GRT INSTITUTE OF ENGINEERING AND TECHNOLOGY, Tiruttani.

(Approved by AICTE, New Delhi Affiliated to Anna University, Chennai.)

Department of Electronics & Communication Engineering

III Year - VIth Semester

WIRELESS COMMUNICATION PPT

REGULATION – 2017

Mobile Radio Propagation

Large-scale Path loss

Wireless Communication

Introduction

- The mobile radio channel places fundamental limitations on the performance of a wireless communication system
- The wireless transmission path may be
  - Line of Sight (LOS)
  - Non line of Sight (NLOS)
- Radio channels are random and time varying
- Modeling radio channels have been one of the difficult parts of mobile radio design and is done in statistical manner
- When electrons move, they create EM waves that can propagate through space.
- By using antennas we can transmit and receive these EM wave
- Microwave ,Infrared visible light and radio waves can be used.

Properties of Radio Waves

- Are easy to generate
- Can travel long distances
- Can penetrate buildings
- May be used for both indoor and outdoor coverage
- Are omni-directional-can travel in all directions
- Can be narrowly focused at high frequencies(>100MHz) using parabolic antenna

Properties of Radio Waves

- Frequency dependence
  - Behave more like light at high frequencies
    - Difficulty in passing obstacle
    - Follow direct paths
    - Absorbed by rain
  - Behave more like radio at lower frequencies
    - Can pass obstacles
    - Power falls off sharply with distance from source
- Subject to interference from other radio waves

Propagation Models

- The statistical modeling is usually done based on data measurements made specifically for
  - the intended communication system
  - the intended spectrum

- They are tools used for:
  - Predicting the average signal strength at a given distance from the transmitter

  - Estimating the variability of the signal strength in close spatial proximity to a particular locations

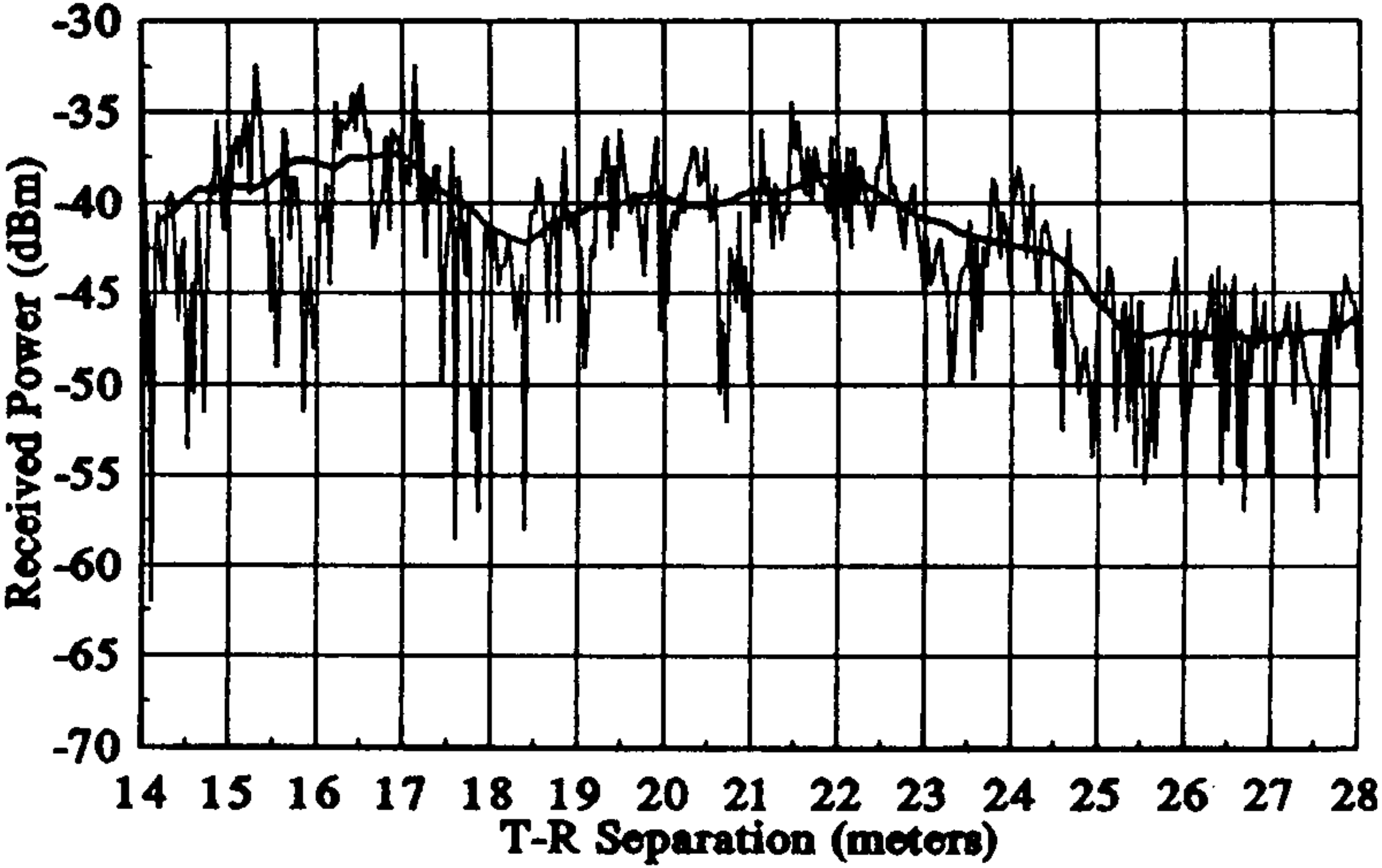
Propagation Models

- Large Scale Propagation Model:
  - Predict the mean signal strength for an arbitrary transmitter-receiver(T-R) separation
  - Estimate radio coverage of a transmitter
  - Characterize signal strength over large T-R separation distances(several 100’s to 1000’s meters)

Propagation Models

- Small Scale or Fading Models:
  - Characterize rapid fluctuations of received signal strength over
    - Very short travel distances(a few wavelengths)
    - Short time durations(on the order of seconds)

Small-scale and large-scale fading



Free Space Propagation Model

- For clear LOS between T-R
  - Ex: satellite & microwave communications

- Assumes that received power decays as a function of T-R distance separation raised to some power.

- Given by Friis free space eqn:

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

'L' is the system loss factor

L >1 indicates loss due to transmission line attenuation, filter losses & antenna losses

L = 1 indicates no loss in the system hardware

- Gain of antenna is related to its effective aperture A_e by

$$G = 4\pi A_e / \lambda^2$$

Free Space Propagation Model

- Effective Aperture A_e is related to physical size of antenna.

  $\lambda = c/f$.
- c is speed of light,
- P_t and P_r must be in same units
- G_t ad G_r are dimensionless

- An isotropic radiator, an ideal radiator which radiates power with unit gain uniformly in all directions, and is often used as reference

- Effective Isotropic Radiated Power (EIRP) is defined as

  EIRP= Pt Gt
- Represents the max radiated power available from a transmitter in direction of maximum antenna gain, as compared to an isotropic radiator

Free Space Propagation Model

- In practice Effective Radiated Power (ERP) is used instead of (EIRP)
- Effective Radiated Power (ERP) is radiated power compared to half wave dipole antennas
- Since dipole antenna has gain of 1.64(2.15 dB) ERP=EIRP-2.15(dB)
- the ERP will be 2.15dB smaller than the EIRP for same Transmission medium

Free Space Propagation Model

- Path Loss (PL) represents signal attenuation and is defined as difference between the effective transmitted power and received power

Path loss PL(dB) = 10 log [Pt/Pr]

= -10 log {GtGr λ^2/(4π)^2d^2}

- Without antenna gains (with unit antenna gains)

PL = - 10 log { λ^2/(4π)^2d^2}

- Friis free space model is valid predictor for P_r for values of d which are in the far-field of transmitting antenna

Free Space Propagation Model

- The far field or Fraunhofer region that is beyond far field distance d_f given as : $d_f=2D^2/\lambda$
- D is the largest physical linear dimension of the transmitter antenna
- Additionally, $d_f \gg D$ and $d_f \gg \lambda$
- The Friis free space equation does not hold for d=0
- Large Scale Propagation models use a close-in distance, d_o, as received power reference point, chosen such that $d_o >= d_f$
- Received power in free space at a distance greater then do

$$Pr(d)=Pr(do)(do/d)^2 \quad d>d_o>d_f$$

Pr with reference to 1 mW is represented as

$$Pr(d)=10\log(Pr(do)/0.001W)+20\log(do/d)$$

Electrostatic,inductive and radiated fields are launched, due to flow of current from anntena.

Regions far away from transmitter electrostatic and inductive fields become negligible and only radiated field components are considered.

Example

- What will be the far-field distance for a Base station antenna with
- Largest dimension D=0.5m
- Frequency of operation fc=900MHz,1800MHz

- For 900MHz
- *λ =3*10^8/900*10^6)=0.33m*
- *df= 2D^2/ λ =2(0.5)^2/0.33=1.5m*

Example

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

solution

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).

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

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

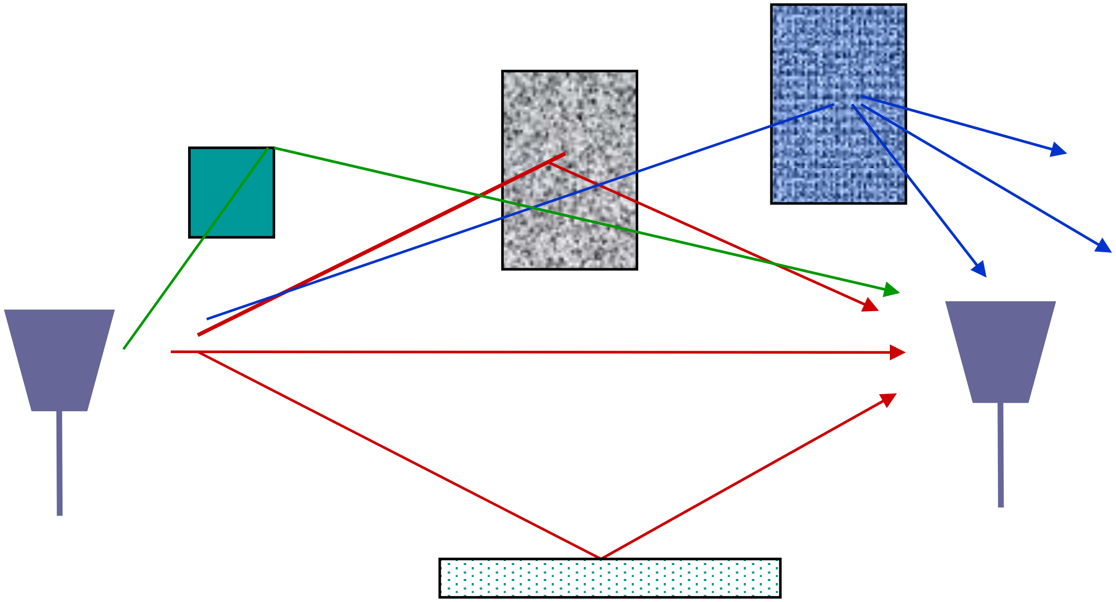
The received power at 10 km can be expressed in terms of dBm using equation (3.9), where $d_0 = 100$ m and $d = 10$ km

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

$$= -64.5\text{ dBm}.$$

Propagation Mechanisms

- Three basic propagation mechanism which impact propagation in mobile radio communication system are:
  - Reflection
  - Diffraction
  - Scattering



Propagation Mechanisms

- Reflection occurs when a propagating electromagnetic wave impinges on an object which has very large dimensions as compared to wavelength e.g. surface of earth , buildings, walls

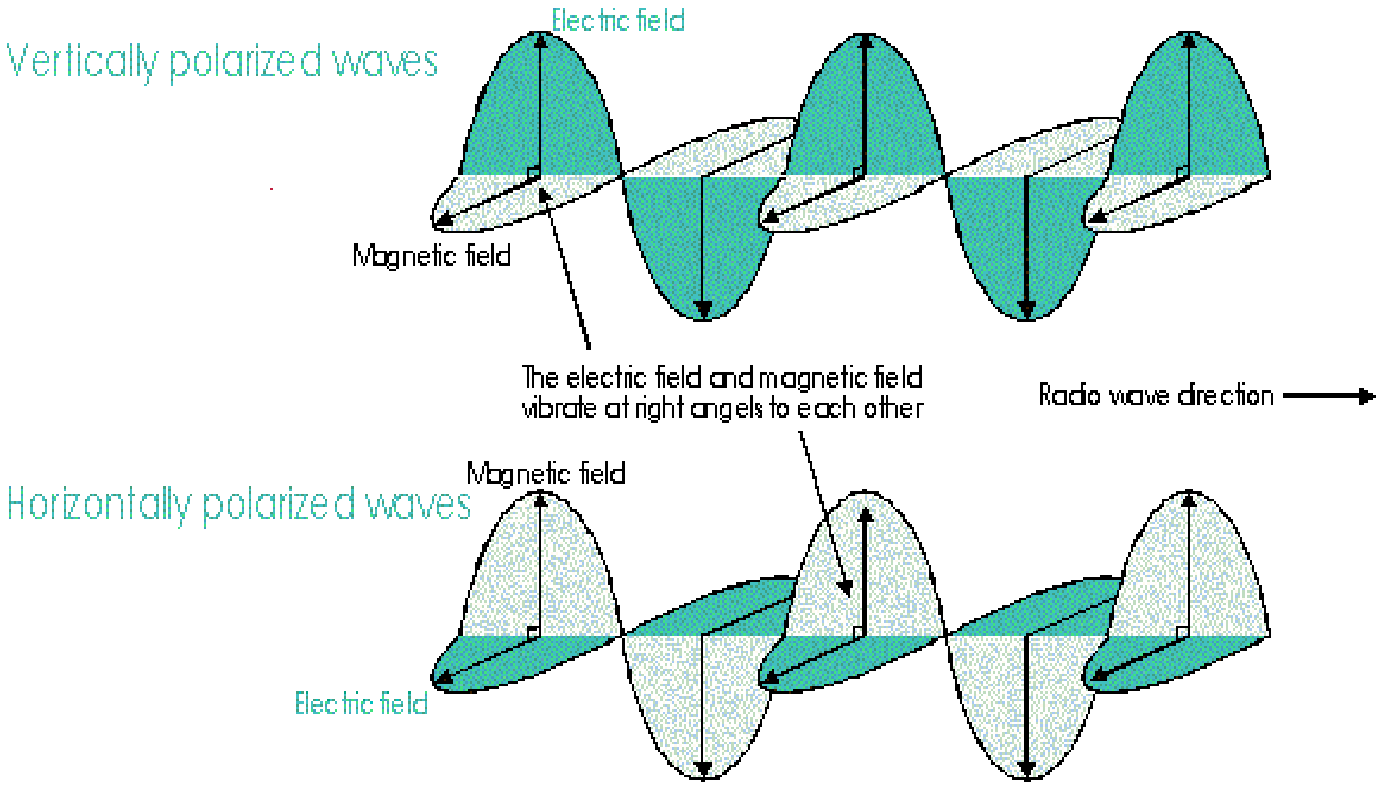
- Diffraction occurs when the radio path between the transmitter and receiver is obstructed by a surface that has sharp irregularities(edges)
  - Explains how radio signals can travel urban and rural environments without a line of sight path

- Scattering occurs when medium has objects that are smaller or comparable to the wavelength (small objects, irregularities on channel, foliage, street signs etc)

Reflection

- Occurs when a radio wave propagating in one medium impinges upon another medium having different electrical properties
- If radio wave is incident on a perfect dielectric
  - Part of energy is reflected back
  - Part of energy is transmitted
- In addition to the change of direction, the interaction between the wave and boundary causes the energy to be split between reflected and transmitted waves
- The amplitudes of the reflected and transmitted waves are given relative to the incident wave amplitude by Fresnel reflection coefficients

Vertical and Horizontal polarization



Reflection- Dielectrics

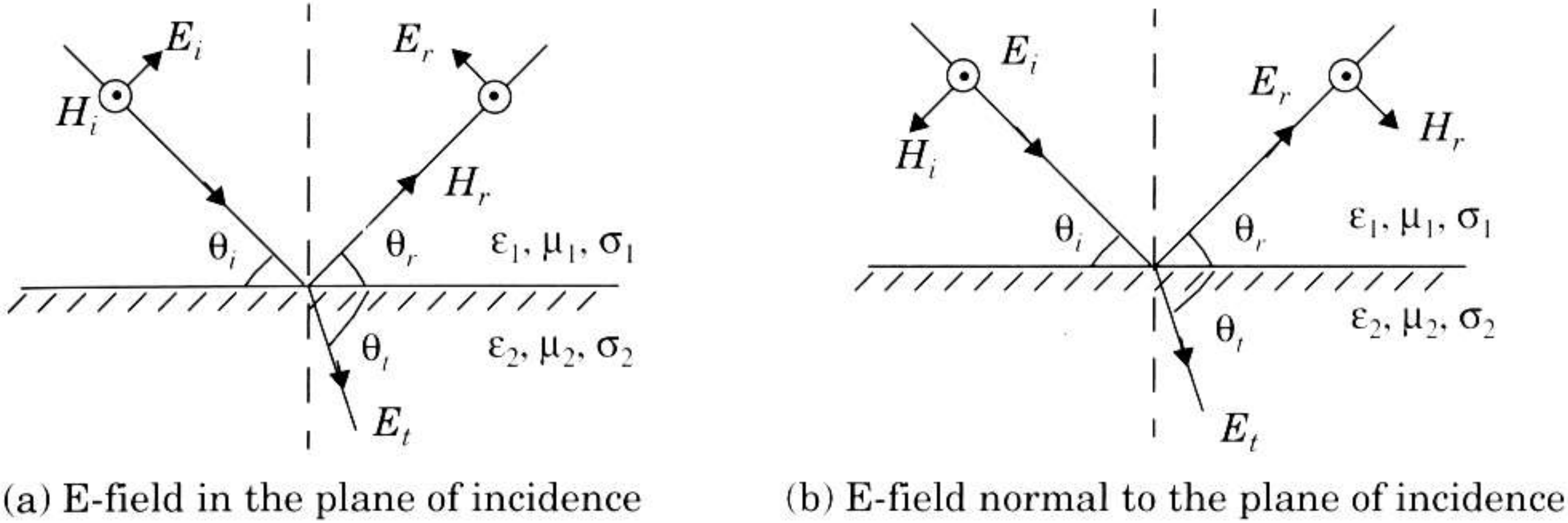


Figure 4.4 Geometry for calculating the reflection coefficients between two dielectrics.

Reflection

- $\Gamma(\parallel) = \frac{E_r}{E_i} = \frac{\eta_2 \sin\theta_t - \eta_1 \sin\theta_i}{\eta_1 \sin\theta_t + \eta_1 \sin\theta_i}$ (Paralell E-field polarization)
- $\Gamma(\perp) = \frac{E_r}{E_i} = \frac{\eta_2 \sin\theta_i - \eta_1 \sin\theta_t}{\eta_1 \sin\theta_i + \eta_1 \sin\theta_t}$ (Perpendicular E-field polarization)
- These expressions express ratio of reflected electric fields to the incident electric field and depend on impedance of media and on angles
- η is the intrinsic impedance given by $\sqrt{(\mu/\varepsilon)}$
- μ=permeability,ε=permittivity

Reflection-Perfect Conductor

- If incident on a perfect conductor the entire EM energy is reflected back
- Here we have $\theta_r = \theta_i$
- $E_i = E_r$ (E-field in plane of incidence)
- $E_i = -E_r$ (E field normal to plane of incidence)
- Γ(parallel)= 1
- Γ(perpendicular)= -1

Reflection - Brewster Angle

- It is the angle at which no reflection occur in the medium of origin. It occurs when the incident angle θ_B is such that the reflection coefficient Γ(parallel) is equal to zero.
- It is given in terms of θ_B as given below

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

- When first medium is a free space and second medium has an relative permittivity of ε_r *then*

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

- Brewster angle only occur for parallel polarization

Ground Reflection(Two Ray) Model

- In mobile radio channel, single direct path between base station and mobile and is seldom only physical means for propagation
- Free space model as a stand alone is inaccurate
- Two ray ground reflection model is useful
  - Based on geometric optics
  - Considers both direct and ground reflected path
- Reasonably accurate for predicting large scale signal strength over several kms that use tall tower height
- Assumption: The height of Transmitter >50 meters

Ground Reflection(Two Ray) Model

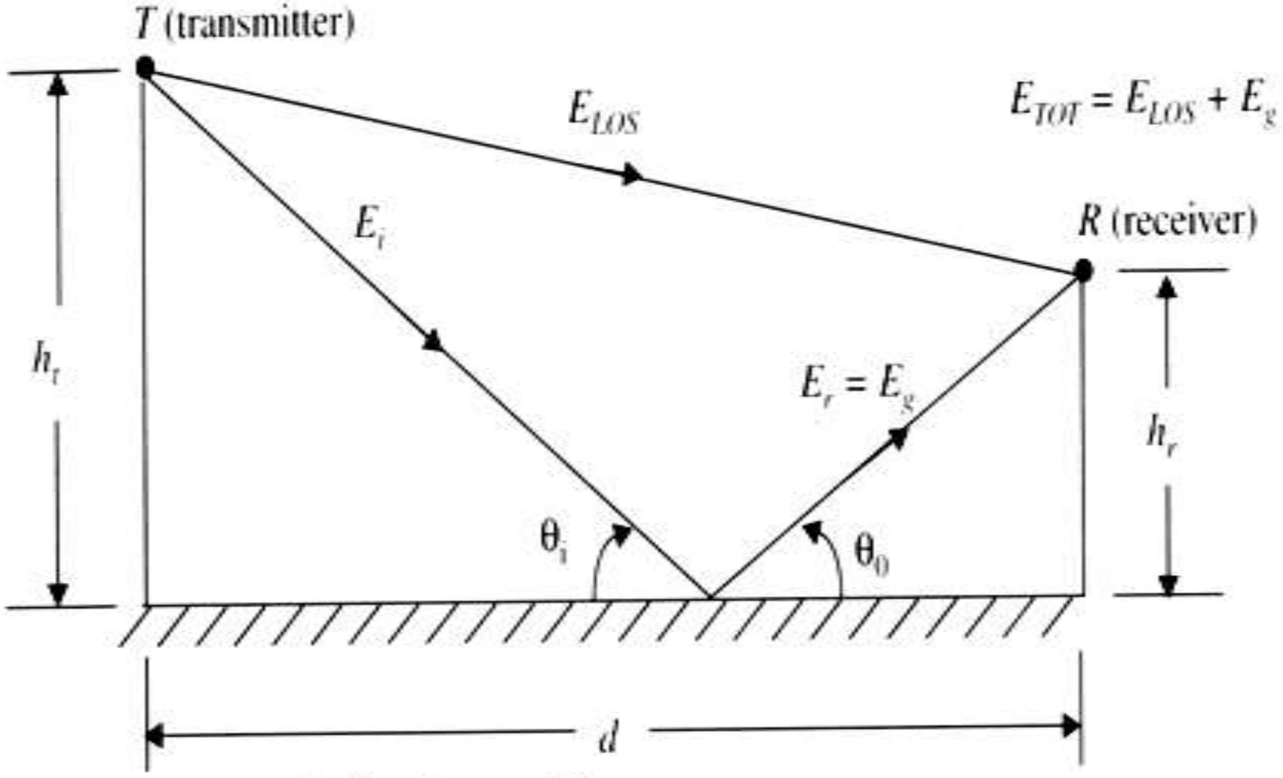


Figure 4.7 Two-ray ground reflection model.

