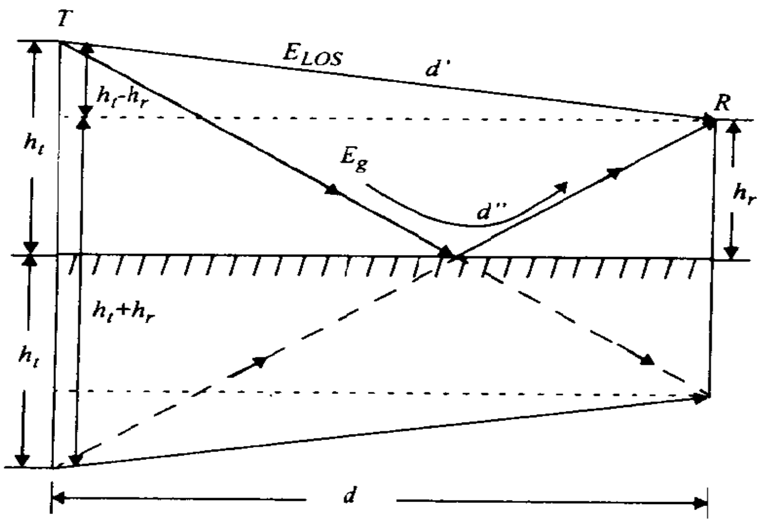


Figure (2) : The method of images is used to find the path difference between the line of sight and the ground reflected paths

Using the method of images, which is demonstrated by the geometry of figure (2) , the path difference , Δ, between the line of sight and the ground reflected paths can be expressed as,

Δ = d″ – d′ = $\sqrt{(h_t + h_r)^2 + d^2} \; - \; \sqrt{(h_t - h_r)^2 + d^2}$ --------------------------- (20)

When the T-R separation distance d is very large compared to $h_t + h_r$, eqn (20) can be simplified using a taylor series approximation,

Δ = d″ – d′ ≈ $\frac{2h_t h_r}{d}$ ---------------------------- (21)

Once the path difference is known, the phase difference Θ_Δ between two E-field components and the time delay τ_d between the arrival of the two components can be easily computed using the following relations

$\Theta_\Delta = \frac{2\pi\Delta}{\lambda} = \frac{\omega_c}{c}$ ---------------------------- (22)

And

$\tau_d = \frac{\Delta}{c} = \frac{\theta_\Delta}{2\pi f_c}$ ------------------------------- (23)

It should be noted that as d becomes large, the difference between the distances d′ and d″ becomes very small, and the amplitude of E_{LOS} and E_g are virtually identical and differ only in phase. That is,

$\left|\frac{E_0 d_0}{d}\right| = \left|\frac{E_0 d_0}{d'}\right| = \left|\frac{E_0 d_0}{d''}\right|$ ------------- (24)

If the received E-field is evaluated at some time, say at t = d″/c, Eqn (19) can be expressed as a phasor sum

$E_{tot}(d, t = \frac{d''}{c}) = \frac{E_0 d_0}{d'} \cos\left(\omega_c\left(\frac{d'' - d'}{c}\right)\right) + (-1)\frac{E_0 d_0}{d''} \cos 0°$ ------ (25)

$= \frac{E_0 d_0}{d'} \Theta_\Delta - \frac{E_0 d_0}{d''}$ --------------------- (26)

$= \frac{E_0 d_0}{d'} [\Theta_\Delta - 1]$ ------------------------ (27)

Where d is the distance over a flat earth between the bases of the transmitter and receiver antennas. Referring the figure (3) which shows how direct and ground reflected rays combine, the electric field (at the receiver) at a distance d from the transmitter can be written as

$| E_{tot}(d)| = \sqrt{(\frac{E_0 d_0}{2})^2 (\cos\theta_\Delta - 1)^2 + (\frac{E_0 d_0}{2})^2 \sin^2\theta_\Delta}$ -------- (28)

Or

$$|E_{tot(d)}| = \frac{E_0 d_0}{d} \sqrt{2 - 2\cos\theta_\Delta} \text{ -------------- (29)}$$

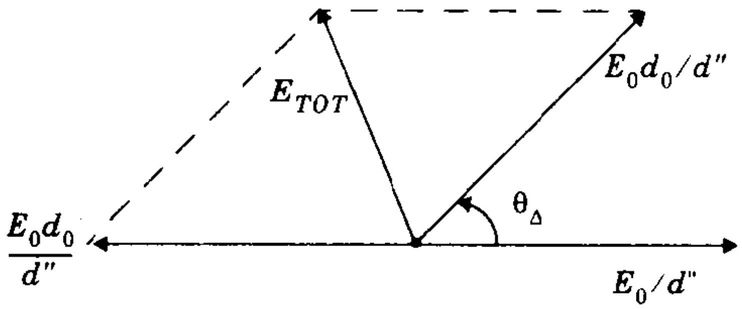


Figure (3): phasor diagram showing the electric field components of the line of sight, ground reflected and total received E –fields, derived from eqn (25)

Using trigonometric identities, eqn (29) can be expressed as,

$$|E_{tot(d)}| = 2\frac{E_0 d_0}{d}\sin\left(\frac{\theta_\Delta}{2}\right) \text{ -------------------- (30)}$$

Eqn (30) may be simplified whenever $\sin(\frac{\theta_\Delta}{2}) = \frac{\theta_\Delta}{2}$. This occurs when $\frac{\theta_\Delta}{2}$ is less than 0.3 radian. Using Eqn (21) and (22),

$$\frac{\theta_\Delta}{2} \approx \frac{2\pi h_t h_r}{\lambda d} < 0.3 \text{ rad --------------------- (31)}$$

