EEM602 Internet of Things

Lecture # 9
(IOT Course wireless Communications fundementals)

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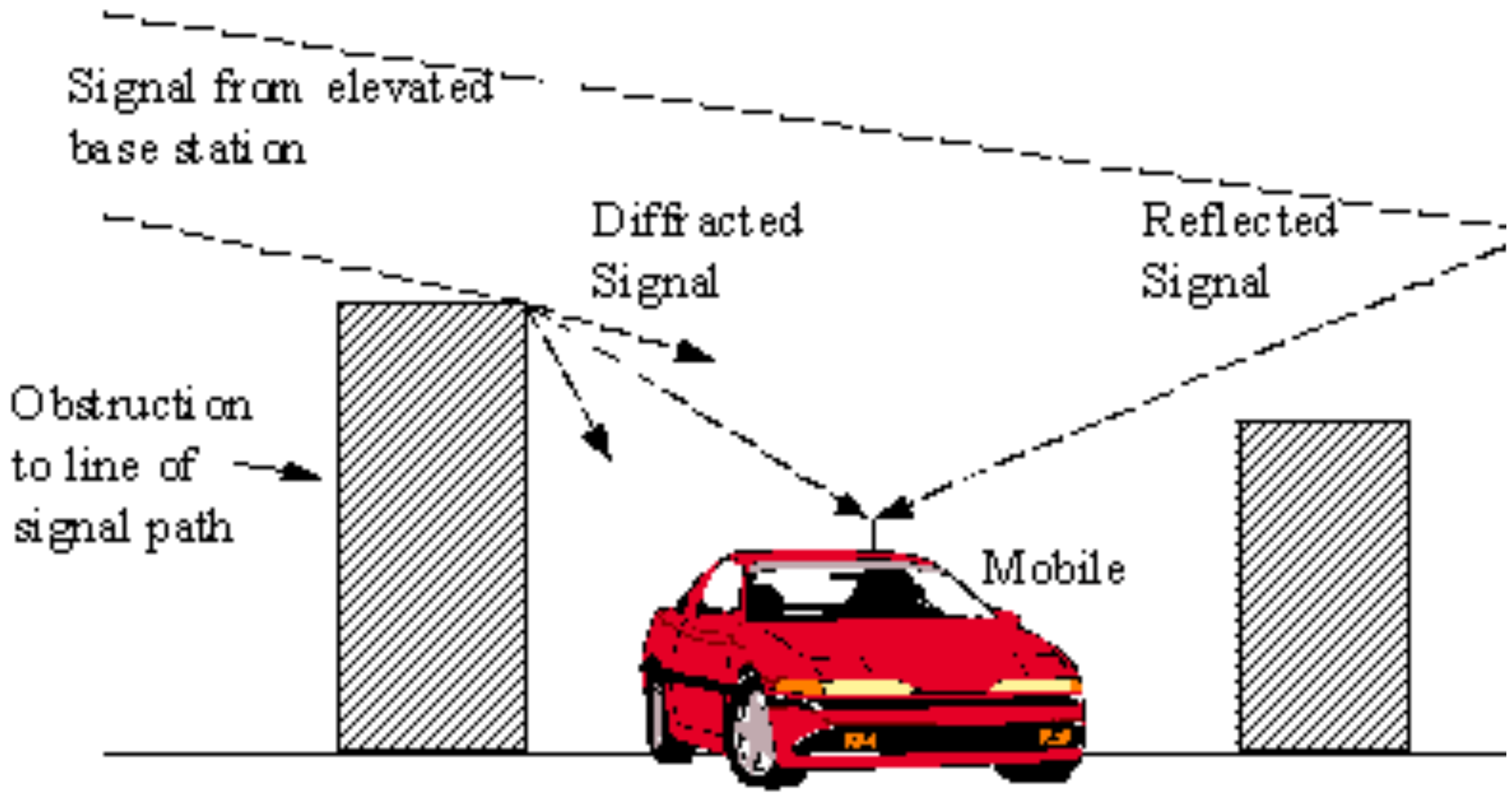


Figure 2 Radio Propagation Effects

Introduction to Radio Wave Propagation

Electromagnetic wave propagation are attributed to **reflection, diffraction, and scattering.**

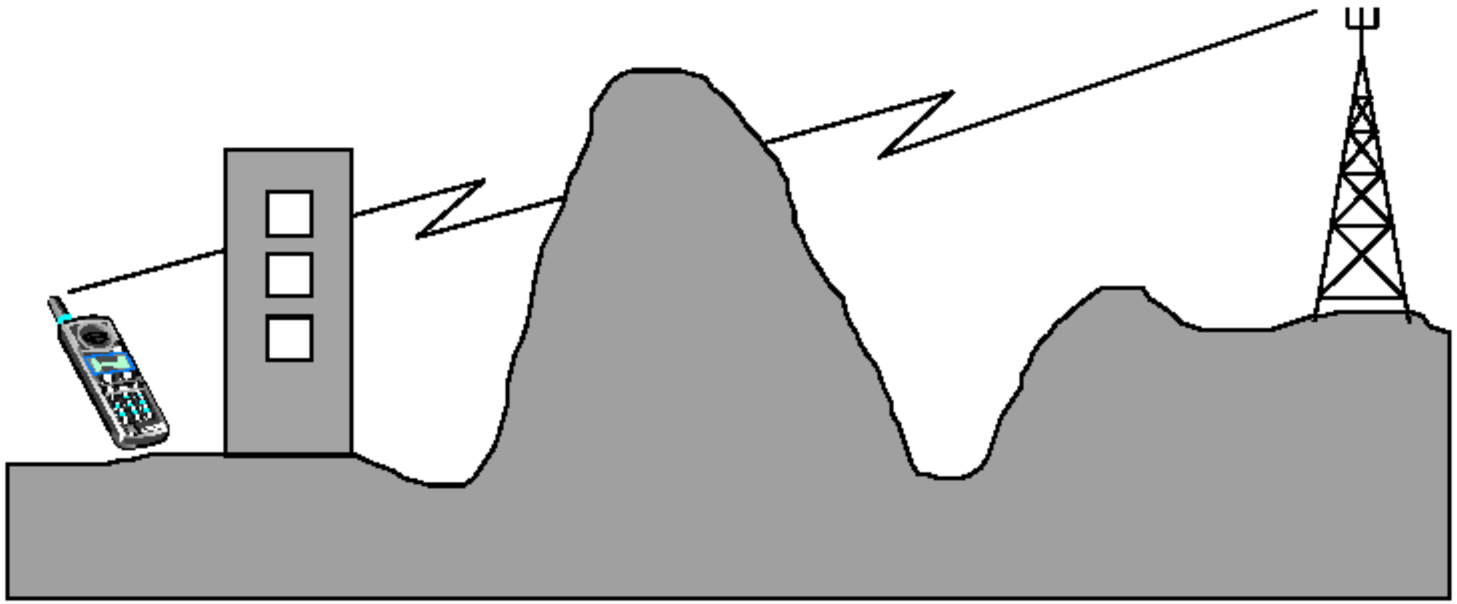
Most cellular radio systems operate in urban areas where there is no direct line-of-sight path between the transmitter and the receiver, and where the presence of high- rise buildings **causes severe diffraction loss.**

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.

Fading Problems

1. Shadowing (Normal fading):

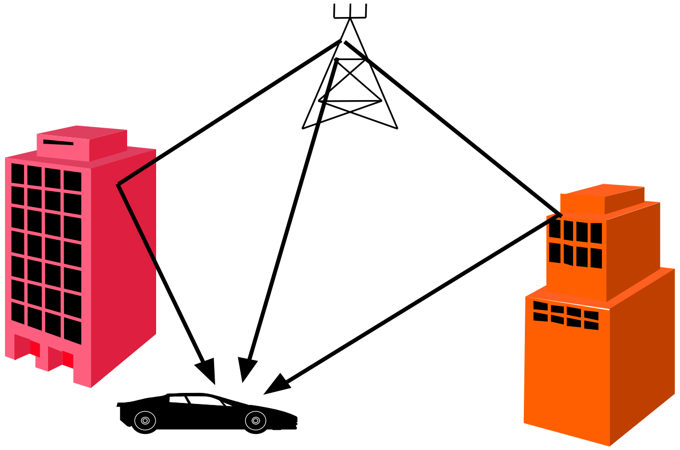
The reason for shadowing is the presence of obstacles like large hills or buildings in the path between the site and the mobile.
The signal strength received fluctuates around a mean value while changing the mobile position resulting in undesirable beats in the speech signal.



Fading Problems

2. Rayleigh Fading (Multi-path Fading)

The received signal is coming from different paths due to a series of reflection on many obstacles. The difference in paths leads to a difference in paths of the received components.



propagation models

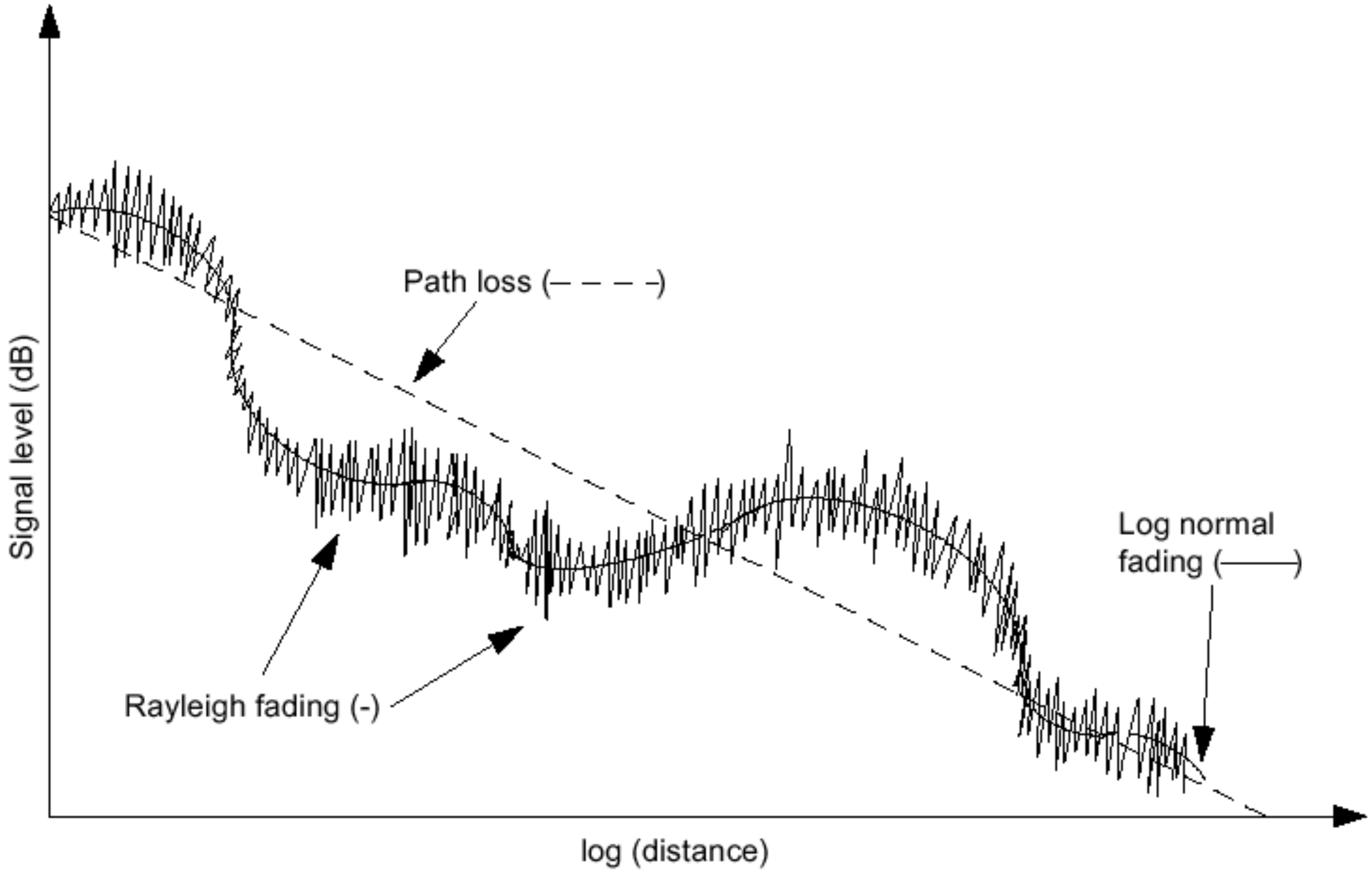
- Propagation models that predict the mean signal strength for an arbitrary transmitter-receiver (T-R) separation distance are useful in estimating the radio coverage area of a transmitter and are called **large-scale propagation models,** since they characterize signal strength over large T-R separation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

In small-scale fading, the received signal power may vary by as much as three or four orders of magnitude **(30 or 40 dB)** when the receiver is moved by only a fraction of a wavelength.

As the mobile moves away from the transmitter over much larger distances, the local average received signal will gradually decrease, and it is this local average signal level that is predicted by **large-scale propagation models.**

Fading Problems



Small-scale and large-scale fading

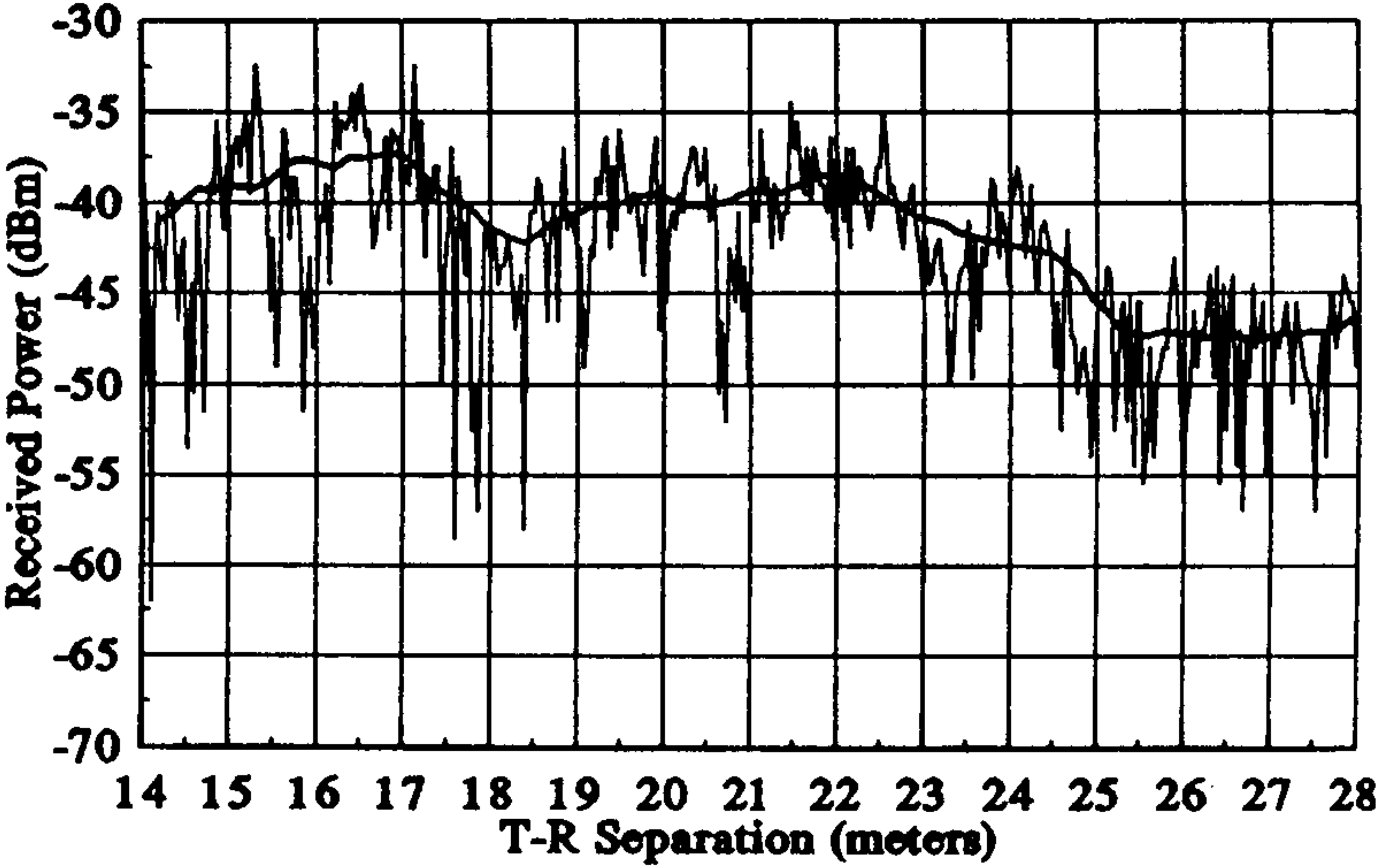
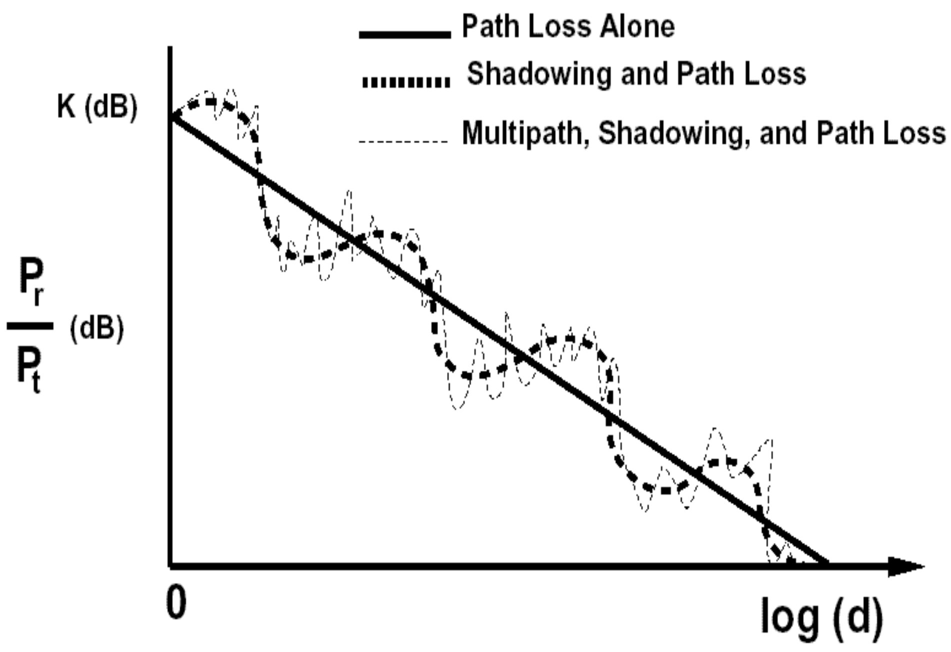


Figure 4.1 Small-scale and large-scale fading.