Ground Reflection(Two Ray) Model

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

let E_o be $|\ \vec{E}\ |$ at reference point d_o then

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

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

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

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

Ground Reflection(Two Ray) Model

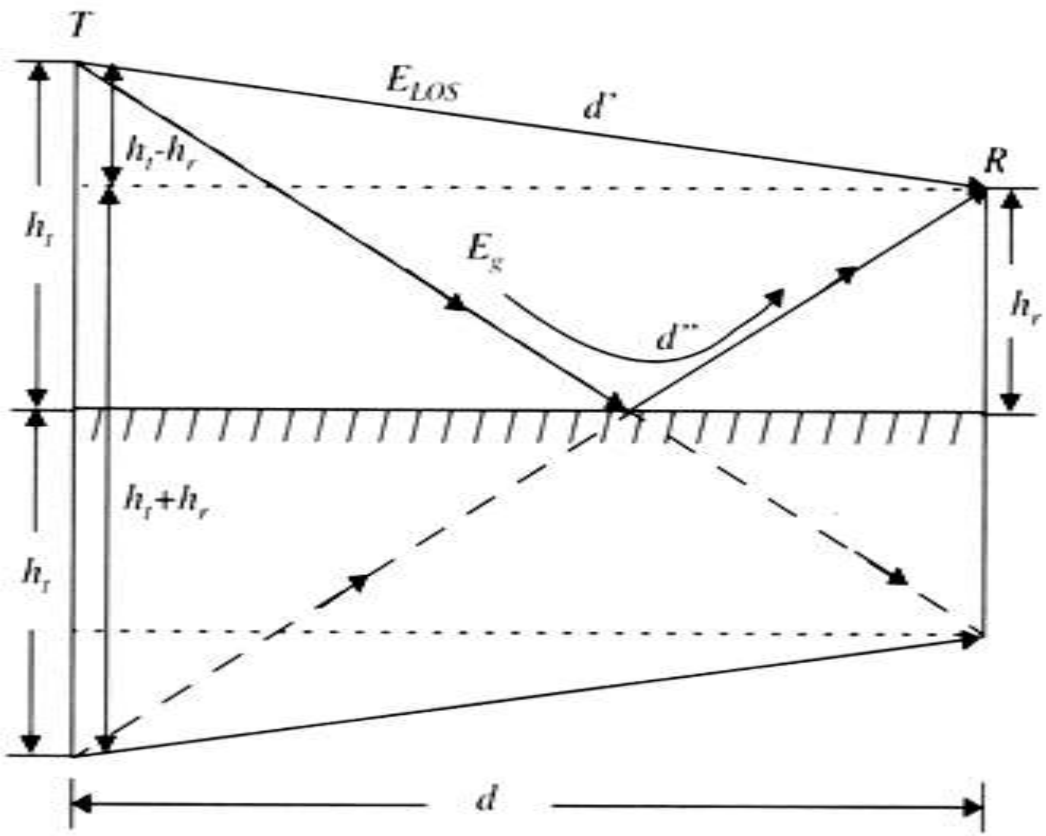


Figure 4.8 The method of images is used to find the path difference between the line-of-sight and the ground reflected paths.

Path Difference

$$
\begin{aligned}
\Delta = d'' - d' &= \sqrt{(h_t + h_r)^2 + d^2} - \sqrt{(h_t - h_r)^2 + d^2} \\
&= d\sqrt{\left(\left(\frac{h_t + h_r}{d}\right)^2 + 1\right)} - d\sqrt{\left(\left(\frac{h_t - h_r}{d}\right)^2 + 1\right)} \\
&\approx d\left(1 + \frac{1}{2}\left(\frac{h_t + h_r}{d}\right)^2\right) - d\left(1 + \frac{1}{2}\left(\frac{h_t - h_r}{d}\right)^2\right) \\
&\approx \frac{1}{2d}\left((h_t + h_r)^2 - (h_t - h_r)^2\right) \\
&\approx \frac{1}{2d}\left(\left(h_t^2 + 2h_t h_r + h_r^2\right) - \left(h_t^2 - 2h_t h_r + h_r^2\right)\right) \\
&\approx \frac{2h_t h_r}{d}
\end{aligned}
$$

Phase difference

$$\theta_\Delta \text{ radians } = \frac{2\pi\Delta}{\lambda} = \frac{2\pi\Delta}{\left(c/f_c \right)} = \frac{\omega_c \Delta}{c}$$

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

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

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

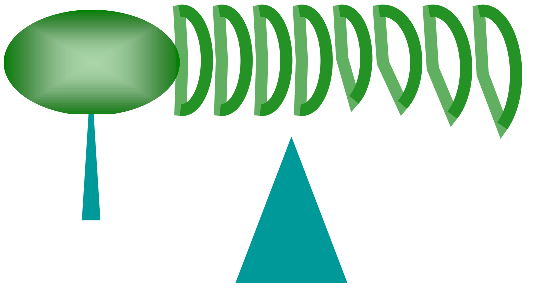
$$\overline{P}_r = P_t G_t G_r \frac{h_t^2 h_r^2}{d^4}$$

Diffraction

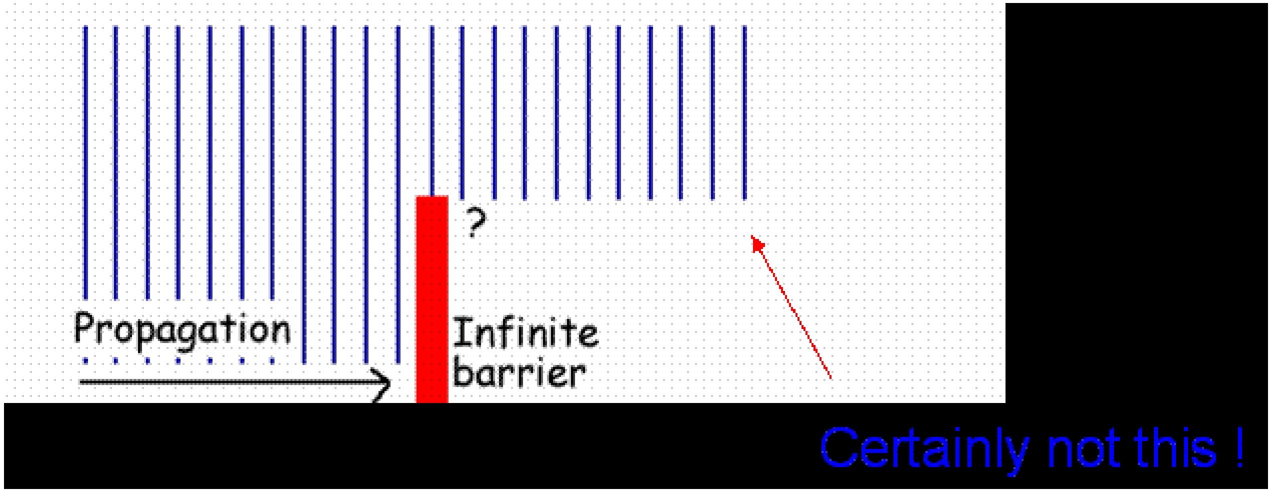
- Diffraction is the bending of wave fronts around obstacles.

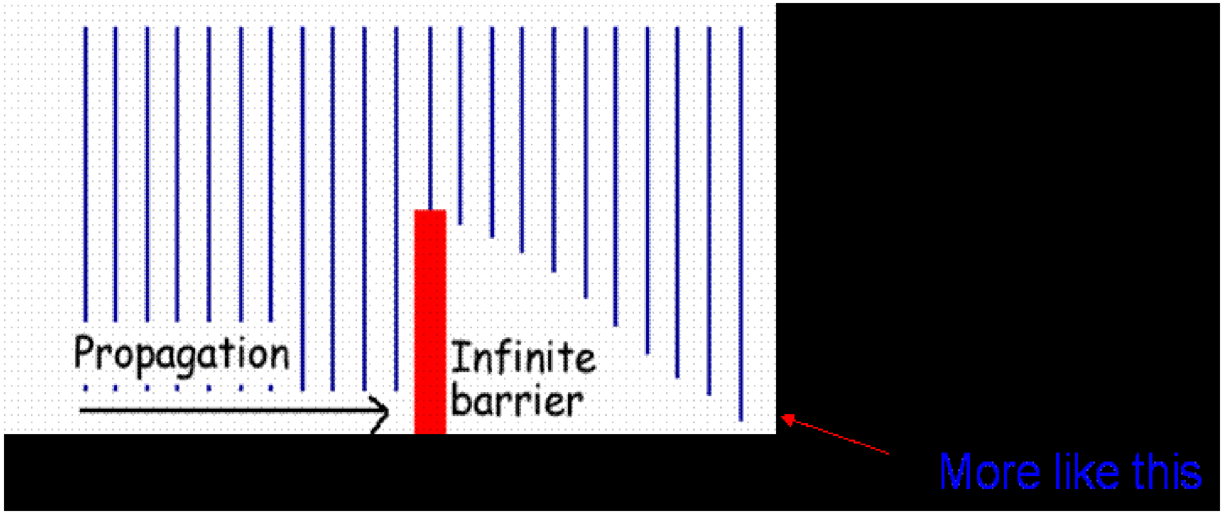
- Diffraction allows radio signals to propagate behind obstructions and is thus one of the factors why we receive signals at locations where there is no line-of-sight from base stations

- Although the received field strength decreases rapidly as a receiver moves deeper into an obstructed (shadowed) region, the diffraction field still exists and often has sufficient signal strength to produce a useful signal.



Diffraction





Knife-edge Diffraction Model

- Estimating the signal attenuation caused by diffraction of radio waves over hills and buildings is essential in predicting the field strength in a given service area.

- As a starting point, the limiting case of propagation over a knife edge gives good in sight into the order of magnitude diffraction loss.

- When shadowing is caused by a single object such as a building, the attenuation caused by diffraction can be estimated by treating the obstruction as a diffracting knife edge

Knife-edge Diffraction Model

Consider a receiver at point *R located in the shadowed region. The field strength at point R is a vector sum of the fields due to all of the secondary Huygens sources in the plane above the knife edge.*

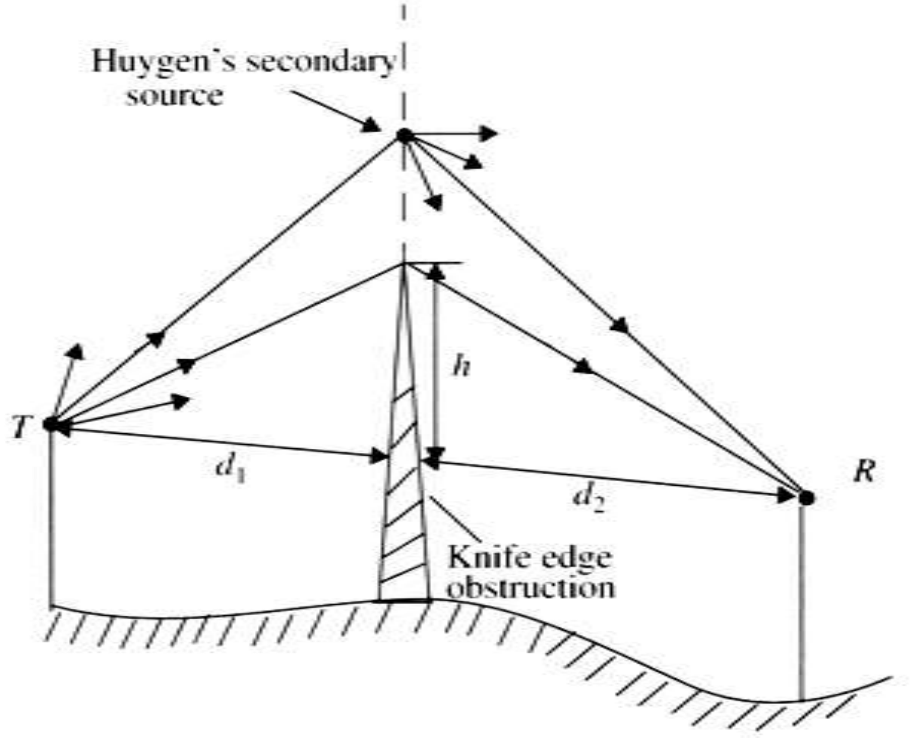


Figure 4.13 Illustration of knife-edge diffraction geometry. The receiver *R* is located in the shadow region.

Knife-edge Diffraction Model

- The difference between the direct path and diffracted path, call *excess path length*

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

- The corresponding phase difference

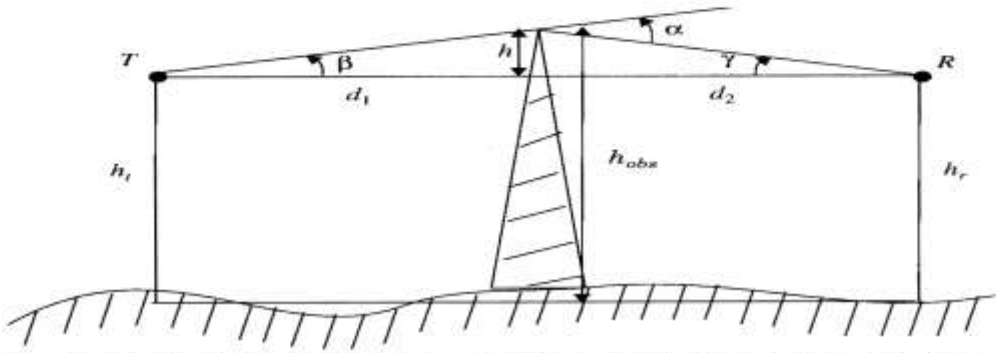
$$\phi = \frac{2\pi\Delta}{\lambda} \approx \frac{2\pi}{\lambda}\frac{h^2}{2}\frac{(d_1 + d_2)}{d_1 d_2}$$

- *Fresnel-Kirchoff* diffraction parameter is used to normalize the phased term and given as

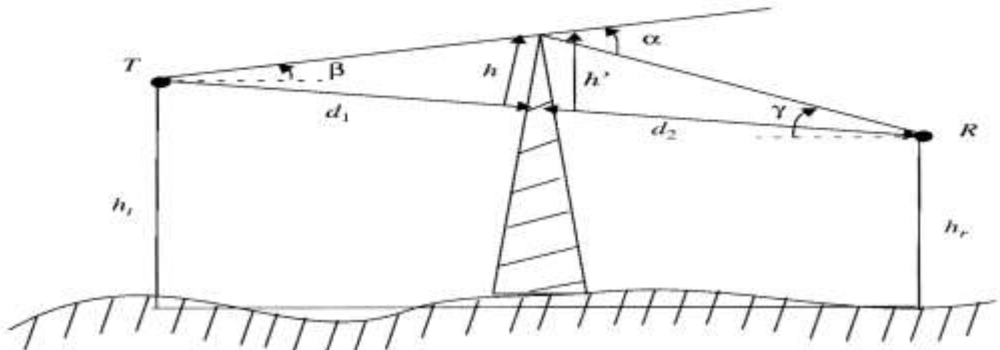
$$v = h\sqrt{\frac{2(d_1 + d_2)}{\lambda d_1 d_2}} = \alpha\sqrt{\frac{2 d_1 d_2}{\lambda(d_1 + d_2)}}$$ Which gives $$\phi = \frac{\pi}{2}v^2$$

- where $\alpha = h(\frac{d_1+d_2}{d_1 d_2})$

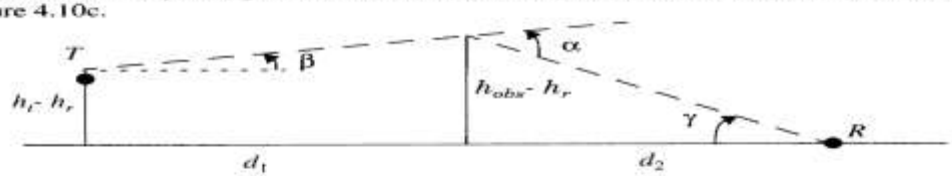
Knife-edge Diffraction Model



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



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



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

Figure 4.10 Diagrams of knife-edge geometry.

Fresnel zones

- Fresnel zones represent successive regions where secondary waves have a path length from the TX to the RX which are $n\lambda/2$ greater in path length than of the LOS path. The plane below illustrates successive Fresnel zones.

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

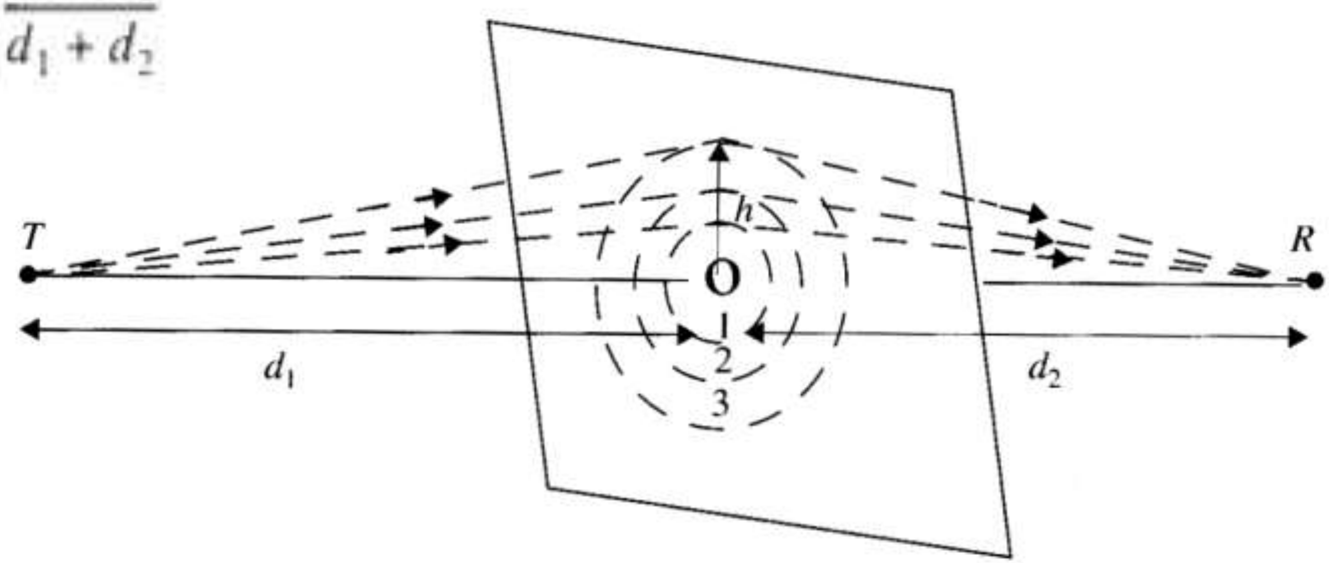
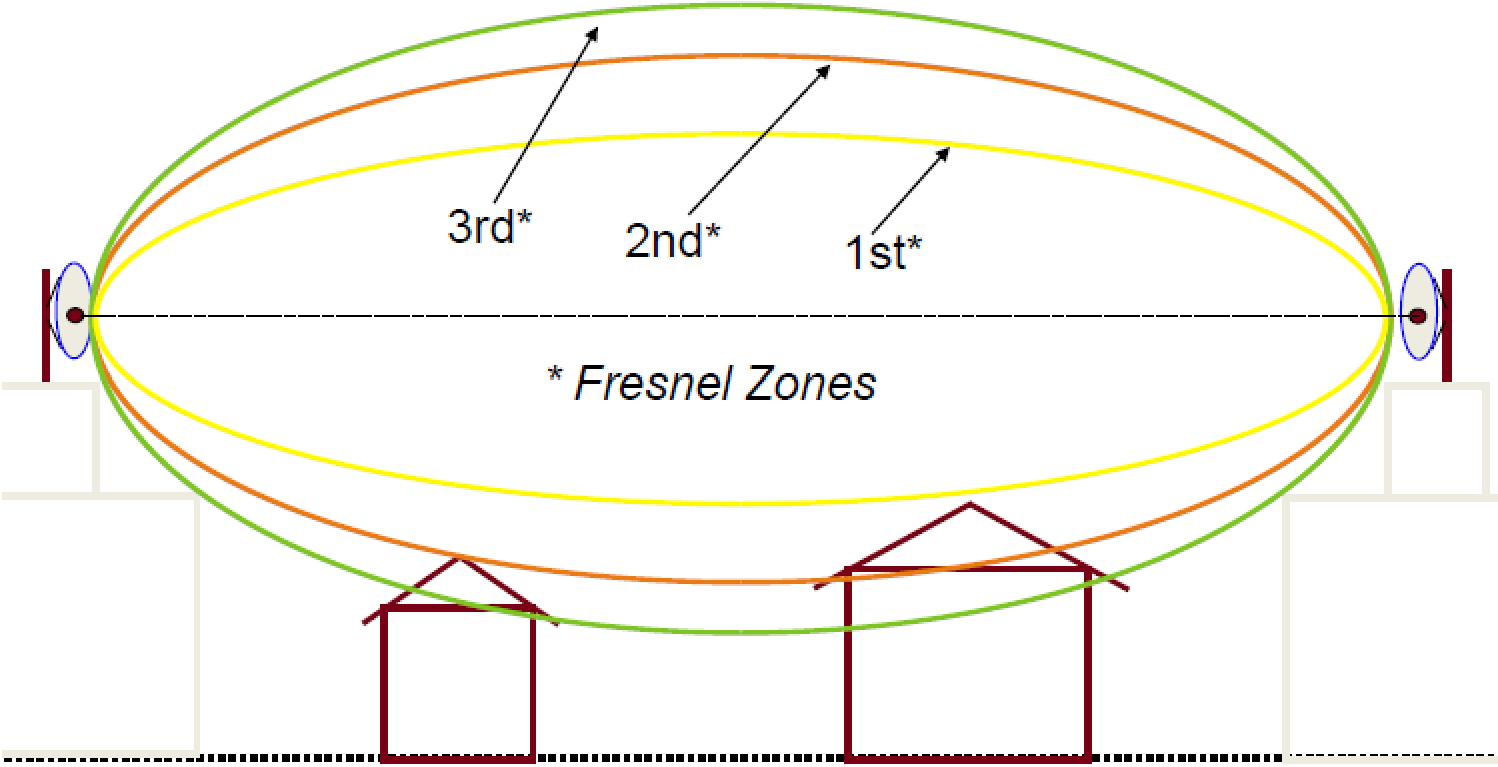


Figure 4.11 Concentric circles which define the boundaries of successive Fresnel zones.