Which implies that eqn (30) may be simplified whenever,

$$d > \frac{20\pi h_t h_r}{3\lambda} = \frac{20 h_t h_r}{\lambda} \text{ -------------------- (32)}$$

Thus, as long as d satisfies (32), the received E-field can be approximated as,

$$E_{tot}(d) \approx 2\frac{E_0 d_0}{d}\frac{2\pi h_t h_r}{\lambda d} \approx \frac{k}{d^2}\ V/m \text{ ---------------- (33)}$$

Where K is a constant related to E_0, the antenna heights, and the wavelength. This asymptotic behaviour is identical for both the E-field in the plane of incidence or normal to the plane of incidence. This free space power received at d is related to the square of the electric field through eqn (34).

$$P_r(d) = P_d A_e = \frac{|E|^2}{120\,\pi} A_e = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2}\ W \text{ ----------------- (34)}$$

Combining eqn (2), (34) and (33), the received power at a distance d from the transmitter for the two ray ground bounce model can be expressed as,

$$P_r = P_t G_t G_r \frac{{h_t}^2 {h_r}^2}{d^4} \text{ ----------------- (35)}$$

As seen from eqn (35) at large distances, the received power falls off with distance raised to the fourth power, or at a rate of 40 dB/decade. This is much more rapid path loss than is experienced in free space. Note also that at large values of d, the receiver power and path loss become independent of frequency. The path loss for the two ray model can be expressed in dB as,

$$PL(dB) = 40 \log d - (10 \log G_t + 10 \log G_r + 20 \log h_t + 20 \log h_r) \text{ ------- (36)}$$

At small T-R separation distances, eqn (19) must be used to compute the total E-field. When eqn (22) is evaluated for $\theta_\Delta = \pi$, then d= $(4h_t h_r)/\lambda$ is where the ground appears in the first Fresnel zone between the transmitter and receiver. The first Fresnel zone distance is a useful parameter in microcell path loss models.

Link Budget design:

Link budgets are, as the name implies, an accounting of the gains and losses that occur in a radio channel between a transmitter and receiver. As this is accounting it is needed to keep track of each loss and each gain that is experienced. Also, to find the noise power PN, it is needed to know the characteristics of the receiver. Figure (4) shows the relationship among link budget variables.

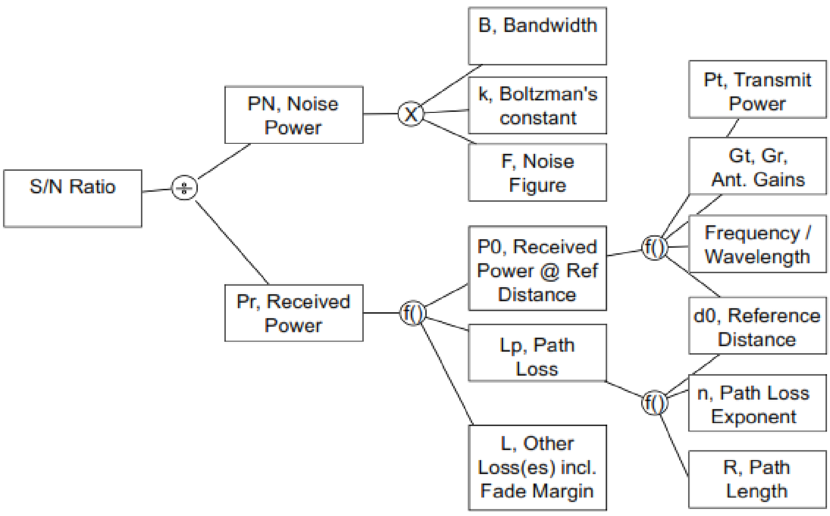


Figure (4) : Relationship among link budget variables

A universal link budget for received power is:

$$P_r \text{ (dBW)} = P_t \text{ (dBW)} + \sum \text{dB (Gains)} - \sum \text{dB (losses)}$$

Considering the simple break point model,

For distances $d < d_{break}$, the power is proportional to d^{-2}. Beyond that point, the power is proportional to d^{-n}, where n typically lies between 3.5 and 4.5. The received power is thus:

$$P_{RX}(d) = P_{Rx}(d_{break})(d/d_{break})^{-n} \qquad \text{for } d > d_{break}$$

Wireless systems, especially mobile systems, suffer from temporal and spatial variations of the transmission channel. In other words, even if the distance is fixed, the received power can change significantly with time, or with the movement of the Mobile Station (MS). The power computed from by using above eqn is only a mean value. If it is used as the basis for the link budget, then the transmission quality will be above the threshold only in approximately 50% of the times and locations. This is completely unacceptable coverage. Therefore it is necessary to add fading margin, which makes sure that the minimum receive power is exceeded at least. The value of the fading margin depends on the amplitude statistics of the fading.

Uplink and downlink are reciprocal, in the sense that the voltage and currents at the currents at the antenna ports are reciprocal. However, the noise figures of BS and MS are typically quite different. As MS have to be produced in quantity, it is desirable to use low cost components, which typically have higher noise figures. Furthermore, battery lifetime considerations dictate that BS can emit more power than MS. Finally BS and MS differ with respect to antenna diversity, how close they are to interferers, etc. Thus the link budgets of uplinks and downlinks are different.

<u>Small Scale fading:</u>

Multipath in the radio channel creates small scale fading effects. The three most important effects are:

1. Rapid changes in signal strength over small travel distance or time interval
2. Random frequency modulation due to varying Doppler shifts on different multipath signals
3. Time dispersion (Echoes) caused by multipath propagation delays.