Basic Ideas: Path Loss, Shadowing, Fading

- Variable decay of signal due to environment, multipaths, mobility



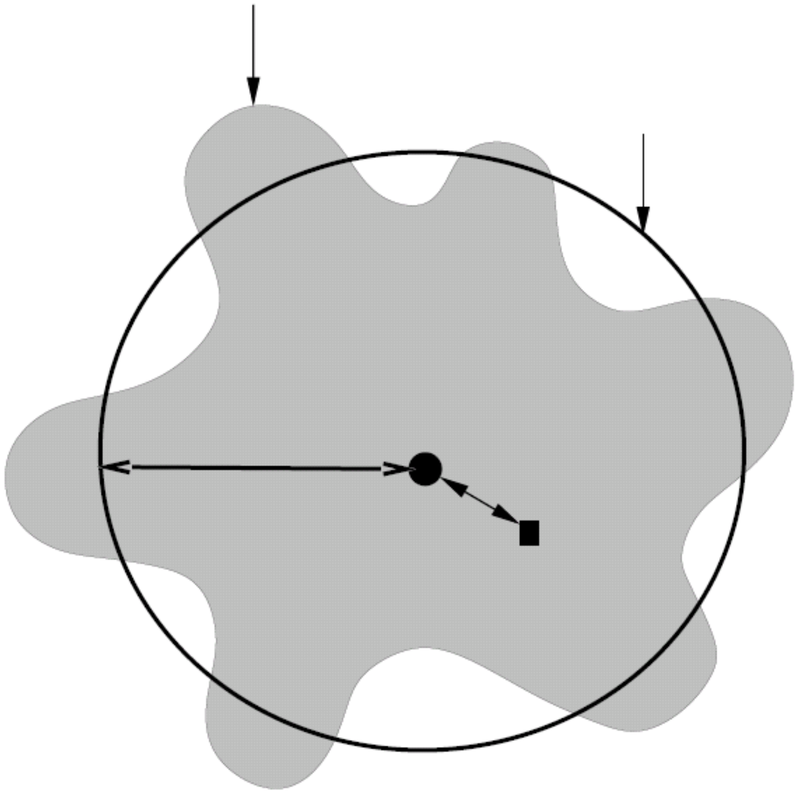
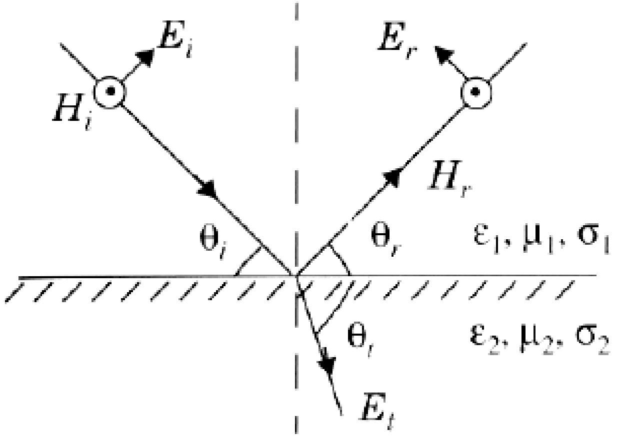


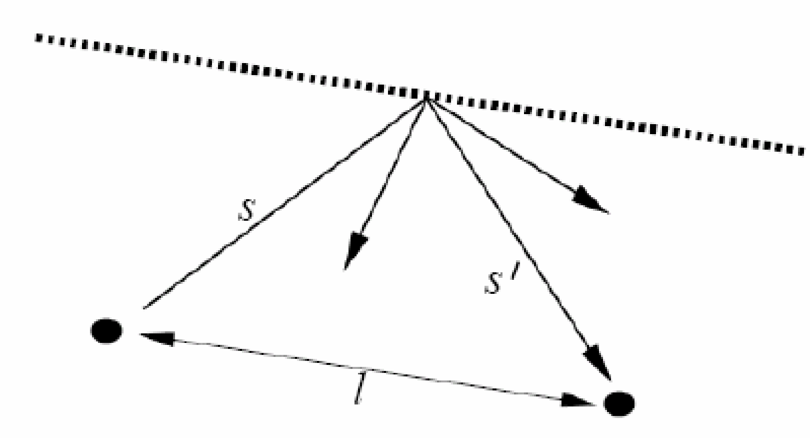
Figure 2.10: Contours of Constant Received Power.

Source: A. Goldsmith book

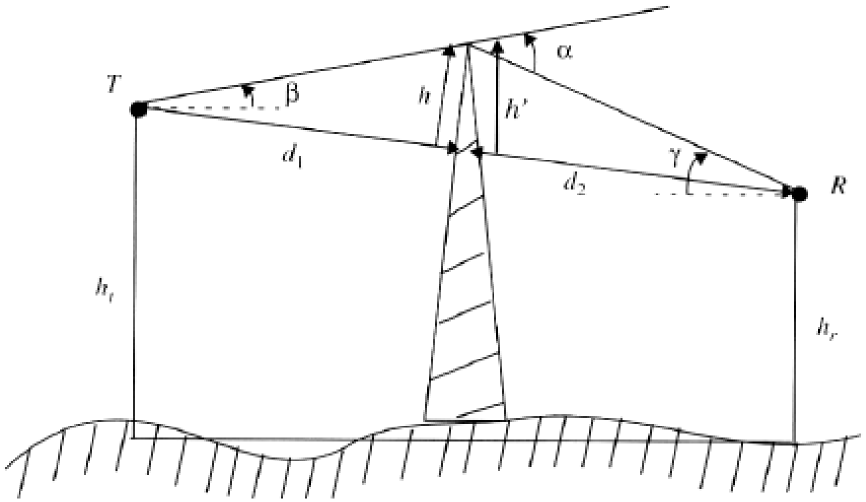
Reflection, Diffraction, Scattering



Reflection/Refraction: large objects (>>λ)



Scattering: small objects, rough surfaces (<λ): foilage, lamposts, street signs



Diffraction/Shadowing: "bending" around sharp edges,

- 900Mhz: λ ~ 30 cm
- 2.4Ghz: λ ~ 13.9 cm
- 5.8Ghz: λ ~ 5.75 cm

Large-scale Fading:
Path Loss, Shadowing

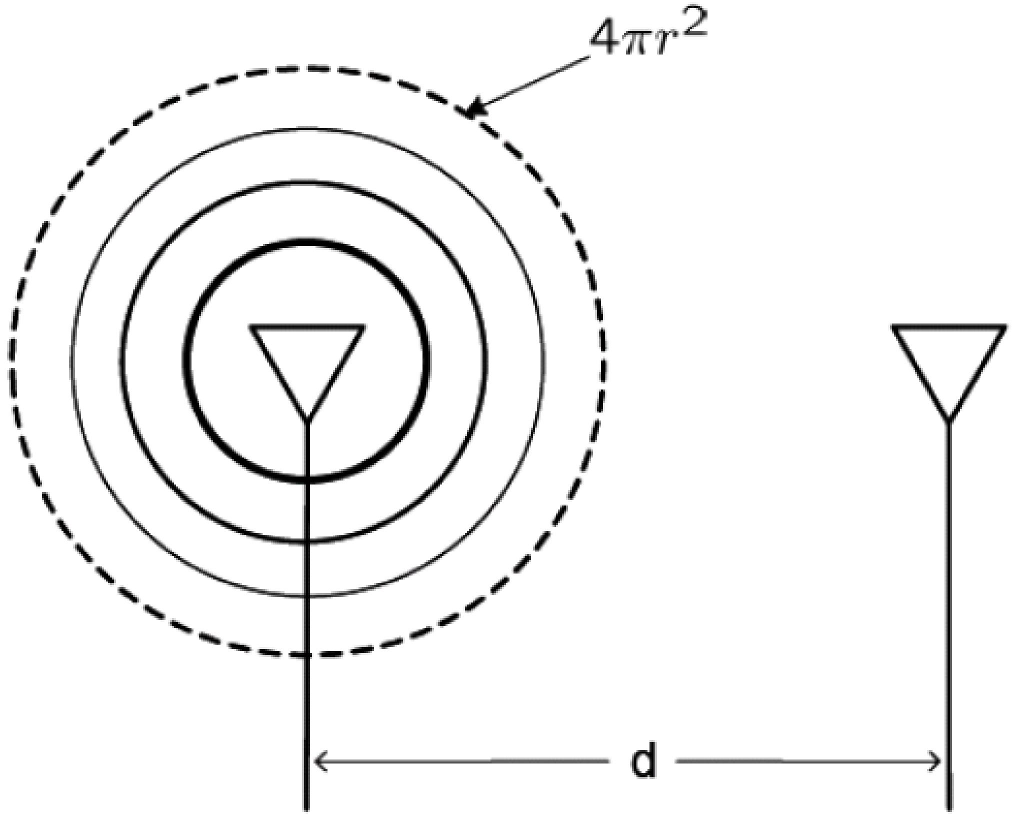
3.2 Free Space Propagation Model

The free space propagation model is used to predict received signal strength **when the transmitter and receiver have a clear, unobstructed line-of-sight path between them.**

the free space model predicts that received power decays as a function of the T-R separation distance raised to some power (i.e. a power law function).

Free-Space-Propagation

- The EM radiation field that decays as 1/d (power decays as $1/d^2$)
-



The free space power received by a receiver antenna which is separated from a radiating transmitter antenna by a distance **d**, is given by the **Friis free space equation,**

$$P_r(d) = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2 L} \qquad (3.1)$$

Where

P_t is the transmitted power,

P_r (d) is the received power which is a of the 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 **($L \geq 1$),** and

λ is the wavelength in meters.

The gain of an antenna is related to its **effective aperture**, *Ae* by

$$G = \frac{4\pi A_e}{\lambda^2} \qquad (3.2)$$

The effective aperture A_e is related to the physical size of the antenna, and λ is related to the carrier frequency by

$$\lambda = \frac{c}{f} = \frac{2\pi c}{\omega_c} \qquad (3.3)$$

where f is the carrier frequency in Hertz, ω_c is the carrier frequency in radians per second, and c is the speed of light given in meters/s

1. The values for **P_t and P_r** must be expressed in the same units,

2. and **G_t and G_r** are dimensionless quantities.

3. The miscellaneous losses **L (L ≥ I)** 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.

4. The Friis free space equation of (3.1) shows that the **received power** falls off as the **square of the T-R separation distance**. This implies that the received power decays with distance at a rate of 20 dB/decade.

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

$$EIRP = P_t G_t \quad (3.4)$$

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

In practice, effective radiated power (ERP) is used instead of EIRP to denote the maximum radiated power as compared to a half-wave dipole antenna (instead of an isotropic antenna).

Since a dipole antenna has a gain of **1.64** (2.15 dB above an isotropic), the ERP will be **2.15 dB** smaller than the EIRP for the same transmission system.

In practice, antenna gains are given in units of **dBi** (dB gain with respect to an isotropic source) or **dBd** (dB gain with respect to a half-wave dipole)

Decibels: dB, dBm, dBi

- **dB** (Decibel) = 10 log $_{10}$ (Pr/Pt)
  Log-ratio of two signal levels. Named after Alexander Graham Bell. For example, a cable has 6 dB loss or an amplifier has 15 dB of gain. System gains and losses can be added/subtracted, especially when changes are in several orders of magnitude.

- **dBm (dB milliWatt)**
  Relative to 1mW, i.e. 0 dBm is 1 mW (milliWatt). Small signals are -ve (e.g. -83dBm).

  Typical 802.11b WLAN cards have +15 dBm (32mW) of output power. They also spec a -83 dBm RX sensitivity (minimum RX signal level required for 11Mbps reception).

  For example, 125 mW is 21 dBm and 250 mW is 24 dBm. (commonly used numbers)

- **dBi (dB isotropic) for EIRP (Effective Isotropic Radiated Power)**
  The gain a given antenna has over a theoretical isotropic (point source) antenna. The gain of microwave antennas (above 1 GHz) is generally given in dBi.

- **dBd (dB dipole)**
  The gain an antenna has over a dipole antenna at the same frequency. A dipole antenna is the smallest, least gain practical antenna that can be made. A dipole antenna has 2.14 dB gain over a 0 dBi isotropic antenna. Thus, a simple dipole antenna has a gain of 2.14 dBi or 0 dBd and is used as a standard for calibration.

  The term dBd (or sometimes just called dB) generally is used to describe antenna gain for antennas that operate under 1GHz (1000Mhz).

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, 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

$$PL\,(\mathrm{dB}) = 10\log\frac{P_t}{P_r} = -10\log\left[\frac{G_t G_r \lambda^2}{(4\pi)^2 d^2}\right] \tag{3.5}$$

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

$$PL\,(\mathrm{dB}) = 10\log\frac{P_t}{P_r} = -10\log\left[\frac{\lambda^2}{(4\pi)^2 d^2}\right] \tag{3.6}$$

The Friis free space model is only a valid predictor 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.

mitter antenna aperture and the carrier wavelength. The Fraunhofer distance is given by

$$d_f = \frac{2D^2}{\lambda} \tag{3.7.a}$$

where D is the largest physical linear dimension of the antenna. Additionally, to be in the far-field region, d_f must satisfy

$$d_f \gg D \tag{3.7.b}$$

and

$$d_f \gg \lambda \tag{3.7.c}$$

a close-in distance, d0, as a known received power reference point.

The received power, ***$P_r(d)$.*** at any distance ***d*** > ***d_0***,

The reference distance must be chosen such that it lies in the far-field region, that is, ***d_0*** > ***d_f***, and ***d_0*** is chosen to be smaller than any practical distance used in the mobile communication system.

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

may be expressed in units of **dBm** or **dBW** by simply taking the logarithm of both sides and multiplying by 10.

For example, if *Pr* is in units of dBm, the received power is given by

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

Example 3.1

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

Solution to Example 3.1

Given:

Largest dimension of antenna, $D = 1$ m

Operating frequency $f = 900$ MHz, $\lambda = c/f = \dfrac{3 \times 10^8 \text{ m/s}}{900 \times 10^6 \text{ Hz}}$ m

Using equation (3.7.a), far-field distance is obtained as

$$d_f = \frac{2(1)^2}{0.33} = 6 \text{ m}$$

Example 3.2

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 to Example 3.2

Given:

Transmitter power, $P_t = 50$ W.

Carrier frequency, $f_c = 900$ MHz

Using equation (3.9),

(a) Transmitter power,

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

(b) Transmitter power,

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

The received power can be determined using equation (3.1).

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

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

The received power at 10 km can be expressed in terms of dBm using equation (3.9), where $d_0 = 100$ m and $d = 10$ km

$$P_r(10\text{ km}) = P_r(100) + 20\log\left[\frac{100}{10000}\right] = -24.5\text{ dBm} - 40\text{ dB}$$
$$= -64.5\text{ dBm}.$$

Practical Link Budget Design using Path Loss Models

• Most radio propagation models are derived using a **combination of analytical** (from a set of measured data) **and empirical methods.** (based on fitting curves)

• **all propagation factors** through actual field measurements **are included**.

•some classical propagation models are now used to predict large-scale coverage for mobile communication systems design.

•Practical path loss estimation techniques are presented next.

Free Space Propagation Model n=2

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

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

Next model is generalized model for any value of n
Log-distance Path Loss Model

1 Log-distance Path Loss Model

average received signal power decreases logarithmically with distance, (theoretical and measurments), whether in outdoor or indoor radio channels.

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

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

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

where :**n** is the path loss exponent, **d_0 is the close-in reference distance** (determined from measurements close to the transmitter), **d** is the **T-R separation distance**.

•Bars denote **the ensemble average** of all possible path loss values for a given ***d***.

• On a **log-log scale plot**, **the modeled path loss is a straight line** with a slope equal to ***10n*** dB per decade.

Path loss at a close-in reference distance

- **(d0)** :**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.

- The reference path loss is calculated using the free space path loss formula given by **friis free space equation** or through field measurements at distance **d0**.

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

Table 3.2 **lists typical path loss exponents obtained in various mobile radio environments.**

Table 3.2 Path Loss Exponents for Different Environments

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

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.

2. Log-normal Shadowing

The log distance path loss model does not consider the fact that the surrounding environmental clutter may be vastly different at two different locations having the same T-R separation.

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

$$PL(d)[dB] = \overline{PL}(d) + X_\sigma = \overline{PL}(d_0) + 10n\log\left(\frac{d}{d_0}\right) + X_\sigma \quad (3.69.a)$$

and

$$P_r(d)[dBm] = P_t[dBm] - PL(d)[dB] \text{ (antenna gains included in } PL(d)) \quad (3.69.b)$$

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

log-normal shadowing. Simply implies that measured signal levels at a specific T-R separation have a Gaussian (normal) distribution about the distance-dependent mean of (3.68),

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

- **d_0**, **n**, **σ** (the standard deviation),

statistically describe the path loss model for an arbitrary location having a specific T-R separation.

- 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.

- **PL(d0)** is obtained from measurements or free space assumption (friis) from the transmitter to d0.

An example of how the path loss exponent is determined from measured data follows.

Figure 3.17 illustrates actual measured data in several cellular radio systems and demonstrates the random variations about the mean path loss (in dB) due to shadowing at specific T-R separations.