Fresnel zones



Diffraction gain

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

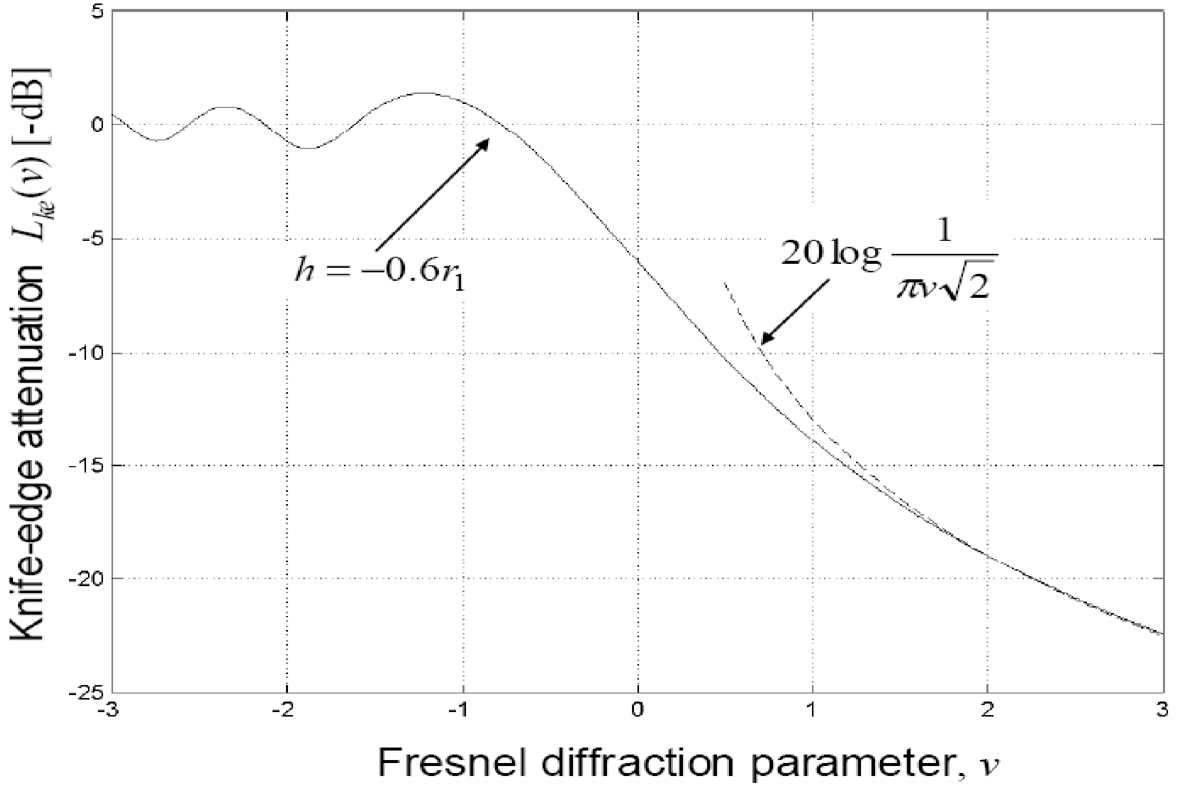
$$G_d(\mathrm{dB}) = 20\log|F(v)|$$

- The electric field strength, *Ed, of a knife edge diffracted wave is given by*

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

- *Eo : is the free space field strength in the absence of both the ground and the knife edge.*
- *F(v): is the complex fresnel integral.*
- *v: is the Fresnel-Kirchoff diffraction parameter*

Graphical Calculation of diffraction attenuation



Numerical solution

- An approximate numerical solution for equation

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

- Can be found using set of equations given below for different values of v

$G_d(dB)$	v
0	≤ -1
$20\log(0.5-0.62v)$	[-1,0]
$20\log(0.5\,e^{-0.95v})$	[0,1]
$20\log(0.4-(0.1184-(0.38-0.1v)^2)^{1/2})$	[1, 2.4]
$20\log(0.225/v)$	> 2.4

Example

Example 4.7

Compute the diffraction loss for the three cases shown in Figure 4.12. Assume $\lambda = 1/3$ m, $d_1 = 1$km, $d_2 = 1$ km, and (a) h = 25 m, (b) $h = 0$, (c) h = –25 m. Compare your answers using values from Figure 4.14, as well as the approximate solution given by Equation (4.61.a)–(4.61.e). For each of these cases, identify the Fresnel zone within which the tip of the obstruction lies.

Given:

$\lambda = 1/3$ m

$d_1 = 1$ km

$d_2 = 1$ km

(a)h = 25 m

Using Equation (4.56), the Fresnel diffraction parameter is obtained as

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

From Figure 4.14, the diffraction loss is obtained as 22 dB.

Using the numerical approximation in Equation (4.61.e), the diffraction loss is equal to 21.7 dB.

The path length difference between the direct and diffracted rays is given by Equation (4.54) as

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

To find the Fresnel zone in which the tip of the obstruction lies, we need to compute n which satisfies the relation $\Delta = n\lambda/2$. For $\lambda = 1/3$ m, and $\Delta = 0/625$ m, we obtain

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

Therefore, the tip of the obstruction completely blocks the first three Fresnel zones.

(b) $h = 0$ m

Therefore, the Fresnel diffraction parameter $v = 0$.

From Figure 4.14, the diffraction loss is obtained as 6 dB.

Using the numerical approximation in Equation (4.61.b), the diffraction loss is equal to 6 dB.

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

(c) $h = -25$ m

Using Equation (4.56), the Fresnel diffraction parameter is obtained as −2.74.

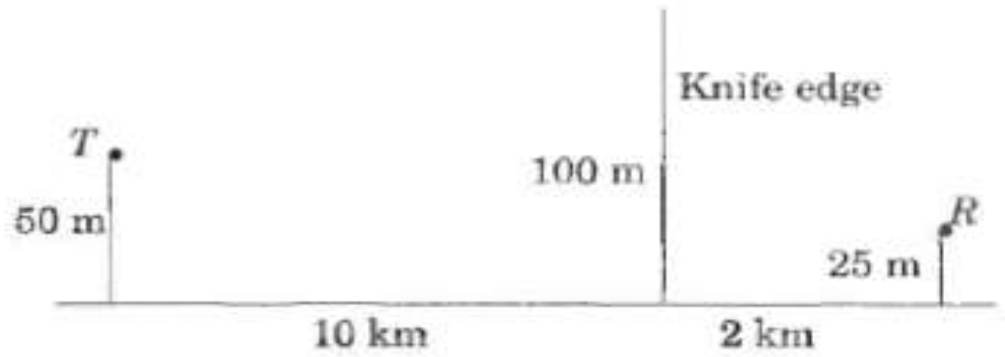
From Figure 4.14, the diffraction loss is approximately equal to 1 dB.

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

Example

Example 4.8

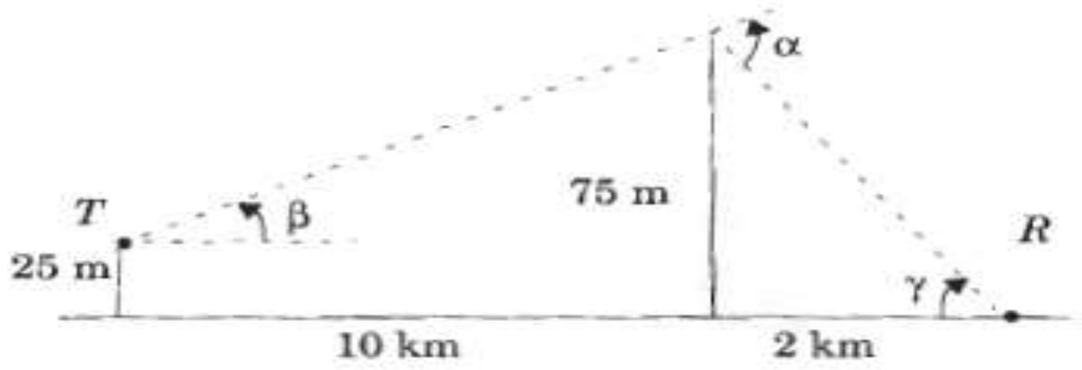
Given the following geometry, determine (a) the loss due to knife-edge diffraction, and (b) the height of the obstacle required to induce 6 dB diffraction loss. Assume $f = 900$ MHz.



Solution

(a) The wavelength $\lambda = \frac{c}{f} = \frac{3 \times 10^8}{900 \times 10^6} = \frac{1}{3}$ m.

Redraw the geometry by subtracting the height of the smallest structure.



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

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

and

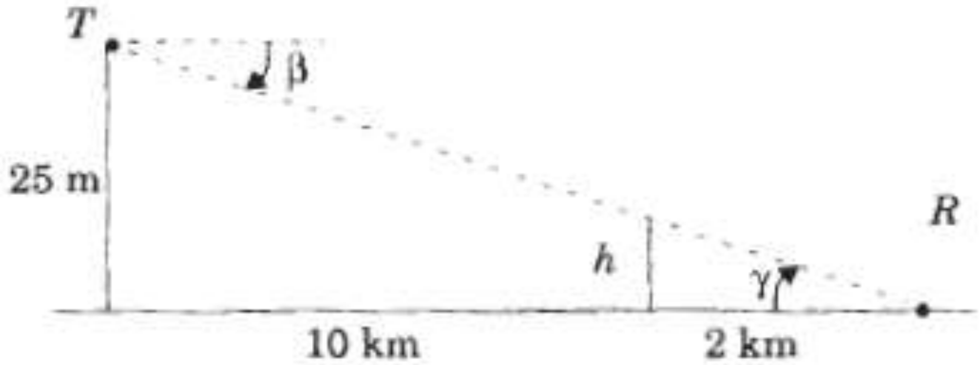
$$\alpha = \beta + \gamma = 2.434° = 0.0424 \text{ rad}$$

Then using Equation (4.56)

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

From Figure 4.14 or (4.61.e), the diffraction loss is 25.5 dB.

(b) For 6 dB diffraction loss, $v = 0$. The obstruction height h may be found using similar triangles ($\beta = \gamma$), as shown below.



It follows that $\frac{h}{2000} = \frac{25}{12000}$, thus $h = 4.16$ m.

Multiple Knife Edge Diffraction

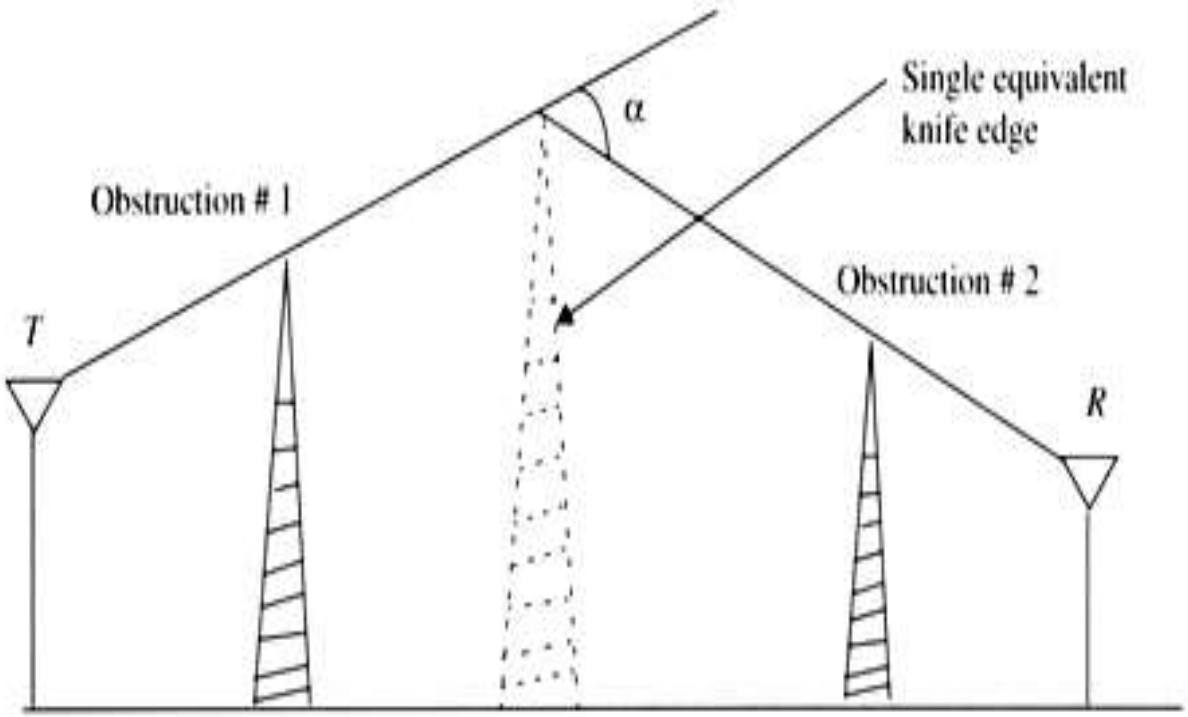


Figure 4.15 Bullington's construction of an equivalent knife edge [from [Bul47] © IEEE].

Scattering

- Scattering occurs when the medium through which the wave travels consists of objects with dimensions that are small compared to the wavelength, and where the number of obstacles per unit volume is large.

- Scattered waves are produced by
  - rough surfaces,
  - small objects,
  - or by other irregularities in the channel.

- Scattering is caused by trees, lamp posts, towers, etc.

Scattering

- Received signal strength is often stronger than that predicted by reflection/diffraction models alone
- The EM wave incident upon a rough or complex surface is **scattered** in **many** directions and provides more energy at a receiver
  - energy that would have been absorbed is instead reflected to the Rx.
- flat surface → EM reflection (one direction)
- rough surface → EM scattering (many directions)

Scattering

- Rayleigh criterion: used for testing surface roughness
- A surface is considered smooth if its min to max protuberance (bumps) h is less than critical height hc

$h_c = \lambda/8 \sin\Theta_i$

- Scattering path loss factor ρ_s is given by

$$\rho_s = \exp[-8[(\pi * \sigma_h * \sin\Theta_i)/ \lambda]^2]$$

Where h is surface height and σ_h is standard deviation of surface height about mean surface height.

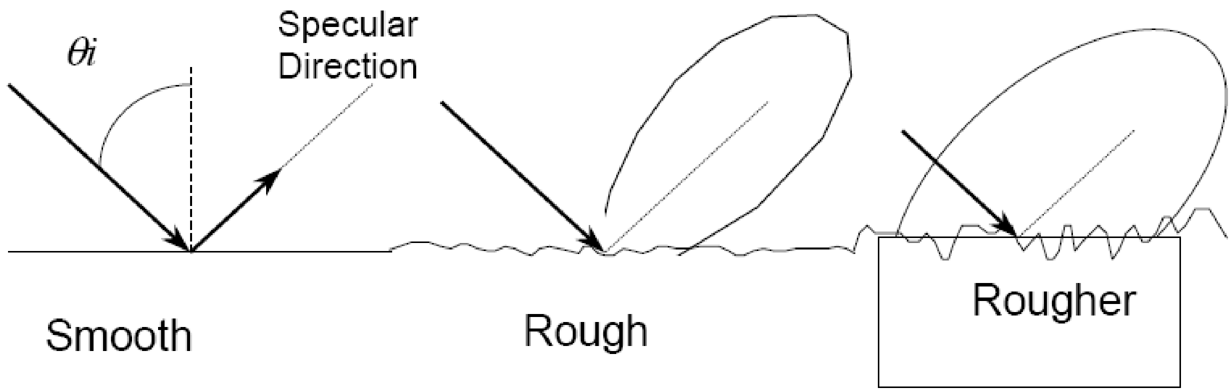
- For rough surface, the flat surface reflection coefficient is multiplied by scattering loss factor ρ_s to account for diminished electric field

- Reflected E-fields for h> h_c for rough surface can be calculated as

$\Gamma_{rough} = \rho_s \Gamma$

Scattering

Rough Surface Scattering



Roughness depends on :

- Surface height range
- Angle of incidence
- Wavelength

14

Outdoor propagation Environment

- Based on the coverage area, the Outdoor propagation environment may be divided into three categories
    1. Propagation in Macro cells
    2. Propagation in Micro cells
    3. Propagation in street Micro cells

Outdoor propagation Environment

Macrocells versus Microcells

	Macrocell	Microcell
Cell Radius	1 to 20 km	0.1 to 1 km
Tx Power	1 to 10 W	0.1 to 1 W
Fading	Rayleigh	Nakgami-Rice
RMS Delay Spread	0.1 to 10 μs	10 to 100ns
Max. Bit Rate	0.3 Mbps	1 Mbps

Outdoor propagation Models

- Outdoor radio transmission takes place over an irregular terrain.
- The terrain profile must be taken into consideration for estimating the path loss

  e.g. trees buildings and hills must be taken into consideration

- Some common models used are
  - Longley Rice Model
  - Okumura Model
  - Hatta model

Longley Rice Model

- Longley Rice Model is applicable to point to point communication.
- It covers 40MHz to 300 GHz
- It can be used in wide range of terrain
- Path geometry of terrain and the refractivity of troposphere is used for transmission path loss calculations
- Geometrical optics is also used along with the two ray model for the calculation of signal strength.
- Two modes
  - Point to point mode prediction
  - Area mode prediction

Longley Rice Model

- Longley Rice Model is normally available as a computer program which takes inputs as
  - Transmission frequency
  - Path length
  - Polarization
  - Antenna heights
  - Surface reflectivity
  - Ground conductivity and dialectic constants
  - Climate factors
  - A problem with Longley rice is that It doesn't take into account the buildings and multipath.

Okumura Model

- In 1968 Okumura did a lot of measurements and produce a new model.
- The new model was used for signal prediction in Urban areas.
- Okumura introduced a graphical method to predict the median attenuation relative to free-space for a quasi-smooth terrain
- The model consists of a set of curves developed from measurements and is valid for a particular set of system parameters in terms of carrier frequency, antenna height, etc.

Okumura Model

- First of all the model determined the free space path loss of link.
- After the free-space path loss has been computed, the median attenuation, as given by Okumura’s curves has to be taken to account
- The model was designed for use in the frequency range 200 up to 1920 MHz and mostly in an urban propagation environment.
- Okumura’s model assumes that the path loss between the TX and RX in the terrestrial propagation environment can be expressed as:

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

Okumura Model

- Estimating path loss using Okumura Model

1. Determine free space loss and $A_{mu}(f, d)$, between points of interest
2. Add $Amu(f, d)$ and correction factors to account for terrain

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

L_{50} = 50% value of propagation path loss (median)

L_F = free space propagation loss

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

$G(h_{te})$ = base station antenna height gain factor

$G(h_{re})$ = mobile antenna height gain factor

G_{AREA} = gain due to environment

Okumura Model

- $A_{mu}(f,d)$ & G_{AREA} have been plotted for wide range of frequencies
- Antenna gain varies at rate of 20dB or 10dB per decade

$G(h_{te}) = 20\log\frac{h_{te}}{200}$ **$10m < h_{te} < 1000m$**

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

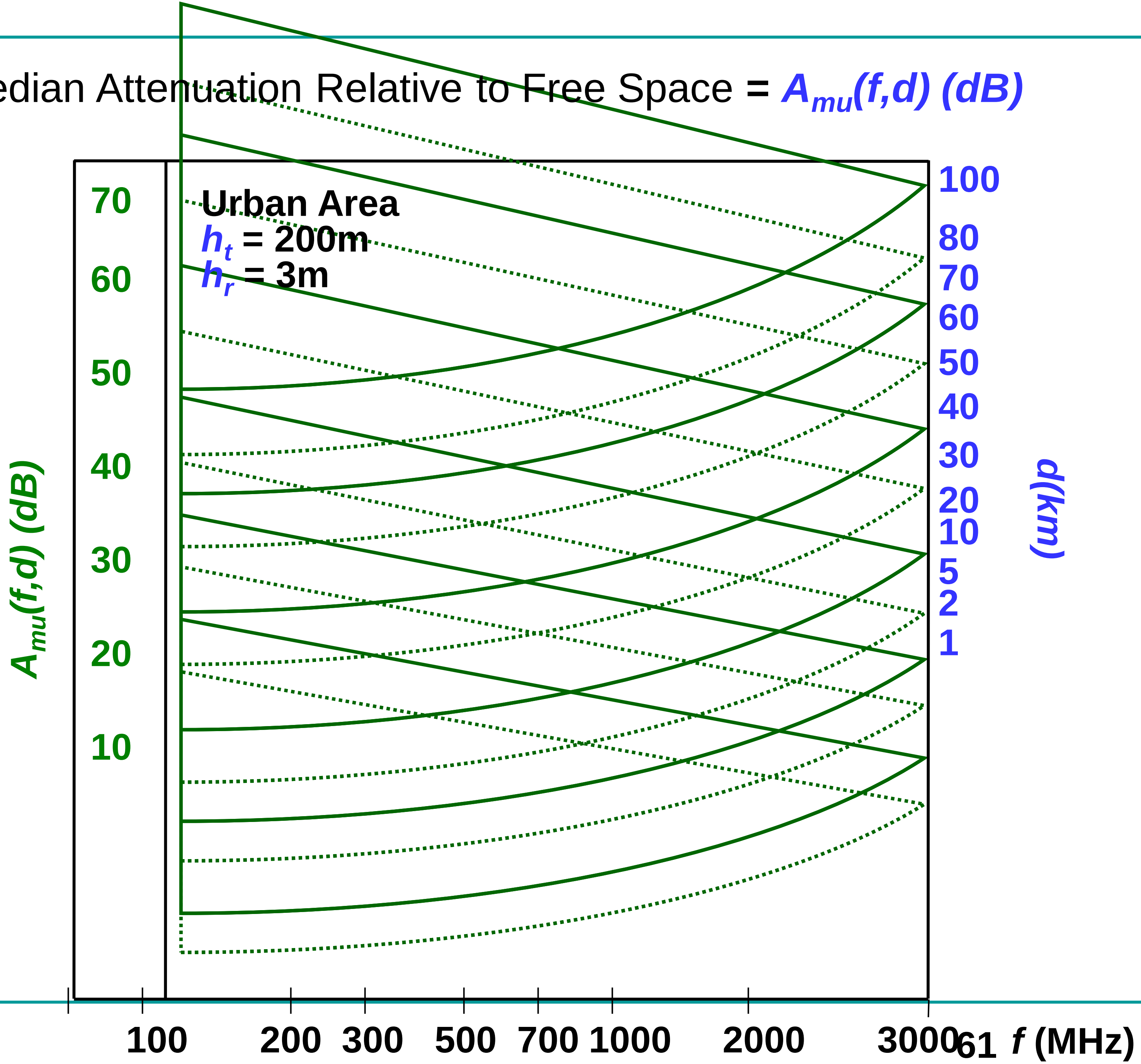
$G(h_{re}) = 20\log\frac{h_{re}}{3}$ **$3m < h_{re} < 10m$**

- **model corrected** for

Δh = terrain undulation height, isolated ridge height

average terrain slope and mixed land/sea parameter

Median Attenuation Relative to Free Space = $A_{mu}(f,d)$ (dB)



Correction Factor G_{AREA}

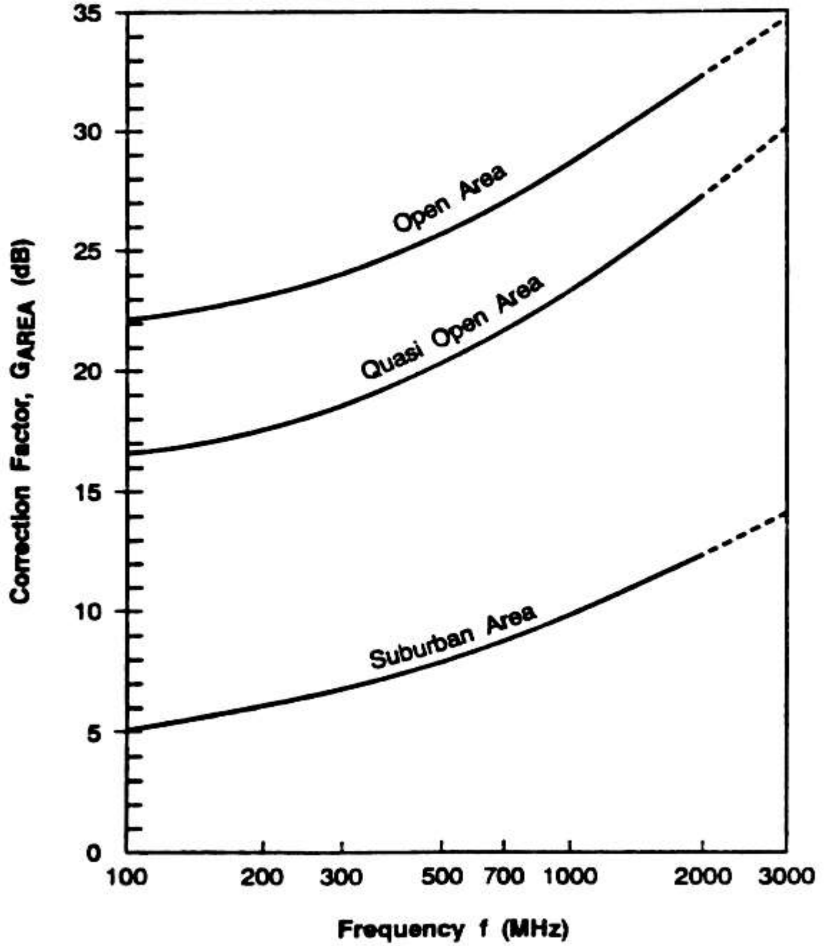


Figure 4.24 Correction factor, G_{AREA}, for different types of terrain [from [Oku68] © IEEE].