In built-up urban areas, fading occurs because the height of the mobile antennas well below the height of surrounding structures, so there is no single line of sight path to the base station. Even when a line of sight exists, multipath still occurs reflections from the ground and surrounding structures. The incoming radio waves arrive from different directions with different propagation delays. The signal received by the mobile at any point in space may consists of a large number of plane waves having randomly distributed amplitudes, phases, and angles of arrival. These multipath components combine vectorially at the receiver antenna, and can cause the signal received by the mobile to distort or fade. Even when a mobile receiver is stationary, the received signal may fade due to movement of surrounding objects in the radio channel. If object in the radio channel are static, and motion is consider to be only due to that of the mobile, then fading is purely a spatial phenomenon. The spatial variation of the resulting signal are seen as temporal variation by the receiver as it moves through the multipath field. Due to the constructive and destructive effects of multipath waves summing a various points in space, a receiver moving at high speed can pass through several fades in a small period of time. In a more serious case, a receiver may stop at a particular location at which the received signal is in a deep fade. Maintaining good communication can then become very difficult, although passing vehicles or people walking in the vicinity of the mobiles can often disturb the field pattern, thereby diminishing the likelihood of the received signal remaining in a deep null for a long period of time. Antenna space diversity can prevent deep fading nulls. Figure (5) shows typical rapid variations in the received signal level due to small scale fading as a receiver is moved over a distance of a few meters.

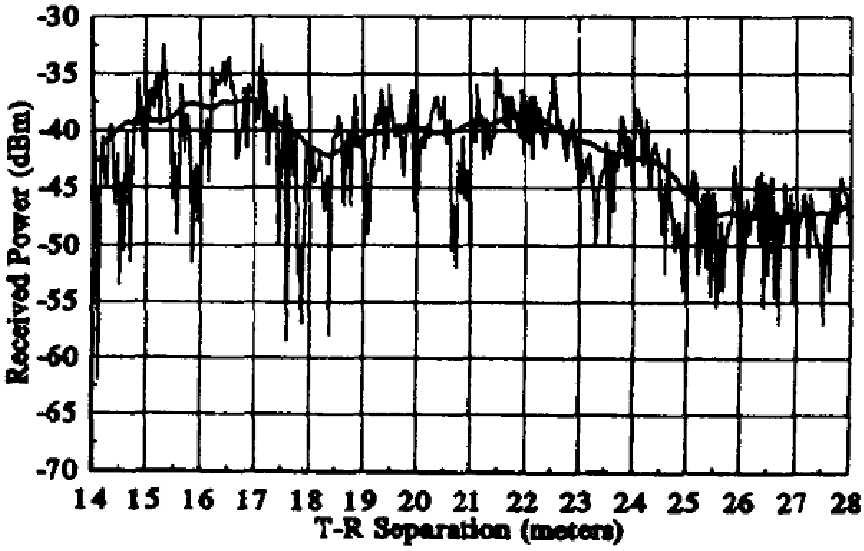


Figure (5) : Small scale and large scale fading

Due to the relative motion between the mobile and the base station, each multipath wave experiences an apparent shift in frequency. The shift in received signal frequency due to motion is called Doppler shift, and is directly proportional to the velocity and direction of motion of the mobile with respect to the direction of arrival of the received multipath wave.

Factors influencing small-scale fading:

Many physical factors in the radio propagation channel influence small scale fading. These include the following:

1. Multipath propagation:

The presence of reflecting objects and scatterers in the channel creates a constantly changing environment that dissipates the signal energy in the amplitude, phase and time. These effects result in multiple versions of the transmitted signal that arrive at the receiving antenna, displaced with respect to one another in time and spatial orientation. The random phase and amplitude of the different multipath components cause fluctuation in signal strength, thereby inducing small scale fading, signal distortion, or both. Multipath propagation often lengthens the time required for the baseband portion of the signal to reach the receiver which can cause signal smearing due to intersymbol interference.

2. Speed of the mobile:

Relative motion between the base station and the mobile results in random frequency modulation due to different Doppler shifts on each of the multipath components. Doppler shift will be positive or negative depending on whether the mobile receiver is moving towards or away from the base station.

3. Speed of the surrounding objects:

If objects in the radio channel are in motion, they induce a time varying Doppler shift on multipath components. If the surrounding objects move at a greater rate than the mobile, then this effect dominates the small scale fading. Otherwise, the motion of surrounding objects may be ignored, and only the speed of the mobile need to be considered. The coherence time defines the "staticness" of the channel, and is directly impacted by the Doppler shift.

4. The transmission bandwidth of the signal:

If the transmitted radio signal bandwidth is greater than the bandwidth of the multipath channel, the received signal will be distorted, but the received signal strength will not fade much over a local area(i.e., the small scale signal fading will not be significant). As will be shown, the bandwidth of the channel can be quantified by the coherence bandwidth which is related to the specific multipath structure of the channel. The coherence bandwidth is a measure of the maximum frequency difference for which signals are still strongly correlated in amplitude. If the transmitted signal has a narrow bandwidth as compared to the channel, the amplitude of the signal will change rapidly, but the signal will not be distorted in time. Thus, the statistics of small scale signal strength and the likelihood of signal smearing appearing over small scale distances are very much related to the specific amplitudes and delays of the multipath channel, as well as the bandwidth of the transmitted signal.