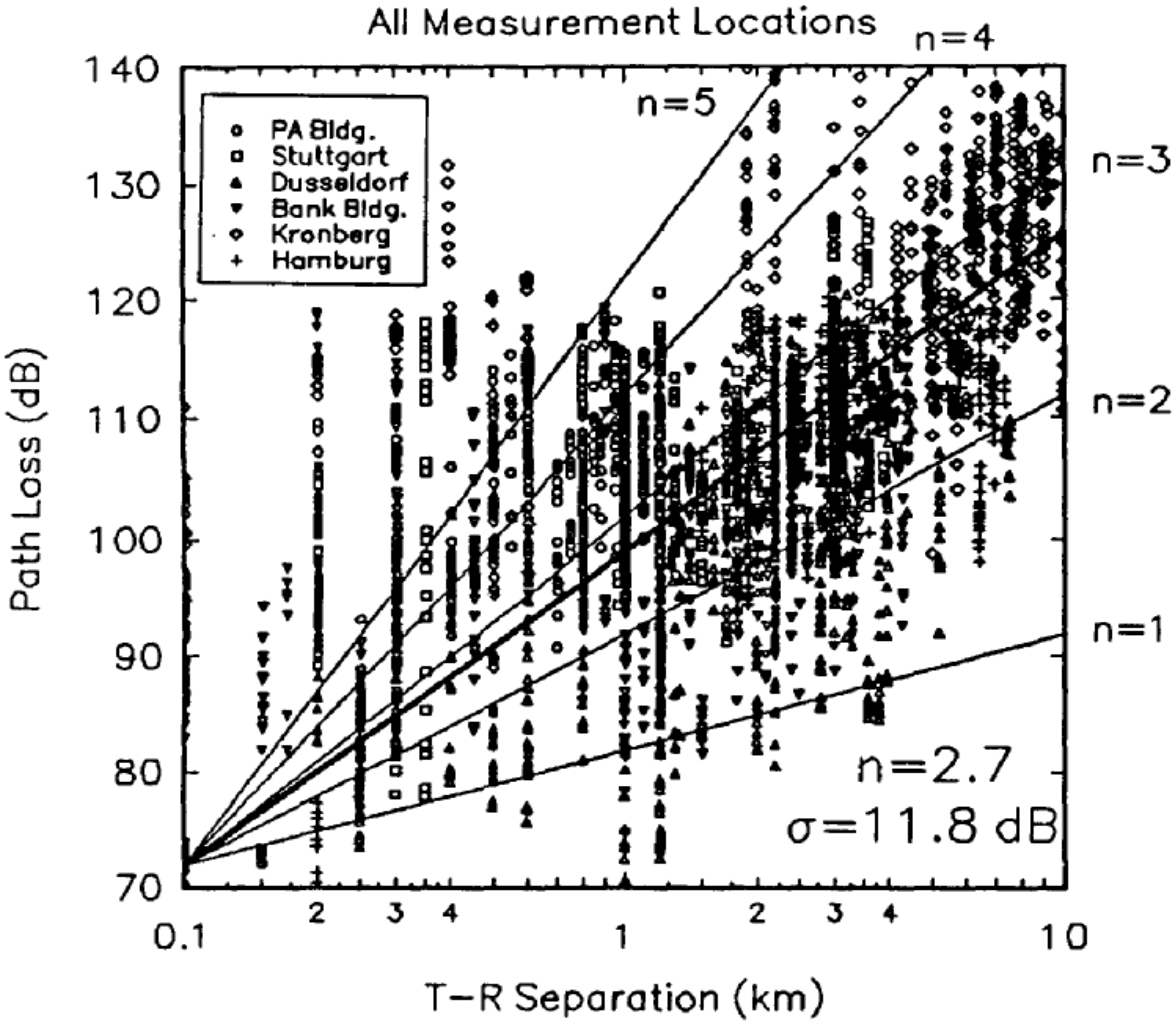


Figure 3.17

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

Example 3.9

Four received power measurements were taken at distances of 100 m, 200 m, 1 km, and 3 km from a transmitter. These measured values are given in the following table. It is assumed that the path loss for these measurements follows the model in equation (3.69.a), where $d_0 = 100$ m: (a) find the minimum mean square error (MMSE) estimate for the path loss exponent, n ; (b) calculate the standard deviation about the mean value; (c) estimate the received power at $d = 2$ km using the resulting model;

Distance from Transmitter	Received Power
100 m	0 dBm
200 m	-20 dBm
1000 m	-35 dBm
3000 m	-70 dBm

Solution to Example 3.9

The MMSE estimate may be found using the following method. Let p_i be the received power at a distance d_i and let $\hat{p}_i$ be the estimate for p_i using the $(d/d_0)^n$ path loss model of equation (3.67). The sum of squared errors between the measured and estimated values is given by

$$J(n) = \sum_{i=1}^{k} (p_i - \hat{p}_i)^2$$

The value of n which minimizes the mean square error can be obtained by equating the derivative of $J(n)$ to zero, and then solving for n.

(a) Using equation (3.68), we find $\hat{p}_i = p_i(d_0) - 10n\log(d_i/100\text{ m})$. Recognizing that $P(d_0) = 0$ dBm, we find the following estimates for $\hat{p}_i$ in dBm:

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

The sum of squared errors is then given by

$$\begin{aligned} J(n) &= (0-0)^2 + (-20-(-3n))^2 + (-35-(-10n))^2 \\ &\quad + (-70-(-14.77n))^2 \\ &= 6525 - 2887.8n + 327.153n^2 \end{aligned}$$

$$\frac{dJ(n)}{dn} = 654.306n - 2887.8.$$

Setting this equal to zero, the value of n is obtained as $n = 4.4$.

(b) The sample variance $\sigma^2 = J(n)/4$ at $n = 4.4$ can be obtained as follows.

$$J(n) = (0+0) + (-20+13.2)^2 + (-35+44)^2 + (-70+64.988)^2$$
$$= 152.36.$$

$$\sigma^2 = 152.36/4 = 38.09$$

therefore

$\sigma = 6.17$ dB, which is a biased estimate. In general, a greater number of measurements are needed to reduce σ^2.

(c) The estimate of the received power at $d = 2$ km is given by

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

A Gaussian random variable having zero mean and $\sigma = 6.17$ could be added to this value to simulate random shadowing effects at $d = 2$ km.

Measurement-Based Propagation Models

Outdoor Propagation Models

- Radio transmission in a mobile communications system often takes place over irregular terrain (landscape).

- The terrain profile of a particular area needs to be taken into account for estimating the path loss. The terrain profile may vary from a simple curved earth profile to a highly mountainous profile.

- The presence of trees, buildings, and other obstacles also must be taken into account. A number of propagation models are available to predict path loss over irregular terrain.

- While all these models aim to predict signal strength at a particular receiving point or in a specific local area (called a sector), the methods vary widely in their approach, complexity, and accuracy.

- Most of these models are based on a systematic interpretation of measurement data obtained in the service area. **Some of the commonly used outdoor propagation models are now discussed**.

Here are number of standard models for computing the mean received signal level. These are well-treated in the literature.

Models are typically developed to approximate system behavior over a given area. Models are developed for different types of areas using extensive measured data. Curve-fitting techniques are used to fit equations to the experimental data. Models are usually developed for the following classifications of areas:

- urban area (built up area such as city centers)
- suburban area (one and two story homes with open spaces)
- open areas (pastures, farms, etc.)

Other models are often included. Examples are

- geographical data bases (USGS data base for the US)
- atmospheric models for scattering

Parameters of interest often include the following:

- transmission frequency
- antenna heights
- surface reflectivity
- path length
- ground dielectric and conductivity constants
- polarization
- terrain effects (ground cover, etc.)

Many different models are possible and most have both strong and weak points to recommend their use.

Okumura Model

The Okumura model is widely used. It is simple to apply and often gives reasonable results. Based a set of curves obtained by curve fitting to measurement results. Typical parameters:

- Frequency range: 150MHz to 2 or 3 GHz
- Distances: 1 km to 100 km
- Base station antenna heights: 30 m to 1000 m

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

where

$$L_F = \text{free space path loss} = -10\log_{10}\left[\left(\frac{\lambda}{4\pi d}\right)^2\right]$$

$A_{mu}(f,d)$ = medium attenuation relative to free space (in graph)

G_{AREA} = terrain correction (in graphs)

$G(h_{re})$ = receiving antenna factor

$G(h_{te})$ = transmitting antenna factor

L_{50} is the 50th percentile (i.e., median) value of propagation path loss,

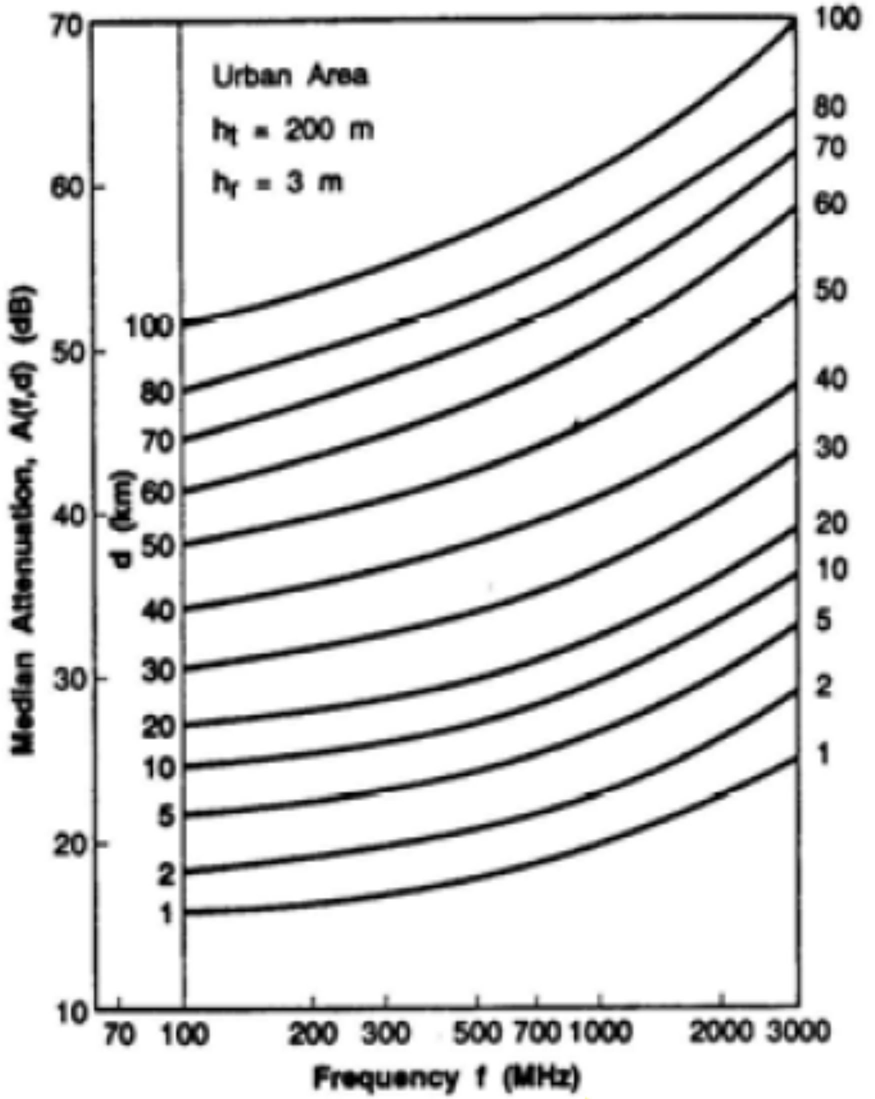


Figure 4.23 Median attenuation relative to free space $(A_{mu}(f,d))$, over a quasi-smooth terrain [from [Oku68] © IEEE].

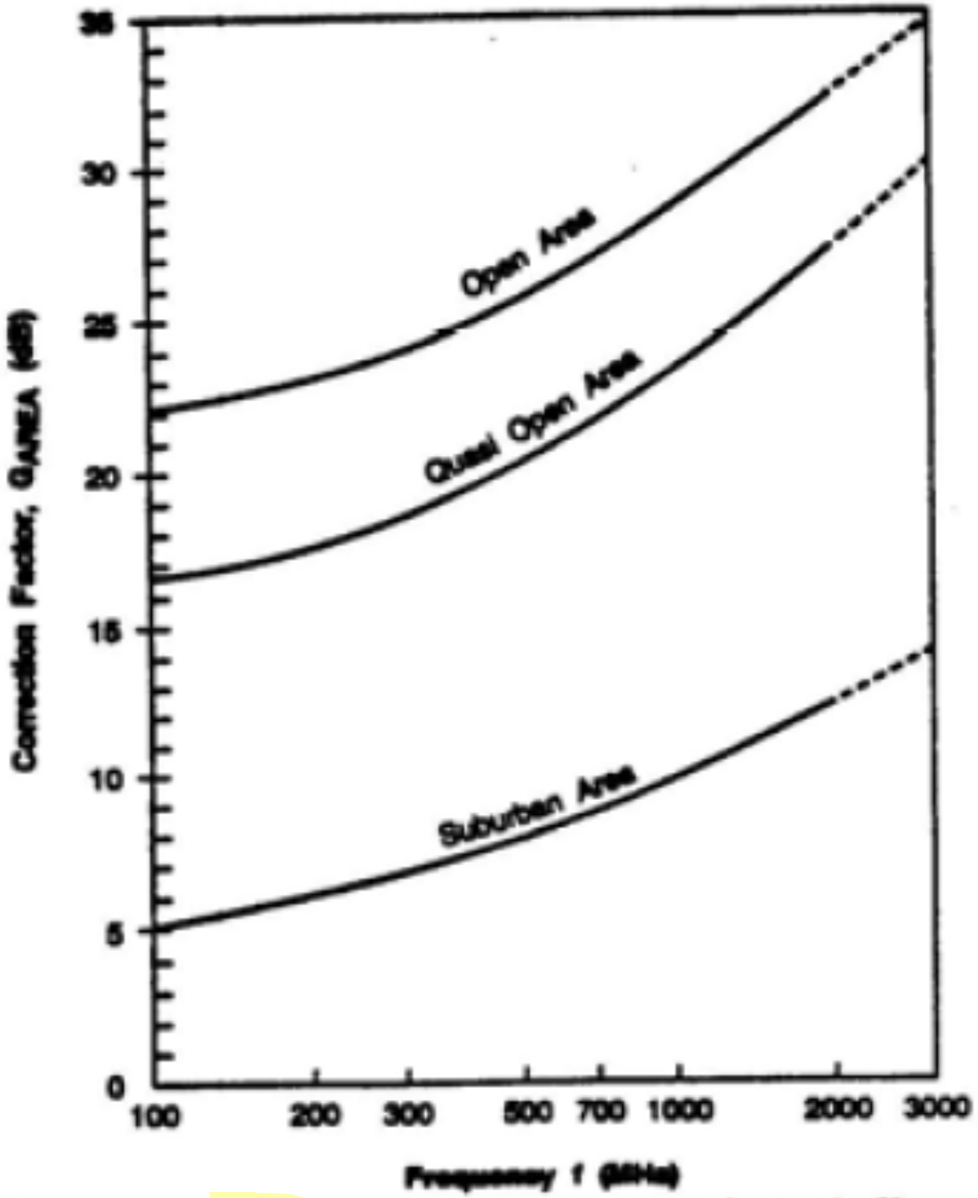


Figure 4.24 Correction factor, G_{AREA}, for different types of terrain [from [Oku68] © IEEE].

Okumura Model

Antenna height correction factors:

$$G(h_{te}) = 20\log_{10}(h_{te}/200), \quad 30\text{ m} < h_{te} < 1000\text{ m}$$

$$G(h_{re}) = 10\log_{10}(h_{re}/3), \quad h_{re} < 3\text{m}$$

$$G(h_{re}) = 20\log_{10}(h_{re}/3), \quad 3\text{ m} < h_{re} < 10\text{ m}$$

Example 3.10

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

Solution to Example 3.10

The free space path loss L_F can be calculated using equation (3.6) as

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

From the Okumura curves

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

and

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

Using equation (3.81.a) and (3.81.c) we have

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

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

Using equation (3.80) the total mean path loss is

$$\begin{aligned} L_{50}(\text{dB}) &= L_F + A_{mu}(f,d) - G(h_{te}) - G(h_{re}) - G_{AREA} \\ &= 125.5 \text{ dB} + 43 \text{ dB} - (-6) \text{ dB} - 10.46 \text{ dB} - 9 \text{ dB} \\ &= 155.04 \text{ dB}. \end{aligned}$$

Therefore, the median received power is

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

Hata Model

An easy to use model that is quite popular is the Hata/Okumura model defined using by following assumptions:

Base station height: between 30 and 200 meters
Carrier frequency: between 150 and 1,500 MHz
Mobile station antenna height: between 1 and 10 meters
Distance from BS to MS: between 1 and 20 meters

For these assumptions, the model on the following page applies. Note that the model is used to calculate path loss. The path is converted to a dB scale and subtracted from the transmitted power expressed in dB.

Hata Model

The path loss (in dB) for urban areas is given in the Hata model as

$$L_{50}(urban) = 69.55 + 26.16\log_{10} f_c - 13.82\log_{10} h_{te} - a(h_{re}) + (44.9 - 6.55\log_{10} h_{te})\log_{10} d$$

For various environments we apply a correction factor for the mobile antenna height. For a small to medium size city

$$a(h_{re}) = (1.1\log_{10} f_c - 0,7)h_{re} - (1.56\log_{10} f_c - 0.8)$$

For a large city the correction factors take the form

$$a(h_{re}) = 8.29(\log_{10} 1.54 h_{re})^2 - 1.1, \qquad f_c < 300 \text{ MHz}$$

$$a(h_{re}) = 3.2(\log_{10} 11.75 h_{re})^2 - 4.97, \qquad f_c > 300 \text{ MHz}$$

Hata Model

For a suburban area the original expression is modified as

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

Finally for open rural areas we have

$$L_{50}(suburban) = L_{50}(urban) - 4.78\left[\log(f_c)\right]^2 + 18.33\log_{10}(f_c) - 40.94$$

Note that the Hata model is a formula and does not have the path specific graphical corrections available in the Okumura model.

Model Accuracy

As previously illustrated, a random variable may be added to account for random fluctuations due to shadowing.

Keep in mind that these models are not very precise and provide only very rough approximations. The approximations are useful however.

Empirical Path Loss: Okamura, Hata, COST231

- Empirical models include effects of path loss, shadowing and multipath.
  - Multipath effects are averaged over several wavelengths: local mean attenuation (LMA)
  - Empirical path loss for a given environment is the average of LMA at a distance d over all measurements
- **Okamura**: based upon Tokyo measurements. 1-100 lm, 150-1500MHz, base station heights (30-100m), median attenuation over free-space-loss, 10-14dB standard deviation.

$$P_L(d)\ \text{dB} = L(f_c, d) + A_{mu}(f_c, d) - G(h_t) - G(h_r) - G_{AREA}$$

- **Hata**: closed form version of Okamura

$$P_{L,urban}(d)\ \text{dB} = 69.55 + 26.16\log_{10}(f_c) - 13.82\log_{10}(h_t) - a(h_r) + (44.9 - 6.55\log_{10}(h_t))\log_{10}(d). \quad (2.31)$$

- **COST 231:** Extensions to 2 GHz

$$P_{L,urban}(d)\,\text{dB} = 46.3 + 33.9\log_{10}(f_c) - 13.82\log_{10}(h_t) - a(h_r) + (44.9 - 6.55\log_{10}(h_t))\log_{10}(d) + C_M, \quad (2.34)$$

Mobile Radio Propagation: Small-Scale Fading and Multipath

Introduction

- **Small-scale fading** is used to describe the rapid fluctuation of the amplitude of a radio signal over a short period of time or travel distance.

- **Fading is caused by** interference between two or more versions of the transmitted signal which arrive at the receiver at slightly different times.

- **Multipath waves** consists of a large number of plane waves having randomly distributed amplitudes, phases, and angles of arrival. that causes the signal to distort or fade.