Example

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

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

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

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

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

- Most widely used model in Radio frequency.
- Predicting the behavior of cellular communication in built up areas.
- Applicable to the transmission inside cities.
- Suited for point to point and broadcast transmission.
- 150 MHz to 1.5 GHz, Transmission height up to 200m and link distance less than 20 Km.

Hata Model

- Hata transformed Okumura's graphical model into an analytical framework.

- The Hata model for urban areas is given by the empirical formula:

$L50_{,\ urban} = 69.55\ dB + 26.16 \log(f_c) - 3.82 \log(h_t) - a(h_r) + (44.9 - 6.55 \log(h_t)) \log(d)$

- Where $L50_{,\ urban}$ is the median path loss in dB.

- The formula is valid for

150 MHz $\le f_c \le$ 1.5GHz,

1 m $\le h_r \le$ 10m, 30m $\le h_t \le$ 200m,

1km $< d <$ 20km

Hata Model

- The correction factor $a(h_r)$ for mobile antenna height hr for a small or medium-sized city is given by:

$a(h_r) = (1.1 \log f_c − 0.7)h_r − (1.56 \log(f_c) − 0.8)$ dB

- For a large city it is given by

$a(h_r) = 8.29[\log(1.54h_r)]^2 − 1.10$ dB for $f_c <= 300$ MHz

$3.20[\log (11.75h_r)]^2 − 4.97$ dB for $f_c >= 300$ MHz

- To obtain path loss for suburban area the standard Hata model is modified as

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

- For rural areas

$L_{50} = L_{50}(urban) - 4.78\log(f_c)^2 - 18.33\log f_c - 40.98$

Indoor Models

- Indoor Channels are different from traditional channels in two ways

  1. The distances covered are much smaller
  2. The variability of environment is much greater for a much small range of Tx and Rx separation.

- Propagation inside a building is influenced by:

  - Layout of the building
  - Construction materials
  - Building Type: office , Home or factory

Indoor Models

- Indoor models are dominated by the same mechanism as out door models:

  - Reflection, Diffraction and scattering

- Conditions are much more variable

  - Doors/Windows open or not
  - Antenna mounting : desk ceiling etc
  - The levels of floor

- Indoor models are classifies as

  - Line of sight (LOS)
  - Obstructed (OBS) with varying degree of clutter

Indoor Models

- Portable receiver usually experience
  - Rayleigh fading for OBS propagation paths
  - Ricean fading for LOS propagation path
- Indoors models are effected by type of building e.g. Residential buildings, offices, stores and sports area etc.
- Multipath delay spread
  - Building with small amount of metal and hard partition have small delay spread 30 to 60ns
  - Building with large amount of metal and open isles have delay spread up to 300ns

Partition losses (same floor)

- Two types of partitions
  1. hard partitions: Walls of room
  2. Soft partitions : Moveable partitions that donot span to ceiling

- Partitions vary widely in their Physical and electrical properties.

- Path loss depend upon the types of partitions

Partition losses (same floor)

Partition Losses (Same Floor)

Material Type	Loss (dB)	Frequency
All metal partition	26	815 MHz
Concrete Block wall	13	1300 MHz
Empty Cardboard boxes	3 – 6 dB	1300 MHz
Dry Plywood (0.75 inches)	1 dB	9.6 GHz
Dry Plywood (0.75 inches)	4 dB	28.8 GHz

Partitions losses (between floors)

- Partition losses between the two floors depend on

1. External dimension and material used for buildings
2. Types of construction used to create floors
3. External surroundings
4. No of windows used
5. Tinting on the windows

- Floor Attenuation Factor (FAF) increases as we increase the no of floors

Partitions losses (between floors)

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

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

Log distance path loss model

- Path loss can be given as

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

where n is path loss exponent and σ is standard deviation

- n and σ depend on the building type.
- Smaller value of σ indicates better accuracy of path loss model

Log distance path loss model

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

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

Ericsson Multiple Break Point Model

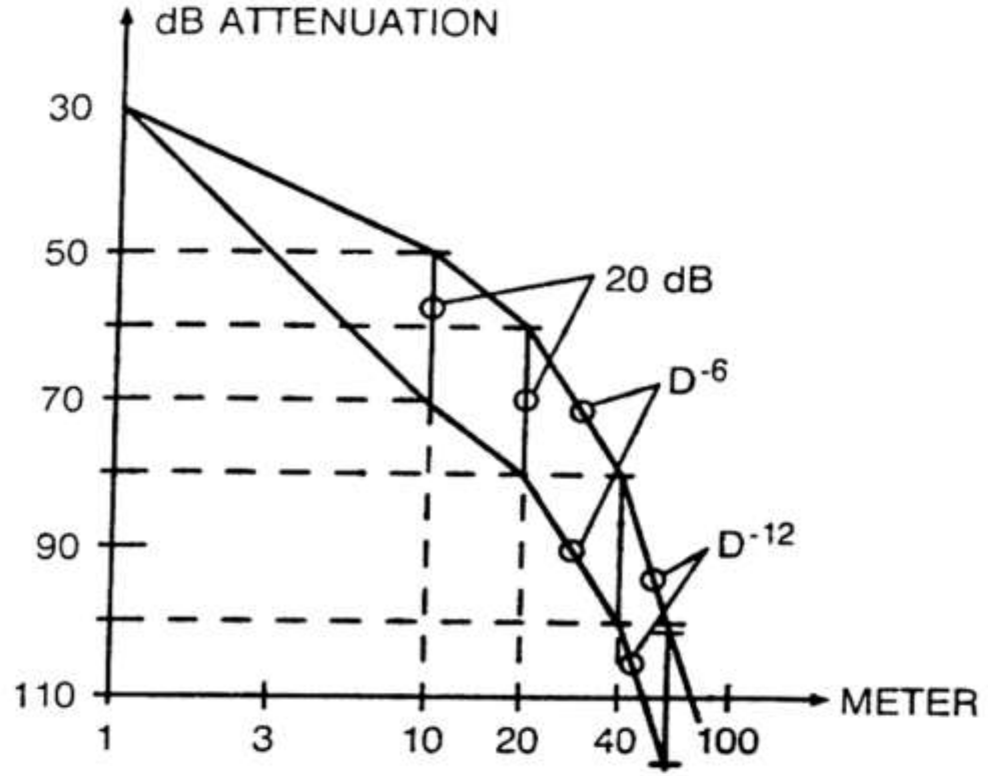


Figure 4.27 Ericsson in-building path loss model [from [Ake88] © IEEE].

Attenuation factor model

- Obtained by measurement in multiple floors building

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

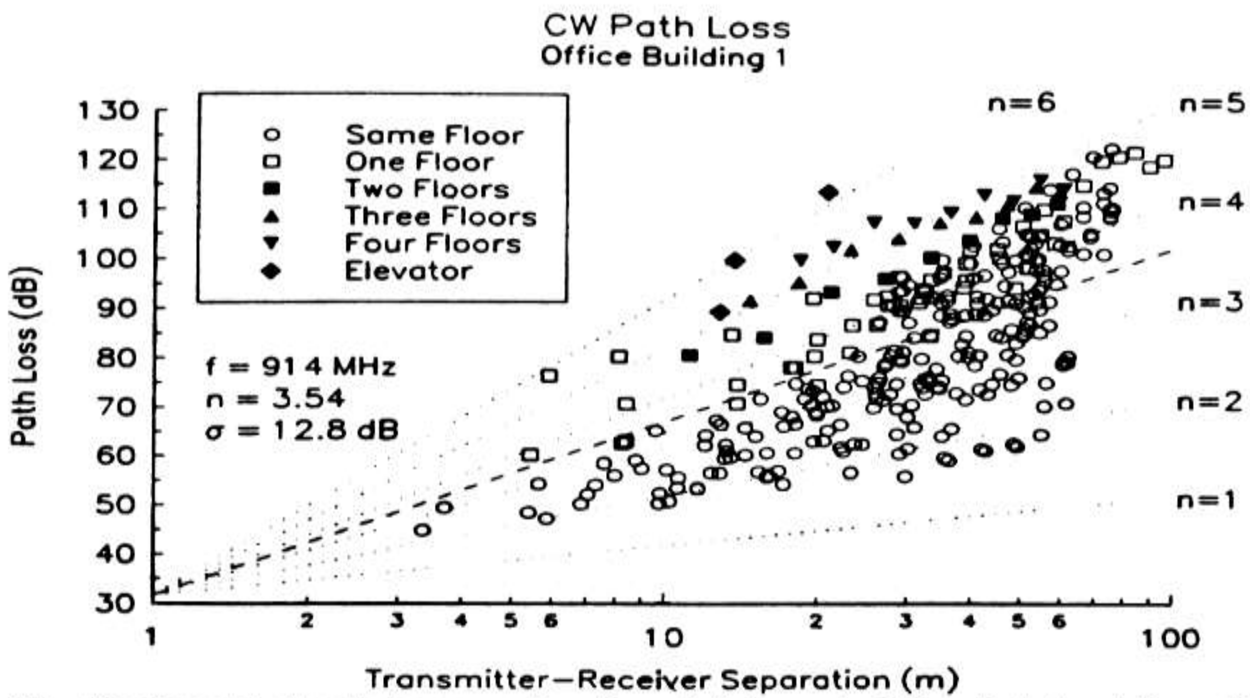


Figure 4.28 Scatter plot of path loss as a function of distance in Office Building 1 [from [Sei92b] © IEEE].

Attenuation factor model

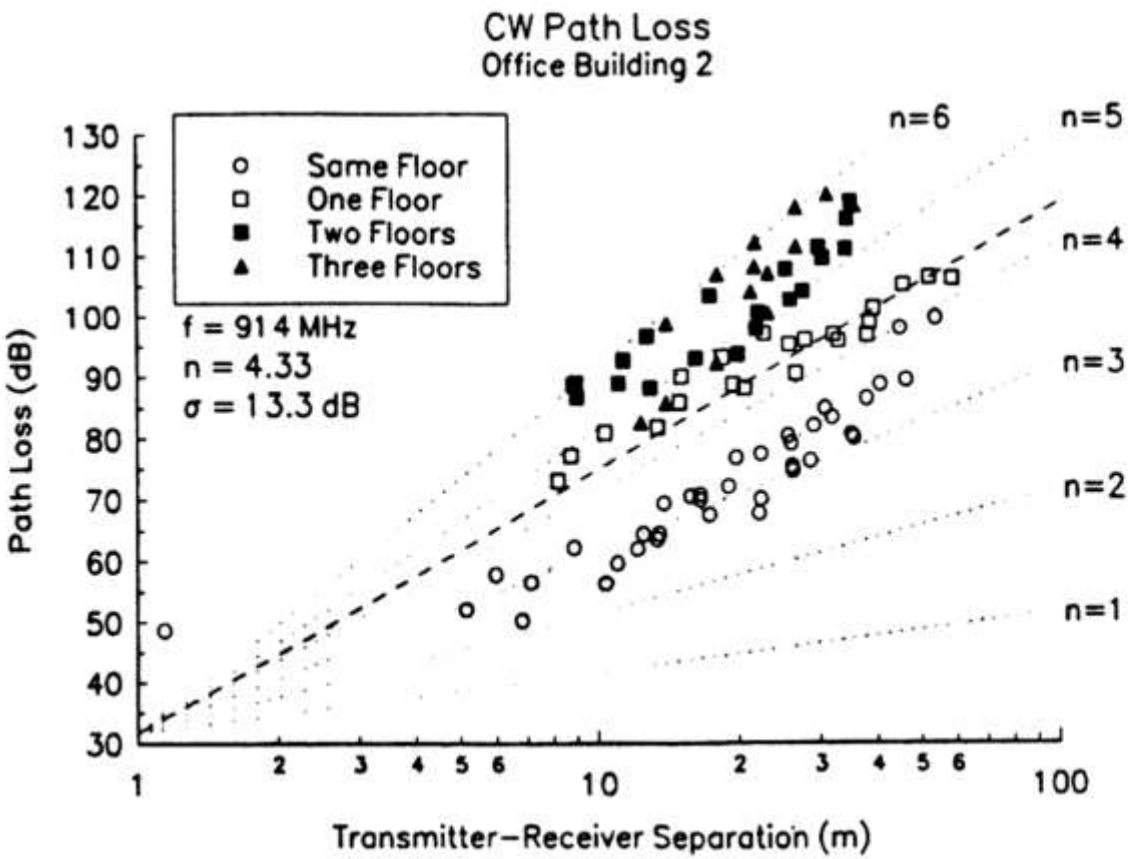


Figure 4.29 Scatter plot of path loss as a function of distance in Office Building 2 [from [Sei92b] © IEEE].

Signal penetration into building

- <u>Effect of frequency</u>

  - Penetration loss decreases with increasing frequency

- <u>Effect of Height</u>

- Penetration loss decreases with the height of building up to some certain height.

  - At lower heights the Urban clutter induces greater attenuation
  - Up to some height attenuation decreases but then again increase after a few floors
  - Increase in attenuation at higher floors is due to the Shadowing effects of adjacent buildings

Large, medium and small scale fading

- Large Scale Fading: Average signal power attenuation/path loss due to motion over large areas.
- Medium scale fading: Local variation in the average signal power around mean average power due to shadowing by local obstructions
- Small scale fading: large variation in the signal power due to small changes in the distance between transmitter and receiver (Also called Rayleigh fading when no LOS available). It is called Rayleigh fading due to the fact that various multipaths at the receiver with random amplitude & delay add up together to render rayleigh PDF for total signal.

Cause of Multipath Fading

- Fading : Fluctuation in the received signal power due to
  - Variations in the received singal amplitude (Different objects present on radio signal path produce attenuation of it’s power as they can scatter or absorb part of the signal power, thus producing a variation of the amplitude
  - Variations in the signal phase
  - Variations in the received signal angle of arrival (different paths travelling different distances may have different phases & angle of arrival)

Causes of Multipath fading Cont..

- Reflections and diffraction from object create many different EM waves which are received in mobile antenna. These waves usually come from many different directions and delay varies.
- In the receiver, the waves are added either constructively or destructively and create a Rx signal which may very rapidly in phase and amplitude depending on the local objects and how mobile moves

Practical examples of small scale multipath fading

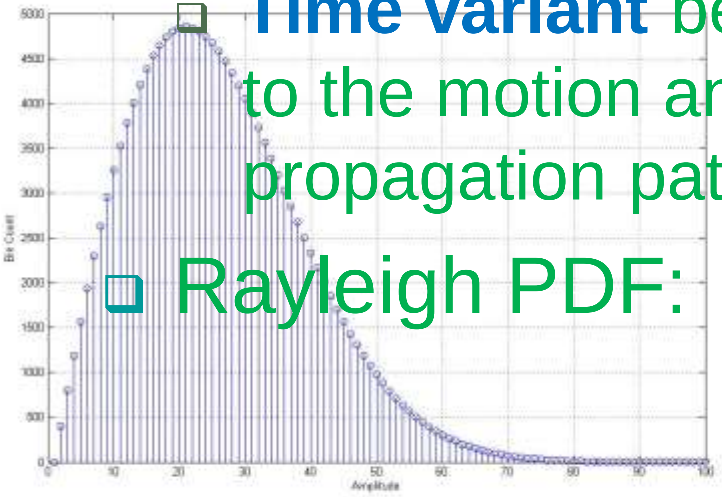
- Common examples of multipath fading are
  - temporary failure of communication due to a severe drop in the channel signal to noise ratio (You may have also experienced this. And you moved a steps away & noted that reception is better. It is due to small scale fading effects. ☺)
  - FM radio transmission experiencing intermittent loss of broadcast when away from station.

Multipath Fading- Most difficult

- Fades of 40 dB or more below local average level are frequent, with successive nulls occurring every half wavelength or so
- Referred to as Rayleigh Fading

Rayleigh Fading Mechanism

- Rayleigh fading manifests in two mechanism
  - **Time spreading** due to multipath (time dispersion)
  - **Time variant** behaviour of the channel due to the motion and subsequent changes in propagation paths
- Rayleigh PDF:



Rayleigh Fading

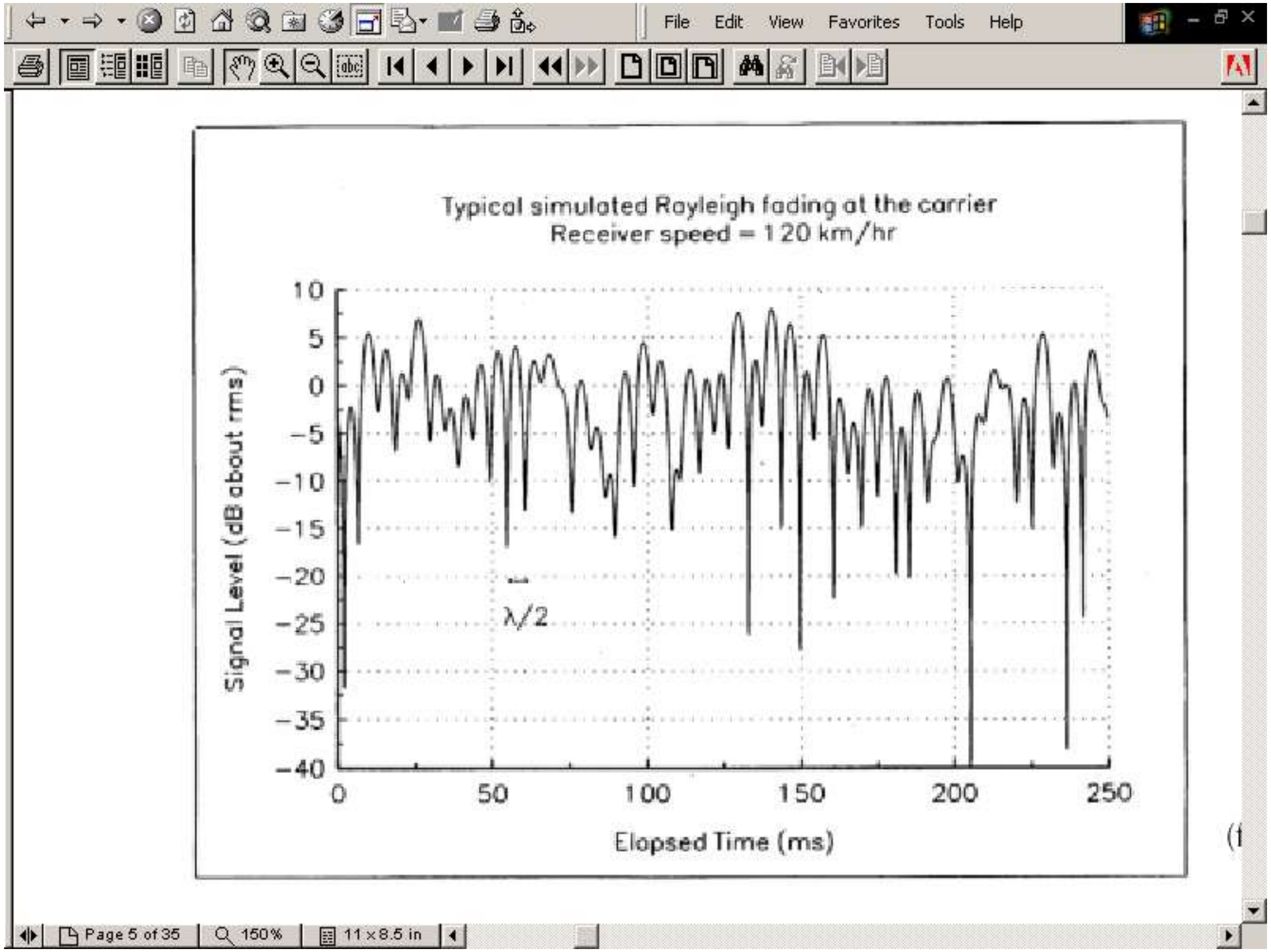
- The Rayleigh pdf is

$$p(r) = \begin{cases} \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right) & for\ r > 0 \\ 0 & otherwise \end{cases}$$