Doppler shift:

Consider a mobile moving at a constant velocity v, along a path segment having a length d between points x and y, while it receives signals from a remote source , as illustrated in figure (6)

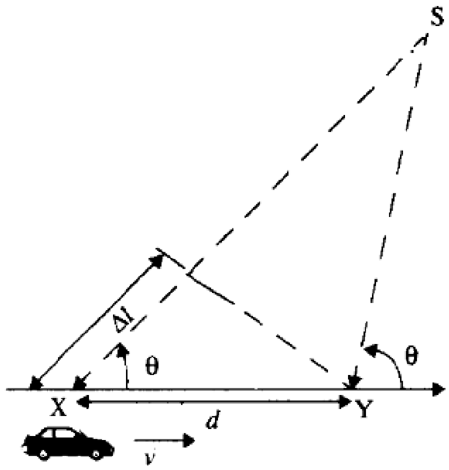


Figure (6) : Illustration of Doppler effect

The difference in path lengths travelled by the wave from source s to the mobile at points x and y is,

$$\Delta l = d \cos \theta \text{ -------------------------------- (37)}$$

$$= v\, \Delta t \cos \theta \text{ ----------------------------- (38)}$$

Where,

Δt is the time required for the mobile to travel from x to y,

θ is assumed to be the same at the points x and y since the source is assumed to be very far away. The phase change in the received signal due to the difference in path lengths is therefore,

$$\Delta\varphi = \frac{2\pi\Delta l}{\lambda} = \frac{2\pi v\Delta t}{\lambda}\cos\theta \text{ --------------------------- (39)}$$

And hence the apparent change in frequency, or doppler shift, is given by f_d, where,

$$f_d = \frac{l}{2\pi} \cdot \frac{\Delta\varphi}{\Delta t} = \frac{v}{\lambda}\cos\Theta \text{ ---------------------------- (40)}$$

eqn (40) relates the Doppler shift to the mobile velocity and the spatial angle between the direction of motion of the mobile and the direction of arrival of the wave. It can be seen from eqn (40) that if the mobile is moving toward the direction of arrival of the wave, the Doppler shift is positive (i.e., the apparent received frequency is increased), and if the mobile is moving away from the direction of arrival of the wave, the Doppler shift is negative (i.e., the apparent frequency is decreased). Multipath components from a CW signal that arrive from different direction contribute to Doppler spreading of the received signal, thus increasing the signal bandwidth.

Parameters of mobile multipath channels:

Many multipath channel parameters are derived from the power delay profile. Power delay profiles are measured using the techniques and are generally represented as plots of relative received power as a function of excess delay with respect to a fixed time delay reference. Power delay profiles are found by averaging instantaneous power delay profile measurements over a local area in order to determine an average small scale power delay profile. Depending on the time resolution of the probing pulse and the type of multipath channels studied, researchers often choose to sample at spatial separations of quarter of a wavelength and over receiver movements no greater than 6m in outdoor channels and no greater than 2m in indoor channels in the 450 MHz – 6 GHz range. This small scale sampling avoids large scale averaging bias in the resulting small scale statistics. Figure (7) shows typical power delay profile plots from outdoor and indoor channels, determined from a large number of closely sampled instantaneous profiles.

Time Dispersion Parameters :

In order to compare different multipath channels and to develop some general designs guidelines for wireless systems, parameters which grossly quantifies the multipath channels are used. The mean excess delay, rms delay spread, and excess delay spread (X dB) are multipath channel parameters that can be determined from the power delay profile. The time dispersive properties of wide band multipath channels are most commonly quantified by their mean excess delay (τ) and rms delay spread (σ_τ). The mean excess delay is the first moment of the power delay profile and is defined to be

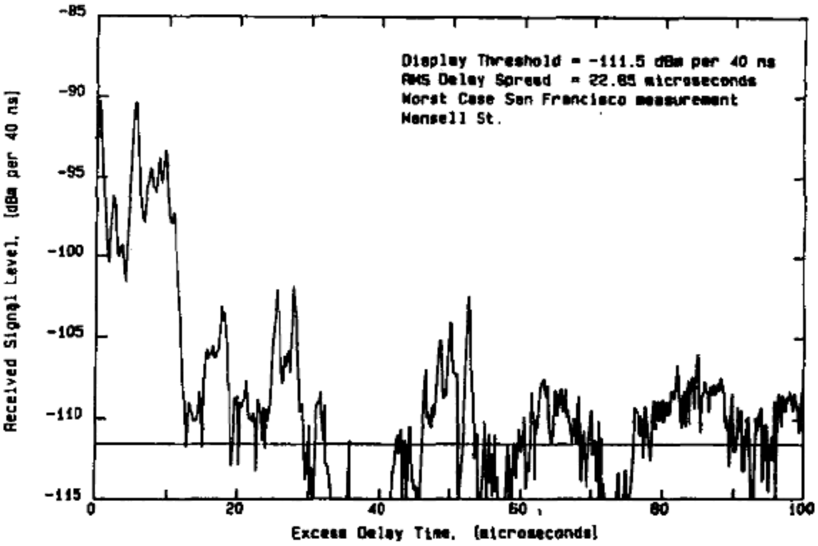
$$\tau = \frac{\sum_k a_k^2 \tau_k}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k)\tau_k}{\sum_k P(\tau_k)} \quad \text{------------------------- (41)}$$

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

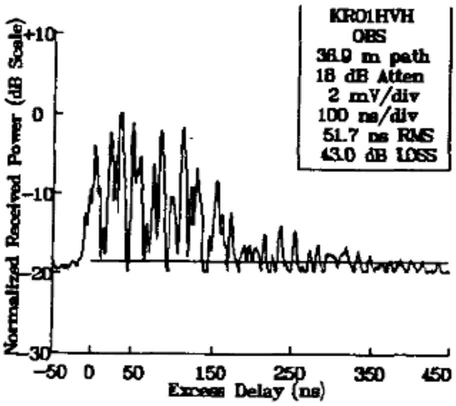
$$\sigma_\tau = \sqrt{\tau^2 - (\tau^2)} \quad \text{----------------------------------- (42)}$$

where

$$\tau^2 = \frac{\sum_k a_k^2 \tau_k^2}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k)\tau_k^2}{\sum_k P(\tau_k)} \quad \text{------------------- (43)}$$