Small-Scale Multipath Propagation

Multipath creates small-scale fading effects such as:

1. **Rapid changes in signal strength over a small travel distance or time interval**

2. **Random frequency modulation due to varying Doppler shifts on different multipath signals.**

3. **Time dispersion caused by multipath propagation delays.**

The physical factors leading to these effects are:

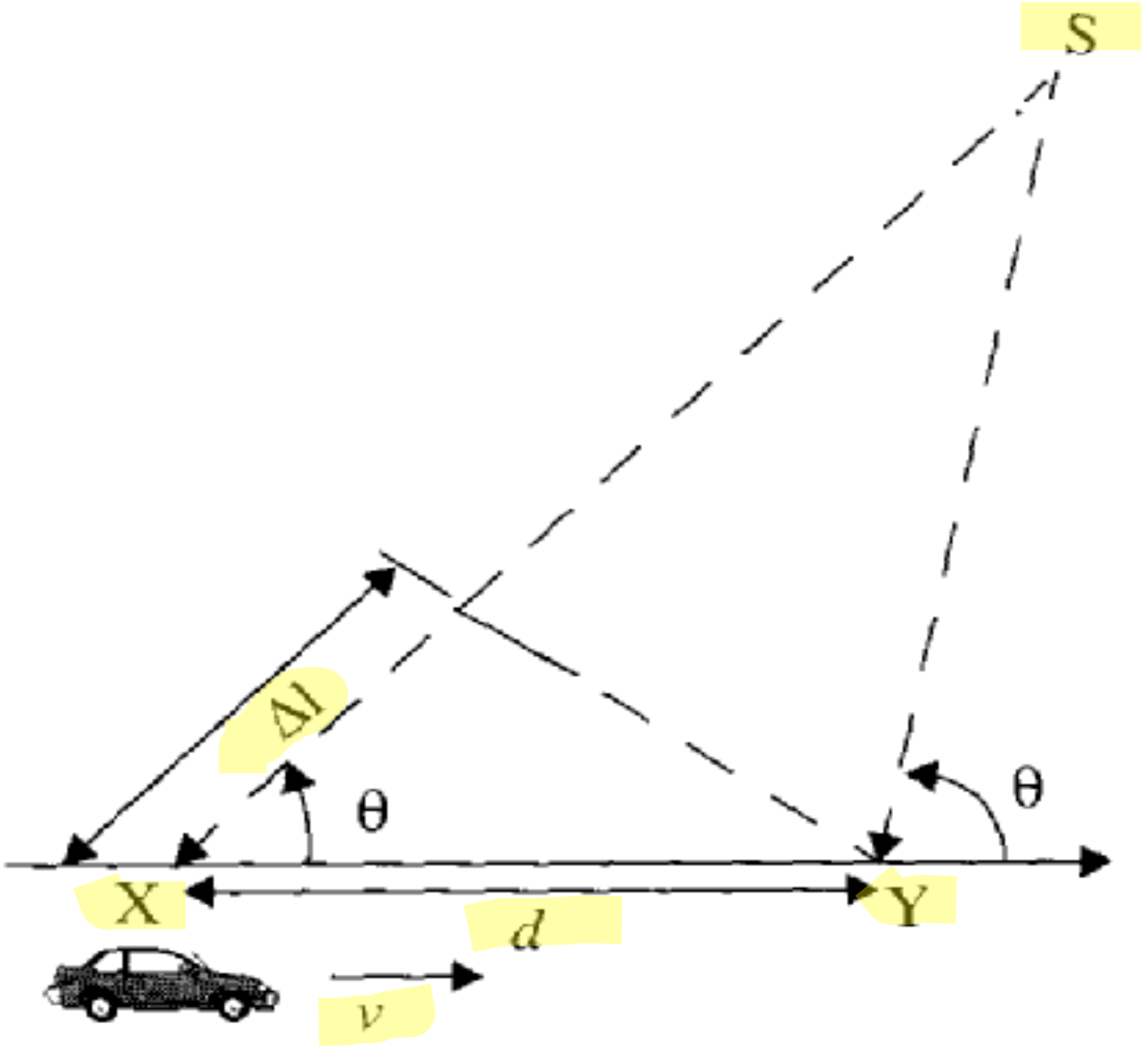
- ➔ The presence of objects in the propagation path (buildings, signs, trees and plants, fixed and moving vehicles, etc.) reflects and/or scatters the incident electromagnetic energy leading to multipath.
- ➔ Relative motion between the transmitter and receiver gives rise to doppler effects.
- ➔ Movement of surrounding objects also gives rise to doppler effects.
- ➔ The bandwidith of the channel, especially if the channel bandwidth is less than the signal bandwidth, leads to time dispersion.

The Doppler shift

- The shift in received signal frequency due to motion

- 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.

Illustration of Doppler effect



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 illustrated in Figure 4.1. The difference in path lengths traveled by the wave from source S to the mobile at points X and Y is $\Delta l = d\cos\theta = v\Delta t\cos\theta$, where Δt is the time required for the mobile to travel from X to Y, and θ is assumed to be the same at 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

$$\Delta\phi = \frac{2\pi\Delta l}{\lambda} = \frac{2\pi v\Delta t}{\lambda}\cos\theta \tag{4.1}$$

and hence the apparent change in frequency, or Doppler shift, is given by f_d, where

$$f_d = \frac{1}{2\pi}\cdot\frac{\Delta\phi}{\Delta t} = \frac{v}{\lambda}\cdot\cos\theta \tag{4.2}$$

- The equation 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.

- **The Doppler shift is positive** (i.e., the apparent received frequency is increased), if the mobile is moving toward the direction of arrival of the wave.

- **The Doppler shift is negative** (i.e. the apparent received frequency is decreased), if the mobile is moving away from the direction of arrival of the wave.

- multipath components from a CW signal which arrive from different directions contribute to Doppler spreading of the received signal, **thus increasing the signal bandwidth.**

Example 4.1

Consider a transmitter which radiates a sinusoidal carrier frequency of 1850 MHz. For a vehicle moving 60 mph, 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 to Example 4.1

Given:

Carrier frequency $f_c = 1850\,\text{MHz}$

Therefore, wavelength $\lambda = c/f_c = \dfrac{3 \times 10^8}{1850 \times 10^6} = 0.162\,\text{m}$

Vehicle speed $v = 60$ mph $= 26.82$ m/s

(a) The vehicle is moving directly towards the transmitter.

The Doppler shift in this case is positive and the received frequency is given by equation (4.2)

$$f = f_c + f_d = 1850 \times 10^6 + \frac{26.82}{0.162} = 1850.00016\ \text{MHz}$$

(b) The vehicle is moving directly away from the transmitter.

The Doppler shift in this case is negative and hence the received frequency is given by

$$f = f_c - f_d = 1850 \times 10^6 - \frac{26.82}{0.162} = 1849.999834 \text{ MHz}$$

(c) The vehicle is moving perpendicular to the angle of arrival of the transmitted signal.

In this case, $\theta = 90°$, $\cos\theta = 0$, and there is no Doppler shift.

The received signal frequency is the same as the transmitted frequency of 1850 MHz.

4.2 Impulse Response Model of a Multipath Channel

- To show that a mobile radio channel may be modeled as **a linear filter with a time varying impulse response**, consider the case where time variation is due strictly to receiver motion in space.

Figure 4.2
The mobile radio channel as a function of time and space.

In Figure 4.2, the receiver moves along the ground at some constant velocity v. For a fixed position d, the channel between the transmitter and the receiver can be modeled as a linear time invariant system. However, due to the different multipath waves which have propagation delays which vary over different spatial locations of the receiver, the impulse response of the linear time invariant channel should be a function of the position of the receiver. That is, the channel impulse response can be expressed as $h(d,t)$. Let $x(t)$ represent the transmitted signal, then the received signal $y(d,t)$ at position d can be expressed as a convolution of $x(t)$ with $h(d,t)$.

$$y(d,t) = x(t) \otimes h(d,t) = \int_{-\infty} x(\tau)h(d,t-\tau)d\tau \qquad (4.3)$$

For a causal system, $h(d,t) = 0$ for $t<0$, thus equation (4.3) reduces to

$$y(d,t) = \int_{-\infty}^{t} x(\tau)h(d,t-\tau)d\tau \qquad (4.4)$$

Since the receiver moves along the ground at a constant velocity v, the position of the receiver can by expressed as

$$d = vt \qquad (4.5)$$

Substituting (4.5) in (4.4), we obtain

t

Since v is a constant, $y(vt,t)$ is just a function of t. Therefore, equation (4.6) can be expressed as

$$y(t) = \int_{-\infty}^{t} x(\tau)h(vt,t-\tau)d\tau = x(t) \otimes h(vt,t) = x(t) \otimes h(d,t) \qquad (4.7)$$

From equation (4.7) it is clear that the mobile radio channel can be modeled as a linear time varying channel, where the channel changes with time and distance.

Since v may be assumed constant over a short time (or distance) interval, we may let $x(t)$ represent the transmitted bandpass waveform, $y(t)$ the received waveform, and $h(t, \tau)$ the impulse response of the time varying multipath radio channel. The impulse response $h(t, \tau)$ completely characterizes the channel and is a function of both t and τ. The variable t represents the time variations due to motion, whereas τ represents the channel multipath delay for a fixed value of t. One may think of τ as being a vernier adjustment of time. The received signal $y(t)$ can be expressed as a convolution of the transmitted signal $x(t)$ with the channel impulse response (see Figure 4.3a).

$$y(t) = \int_{-\infty}^{\infty} x(\tau)h(t, \tau)d\tau = x(t) \otimes h(t, \tau) \tag{4.8}$$

If the multipath channel is assumed to be a bandlimited bandpass channel, which is reasonable, then $h(t,\tau)$ may be equivalently described by a complex baseband impulse response $h_b(t, \tau)$ with the input and output being the complex envelope representations of the transmitted and received signals, respectively (see Figure 4.3b). That is,

$$r(t) = c(t) \otimes \frac{1}{2}h_b(t, \tau) \tag{4.9}$$

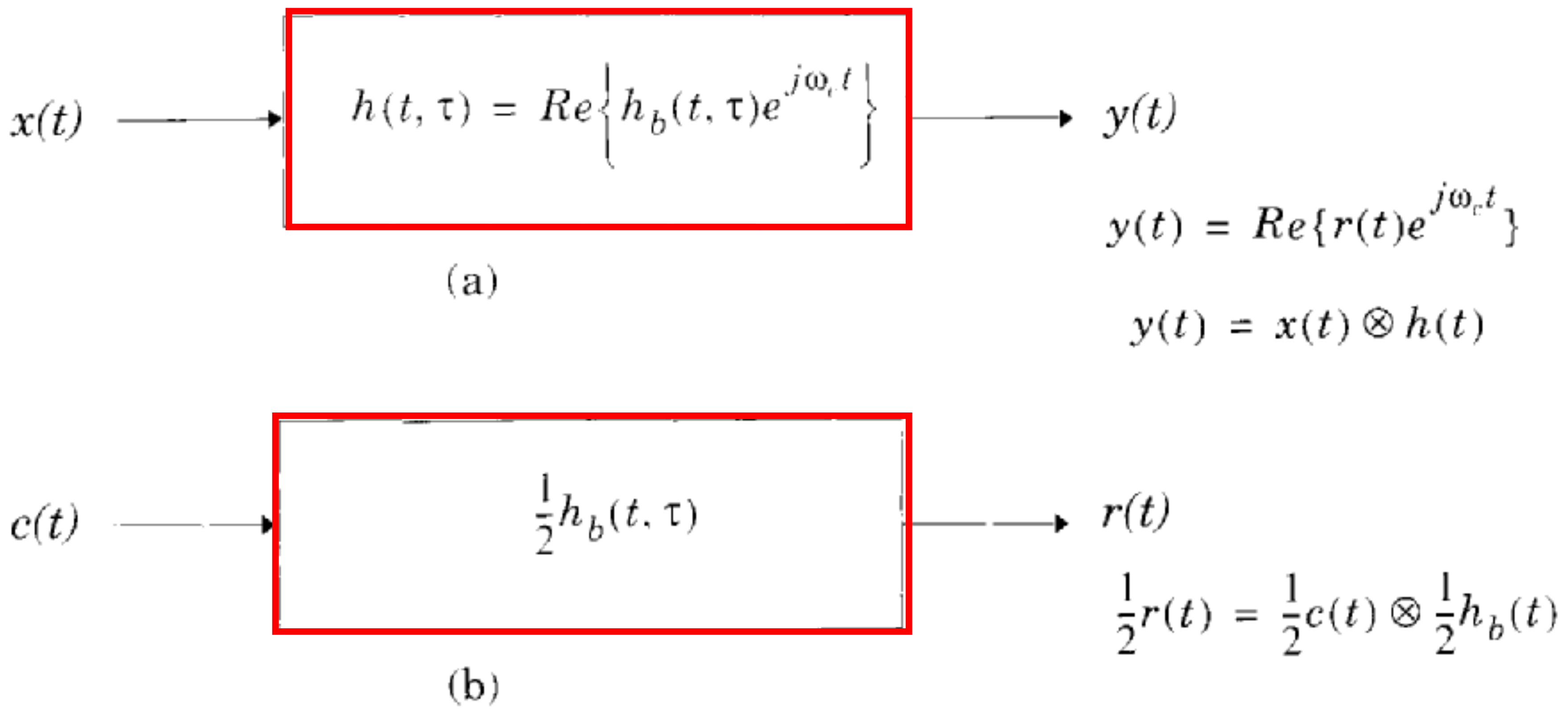


Figure 4.3
(a) Bandpass channel impulse response model.
(b) Baseband equivalent channel impulse response model.

where $c(t)$ and $r(t)$ are the complex envelopes of $x(t)$ and $y(t)$, defined as

$$x(t) = Re\{c(t)\exp(j2\pi f_c t)\} \tag{4.10}$$

$$y(t) = Re\{r(t)\exp(j2\pi f_c t)\} \tag{4.11}$$

The factor of $1/2$ in equation (4.9) is due to the properties of the complex envelope, in order to represent the passband radio system at baseband. The low-pass characterization removes the high frequency variations caused by the carrier, making the signal analytically easier to handle. It is shown by Couch [Cou93] that the average power of a bandpass signal $\overline{x^2(t)}$ is equal to $\frac{1}{2}\overline{|c(t)|^2}$, where the overbar denotes ensemble average for a stochastic signal, or time average for a deterministic or ergodic stochastic signal.

It is useful to discretize the multipath delay axis τ of the impulse response into equal time delay segments called *excess delay bins*, where each bin has a time delay width equal to $\tau_{i+1} - \tau_i$, where τ_0 is equal to 0, and represents the first arriving signal at the receiver. Letting i = 0, it is seen that $\tau_1 - \tau_0$ is equal to the time delay bin width given by $\Delta\tau$. For convention, $\tau_0 = 0$, $\tau_1 = \Delta\tau$, and $\tau_i = i\Delta\tau$, for $i = 0$ to $N-1$, where N represents the total number of possible equally-spaced multipath components, including the first arriving component. Any number of multipath signals received within the i th bin are represented by a single resolvable multipath component having delay τ_i. This technique of quantizing the delay bins determines the time delay resolution of the channel model, and the useful frequency span of the model can be shown to be $1/(2\Delta\tau)$. That is, the model may be used to analyze transmitted signals having bandwidths which are less than $1/(2\Delta\tau)$. Note that $\tau_0 = 0$ is the excess time delay of the first arriving multipath component, and neglects the propagation delay between the transmitter and receiver. *Excess delay* is the relative delay of the i th multipath component as compared to the first arriving component and is given by τ_i. The *maximum excess delay* of the channel is given by $N\Delta\tau$.

Since the received signal in a multipath channel consists of a series of attenuated, time-delayed, phase shifted replicas of the transmitted signal, the baseband impulse response of a multipath channel can be expressed as

$$h_b(t, \tau) = \sum_{i=0}^{N-1} a_i(t, \tau)\exp[j(2\pi f_c\tau_i(t) + \phi_i(t, \tau))]\delta(\tau - \tau_i(t)) \tag{4.12}$$

where $a_i(t, \tau)$ and $\tau_i(t)$ are the real amplitudes and excess delays, respectively, of i th multipath component at time t [Tur72]. The phase term $2\pi f_c\tau_i(t) + \phi_i(t, \tau)$ in (4.12) represents the phase shift due to free space propagation of the i th multipath component, plus any additional phase shifts which are encountered in the channel. In general, the phase term is simply represented by encountered in the channel. In general, the phase term is simply represented by a single variable $\theta_i(t, \tau)$ which lumps together all the mechanisms for phase shifts of a single multipath component within the ith excess delay bin. Note that some excess delay bins may have no multipath at some time t and delay τ_i, since $a_i(t, \tau)$ may be zero. In equation (4.12), N is the total possible number of multipath components (bins), and $\delta(\bullet)$ is the unit impulse function which determines the specific multipath bins that have components at time t and excess delays τ_i. Figure 4.4 illustrates an example of different snapshots of $h_b(t, \tau)$, where t varies into the page, and the time delay bins are quantized to widths of $\Delta\tau$.

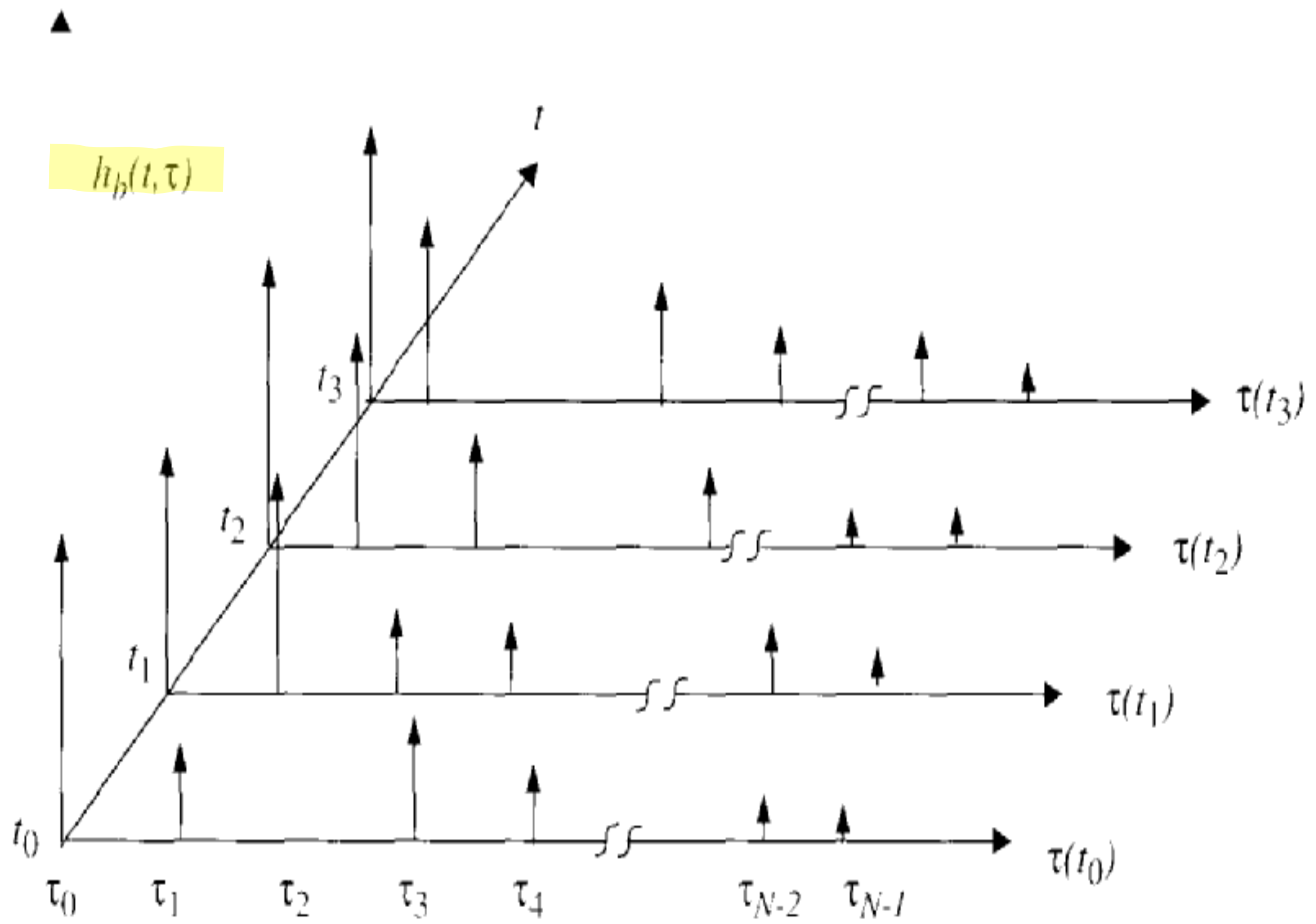


Figure 4.4

An example of the time varying discrete-time impulse response model for a multipath radio channel.