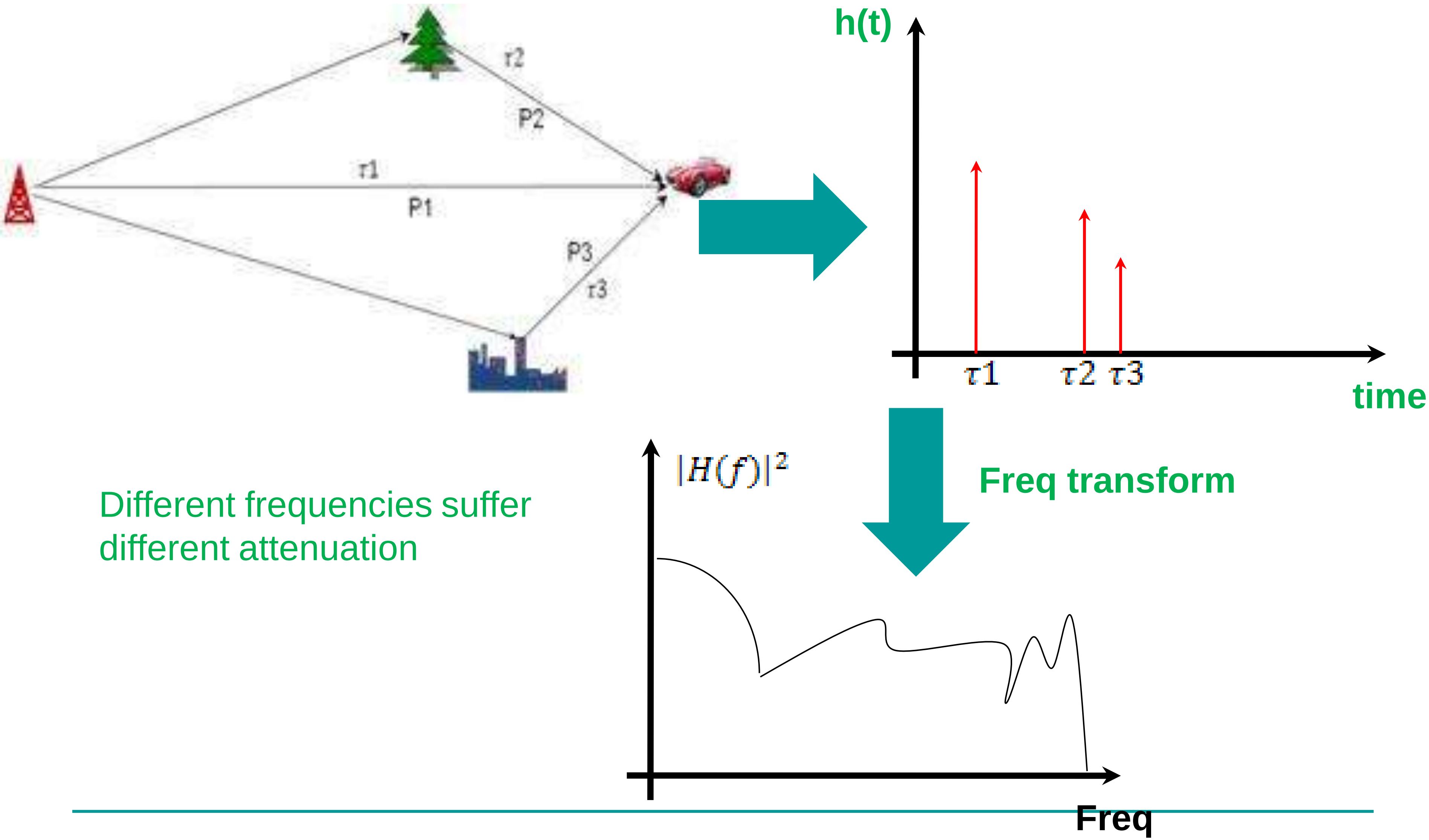
$Where\ r\ is\ the\ envelope\ amplitude\ of\ Rx\ signal\ \&$
$2\sigma^2\ is\ the\ power\ of\ the\ signal$

With Rayleigh Fading

Rayleigh Fading waveform envelope

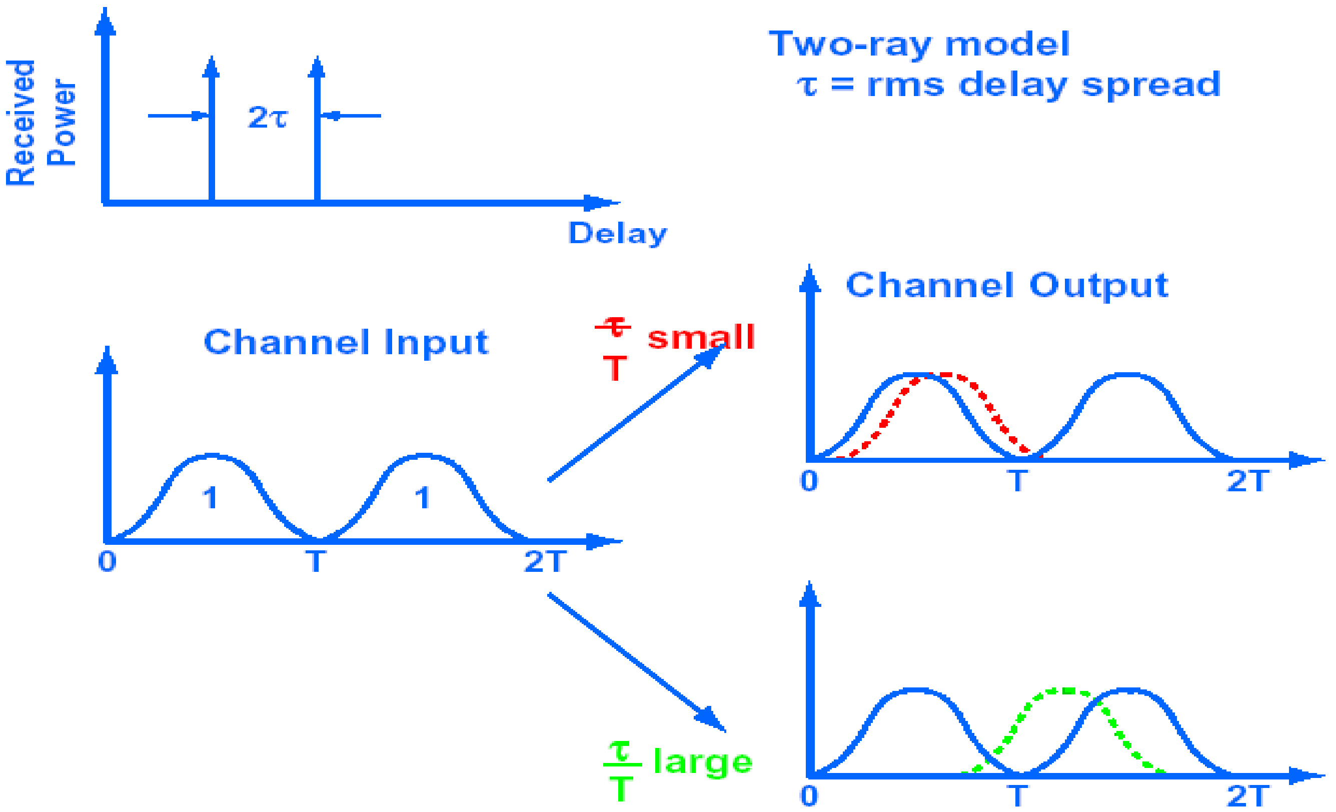


Time Dispersion phenomenon



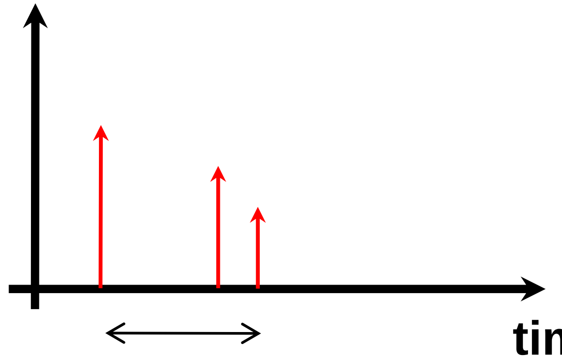
Different frequencies suffer different attenuation

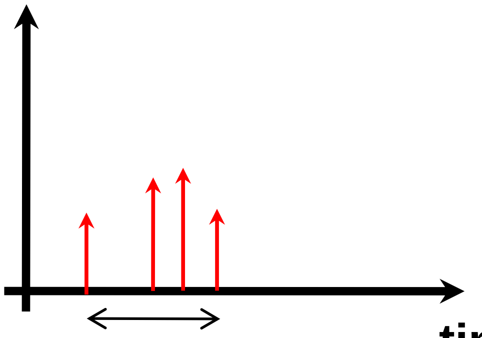
Delay Spread – Time Domain interpretation



- $\frac{\tau}{T}$ small ⇒ negligible intersymbol interference
- $\frac{\tau}{T}$ large ⇒ significant intersymbol interference, which causes an irreducible error floor

Delay Spread





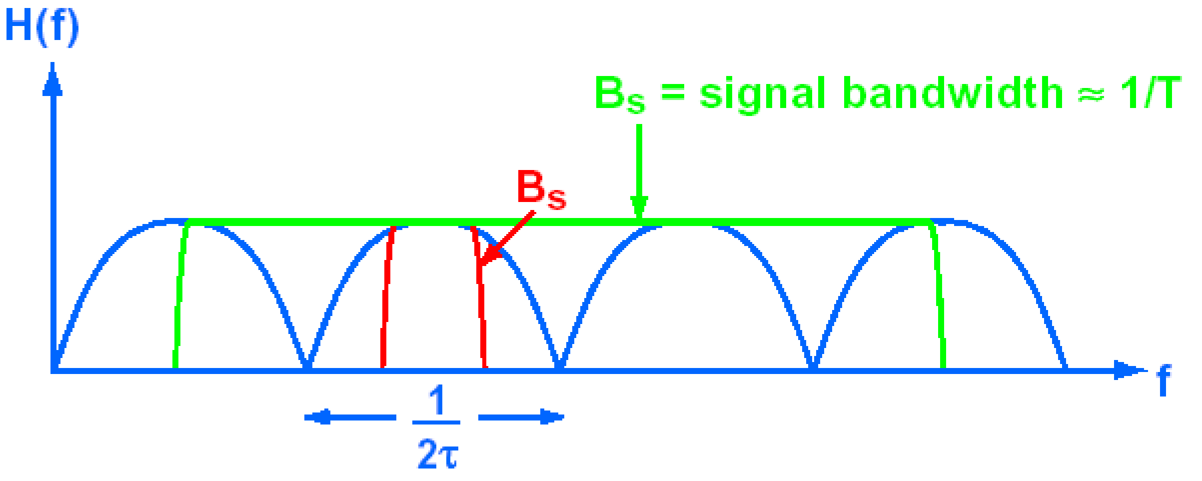
- Multiple impulses of varying power correspond to various multipaths. This time dispersion is also referred to as multipath delay spread.
- Delay between first significant path & last significant paths is loosely termed as channel excess delay spread.
- Two totally different channels can have same excess delay spread.
- A better measure of delay spread is rms delay spread σ_τ
  - L is the number of paths & β_i is the amplitude of the path i arriving at time τ_i

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

$$\overline{\tau^2} = \frac{\sum_{i=1}^{L} \tau_i^{\,2} \beta_i^{\,2}}{\sum_{i=1}^{L} \beta_i^{\,2}}$$

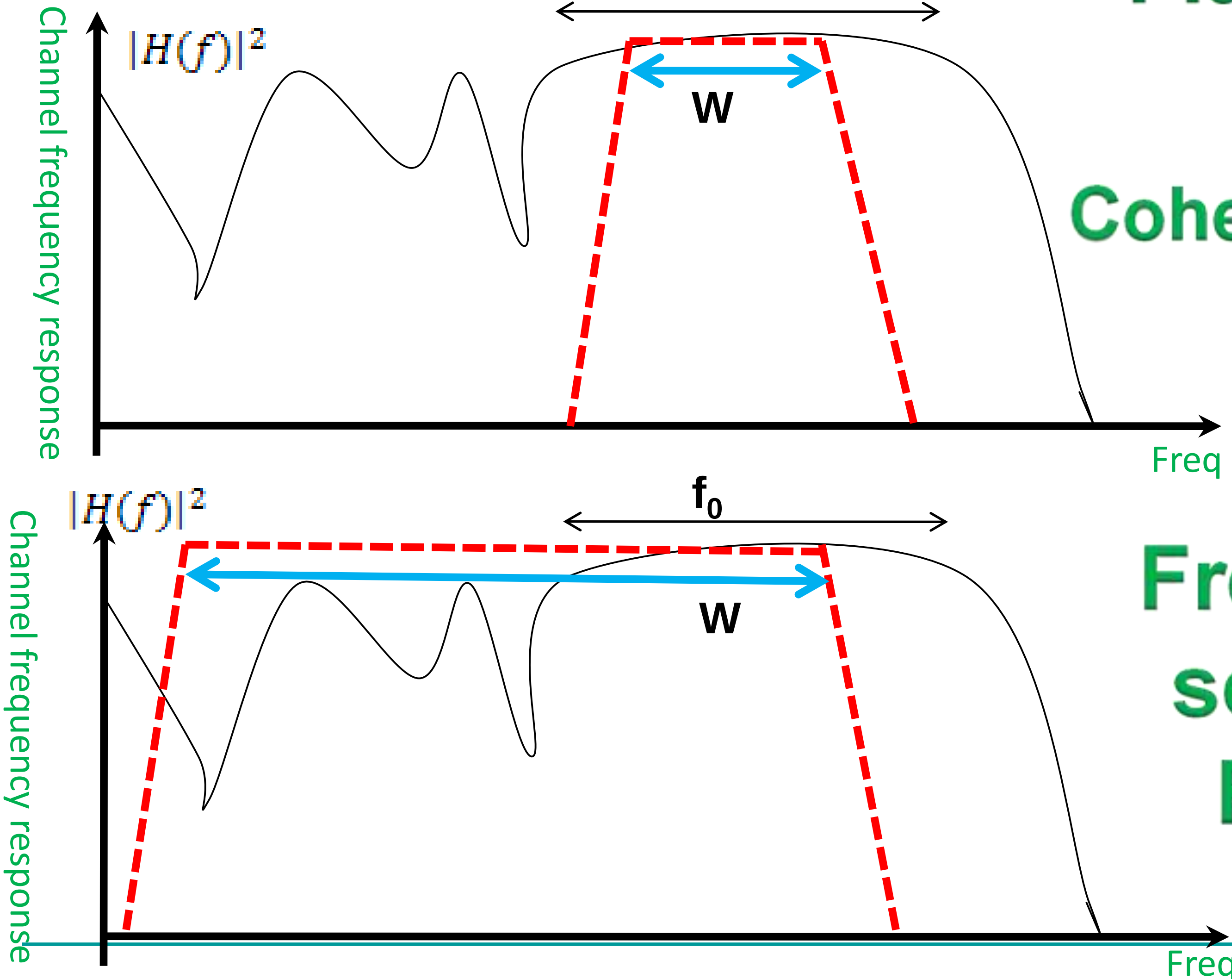
$\overline{\tau^2}$ is the second moment

Delay Spread- Freq. Domain Interpretation



- $\frac{\tau}{T}$ small ➡ flat fading
- $\frac{\tau}{T}$ large ➡ frequency-selective fading

Time spreading : Coherence Bandwidth



Flat Fading

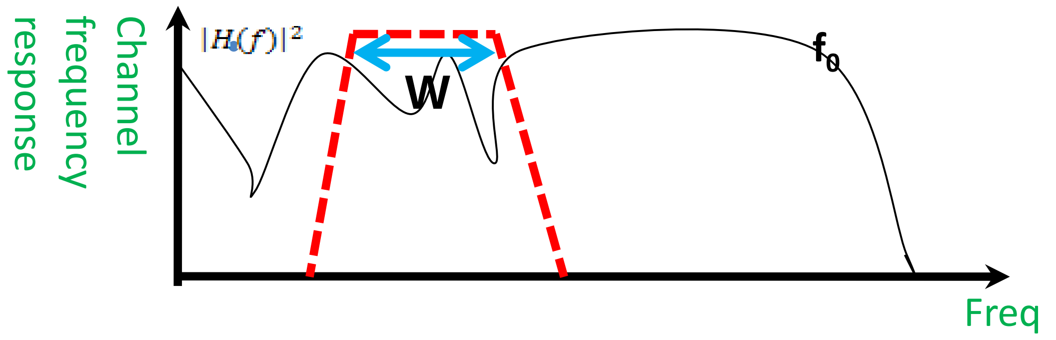
$W < f_0$

Coherence BW = f_0

Frequency selective Fading

$W > f_0$

More on flat fading



Condition f0 > W does not guarantee flat fading. As shown above, frequency nulls (frequency selective fading) may be there occasionally even though f0 > W.

Similarly, frequency selective fading channel may also show flat fading sometimes.

Bit Rate Limitations by Delay Spread

- QPSK modulation
- Bit error rate is 10^{-4}

	τ	Maximum Bit Rate
Mobile (rural)	25 μs	8 kbps
Mobile (city)	2.5 μs	80 kbps
Microcells	500 ns	400 kbps
Large Building	100 ns	2 Mbps

Coherence Bandwidth and delay spread

- There is no exact relationship between Coherence bandwidth and delay spread. For at least 0.9 correlation for channel's complex frequency transfer function, Coherence bandwidth f_0 is approximated by following relation: $$f_0 \approx \frac{1}{50\sigma_\tau}$$ **Where σ_τ is r.m.s. delay spread**

- For dense scatterer model which is useful for urban surroundings, coherence bandwidth is defined as assuming at least 0.5 correlation: $$f_0 \approx \frac{0.276}{\sigma_\tau}$$

- Another popular approximation assuming at least 0.5 correlation: $$f_0 \approx \frac{1}{5\sigma_\tau}$$

Effects of Flat & frequency selective fading

- Flat fading
  - Reduces SNR forcing various mitigation techniques to handle that. Not such a bad thing.
- Frequency selecting fading
  - ISI distortion (need equalizer in receiver)
  - Pulse mutilation
  - Irreducible BER

Summary of Time dispersion

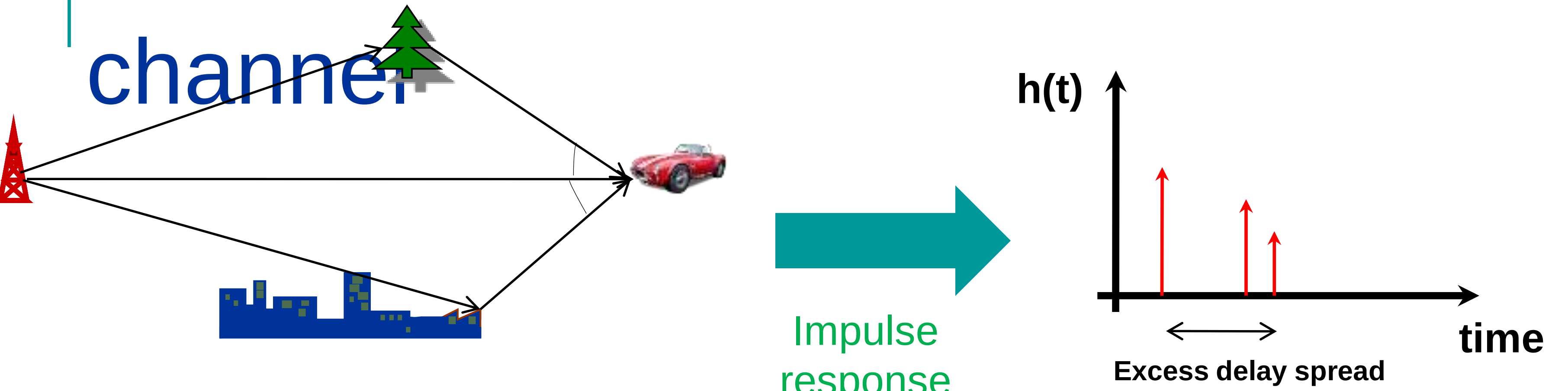
Small scale fading

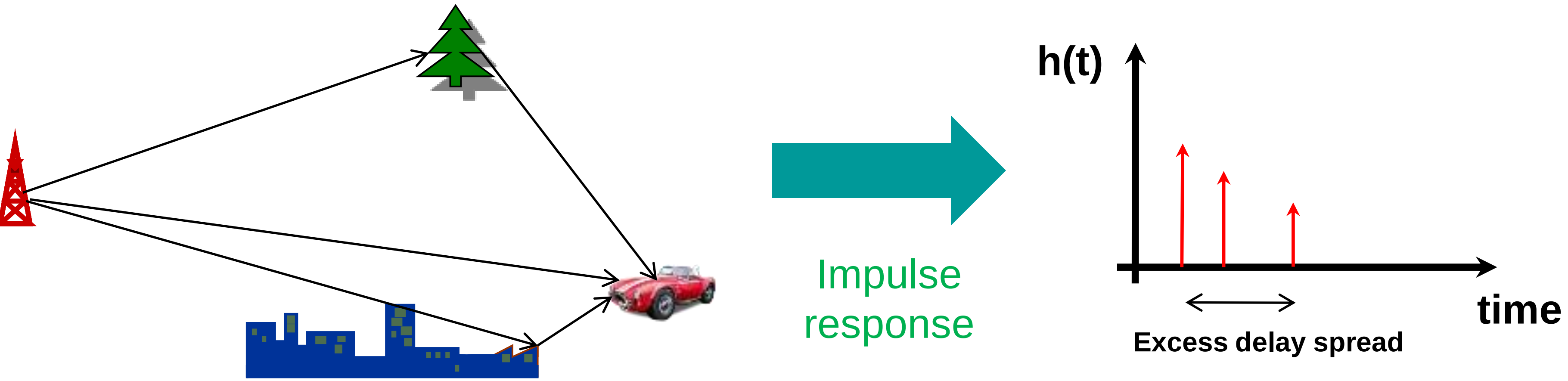
(based on multipath delay spread)

Flat Fading
BW of signal < BW of channel
Or
Delay Spread < Symbol period

Frequency selective Fading
BW of signal > BW of channel
Or
Delay Spread > Symbol period

Time variant behavior of the channel





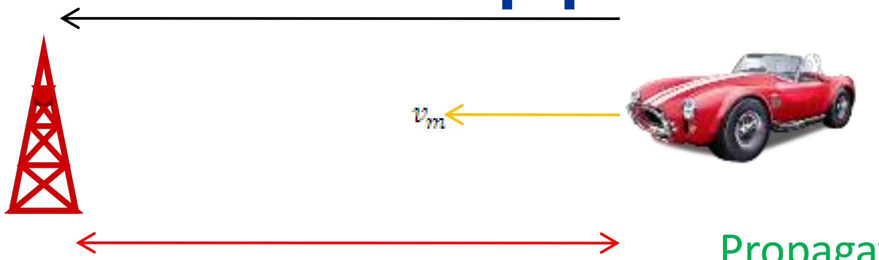
Relative movement between transmitter and receiver or objects between those causes variation in channel's characteristics over time. This happens due to propagation path change over time. Relative movement also creates frequency spreading due to Doppler effect

Time Variance

- Variance in channel conditions over time is an important factor when designing a mobile communication system.
- If fast variations happen, it can lead to severe pulse distortion and loss of SNR subsequently causing irreducible BER.

t)

Basic Doppler effect



$$\tau(t) = \frac{d(t)}{c} = \frac{d0 - v_m.t}{c} = \tau 0 - \frac{v_m.t}{c}$$

c is the light velocity and v_m is the car speed

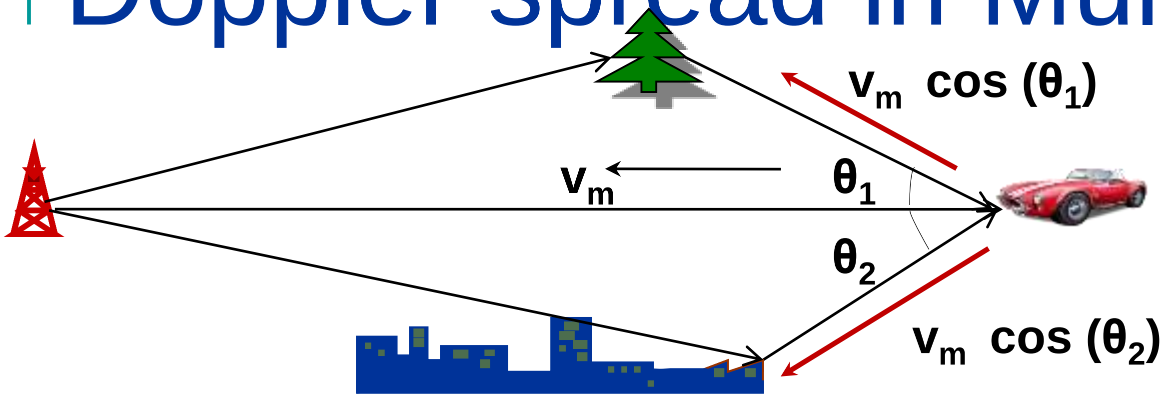
Propagation time is a function of time due to mobile car.