(a)



(b)

Figure (7) : Measured multipath power delay profiles:
(a) From a 900 MHz cellular system in San Fransico
(b) Inside a grocery store at 4 GHz

These delays are measured relative to the first detectable signal arriving at the receiver at $\tau_0 = 0$. Eqn (41) – Eqn (43) do not rely on the absolute power level $P(\tau_k)$ only the relative amplitudes of the multipath components within $P(\tau_k)$. typically values of rms delay spread are on the order of microseconds in outdoor mobile radio channels and on the order of nanoseconds in indoor radio channels. Table (1) shows the typical measured values of rms delay spread.

Environment	**Frequency (MHz)**	**RMS delay spread (σ_τ)**	**Notes**
Urban	910	1300 ns avg. 600 ns st. dev. 3500 ns max.	New York City
Urban	892	10 – 25 μs	Worst case san Francisco
Sub urban	910	200 – 310 μs	Averaged typical case
Sub urban	910	1960-2110 ns	Averaged extreme case
Indoor	1500	10-50 ns 25 ns median	Office building
Indoor	850	270 ns max.	Office building
Indoor	1900	70-94 ns avg. 1470 ns max.	Three san Francisco buildings

Table (1) : Typical measured values of RMS delay spread

It is important to note that the rms delay spread and mean excess delay are defined from a single power delay profile which is the temporal or spatial average of consecutive impulse response measurements collected and averaged over a local area. Typically, many measurements are made at many local areas in order to determine a statistical range of channel parameters for a mobile communication system over a large scale area.

The maximum excess delay (X dB) of the power delay profile is defined to be the time delay during which multipath energy falls to X dB below the maximum. In other words, the maximum excess is defined as $\tau_X - \tau_0$, where τ_0 is the first arriving signal and τ_X is the maximum delay at which a multipath component is within X dB of the strongest arriving multipath signal (which does not necessarily arrived at τ_0). Figure (8) illustrates the computation of the maximum excess delay for multipath components within 10 dB of the maximum. The maximum excess delay (X dB) defines the temporal extent of the multipath that is above a particular threshold. The value of τ_X is sometimes called the excess delay spread of a power delay profile, but in all cases must be specified with the threshold that relates the multipath noise floor to the maximum received multipath component.

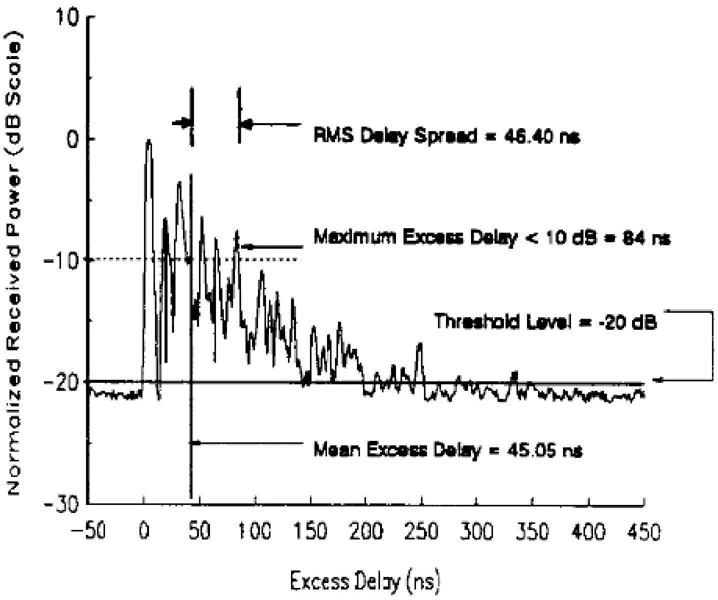


Figure (8) : Example of an indoor power delay profile; rms delay spread, mean excess delay, maximum excess delay(10 dB) and threshold level

In practice, values for τ, τ^2 and σ_τ depend on the choice of noise threshold used to process $P(\tau)$. The noise threshold is used to differentiate between received multipath components and thermal noise. If the noise threshold is set too low, then noise will be processed as multipath, thus giving rise to values of τ, τ^2 and σ_τ that are artificially high.

It should be noted that the power delay profile and the magnitude frequency response of a mobile radio channel are related through the fourier transform. It is therefore possible to obtain an equivalent description of the channel in the frequency domain using its frequency response characteristics. Analogous to the delay spread parameters in the time domain, coherence bandwidth is used to characterize the channel in the frequency domain. The rms delay spread and coherence bandwidth are inversely proportional to one another, although their exact relationship is a function of the exact multipath structure.