If the channel impulse response is assumed to be time invariant, or is at least wide sense stationary over a small-scale time or distance interval, then the channel impulse response may be simplified as

$$h_b(\tau) = \sum_{i=0}^{N-1} a_i \exp(-j\theta_i)\,\delta(\tau - \tau_i) \tag{4.13}$$

When measuring or predicting $h_b(\tau)$, a probing pulse $p(t)$ which approximates a delta function is used at the transmitter. That is,

$$p(t) \approx \delta(t - \tau) \tag{4.14}$$

is used to sound the channel to determine $h_b(\tau)$.

Po

For small-scale channel modeling, the *power delay profile* of the channel is found by taking the spatial average of $|h_b(t;\tau)|^2$ over a local area. By making several local area measurements of $|h_b(t;\tau)|^2$ in different locations, it is possible to build an ensemble of power delay profiles, each one representing a possible small-scale multipath channel state [Rap91a].

Based on work by Cox [Cox72], [Cox75], if $p(t)$ has a time duration much smaller than the impulse response of the multipath channel, $p(t)$ does not need to be deconvolved from the received signal $r(t)$ in order to determine relative multipath signal strengths. The received power delay profile in a local area is given by

$$P(t;\tau) \approx k|h_b(t;\tau)|^2 \tag{4.15}$$

and many snapshots of $|h_b(t;\tau)|^2$ are typically averaged over a local (small-scale) area to provide a single time-invariant multipath power delay profile $P(\tau)$. The gain k in equation (4.15) relates the transmitted power in the probing pulse $p(t)$ to the total power received in a multipath delay profile.

4.4 Parameters of Mobile Multipath Channels

- Many multipath channel parameters are derived from the 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 a quarter of a wavelength and over receiver movements no greater than 6 m in outdoor channels and no greater than 2 m in indoor channels in the 450 MHz - 6 GHz range. This small-scale sampling avoids averaging bias in the resulting small-scale statistics.

- Figure 4.9 shows typical power delay profile plots from outdoor and indoor channels, determined from a large number of closely sampled instantaneous profiles.

4.4.1 Time Dispersion Parameters

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

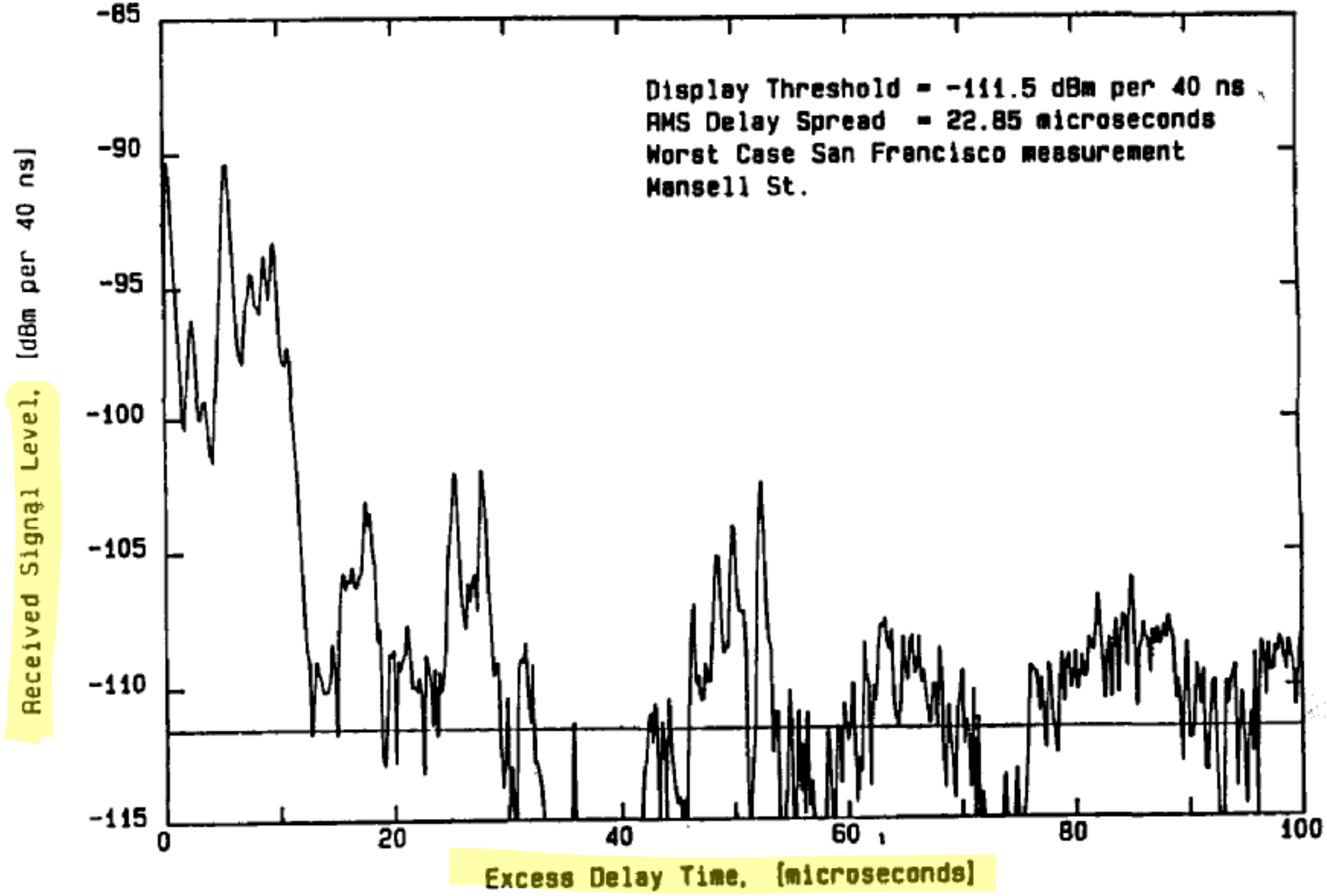
$$\bar{\tau} = \frac{\sum_k a_k^2 \tau_k}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k)\tau_k}{\sum_k P(\tau_k)} \tag{4.35}$$

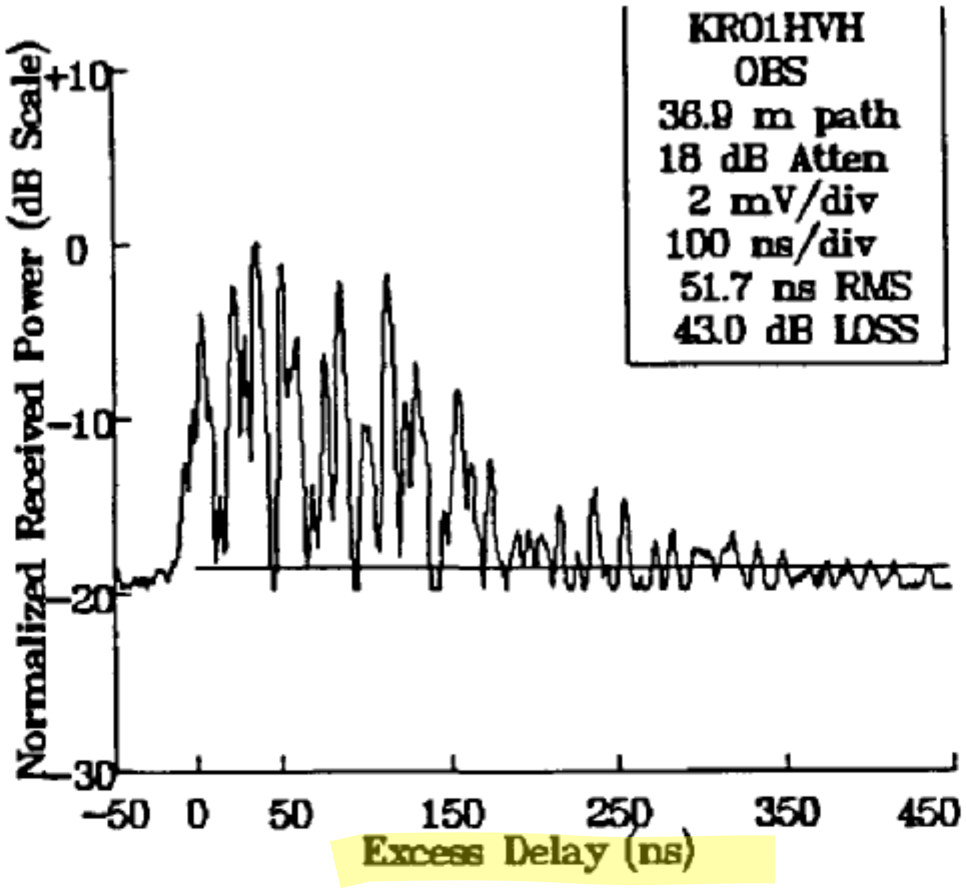
The rms delay spread is the square root of the second central moment of the power delay profile and is defined to be

$$\sigma_\tau = \sqrt{\overline{\tau^2} - (\bar{\tau})^2} \tag{4.36}$$

where

$$\overline{\tau^2} = \frac{\sum_k a_k^2 \tau_k^2}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k)\tau_k^2}{\sum_k P(\tau_k)} \tag{4.37}$$


Display Threshold = -111.5 dBm per 40 ns
RMS Delay Spread = 22.85 microseconds
Worst Case San Francisco measurement
Mansell St.
-85
-90
-95
-100
-105
-110
-115
0
20
40
60
80
100
Received Signal Level, [dBm per 40 ns]
Excess Delay Time, [microseconds]



(b)

Figure 4.9

Measured multipath power delay profiles

a) From a 900 MHz cellular system in San Francisco [From [Rap90] © IEEE].

b) Inside a grocery store at 4 GHz [From [Haw91] © IEEE].

Timer Dispersion Parameters

Determined from a power delay profile.

Mean excess delay( ): $\bar{\tau}$

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

Rms delay spread (σ_τ):

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

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

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 delay 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 arrive at τ_0). Figure 4.10 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 a threshold that relates the multipath noise floor to the maximum received multipath component.

- Table 4.1 shows the typical measured values of rms delay spread.

- Typical 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.

- 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.

Table 4.1 Typical Measured Values of RMS Delay Spread

Environment	Frequency (MHz)	RMS Delay Spread (σ_τ)	Notes	Reference
Urban	910	1300 ns avg. 600 ns st. dev. 3500 ns max.	New York City	[Cox75]
Urban	892	10-25 μs	Worst case San Francisco	[Rap90]
Suburban	910	200-310 ns	Averaged typical case	[Cox72]
Suburban	910	1960-2110 ns	Averaged extreme case	[Cox72]
Indoor	1500	10-50 ns 25 ns median	Office building	[Sal87]
Indoor	850	270 ns max.	Office building	[Dev90a]
Indoor	1900	70-94 ns avg. 1470 ns max.	Three San Francisco buildings	[Sei92a]

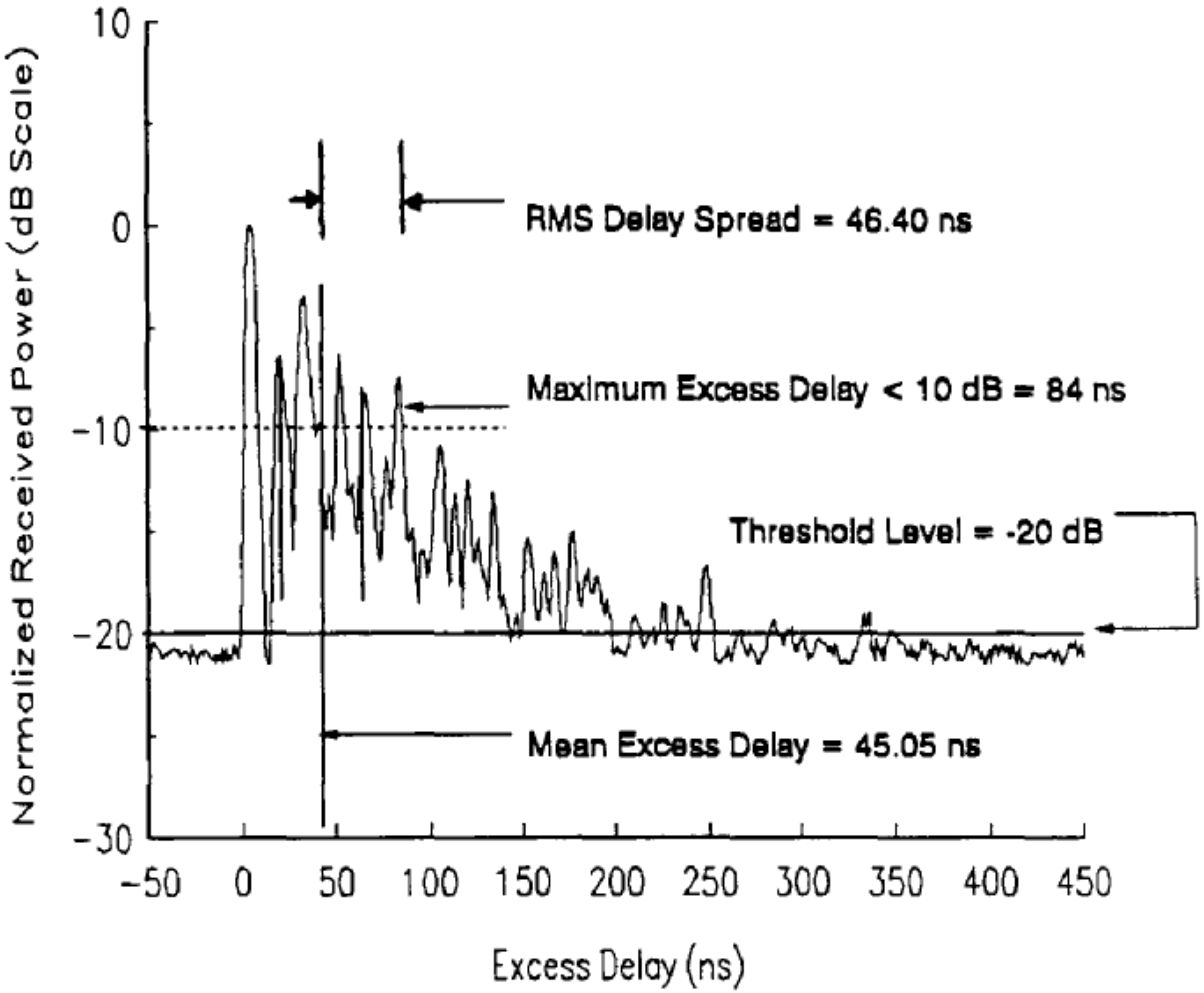
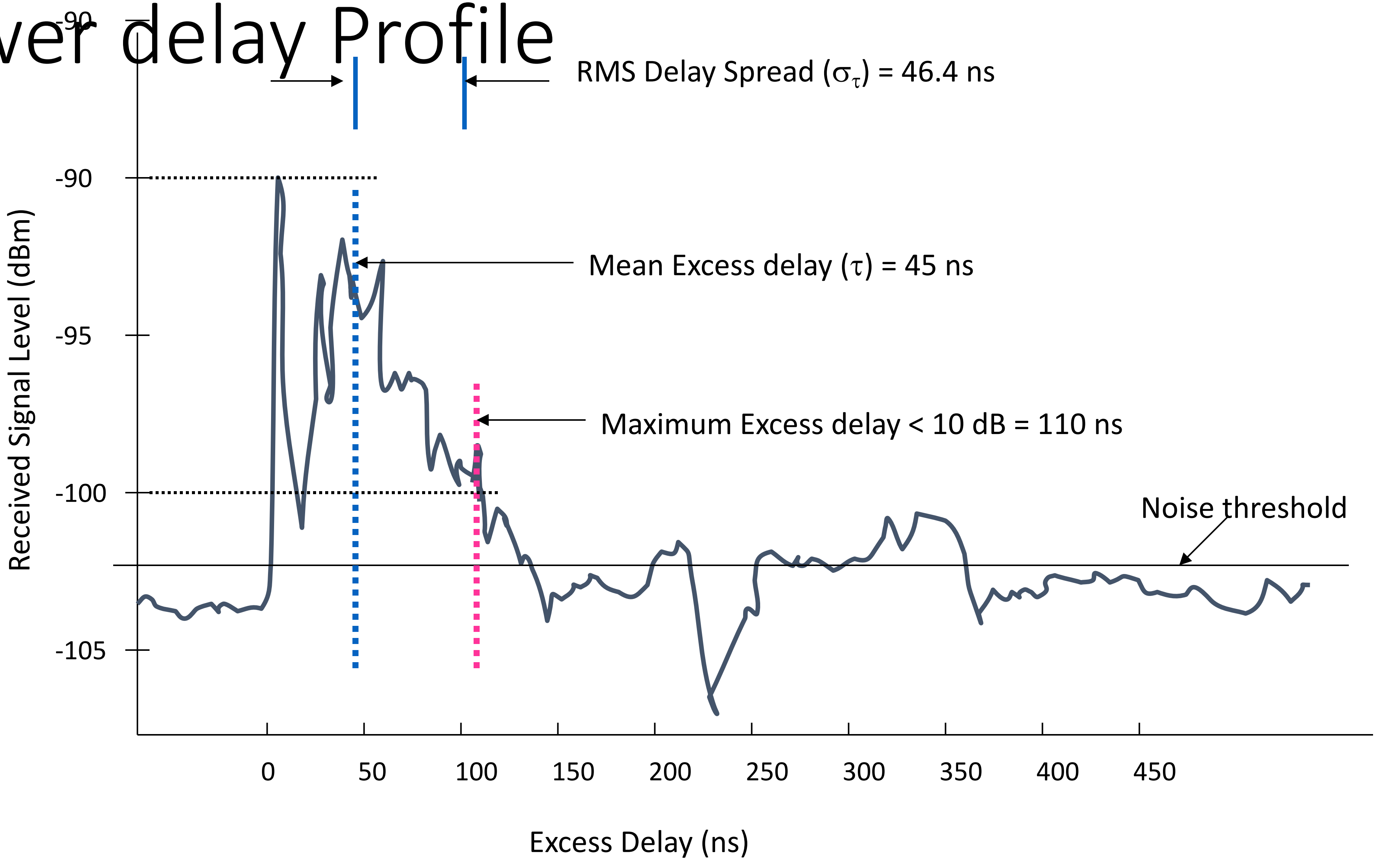


Figure 4.10
Example of an indoor power delay profile; rms delay spread, mean excess delay, maximum excess delay (10 dB), and threshold level are shown.

Power delay Profile



Example 4.4

Calculate the mean excess delay, rms delay spread, and the maximum excess delay (10 dB) for the multipath profile given in the figure below. Estimate the 50% coherence bandwidth of the channel. Would this channel be suitable for AMPS or GSM service without the use of an equalizer?