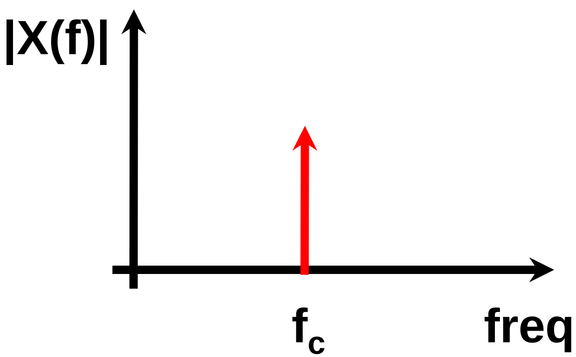
$$transmitted\ signal: \cos(2\pi f_c t)$$

$$Received\ signal: \cos[2\pi f_c(t - \tau(t))]$$
$$= \cos\left[2\pi f_c\left(t - \tau 0 + \frac{v_m.t}{c}\right)\right]$$
$$= \cos\left[2\pi(f_c + f_d)t - \emptyset\right]$$

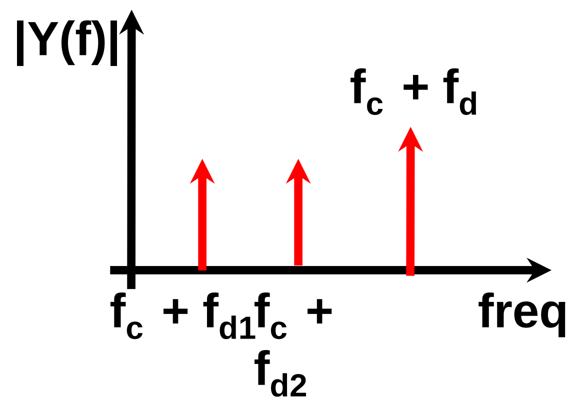
$$Doppler\ frequency\ shift\ f_d = \frac{v_m}{c}.f_c$$

Doppler spread in Multipath



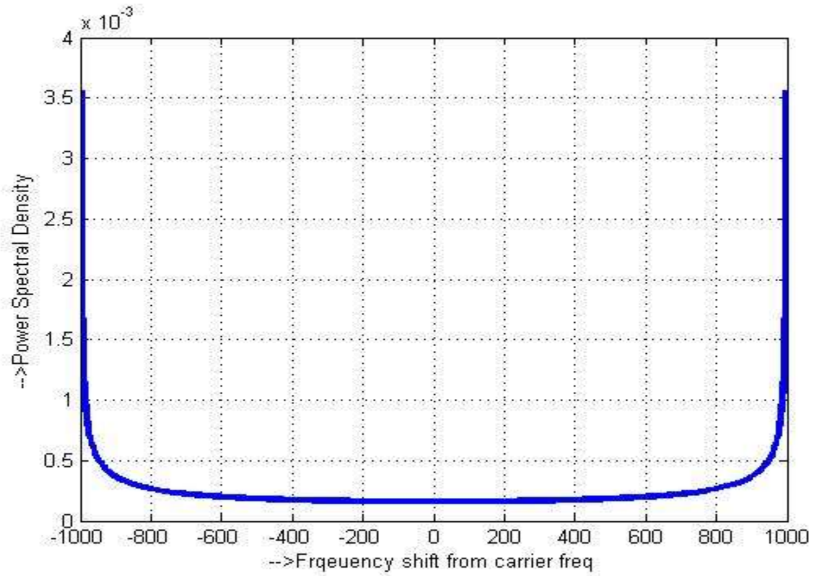


After passing through multipath channel



Due to multipaths, a single sinusoid by base station is perceived as summation of 3 sinusoids f_c+f_{d1}, f_c+f_{d2} and f_c+f_d , where f_d is maximum doppler frequency = $f_c*(v_m/d)$. Due to different arrivals of angle due to multipaths, perceived velocity is different for multipaths.

Doppler Spectrum



Imagine now multiple paths with different angles of arrival causing amagamalation of various frequencies between $f_c + f_d$ & $f_c - f_d$.

A popular model assumes that distribution of angle of arrival is distributed uniformly between 0 & 2π which leads to following spectrum

$$D(f) = \frac{1}{2\pi f_d \sqrt{1 - (\frac{f}{f_d})^2}}$$

This is called classical Doppler spectrum & shows how a single sinusoid ends up having a broad spectrum due to multipath & relative motion between Tx and Rx.

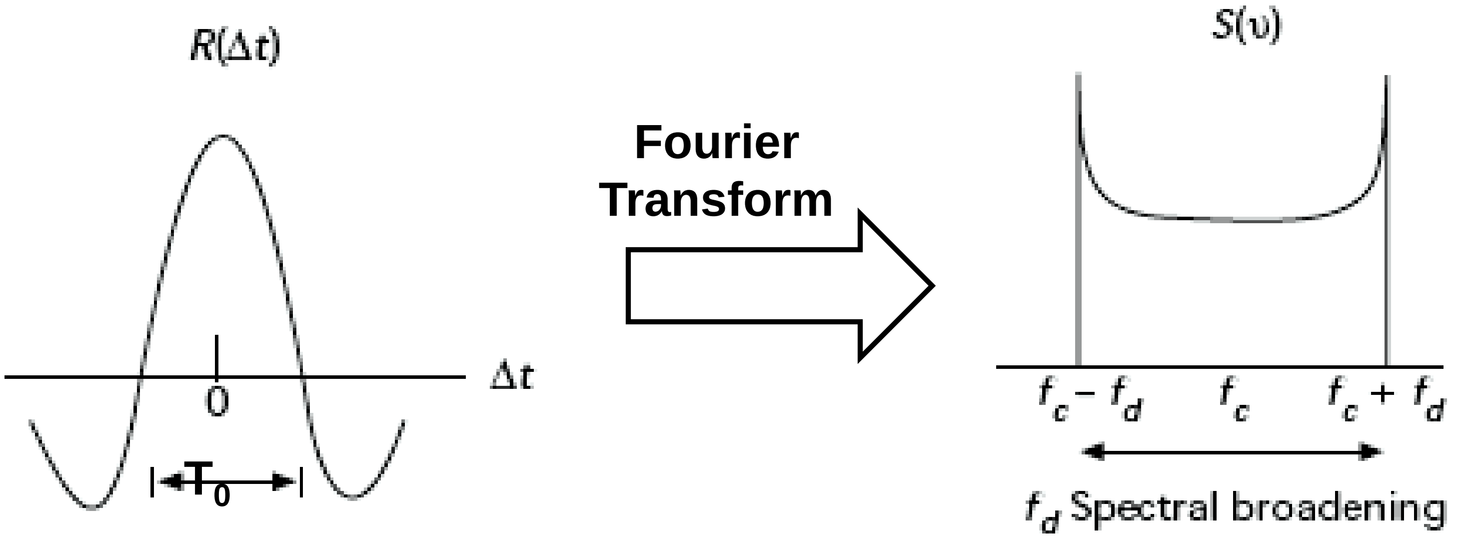
Time variant Channel: Coherence Time

- **Maximum doppler frequency is an important measure of time variance of channel characteristics. It depends on relative speed of any movement between Tx & Rx and the carrier frequency**
- **Coherence time: Approximate time duration over which the channel's response remains invariant**

$$T_0 \approx \frac{1}{f_d}$$

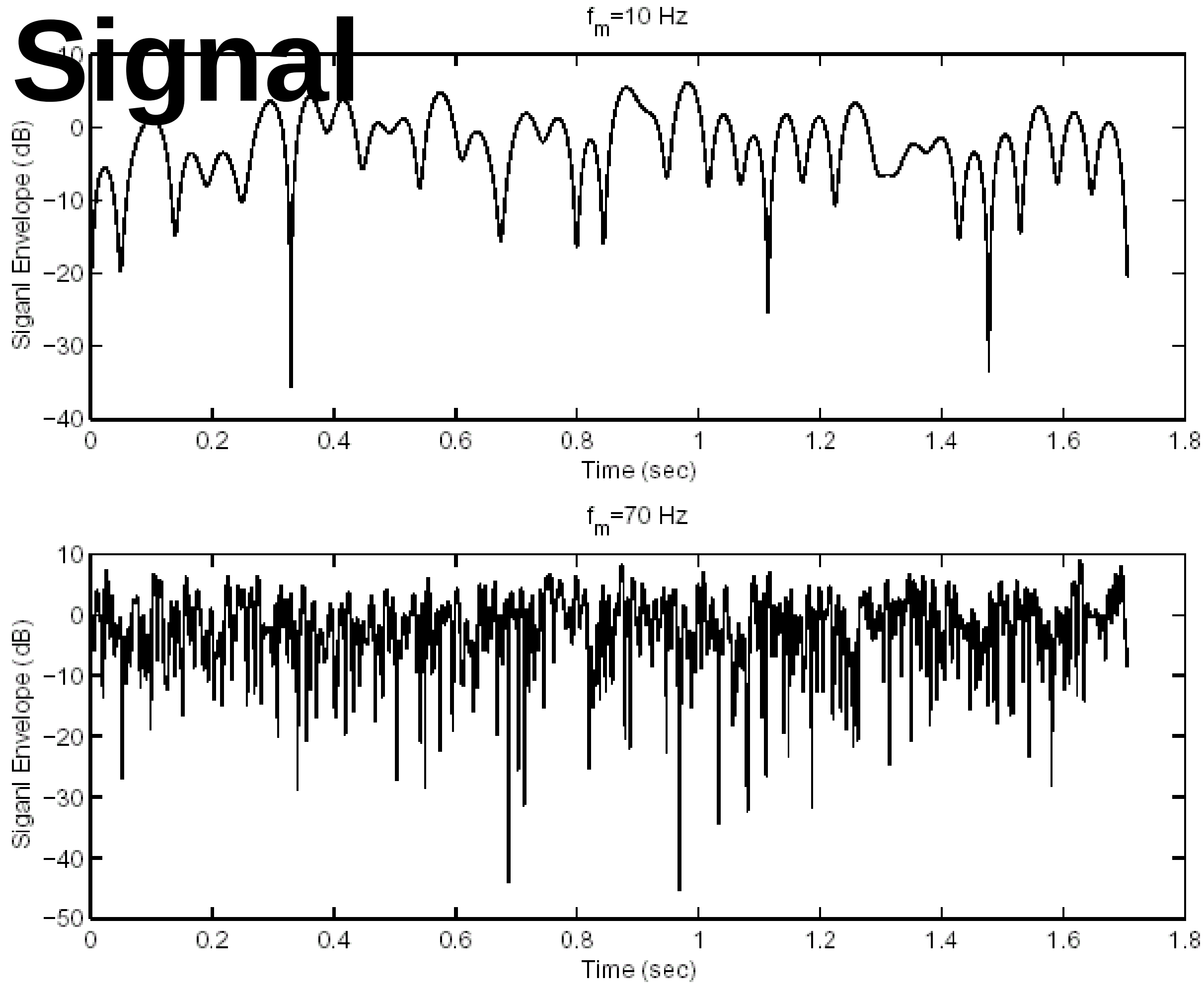
- **Where is Maximum Doppler Frequency**

Frequency Dual



Function $R(\Delta t)$ denotes space time correlation for the channel response to a sinusoid . So this indicates the amount of correlation between two sinusoids sent at different times t_1 & t_2 .

Waveform of Rayleigh Fading Signal



Rayleigh fading envelopes of signals with different maximum Doppler frequencies at carrier frequency of 800 MHz

Time Variance : Fast Fading

Fast Fading :

$T_0 < T_s$ Where T_s : Transmitted Symbol time

Or

$f_d > W$ Where W: Transmitted bandwidth

Above relationship means that channel changes drastically many times while a symbol is propagating;

Only highly mobile systems (~500 Km/Hr) will have fd ~1 kHz so systems having signalling rate of that order will be fast fading.

Impact of fast fading:

- Severe distortion of baseband pulse leading to detection problems
- Loss in SNR
- Synchronization problems (e.g. Failure of PLL)

Time variance: Slow Fading

Slow Fading :

$T_0 > T_s$ where T_s : Transmitted Symbol time

Or

$f_d < W$ where W: Transmitted bandwidth

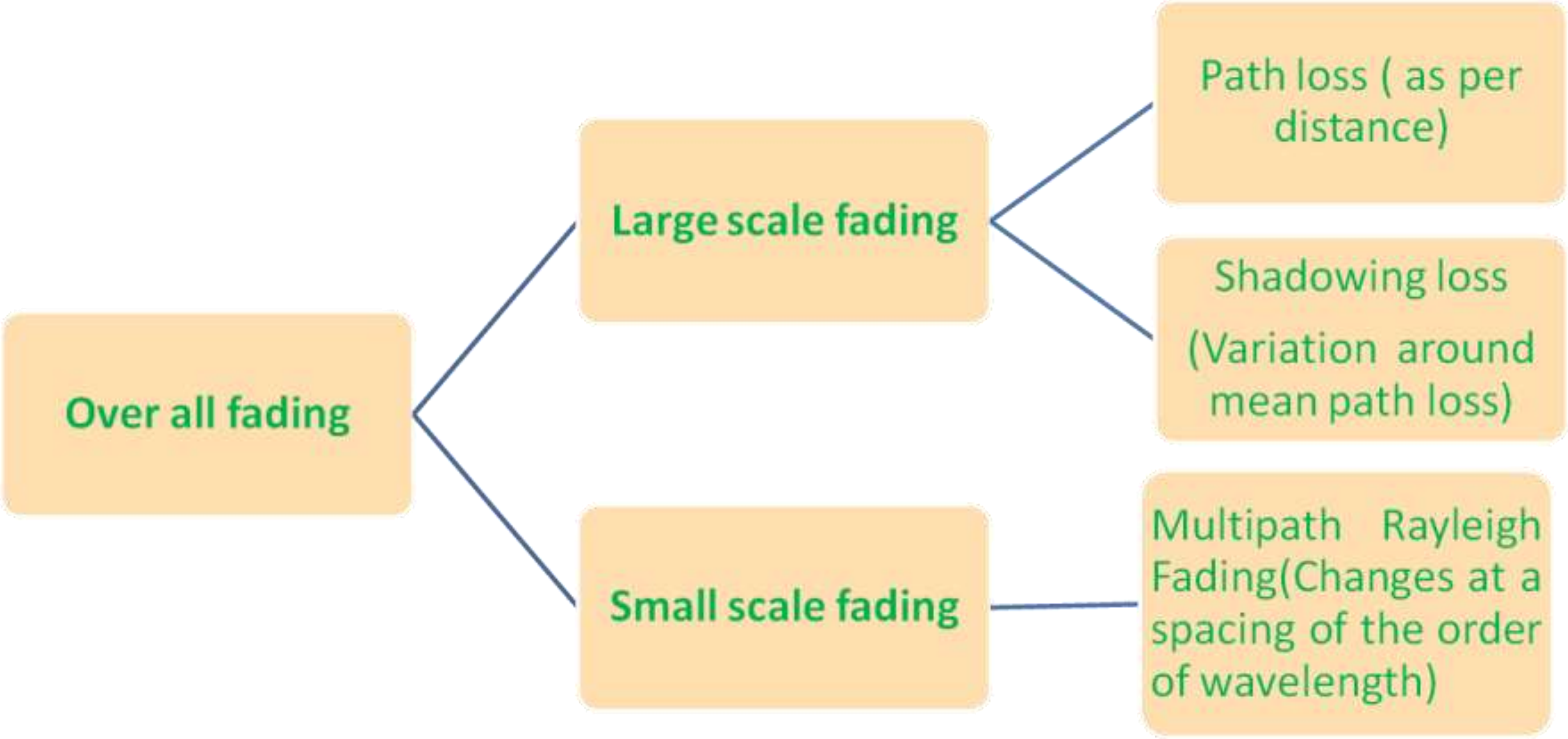
Above relationship means that channel does not change drastically during symbol duration

Most of the modern communication systems are slow fading channels

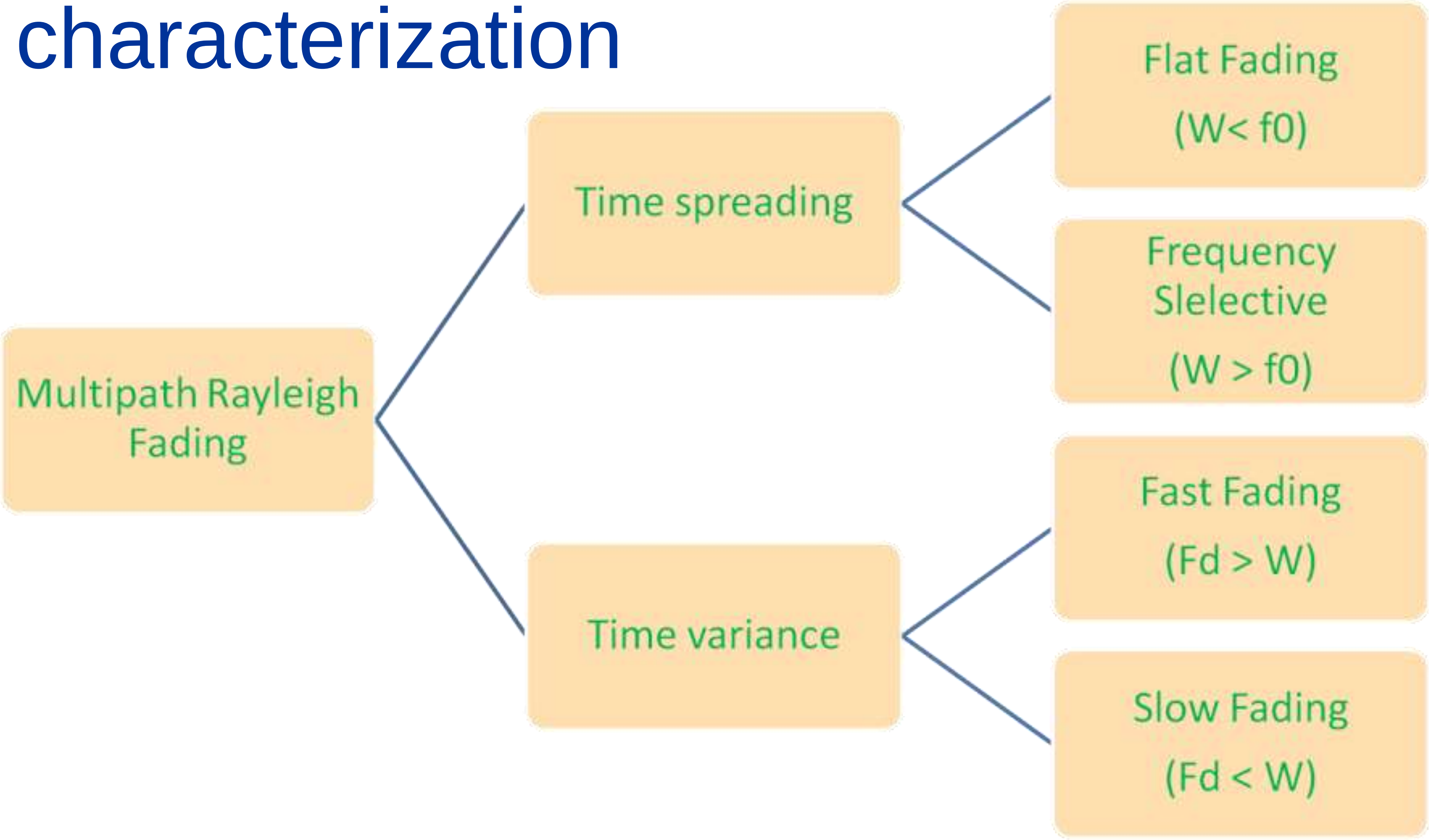
Impact of fast fading:

- ❑Loss in SNR

Summary of Overall Fading



Summary of Multipath Fading characterization



Multiple Access Techniques for Wireless Communication

FDMA

TDMA

SDMA

PDMA

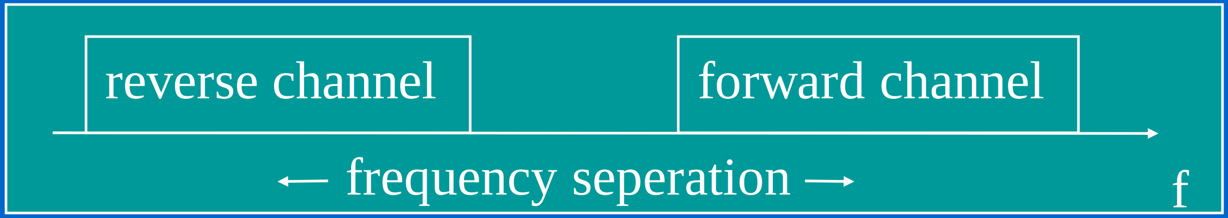
Click to add text

Introduction

- many users at same time
- share a finite amount of radio spectrum
- high performance
- duplexing generally required
- frequency domain
- time domain