Coherence Bandwidth:

While the delay spread is a natural phenomenon caused by the reflected and scattered propagation path in radio channel, the coherence bandwidth, B_c is a defined relation derived from the rms delay spread. Coherence bandwidth is a statistical measure of the range of frequencies over which the channel can be considered "flat"(i.e., a channel which passes all spectral components with approximately equal gain and linear phase). In other words, coherence bandwidth is the range of frequencies over which two frequency components have a strong potential for amplitude correlation. Two sinusoids with frequency separation greater than B_c are affected quite differently by the channel. If the coherence bandwidth is defined as the bandwidth over which the frequency correlation function is above 0.9, then the coherence bandwidth is approximately,

$$B_c = \frac{1}{50\sigma_\tau} \text{ ----------------------------- } (44)$$

If the definition is relaxed so that the frequency correlation function is above 0.5, then the coherence bandwidth is approximately,

$$B_c = \frac{1}{5\sigma_\tau} \text{ ----------------------------- } (45)$$

It is important to note that an exact relationship between coherence bandwidth and rms delay spread is a function of specific channel impulse responses and applied signals. In general, spectral analysis techniques and simulation are required to determine the exact impact that time varying multipath has on a particular transmitted signal. For this reason, accurate multipath channel models must be used in the design of specific modems for wireless applications.

Doppler spread and coherence time:

Delay spread and coherence bandwidth are parameters which describe the time dispersive nature of the channel in a local area. However, they do not offer information about the time varying nature of the channel caused by either relative motion between the mobile and base station, or by movement of objects in the channel. Doppler spread and coherence time are parameters which describe the time varying nature of the channel in a small scale region.

Doppler spread B_d is a measure of the spectral broadening caused by 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. When a pure sinusoidal tone of frequency f_c is transmitted, the received signal spectrum, called the Doppler spectrum, will have components in the range $f_c - f_d$. to $f_c + f_d$, where f_d is Doppler shift. The amount of spectral broadening depends on f_d which is a function of the relative velocity of the mobile, and the angle Θ between the direction of motion of the mobile and direction of arrival of the scattered waves. If the baseband signal bandwidth is much greater than B_d, the effects of Doppler spread are negligible at the receiver. This is a slow fading channel.

Coherence Time T_c, is the time domain dual of Doppler spread and is used to characterize the time varying nature of the frequency dispersiveness of the channel in the time domain. The Doppler spread and coherence time are inversely proportional to one another. That is,

$$T_c = 1/f_m \text{ ----------------------------- } (46)$$

Coherence time is actually a statistical measure of the time duration over which the channel impulse response is essentially invariant, and quantifies the similarity of the channel response is essentially invariant, and quantifies the similarity of the channel response at different times. In other words, coherence time is the time duration over which two received signals have a strong potential for amplitude correlation. If the reciprocal bandwidth of the baseband signal is greater than the coherence time of the channel, then the channel will change during the transmission of the baseband message, thus causing distortion at the receiver. 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,

$$T_c = \frac{9}{16\,\pi f_m} \text{ -------------------------- } (47)$$

Where

$f_m = v/\lambda$ is maximum Doppler shift,

In practice, eqn (46) suggests a time duration during which a signal may fluctuate wildly and eqn (47) is often too restrictive. A popular rule of thumb for modern communication is to define the coherence time as the geometric mean of equation (46) and (47). That is,

$$T_c = \sqrt{\frac{9}{16\pi\, f_m^2}} = \frac{0.423}{f_m} \text{ ------------ } (48)$$

The definition of coherence time implies that two signals arriving with a time separation greater than T_c are affected differently by the channel.

Types of small scale fading:

Type of fading experienced by a signal propagating through a mobile radio channel depends on the nature of the transmitted signal with respect to the characteristics of the channel. Depending on the relation between the signal parameters and the channel parameters , different transmitted signals will undergo different types of fading. The time dispersion and frequency dispersion mechanisms in a mobile radio channel lead to four possible distinct effects, which are manifested depending on the nature of the transmitted signal, the channel and the velocity. While multipath delay spread leads to time dispersion and frequency selective fading, Doppler spread leads to frequency dispersion an frequency dispersion and time selective fading. The two propagations mechanisms are independent of another. Figure (9) shows a tree of the four different types of fading.