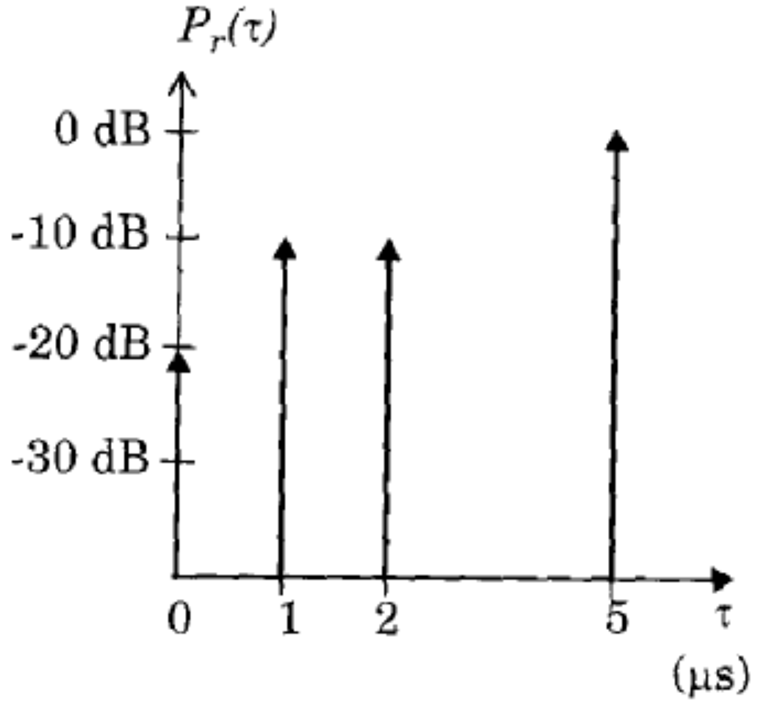


Figure E4.4

Solution to Example 4.4

The rms delay spread for the given multipath profile can be obtained using equations (4.35) — (4.37). The delays of each profile are measured relative to the first detectable signal. The mean excess delay for the given profile

$$\bar{\tau} = \frac{(1)(5) + (0.1)(1) + (0.1)(2) + (0.01)(0)}{[0.01 + 0.1 + 0.1 + 1]} = 4.38\mu s$$

The second moment for the given power delay profile can be calculated as

$$\overline{\tau^2} = \frac{(1)(5)^2 + (0.1)(1)^2 + (0.1)(2)^2 + (0.01)(0)}{1.21} = 21.07\mu s^2$$

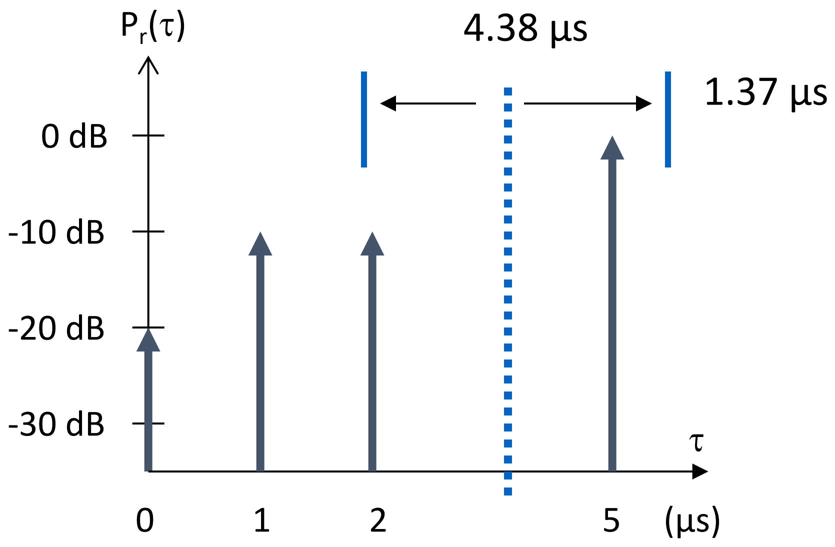
Therefore the rms delay spread, $\sigma_\tau = \sqrt{21.07 - (4.38)^2} = 1.37\mu s$

The coherence bandwidth is found from equation (4.39) to be

$$B_c \approx \frac{1}{5\sigma_\tau} = \frac{1}{5(1.37\mu s)} = 146\ \text{kHz}$$

Since B_c is greater than 30 kHz, AMPS will work without an equalizer. However, GSM requires 200 kHz bandwidth which exceeds B_c, thus an equalizer would be needed for this channel.

Example (Power delay profile)

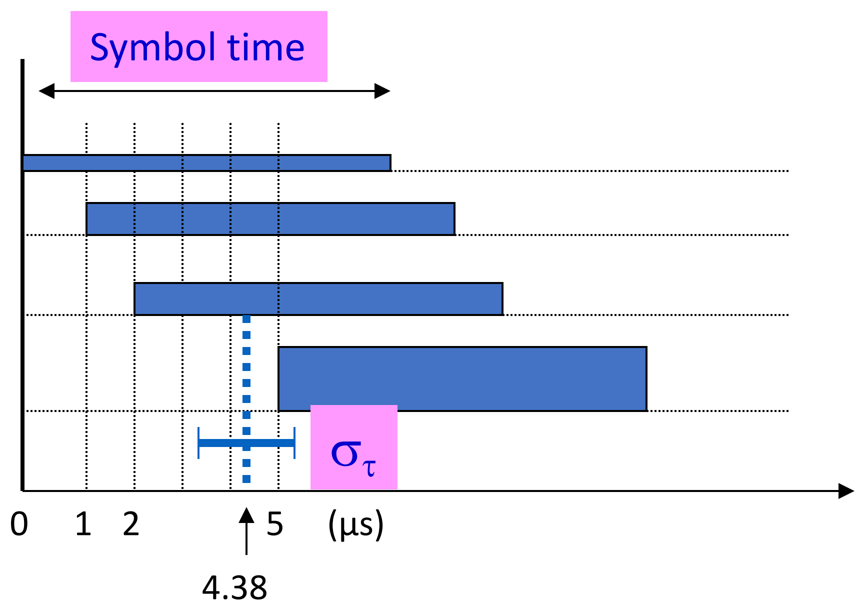


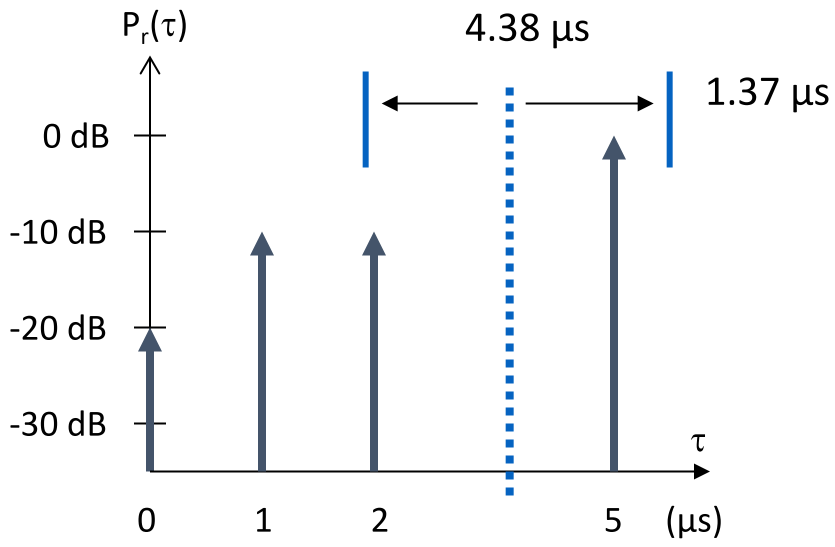
$$\bar{\tau} = \frac{(1)(5)+(0.1)(1)+(0.1)(2)+(0.01)(0)}{[0.01+0.1+0.1+1]} = 4.38\mu s$$

$$\bar{\tau^2} = \frac{(1)(5)^2+(0.1)(1)^2+(0.1)(2)^2+(0.01)(0)^2}{[0.01+0.1+0.1+1]} = 21.07\mu s^2$$

$$\sigma_\tau = \sqrt{21.07-(4.38)^2} = 1.37\mu s$$

Inter Symbol Interference





Symbol time > 10* σ_τ --- No equalization required

Symbol time < 10* σ_τ --- Equalization will be required to deal with ISI

In the above example, symbol time should be more than 14µs to avoid ISI.
This means that link speed must be less than 70Kbps (approx)

2.5 Multipath Delay Spread

Multipath occurs when signals arrive at the receiver directly from the transmitter and, indirectly, due to transmission through objects or reflection. The amount of signal reflection depends on factors such as angle of arrival, carrier frequency, and polarization of incident wave. Because the path lengths are different between the direct path and the reflected path(s), different signal paths could arrive at the receiver at different times over different distances. Figure 2.3 illustrates the concept. An impulse is transmitted at time 0; assuming that there are a multitude of reflected paths present, a receiver approximately 1 km away should detect a series of pulses, or *delay spread.*

If the time difference Δt is significant compared to one symbol period, *intersymbol interference* (ISI) can occur. In other words, symbols arriving significantly earlier or later than their own symbol periods can corrupt preceding or trailing symbols. For a fixed-path difference and a given delay spread, a higher data rate system is more likely to suffer ISI due to delay spread. For a fixed data rate system, a propagation environment with longer path differences (and thus higher delay spread) is more likely to cause ISI.

By: Dr.Mohab Mangoud

2. Coherence Bandwidth

- While the delay spread is a natural phenomenon caused by multipaths in the radio channel, the coherence bandwidth, **B**, 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).

- **Two sinusoids with frequency separation greater than B 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

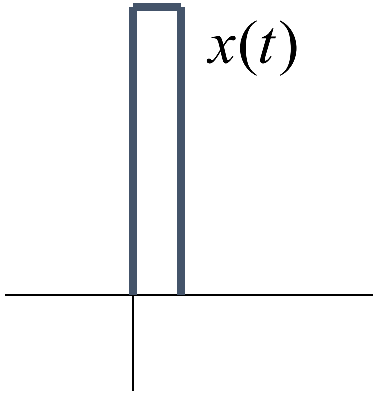
$$B_c \approx \frac{1}{50\sigma_\tau} \tag{4.38}$$

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

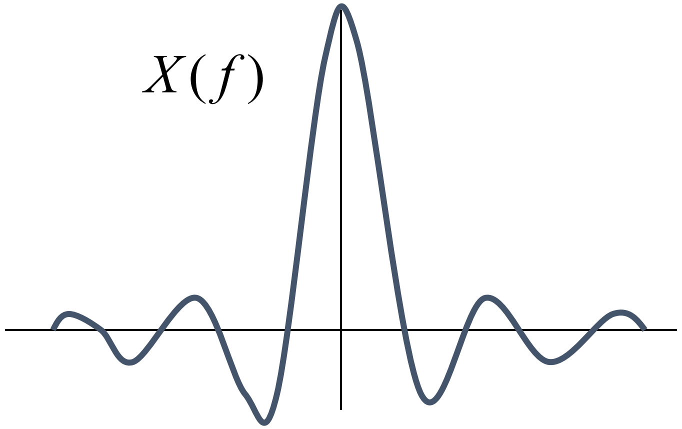
$$B_c \approx \frac{1}{5\sigma_\tau} \tag{4.39}$$

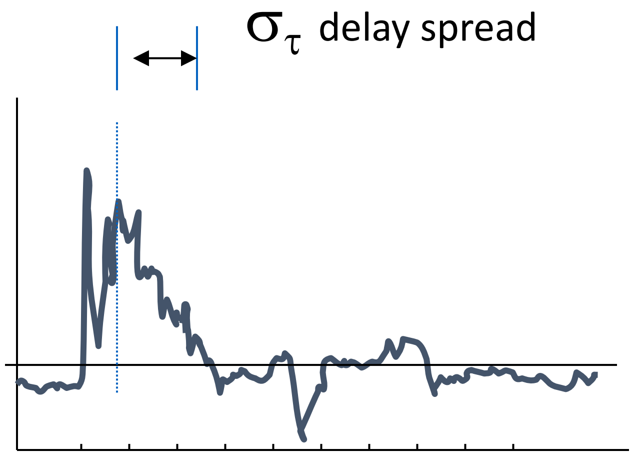
Coherence Bandwidth

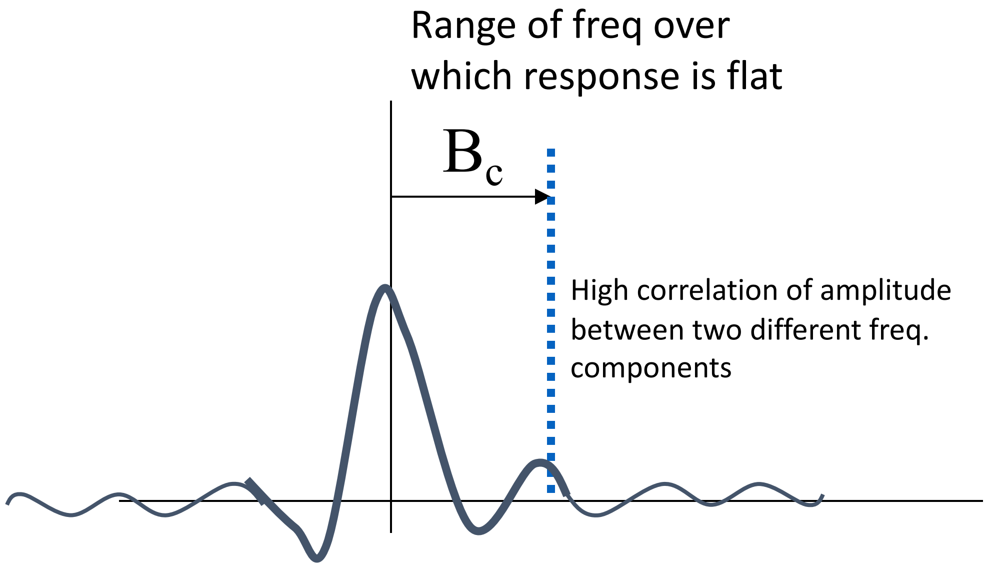
Time domain view



Freq. domain view







RMS delay spread and coherence BW

$$B_c \alpha \frac{1}{\sigma_\tau}$$

- RMS delay spread and coherence b/w (B_c) are inversely proportional

The picture can't be displayed.

$$B_c \approx \frac{1}{50.\sigma_\tau}$$

For 0.9 correlation

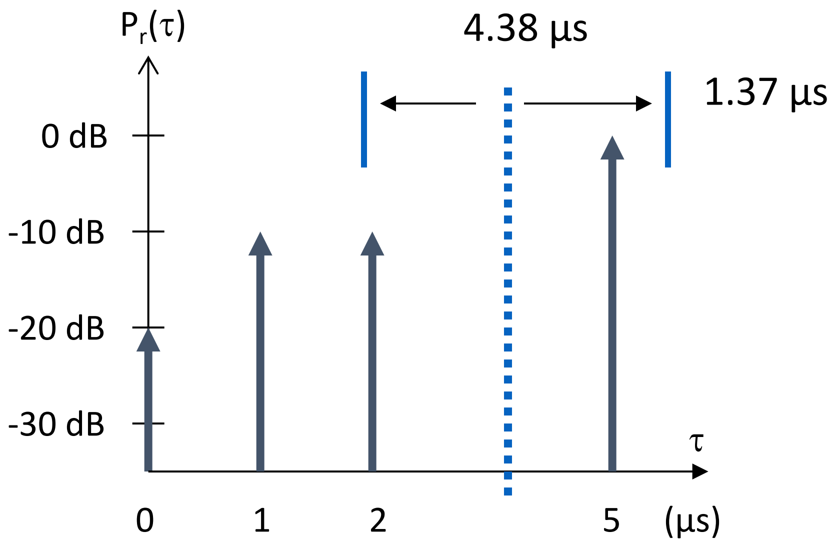
$$B_c \approx \frac{1}{5.\sigma_\tau}$$

For 0.5 correlation

Coherence Bandwidth

- Example:
  - For a multipath channel, σ is given as 1.37μs.
  - The 50% coherence bandwidth is given as: $1/5\sigma$ = 146kHz.
    - This means that, for a good transmission from a transmitter to a receiver, the range of transmission frequency (channel bandwidth) should not exceed 146kHz, so that all frequencies in this band experience the same channel characteristics.
    - Equalizers are needed in order to use transmission frequencies that are separated larger than this value.
    - This coherence bandwidth is enough for an AMPS channel (30kHz band needed for a channel), but is not enough for a GSM channel (200kHz needed per channel).

Revisit Example (Power delay profile)



$$\bar{\tau} = 4.38 \mu s$$

$$\sigma_\tau = 1.37 \mu s$$

$$\bar{\tau^2} = 21.07 \mu s^2$$

$$(50\% - coherence) B_c \approx \frac{1}{5.\sigma_\tau} = 146 kHz$$

Signal bandwidth for Analog Cellular = 30 KHz
Signal bandwidth for GSM = 200 KHz

Doppler spread and coherence time

- 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 or the 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 and coherence time

Coherence time definition implies that two signals arriving with a time separation greater than T_C are affected differently by the channel.

- Doppler spread and coherence time (T_c) are inversely proportional

The picture can't be displayed.

$$T_c \alpha \frac{1}{f_m}$$

f_m is the max doppler shift

$$T_c \approx \frac{9}{16\pi f_m}$$

For 0.5 correlation

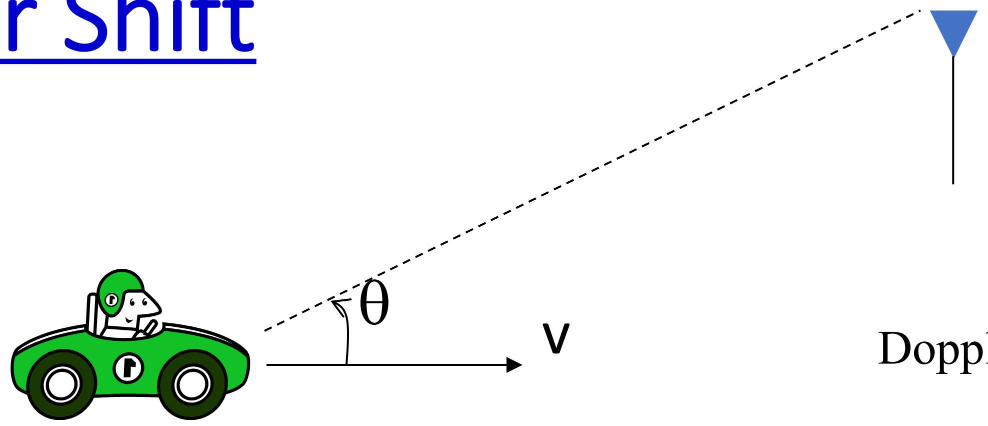
$$T_c \approx \frac{0.423}{f_m}$$

Rule of thumb

Doppler Spread

- Measure of spectral broadening caused by motion
- We know how to compute Doppler shift: f_d
- Doppler spread: B_D, is defined as the maximum Doppler shift: $f_m = v/\lambda$
- If the baseband signal bandwidth (B_S) is much greater than (B_D) then effect of Doppler spread is negligible at the receiver.