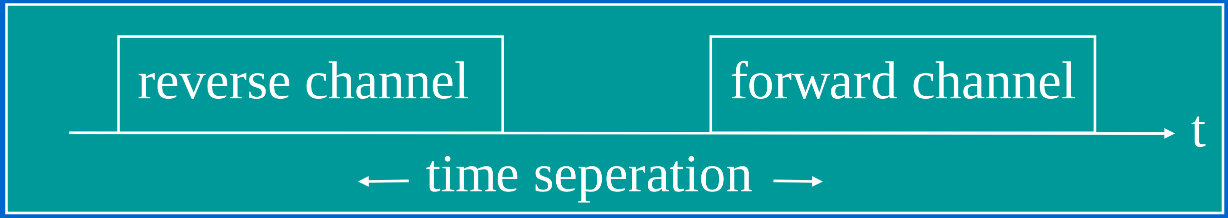
Frequency division duplexing (FDD)

- two bands of frequencies for every user
- forward band
- reverse band
- duplexer needed
- frequency seperation between forward band and reverse band is constant



Time division duplexing (TDD)

- uses time for forward and reverse link
- multiple users share a single radio channel
- forward time slot
- reverse time slot
- no duplexer is required



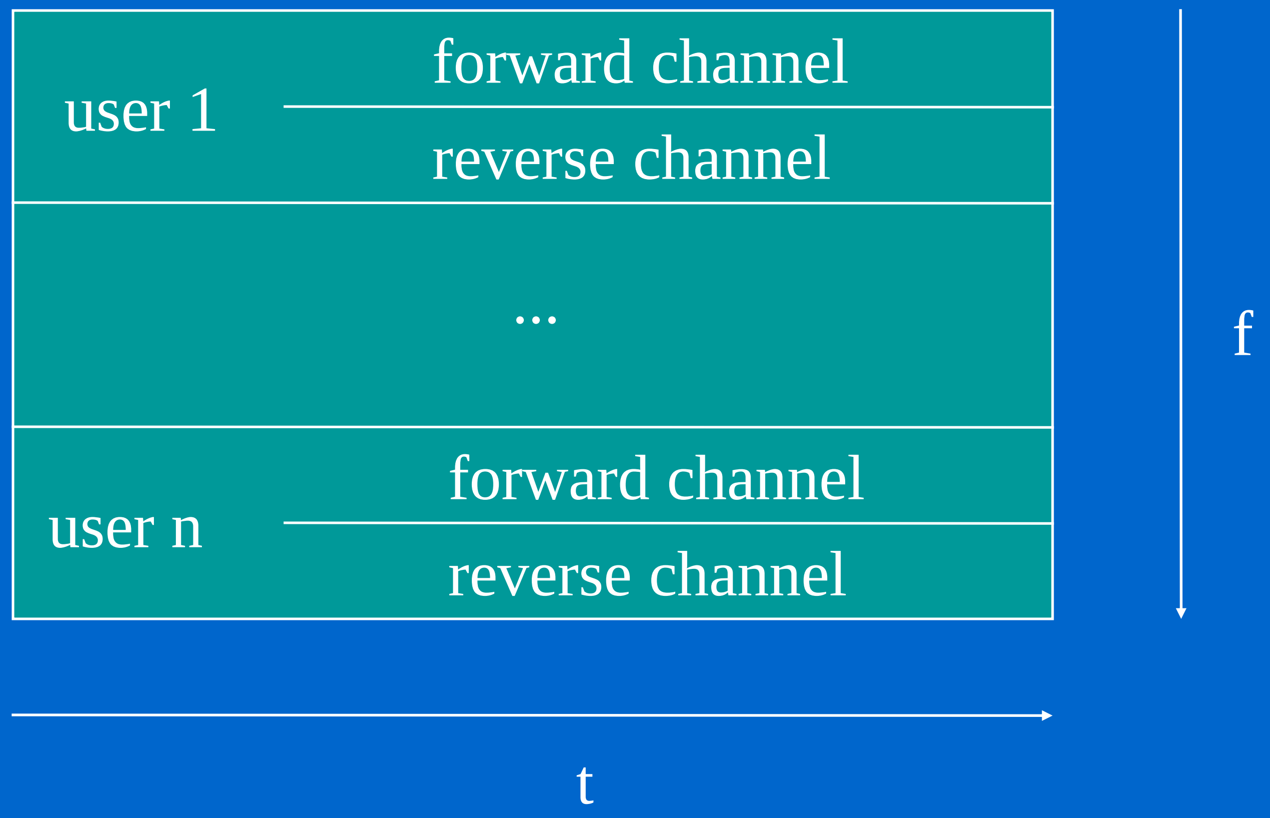
Multiple Access Techniques

- Frequency division multiple access (FDMA)
- Time division multiple access (TDMA)
- Code division multiple access (CDMA)
- Space division multiple access (SDMA)
- grouped as:
- narrowband systems
- wideband systems

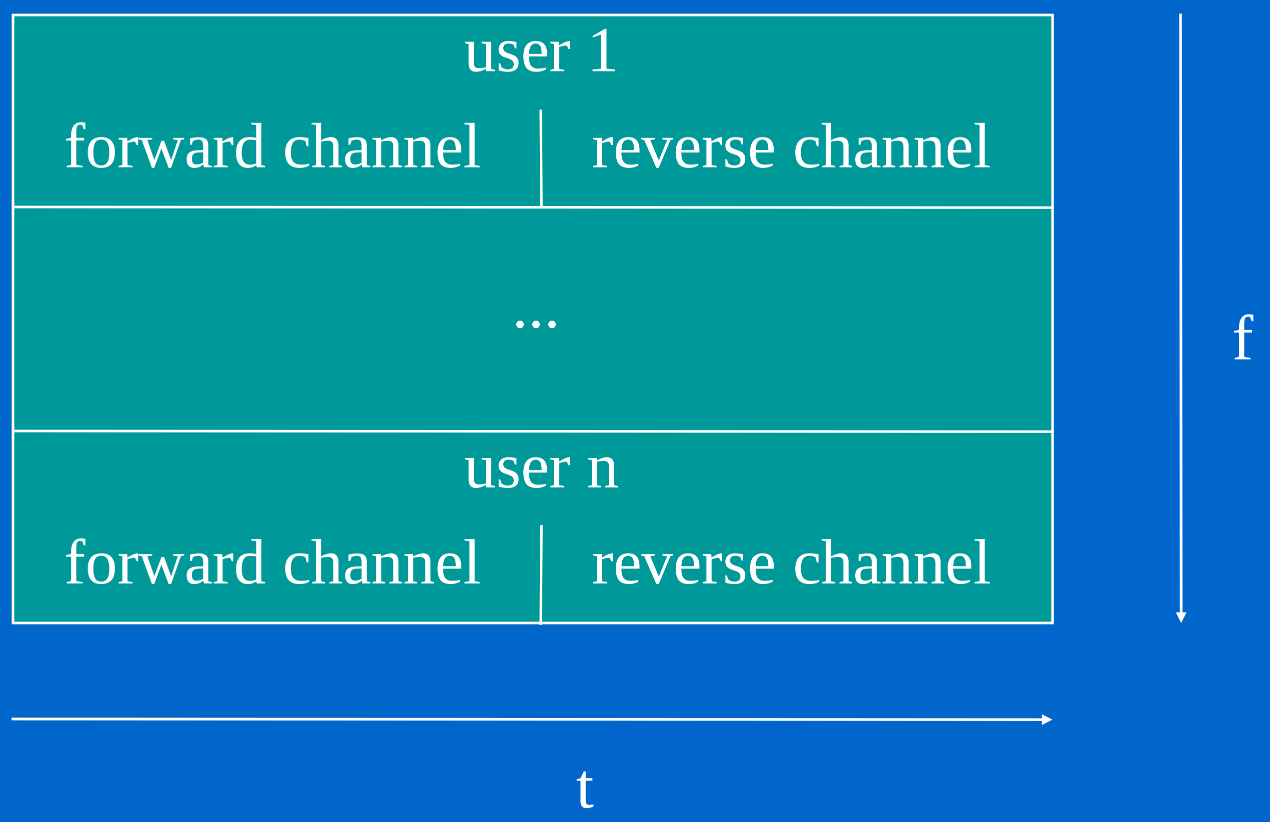
Narrowband systems

- large number of narrowband channels
- usually FDD
- Narrowband FDMA
- Narrowband TDMA
- FDMA/FDD
- FDMA/TDD
- TDMA/FDD
- TDMA/TDD

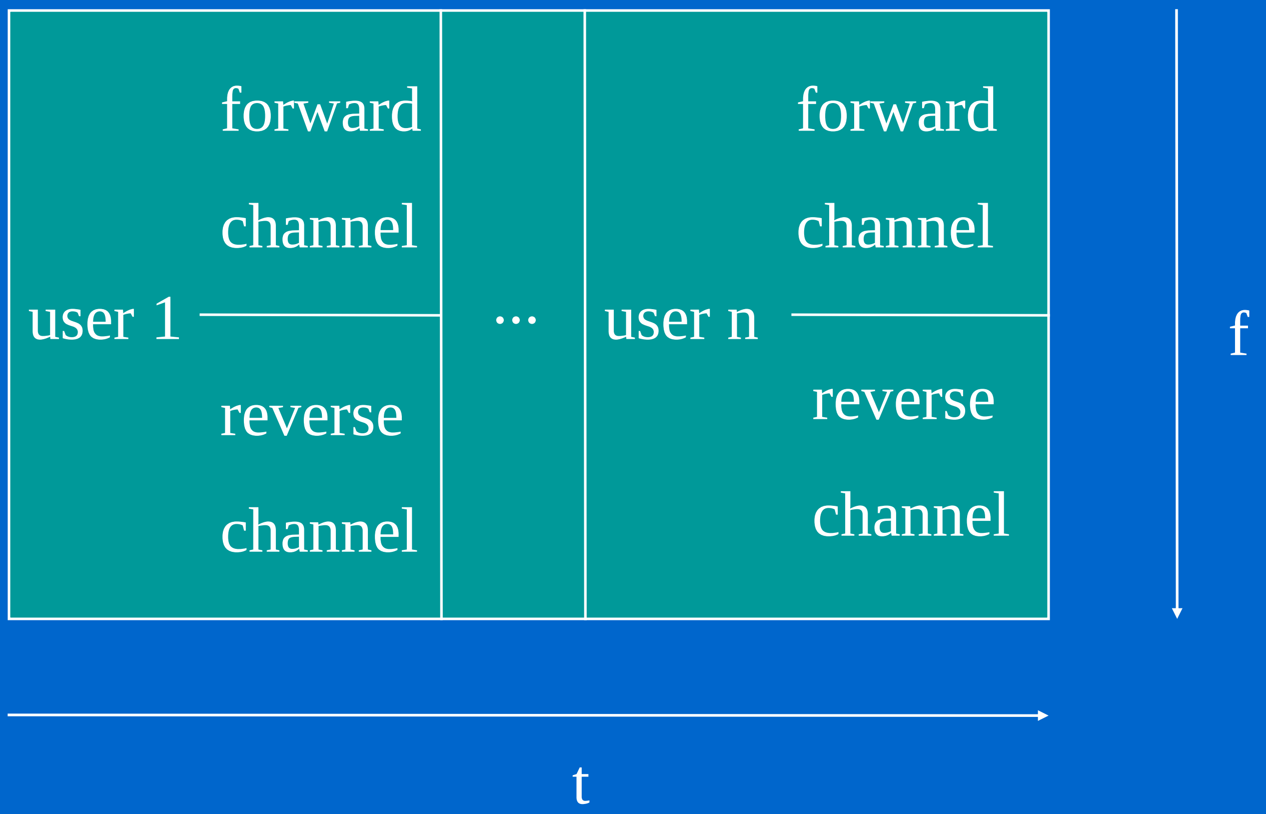
Logical separation FDMA/FDD



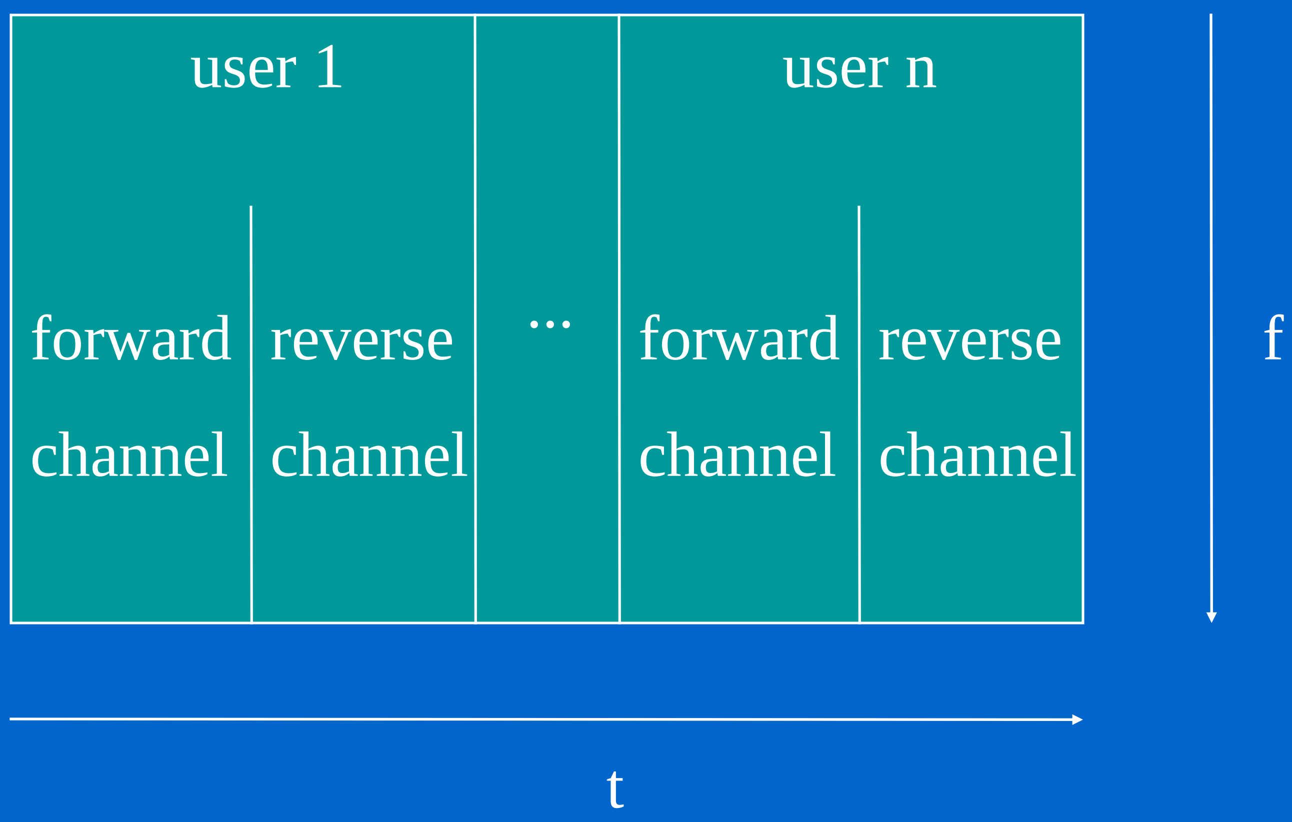
Logical separation FDMA/TDD



Logical separation TDMA/FDD



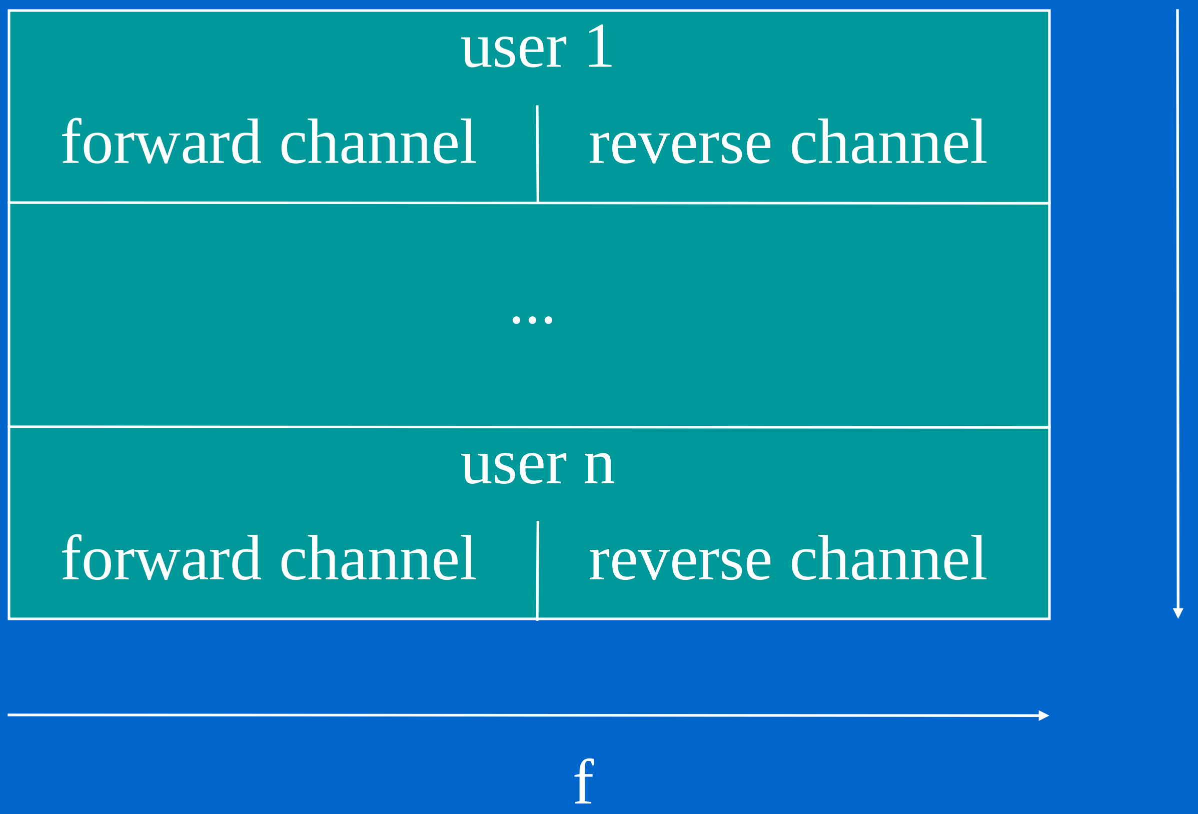
Logical separation TDMA/TDD



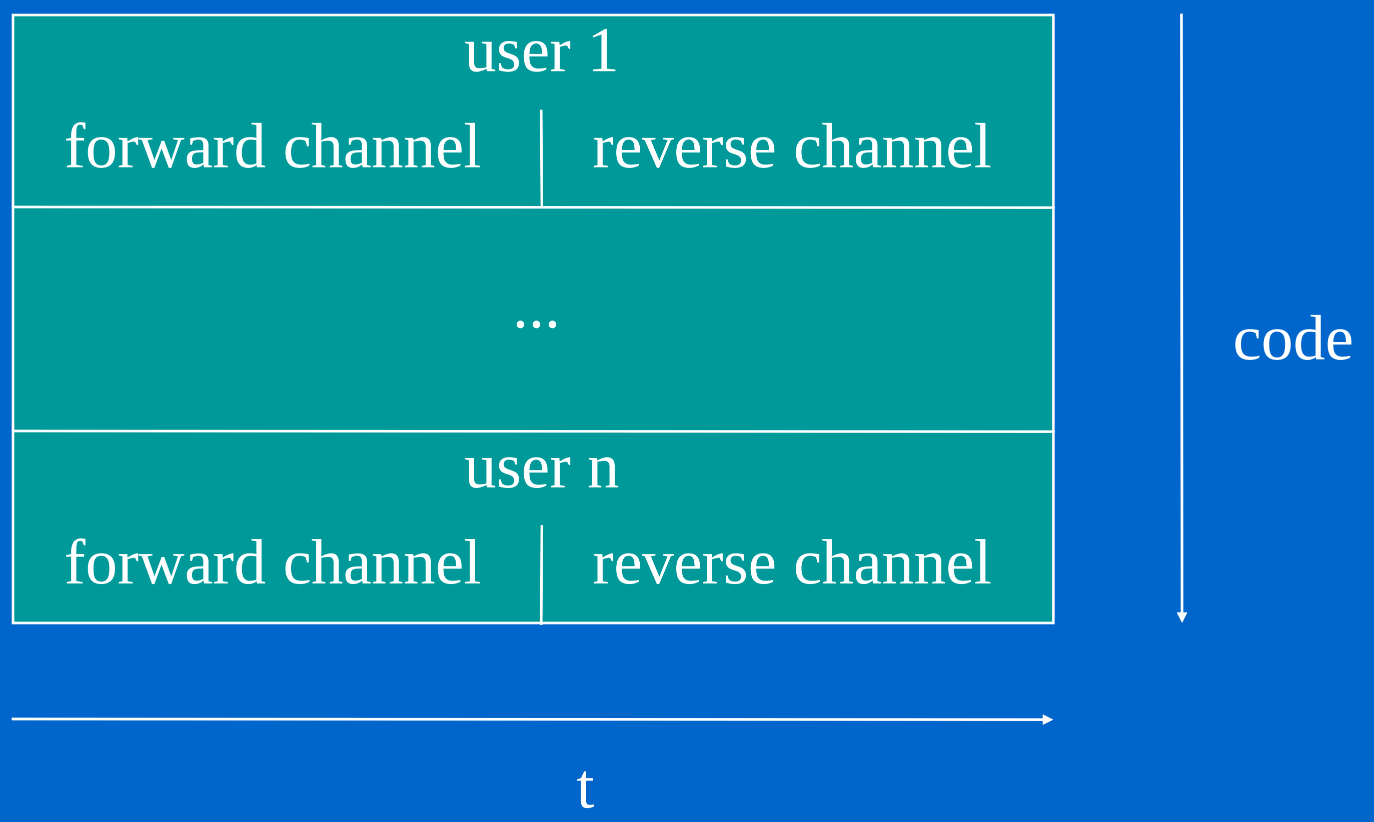
Wideband systems

- large number of transmitters on one channel
- TDMA techniques
- CDMA techniques
- FDD or TDD multiplexing techniques
- TDMA/FDD
- TDMA/TDD
- CDMA/FDD
- CDMA/TDD

Logical separation CDMA/FDD



Logical separation CDMA/TDD



Multiple Access Techniques in use

Cellular System	Multiple Access Technique
Advanced Mobile Phone System (AMPS)	FDMA/FDD
Global System for Mobile (GSM)	TDMA/FDD
US Digital Cellular (USDC)	TDMA/FDD
Digital European Cordless Telephone (DECT)	FDMA/TDD
US Narrowband Spread Spectrum (IS-95)	CDMA/FDD

Frequency division multiple access FDMA

- one phone circuit per channel
- idle time causes wasting of resources
- simultaneously and continuously transmitting
- usually implemented in narrowband systems
- for example: in AMPS is a FDMA bandwidth of 30 kHz implemented

FDMA compared to TDMA

- fewer bits for synchronization
- fewer bits for framing
- higher cell site system costs
- higher costs for duplexer used in base station and subscriber units
- FDMA requires RF filtering to minimize adjacent channel interference