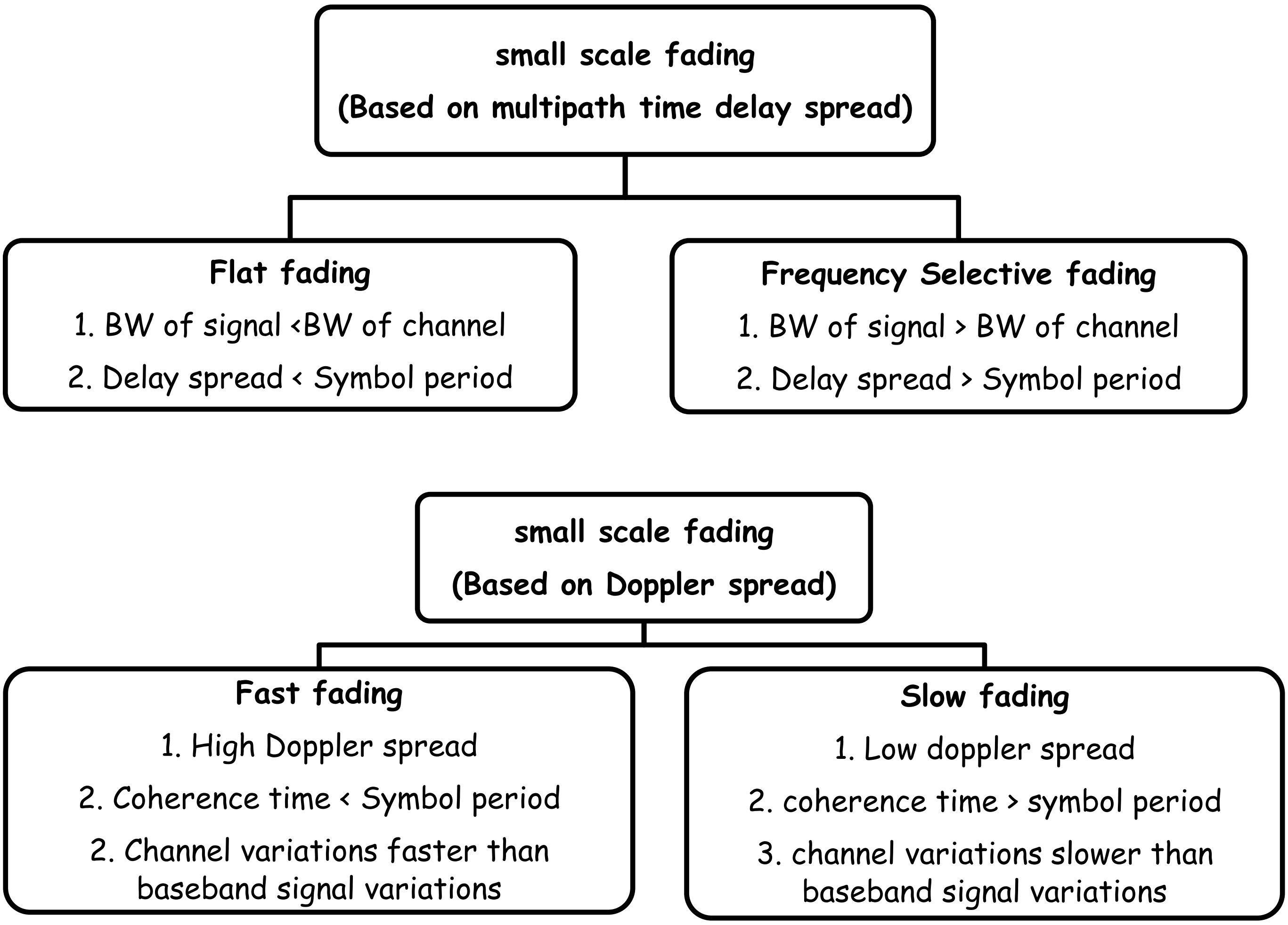


Figure (9): Types of small scale fading

Fading effects due to multipath time delay spread:

Time dispersion due to multipath causes the transmitted signal to undergo either flat or frequency selective fading.

Flat fading:

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. This type of fading is historically the most common type of fading. In flat fading, the multipath structure of the channel is such that the spectral characteristics of the transmitted signal are preserved at the receiver. However, the strength of received signals changes with time due to the fluctuations in the gain of the signal caused by multipath. The characteristics of a flat fading channel are illustrated in figure (10).

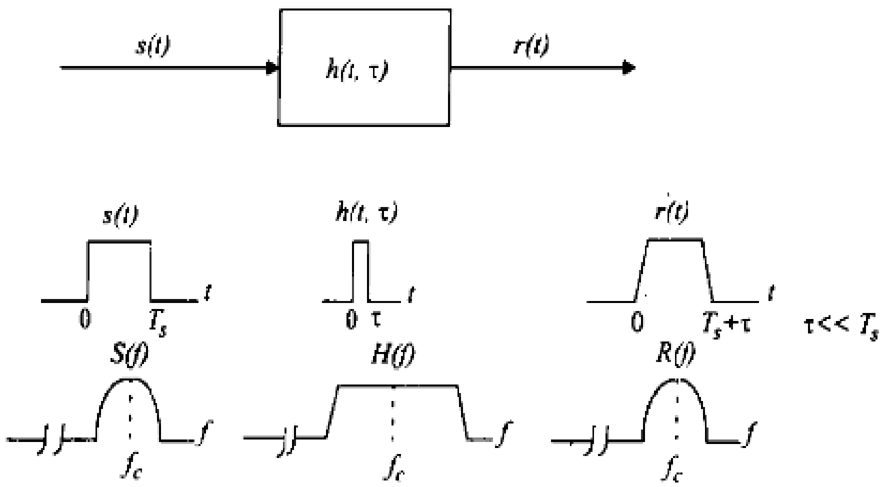


Figure (10): Flat fading channel characteristics

It can be seen from figure (10) that if a channel gain changes over time a change of amplitude occurs in the received signals. Overtime, the received signals r(t) varies in gain but the spectrum of the transmission is preserved. In flat fading channels, the reciprocal bandwidth of the transmitted signal is much larger than the multipath time delay spread of the channel, and $h_b(t,\tau)$ can be approximated as having no excess delay. (i.e., a single delta function with τ =0). Flat fading channels are also known as amplitude varying channels and are sometimes referred to as narrow band channel, since the bandwidth of the applied signal is narrow as compared to the channel flat fading bandwidth. Typical falt fading channels cause deep fades, and thus may require 20 or 30 dB more transmitter power to achieve low bit error rates during time of deep fades as compared to systems operating over non fading channels. The distribution of the instantaneous gain of flat fading channels is important for designing ratio links, and the most common amplitude distribution is Rayleigh distribution. The Rayleigh flat fading channel mode assumes that the channel induce an amplitude which varies in time according to the Rayleigh distribution.

To summarize, a signal undergoes flat fading if

$$B_s \ll B_c \text{ ------------------------------------ } (49)$$

And

$$T_s \gg \sigma_T \text{ ------------------------------------ } (50)$$

Where
T_s is the reciprocal bandwidth
B_s is the bandwidth of the transmitted modulation
σ_T is rms delay spread of the channel
B_c is coherence bandwidth of the channel

Frequency selective fading :
If the channel possess 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. Under such conditions, the channel impulse response has a multipath delay spread which is greater than the reciprocal bandwidth of the transmitted waveform. When this occurs, the received signals includes multiple versions of the transmitted waveform which are attenuated (faded) and delayed in time, and hence the recived signals is distorted. Frequency selective fading is due to time dispersion of the transmitted symbols within the channel. Thus the channel includes intersymbol interference (ISI). Viewed in the frequency domain, certain frequency compnents in the recived signals spectrum have greater gains than others.