Doppler Shift



Doppler shift $\Delta f = \dfrac{v\cos\theta}{\lambda}$

Example

- Carrier frequency f_c = 1850 MHz (i.e. λ = 16.2 cm)
- Vehicle speed v = 60 mph = 26.82 m/s

- If the vehicle is moving directly towards the transmitter

$$\Delta f = \frac{26.82}{0.162} = 165 Hz$$

- If the vehicle is moving perpendicular to the angle of arrival of the transmitted signal

$$\Delta f = 0$$

Types of Small-Scale Fading

The 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.

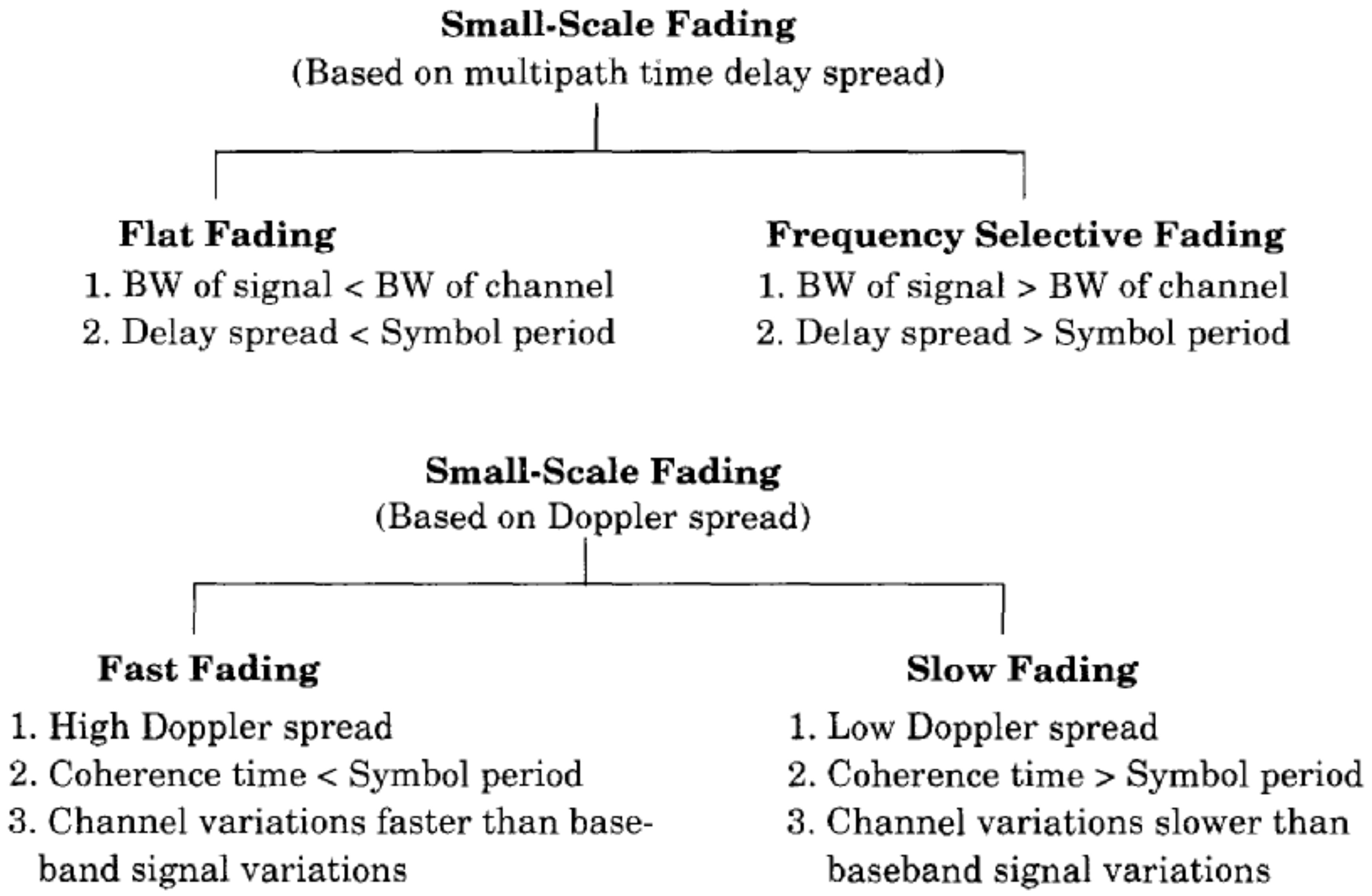
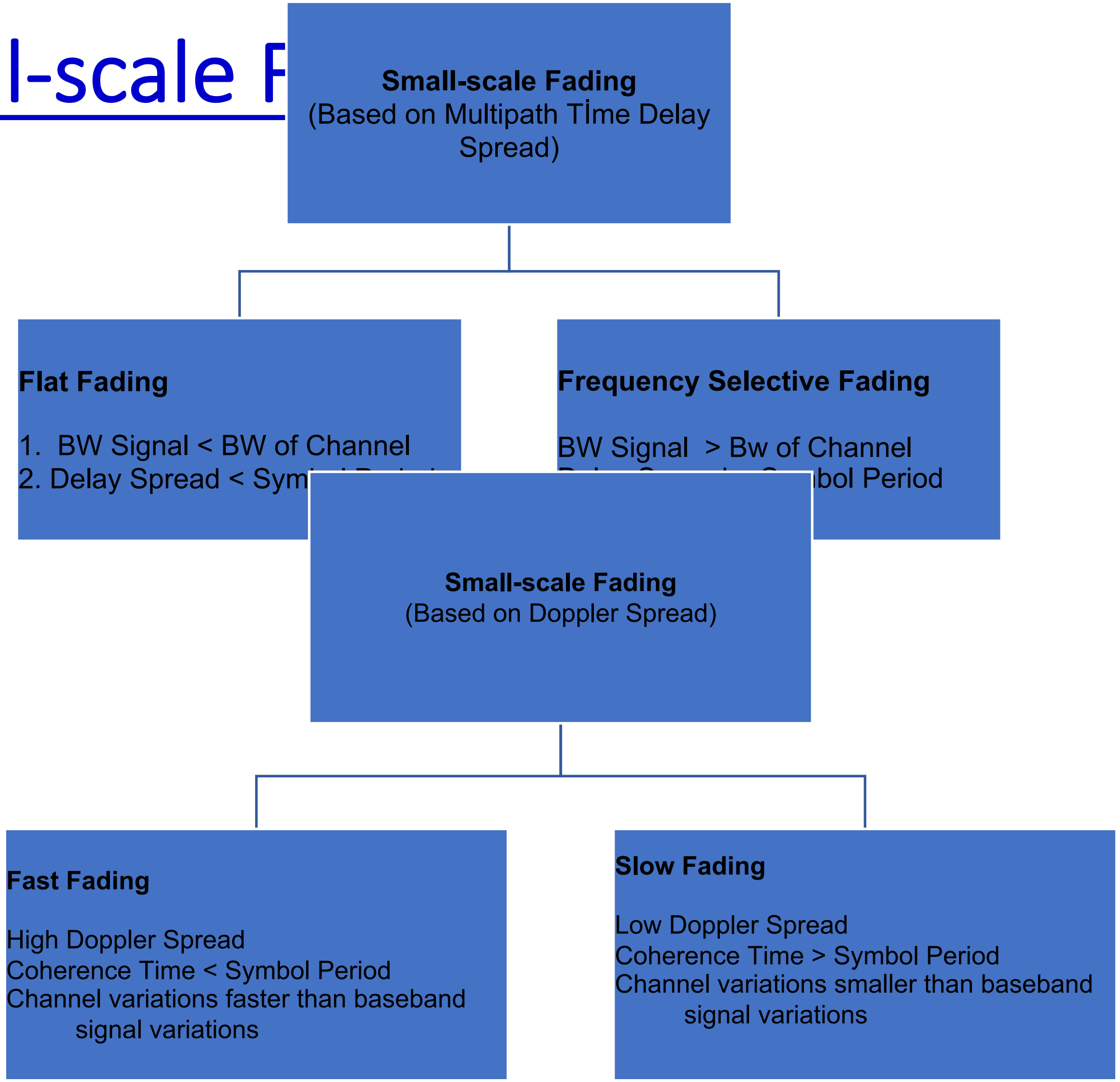
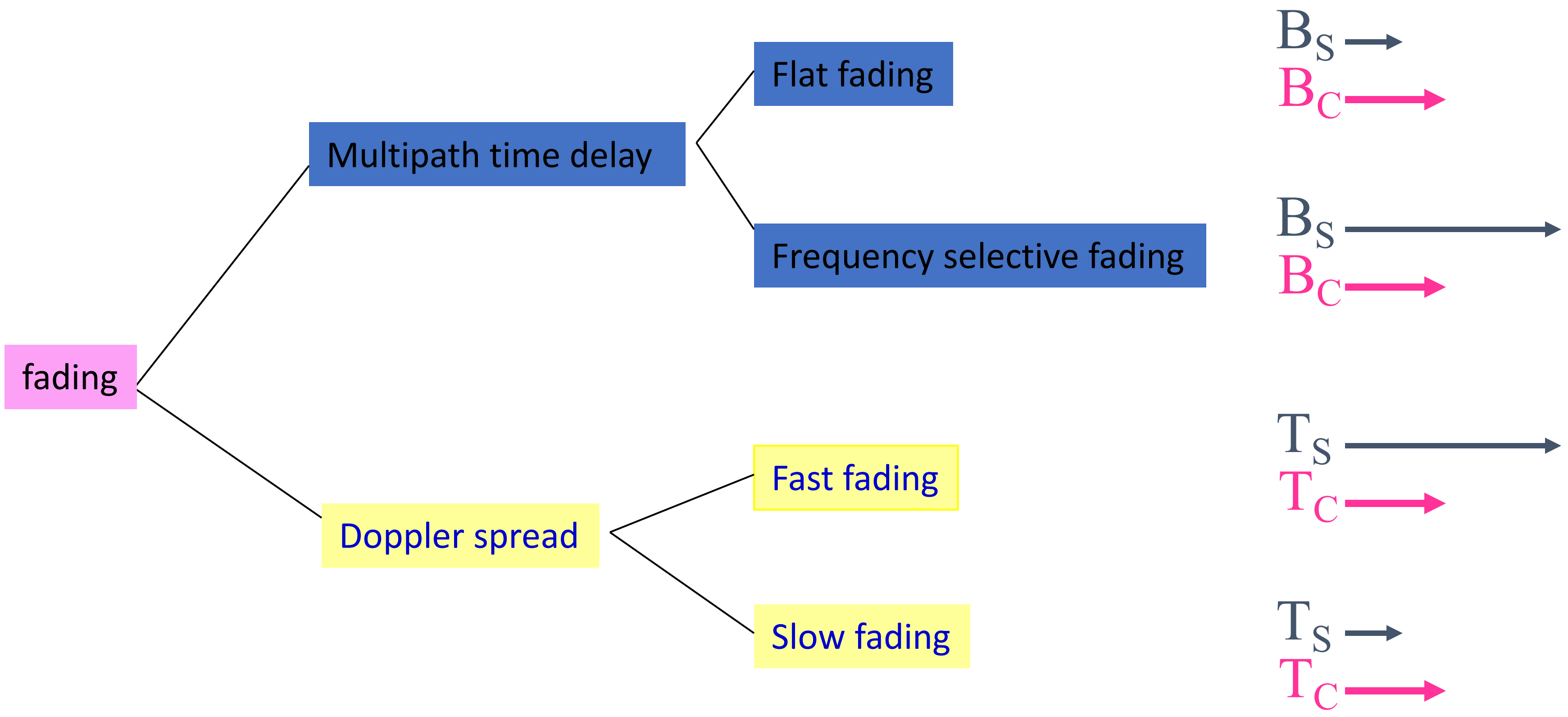


Figure 4.11
Types of small-scale fading.

Types of Small-scale F



Small scale fading



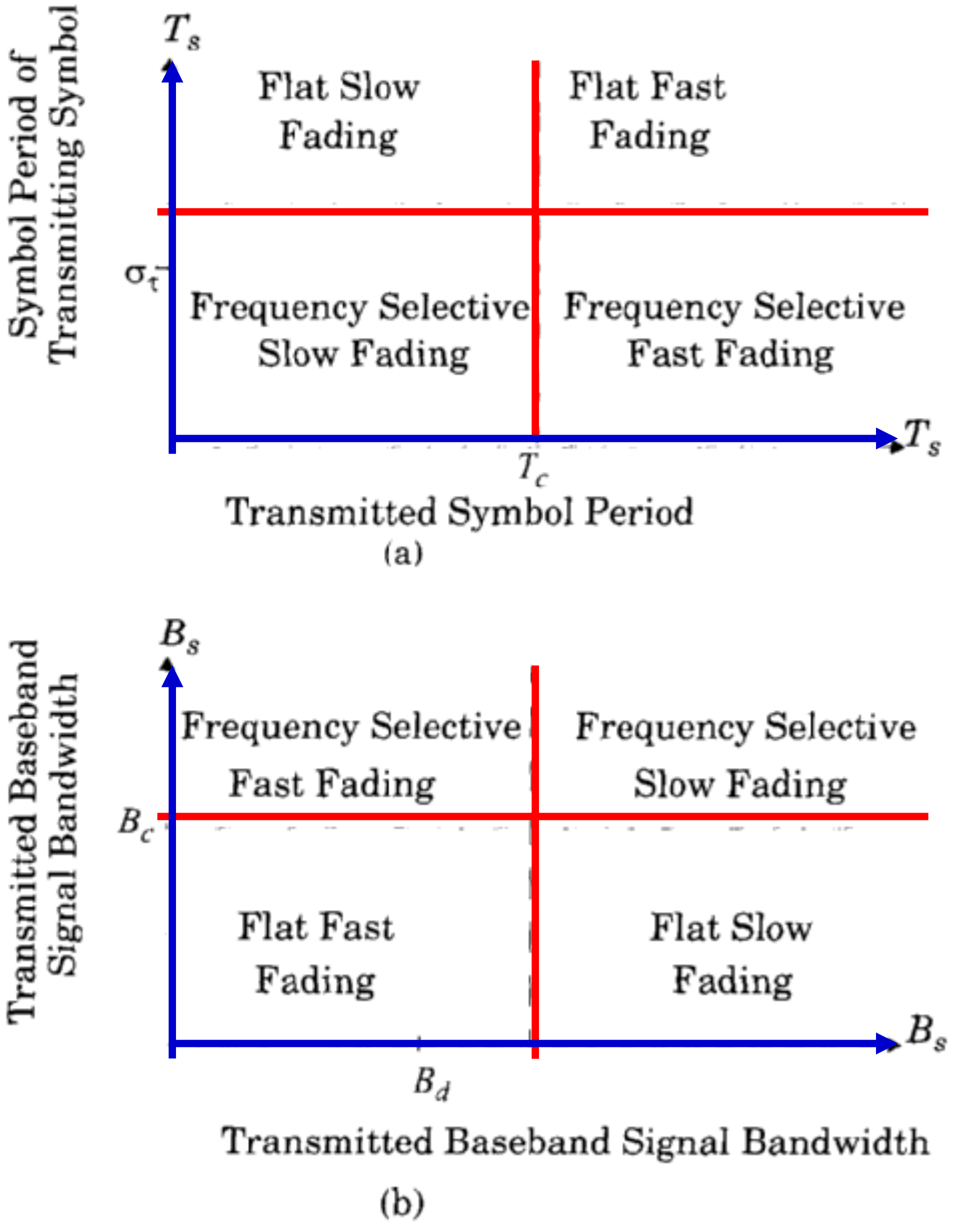


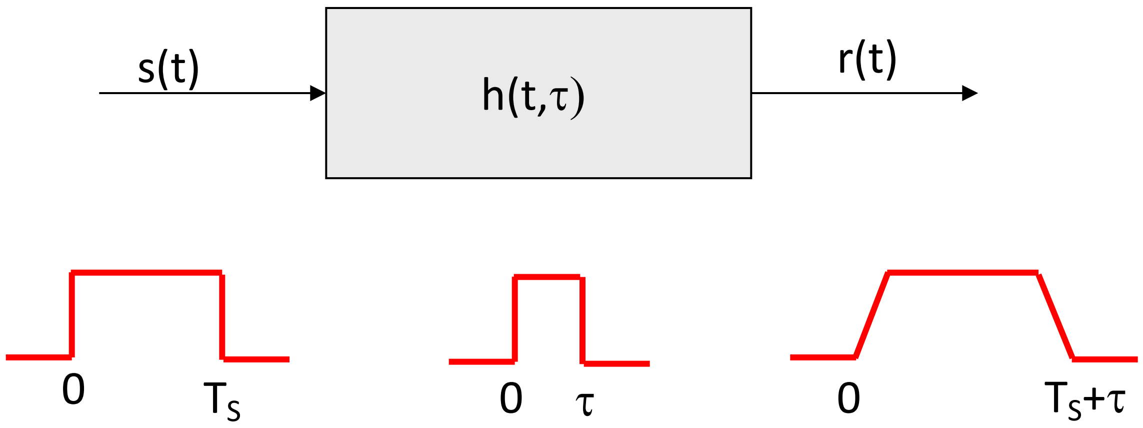
Figure 4.14

Matrix illustrating type of fading experienced by a signal as a function of
(a) symbol period
(b) baseband signal bandwidth.

Flat Fading

- Occurs when the amplitude of the received signal changes with time
  - For example according to Rayleigh Distribution
- 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.
- May cause deep fades.
  - Increase the transmit power to combat this situation.