Frequency selective fading channels are much more difficult to made models than flat fading channels since each multipath signals must be modelled and the signal must be considered to be a linear filter. It is for this reason that wide band multipath measurements are made, and models are developed from these measurements. When analysing mobile communication systems, statistical impulse response model such as the two way Rayleigh fading models (which considers the impulse response to be made up of two delta functions which independently fade and have sufficient time delay between them to induce frequency selective fading upon the applied signal), or computer generated are measured impulse responses, are generally used for analysing frequency selective small scale fading. Figure (11) illustrates the characteristics of a frequency selective fading channel.

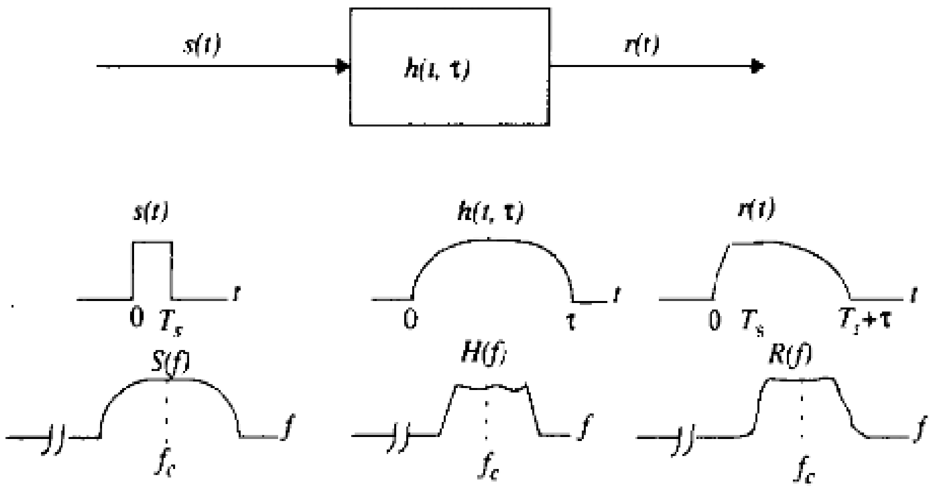


Figure (11) : Frequency selective fading channel characteristics

For frequency selective fading, the spectrum S(f) of the transmitted signal has a bandwidth which is greater than the coherence bandwidth B_c of the channel. Viewed in the frequency domain, the channel becomes frequency selective, where the gain is different for different frequency components. frequency selective fading is caused by multipath delay which approach or exceed the symbol period of the transmitted symbol. frequency selective fading channels are also known as wide band channels since the bandwidth of the signal S(t) is wider than the bandwidth of channel impulse response. As time varies, the channel varies in gain and phase across the spectrum of S(t), resulting in time varying in distortion in the received signals r(t).

To summarize, a channel undergoes frequency selective fading if

$$B_s > B_c \quad \text{-------------------------------------} \quad (51)$$

And

$$T_s < \sigma_T \quad \text{---------------------------------------} \quad (52)$$

A common role of thumb is that a channel is flat fading if $T_s \geq 10\ \sigma_T$ and a channel is frequency selected if $T_s \leq 10\ \sigma_T$, although this is dependent on the specific type of modulation used.

Fading effects due to Doppler spread:

Fast fading:

Depending on how rapidly the transmitted baseband signal changes as compared to the rate of change of the channel, a channel may classified either as a fast fading or slow channel. In a fast fading channel, 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 transmitted signal. This causes frequency dispersion (also called as time selective fading) due to Doppler spreading, which leads to the signal distortion. Viewed in the frequency domain, signal distortion due to the fast fading increases with increase in Doppler spread relative to the bandwidth of the transmitted signal. Therefore, a signals undergoes fast fading if

$$T_s > T_c \quad (53)$$

And

$$B_s < B_d \quad (54)$$

It should be noted that when a channel is specified as a fast or slow fading channel, it does not specify whether the channel is flat fading or frequency selective in nature. Fast fading only deals with the rate of change of the channel due to motion. In the case of the flat fading channel, it can be approximated that the impulse response to be simply delta function(no time delay). Hence, a flat fading, fast fading channel is a channel in which the amplitude of a delta function varies faster than the rate of change of the transmitted base band signal. In case of a frequency selective, fast fading channel, the amplitudes, phases, and time delays of any one of the multipath components vary faster than the rate of change of the transmitted signal. In practice, fast fading only occurs for very low data rates.

Slow fading:

In a slow fading channel, the channel impulse response changes at a rate much slower than the transmitted baseband signal S(t). In this case, the channel may be assumed to be static over one or several reciprocal bandwidth intervals. In the frequency domain, this implies that the Doppler spread of the channel is much less than the bandwidth of the baseband signal. Therefore, a signal undergoes slow fading if

$$T_s \ll T_c \quad (55)$$

And

$$B_s >> B_d \text{ ---------------------------------- } (56)$$

It should be clear that the velocity of the mobile (or velocity of subjects in the channel) and the baseband signalling determines whether the signals undergoes fast fading and slow fading.

The relation the various multipath parameters and the type of fading experienced by the signal are summarized in figure (11). Fast and slow fading is different from small scale and large scale fading. It should be emphasized that fast and slow fading deals with the relationship between the time rate of change in the channel and the transmitted signal and not with propagation path loss models.