Flat Fading



$\tau << T_S$

Occurs when:
$B_S << B_C$
and
$T_S >> \sigma_\tau$

B_C: Coherence bandwidth
B_S: Signal bandwidth
T_S: Symbol period
σ_τ: Delay Spread

Flat Fading

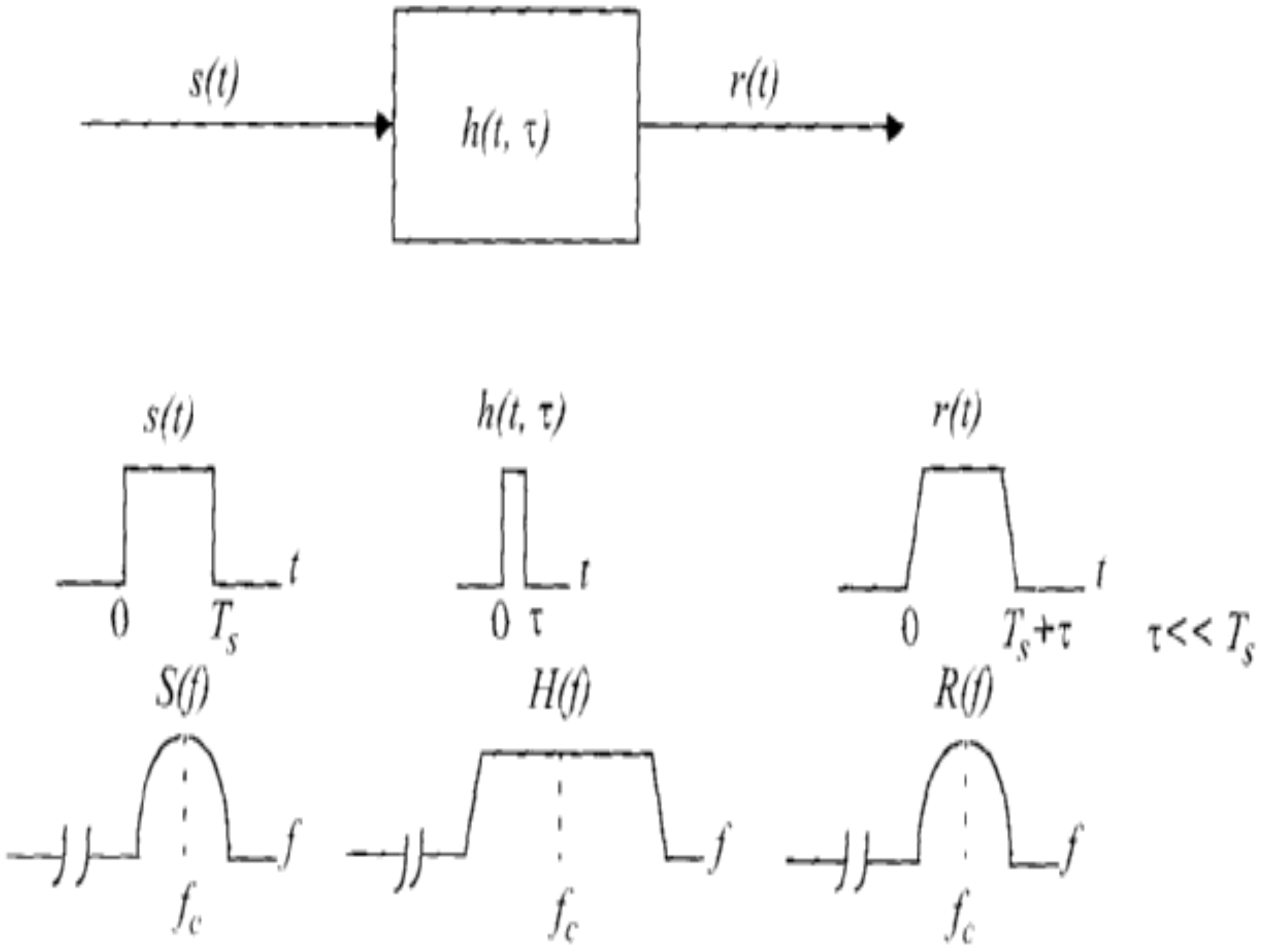
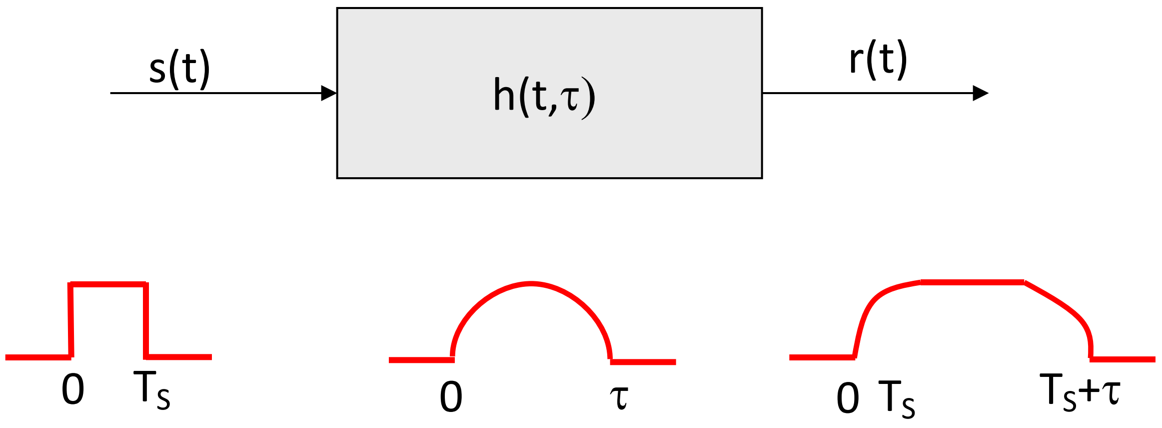


Figure 4.12
Flat fading channel characteristics.

Frequency Selective Fading

- 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.

Frequency Selective Fading



$\tau >> T_S$

Causes distortion of the received baseband signal

Causes Inter-Symbol Interference (ISI)

Occurs when:
$B_S > B_C$
and
$T_S < \sigma_\tau$

As a rule of thumb: $T_S < \sigma_\tau$

Frequency Selective Fading

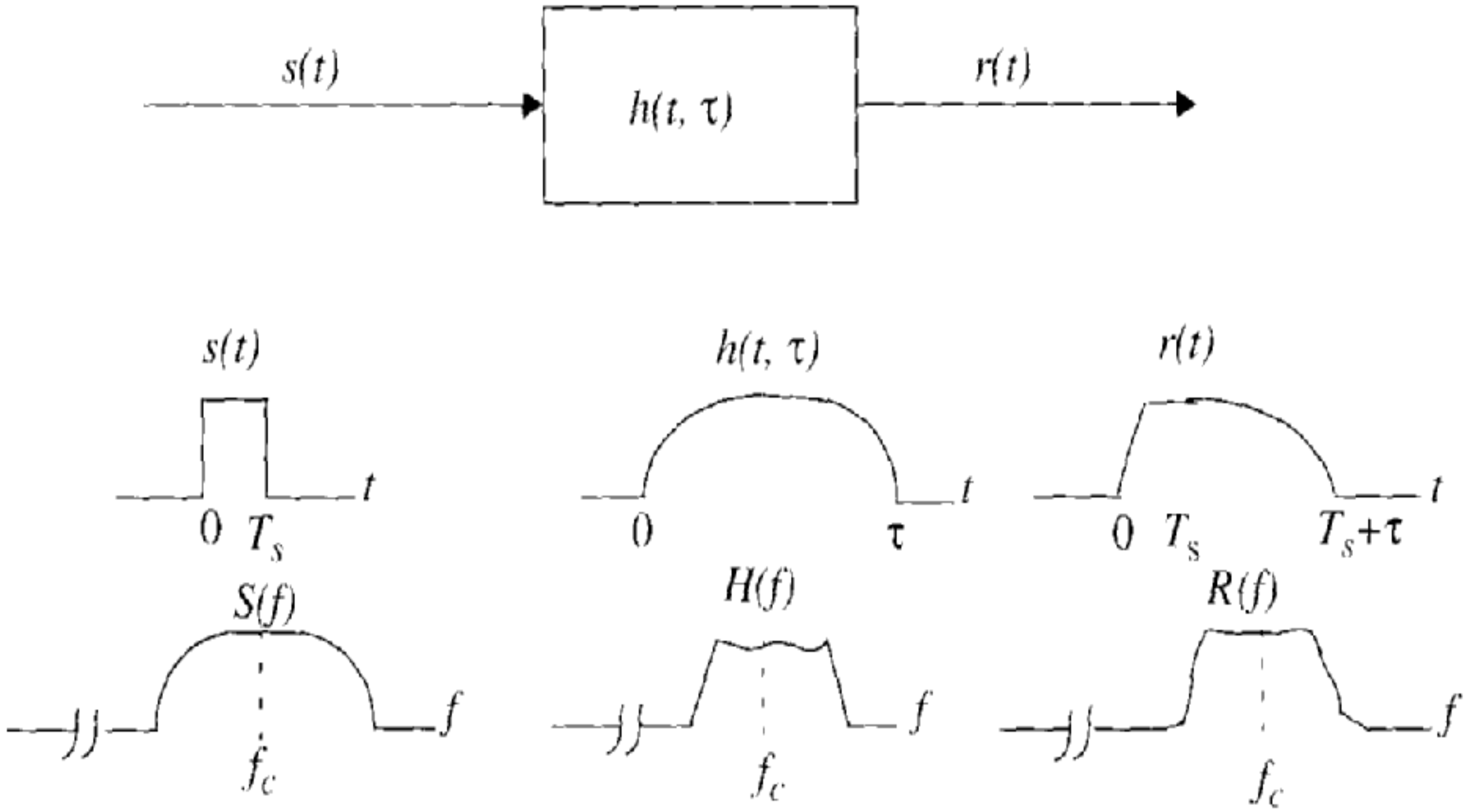


Figure 4.13

Frequency selective fading channel characteristics.

Fast Fading

- Due to Doppler Spread
  - Rate of change of the channel characteristics
    is **larger** than the
    Rate of change of the transmitted signal
  - The channel changes during a symbol period.
  - The channel changes because of receiver motion.
  - Coherence time of the channel is smaller than the symbol period of the transmitter signal

Occurs when:
$B_S < B_D$
and
$T_S > T_C$

B_S: Bandwidth of the signal
B_D: Doppler Spread
T_S: Symbol Period
T_C: Coherence Bandwidth

Slow Fading

- Due to Doppler Spread
  - Rate of change of the channel characteristics
    is **much smaller** than the
    Rate of change of the transmitted signal

Occurs when:
$B_S >> B_D$
and
$T_S << T_C$

B_S: Bandwidth of the signal
B_D: Doppler Spread
T_S: Symbol Period
T_C: Coherence Bandwidth

Fading Distributions

- Describes how the received signal amplitude changes with time.
  - Remember that the received signal is combination of multiple signals arriving from different directions, phases and amplitudes.
  - With the received signal we mean the baseband signal, namely the envelope of the received signal (i.e. r(t)).
- Its is a statistical characterization of the multipath fading.
- Two distributions
  - Rayleigh Fading
  - Ricean Fading

Rayleigh and Ricean Distributions

- Describes the received signal envelope distribution for channels, where all the components are non-LOS:
    - i.e. there is no line-of–sight (LOS) component.
- Describes the received signal envelope distribution for channels where one of the multipath components is LOS component.
    - i.e. there is one LOS component.

Rayleigh

Rayleigh distribution has the probability density function (PDF) given by:

$$p(r) = \begin{cases} \dfrac{r}{\sigma^2} e^{\left(-\frac{r^2}{2\sigma^2}\right)} & (0 \le r \le \infty) \\ 0 & (r < 0) \end{cases}$$

σ^2 is the time average power of the received signal before envelope detection.
σ is the rms value of the received voltage signal before envelope detection

Remember: $\overline{P}$ (average power) $\propto V_{rms}^2$ (see end of slides 5)

Rayleigh

The probability that the envelope of the received signal does not exceed a specified value of R is given by the CDF:

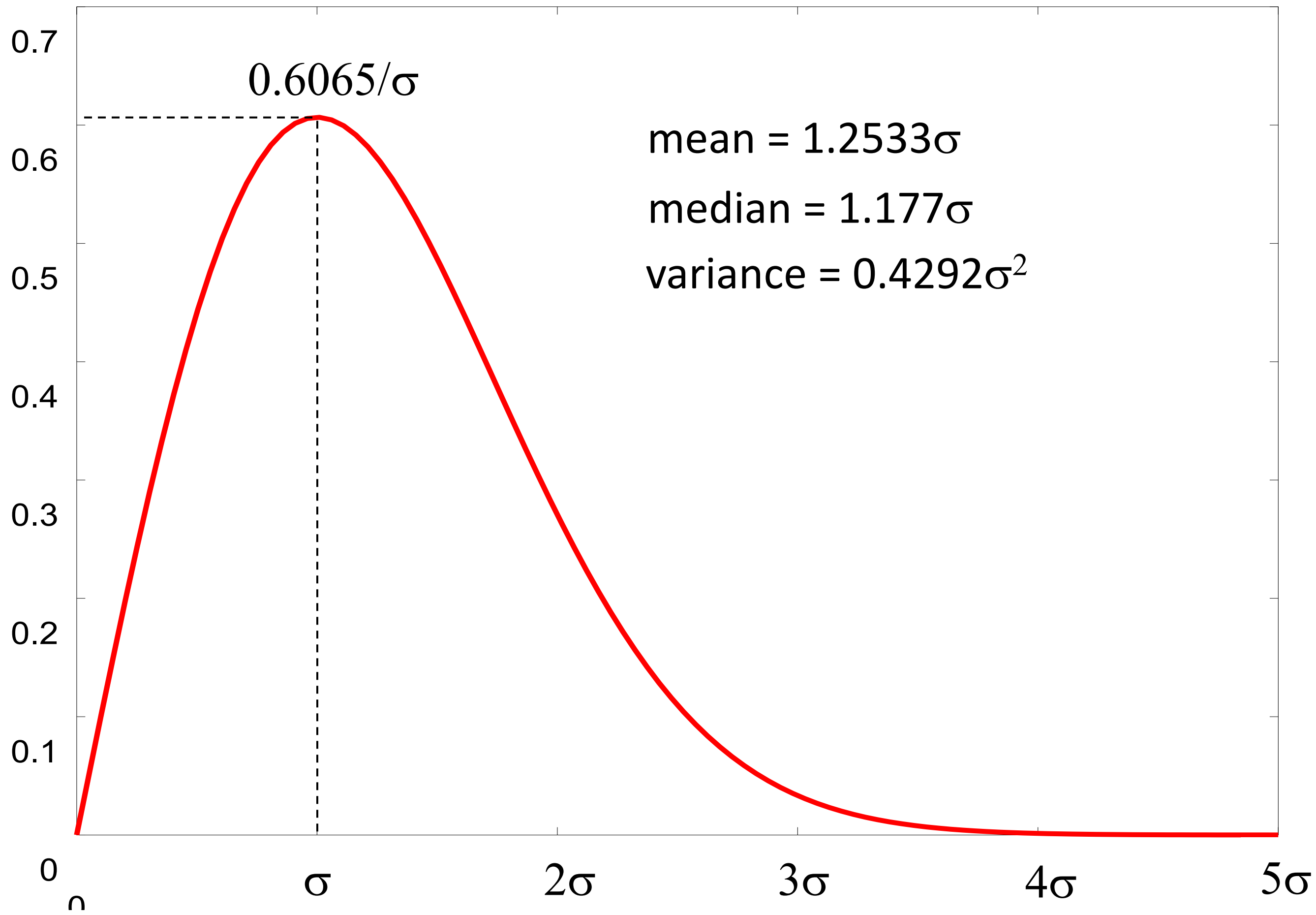
$$P(R) = P_r(r \le R) = \int_0^R p(r)dr = 1 - e^{-\frac{R^2}{2\sigma^2}}$$

$$r_{mean} = E[r] = \int_0^\infty rp(r)dr = \sigma\sqrt{\frac{\pi}{2}} = 1.2533\sigma$$

$$r_{median} = 1.177\sigma \quad \text{found by solving } \frac{1}{2} = \int_0^{r_{median}} p(r)dr$$

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

Rayleigh PDF



Ricean Distribution

- When there is a stationary (non-fading) LOS signal present, then the envelope distribution is Ricean.
- The Ricean distribution degenerates to Rayleigh when the dominant component fades away.

How do systems handle fading problem?

Analog

- Narrowband transmission

GSM

- Adaptive channel equalization
- Channel estimation training sequence

DECT

- Use the handset only in small cells with small delay spreads
- Diversity and channel selection can help a little bit (pick a channel where late reflections are in a fade)

IS95 Cellular CDMA

- Rake receiver separately recovers signals over paths with excessive delays

Digital Audio Broacasting

- OFDM multi-carrier modulation: The radio channel is split into many narrowband (ISI- free) subchannels

4.2 If a particular modulation provides suitable BER performance whenever $\sigma/T_s \le 0.1$, determine the smallest symbol period T_s (and thus the greatest symbol rate) that may be sent through RF channels shown in Figure P4.2, without using an equalizer.

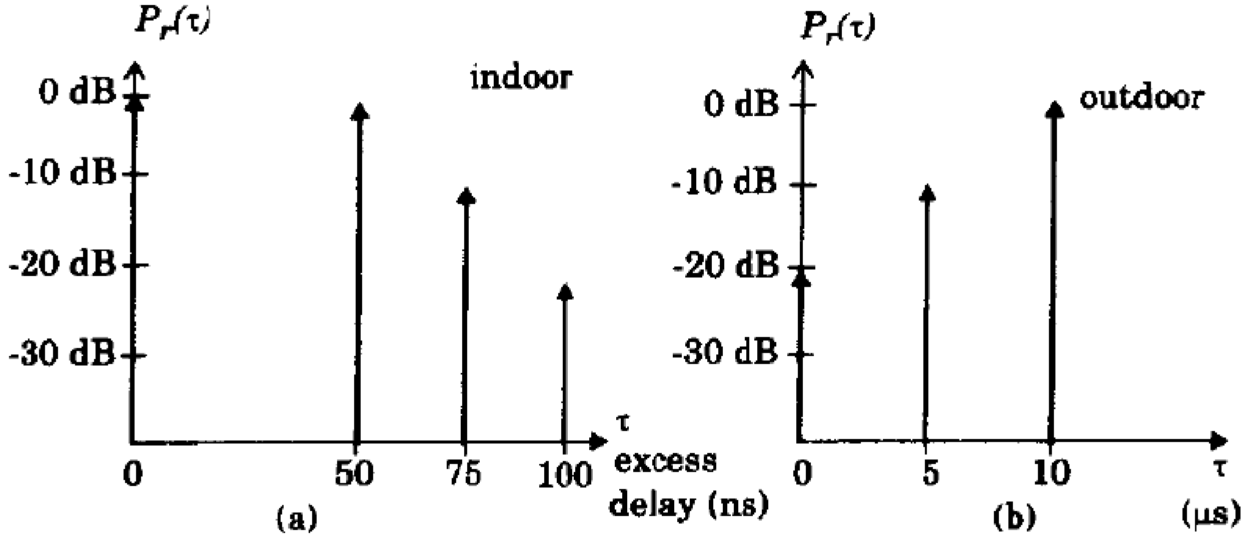


Figure P4.2: Two channel responses for Problem 4.2

5.6 For (a), $\bar{\tau} = \frac{1\times0 + 1\times50 + 0.1\times75 + 0.01\times100}{1+1+0.1+0.01} \doteq 27.725\,(ns)$

$$\overline{\tau^2} = \frac{1\times0 + 1\times50^2 + 0.1\times75^2 + 0.01\times100^2}{1+1+0.1+0.01} \doteq 1498.8\,(ns^2)$$

$\Rightarrow$ the rms delay spread $\sigma_\tau = \sqrt{1498.8 - 27.725^2} \doteq 27\,(ns)$

Since $\frac{\sigma_\tau}{T_s} \leq 0.1$, $T_s \geq 10\sigma_\tau = 270\,ns$

$\Rightarrow$ Smallest symbol period $T_{s\,min} = 270\,ns$

greatest data rate $R_{max} = \frac{1}{T_{s\,min}} \doteq 3.7\,Mbps$

For (b), $\bar{\tau} = \frac{0.01\times0 + 0.1\times5 + 1\times10}{0.01+0.1+1} \doteq 9.46\,(\mu s)$

$$\overline{\tau^2} = \frac{0.01\times0 + 0.1\times5^2 + 1\times10^2}{0.01+0.1+1} \doteq 92.34\,(\mu s^2)$$

$$\sigma_\tau = \sqrt{\overline{\tau^2} - (\bar{\tau})^2} = \sqrt{92.34 - (9.46)^2} \doteq 1.688\,(\mu s)$$

$T_{s\,min} = 10\,\sigma_\tau = 16.88\,(\mu s)$ $\quad R_{max} = \frac{1}{T_{s\,min}} \doteq 59.25\,Kbps$