Nonlinear Effects in FDMA

- many channels - same antenna
- for maximum power efficiency operate near saturation
- near saturation power amplifiers are nonlinear
- nonlinearities causes signal spreading
- intermodulation frequencies

Nonlinear Effects in FDMA

- IM are undesired harmonics
- interference with other channels in the FDMA system
- decreases user C/I - decreases performance
- interference outside the mobile radio band: adjacent-channel interference
- RF filters needed - higher costs

Number of channels in a FDMA system

$$N = \frac{B_t - B_{guard}}{B_c}$$

- N … number of channels
- B_t … total spectrum allocation
- B_{guard} … guard band
- B_c … channel bandwidth

Example: Advanced Mobile Phone System

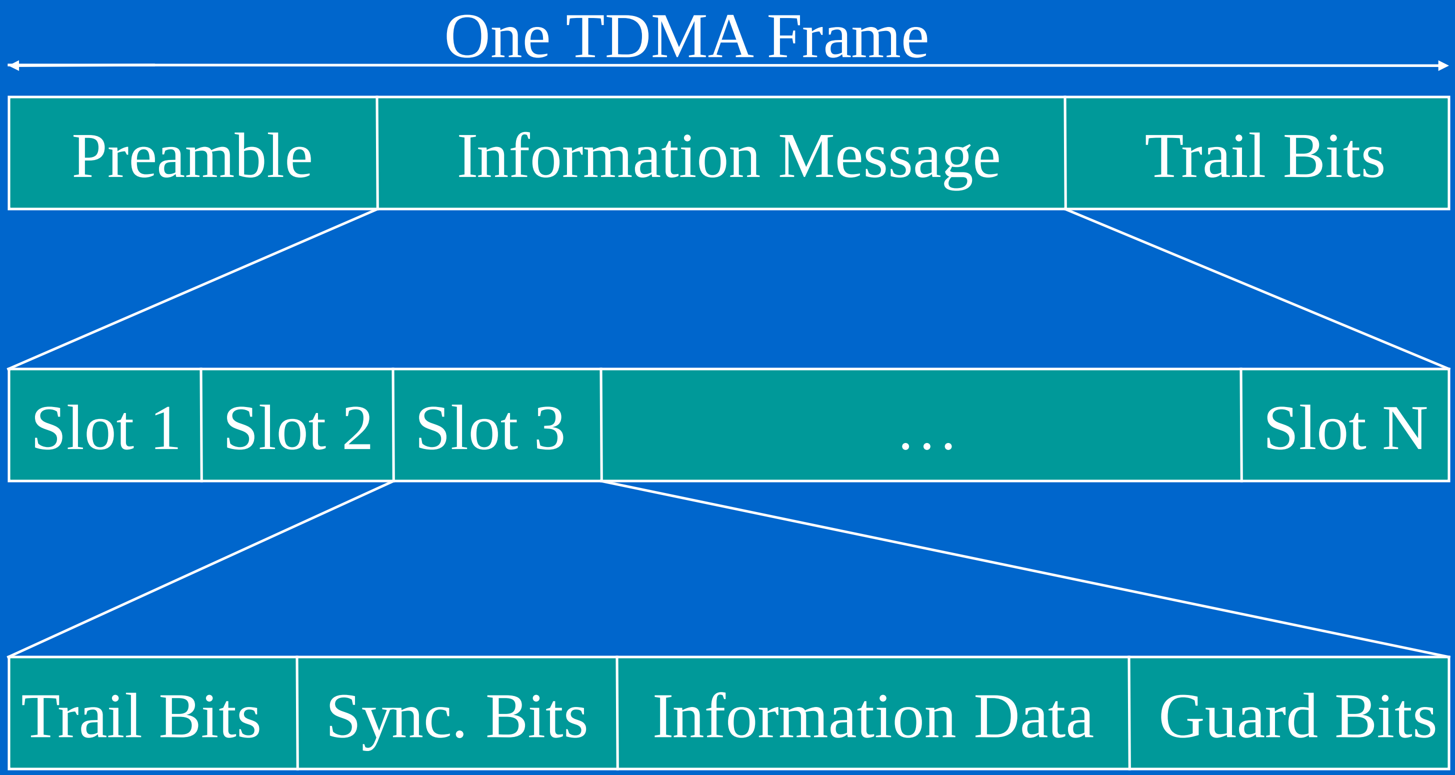
- AMPS
- FDMA/FDD
- analog cellular system
- 12.5 MHz per simplex band - B_t
- B_{guard} = 10 kHz ; B_c = 30 kHz

$$N = \frac{12.5E6 - 2*(10E3)}{30E3} = 416 \text{ channels}$$

Time Division Multiple Access

- time slots
- one user per slot
- buffer and burst method
- noncontinuous transmission
- digital data
- digital modulation

Repeating Frame Structure



The frame is cyclically repeated over time.

Features of TDMA

- a single carrier frequency for several users
- transmission in bursts
- low battery consumption
- handoff process much simpler
- FDD : switch instead of duplexer
- very high transmission rate
- high synchronization overhead
- guard slots necessary

Number of channels in a TDMA system

$$N = \frac{m*(B_{tot} - 2*B_{guard})}{B_c}$$

- N … number of channels
- m … number of TDMA users per radio channel
- B_{tot} … total spectrum allocation
- B_{guard} … Guard Band
- B_c … channel bandwidth

Example: Global System for Mobile (GSM)

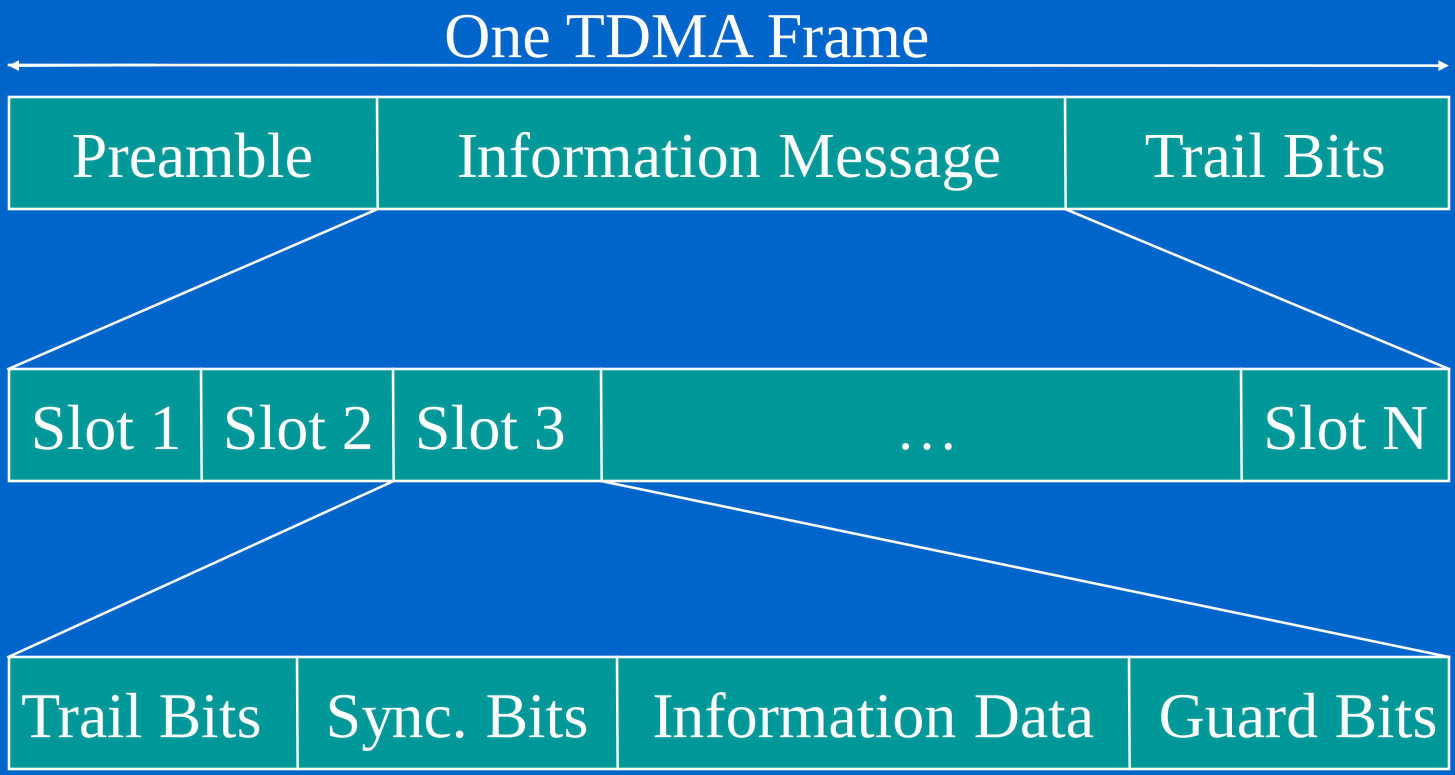
- TDMA/FDD
- forward link at B_{tot} = 25 MHz
- radio channels of B_c = 200 kHz
- if m = 8 speech channels supported, and
- if no guard band is assumed :

$$N = \frac{8*25E6}{200E3} = 1000 \text{ simultaneous users}$$

Efficiency of TDMA

- percentage of transmitted data that contain information
- frame efficiency ♒$_f$
- usually end user efficiency < ♒$_f$,
- because of source and channel coding
- How get ♒$_f$?

Repeating Frame Structure



The frame is cyclically repeated over time.

Efficiency of TDMA

$$b_{OH} = N_r*b_r + N_t*b_p + N_t*b_g + N_r*b_g$$

- b_{OH} … number of overhead bits
- N_r … number of reference bursts per frame
- b_r … reference bits per reference burst
- N_t … number of traffic bursts per frame
- b_p … overhead bits per preamble in each slot
- b_g … equivalent bits in each guard time intervall

Efficiency of TDMA

$$b_T = T_f * R$$

- b_T … total number of bits per frame
- T_f … frame duration
- R … channel bit rate

Efficiency of TDMA

$$\eta_f = (1 - b_{OH}/b_T) * 100\%$$

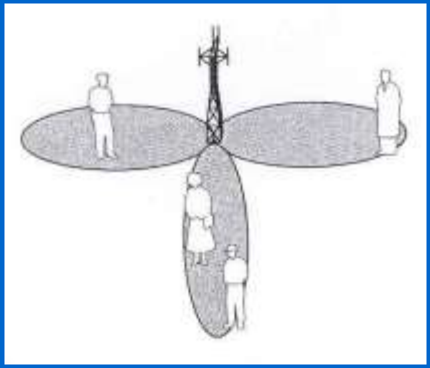
- η_f … frame efficiency
- b_{OH} … number of overhead bits per frame
- b_T … total number of bits per frame

Space Division Multiple Access

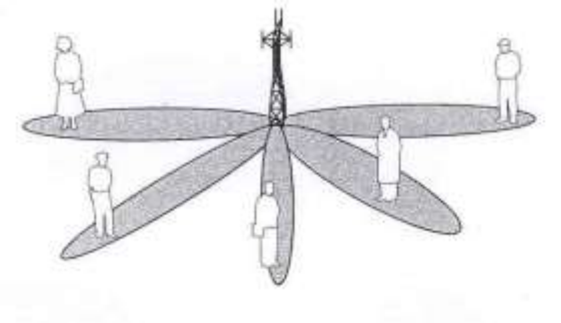
- Controls radiated energy for each user in space
- using spot beam antennas
- base station tracks user when moving
- cover areas with same frequency:
- TDMA or CDMA systems
- cover areas with same frequency:
- FDMA systems

Space Division Multiple Access

- primitive applications are "Sectorized antennas"



- in future adaptive antennas simultaneously steer energy in the direction of many users at once



Reverse link problems

- general problem
- different propagation path from user to base
- dynamic control of transmitting power from each user to the base station required
- limits by battery consumption of subscriber units
- possible solution is a filter for each user

Solution by SDMA systems

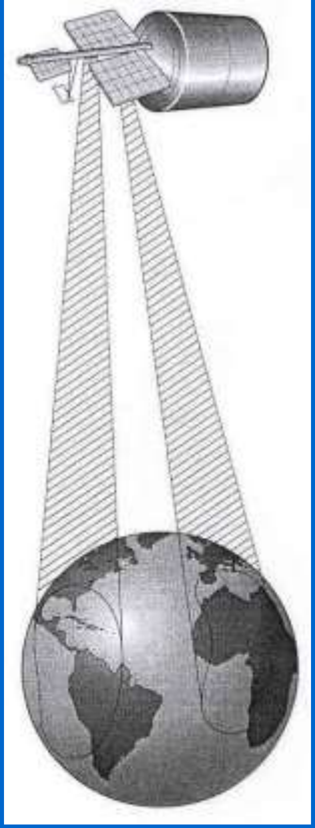
- adaptive antennas promise to mitigate reverse link problems
- limiting case of infinitesimal beamwidth
- limiting case of infinitely fast track ability
- thereby unique channel that is free from interference
- all user communicate at same time using the same channel

Disadvantage of SDMA

- perfect adaptive antenna system: infinitely large antenna needed
- compromise needed

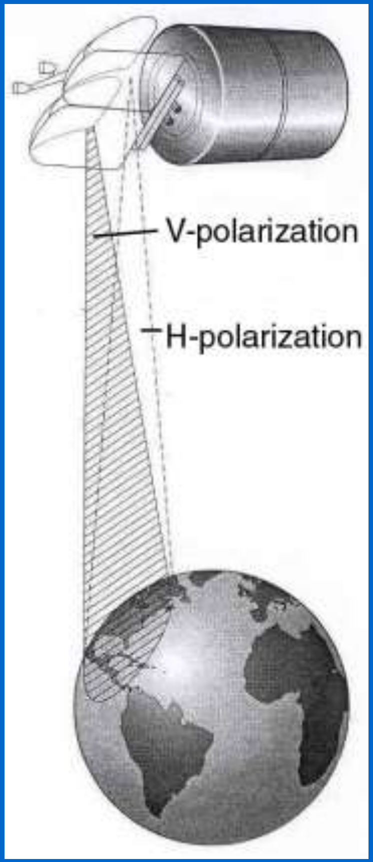
SDMA and PDMA in satellites

- INTELSAT IVA
- SDMA dual-beam receive antenna
- simultaneously access from two different regions of the earth



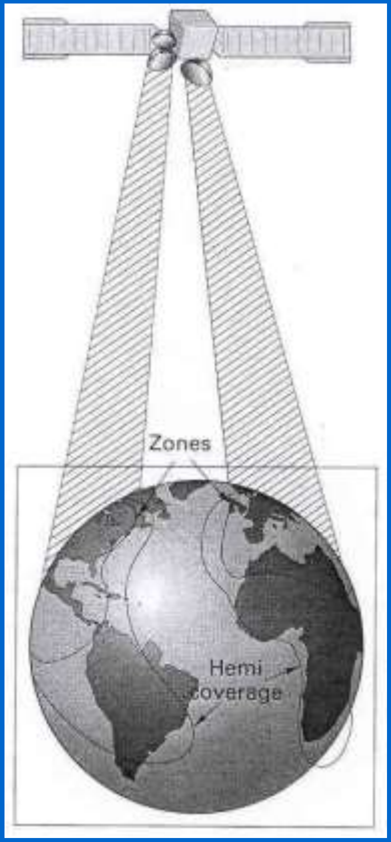
SDMA and PDMA in satellites

- COMSTAR 1
- PDMA
- separate antennas
- simultaneously access from same region



SDMA and PDMA in satellites

- INTELSAT V
- PDMA and SDMA
- two hemispheric coverages by SDMA
- two smaller beam zones by PDMA
- orthogonal polarization



Capacity of Cellular Systems

- channel capacity: maximum number of users in a fixed frequency band
- radio capacity : value for spectrum efficiency
- reverse channel interference
- forward channel interference
- How determine the radio capacity?

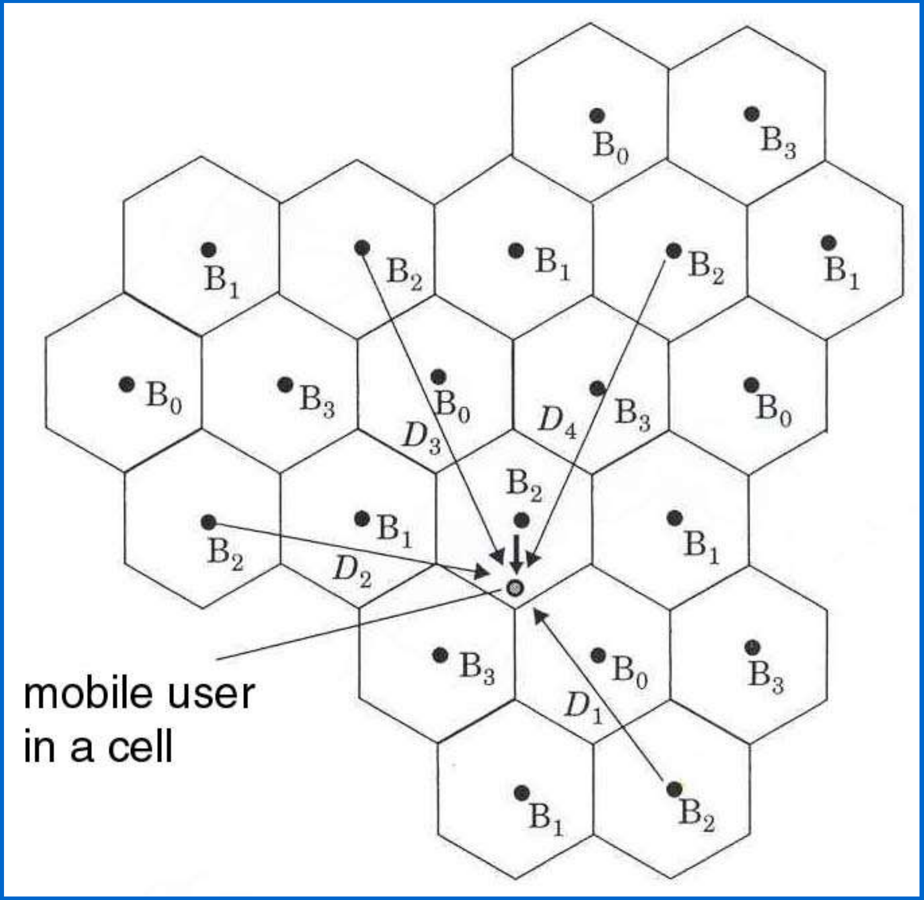
Co-Channel Reuse Ratio Q

$$Q=D/R$$

- Q … co-channel reuse ratio
- D … distance between two co-channel cells
- R … cell radius

Forward channel interference

- cluster size of 4
- D_0 … distance serving station to user
- D_K … distance co-channel base station to user



Carrier-to-interference ratio C/I

- M closest co-channels cells cause first order interference

$$\frac{C}{I} = \frac{D_0^{-n_0}}{\sum_{k=1}^{M} D_K^{-n_k}}$$

- n_0 … path loss exponent in the desired cell
- n_k … path loss exponent to the interfering base station

Carrier-to-interference ratio C/I

- Assumption:
- just the 6 closest stations interfere
- all these stations have the same distance D
- all have similar path loss exponents to n_0

$$\frac{C}{I} = \frac{D_0^{-n}}{6 * D^{-n}}$$

Worst Case Performance

- maximum interference at $D_0 = R$
- $(C/I)_{min}$ for acceptable signal quality
- following equation must hold:

$$1/6 * (R/D)^{-n} \geqq (C/I)_{min}$$

Co-Channel reuse ratio Q

$$Q = D/R = (6*(C/I)_{min})^{1/n}$$

- D … distance of the 6 closest interfering base stations
- R … cell radius
- $(C/I)_{min}$ … minimum carrier-to-interference ratio
- n … path loss exponent

Radio Capacity m

$$m = \frac{B_t}{B_c * N} \text{ radio channels/cell}$$

- Bt … total allocated spectrum for the system
- Bc … channel bandwidth
- N … number of cells in a complete frequency reuse cluster

Radio Capacity m

- N is related to the co-channel factor Q by:

$$Q = (3*N)^{1/2}$$

$$m = \frac{B_t}{B_c * (Q^2/3)} = \frac{B_t}{B_c * \left(\frac{6}{3^{n/2}} * \left(\frac{C}{I} \right)_{min} \right)^{2/n}}$$

Radio Capacity m for n = 4

$$m = \frac{B_t}{B_c * \sqrt{2/3 * (C/I)_{min}}}$$

- m … number of radio channels per cell
- $(C/I)_{min}$ lower in digital systems compared to analog systems
- lower $(C/I)_{min}$ imply more capacity
- exact values in real world conditions measured

Compare different Systems

- each digital wireless standard has different $(C/I)_{min}$
- to compare them an equivalent (C/I) needed
- keep total spectrum allocation B_t and number of rario channels per cell m constant to get $(C/I)_{eq}$:

Compare different Systems

$$\left(\frac{C}{I}\right)_{eq} = \left(\frac{C}{I}\right)_{min} * \left(\frac{B_c}{B_c'}\right)^2$$

- B_c … bandwidth of a particular system
- $(C/I)_{min}$ … tolerable value for the same system
- B_c' … channel bandwidth for a different system
- $(C/I)_{eq}$ … minimum C/I value for the different system

C/I in digital cellular systems

$$\frac{C}{I} = \frac{E_b * R_b}{I} = \frac{E_c * R_c}{I}$$

- R_b … channel bit rate
- E_b … energy per bit
- R_c … rate of the channel code
- E_c … energy per code symbol

C/I in digital cellular systems

- combine last two equations:

$$\frac{(C/I)}{(C/I)_{eq}} = \frac{(E_c * R_c)/I}{(E_c' * R_c')/I'} = \left(\frac{B_c'}{B_c}\right)^2$$

- The sign ' marks compared system parameters

C/I in digital cellular systems

- Relationship between R_c and B_c is always linear ($R_c/R_c' = B_c/B_c'$)
- assume that level I is the same for two different systems ($I' = I$) :

$$\frac{E_c}{E_c'} = \left(\frac{B_c'}{B_c}\right)^3$$

Compare C/I between FDMA and TDMA

- Assume that multichannel FDMA system occupies same spectrum as a TDMA system
- FDMA : $C = E_b * R_b$; $I = I_0 * B_c$
- TDMA : $C' = E_b * R_b'$; $I' = I_0 * B_c'$
- E_b … Energy per bit
- I_0 … interference power per Hertz
- R_b … channel bit rate
- B_c … channel bandwidth

Example

- A FDMA system has 3 channels , each with a bandwidth of 10kHz and a transmission rate of 10 kbps.
- A TDMA system has 3 time slots, a channel bandwidth of 30kHz and a transmission rate of 30 kbps.
- What’s the received carrier-to-interference ratio for a user ?

Example

- In TDMA system C'/I' be measured in 333.3 ms per second - one time slot

$$C' = E_b*R_b' = 1/3*(E_b*10E4 \text{ bits}) = 3*R_b*E_b=3*C$$
$$I' = I0*B_c' = I0*30kHz = 3*I$$

- In this example FDMA and TDMA have the same radio capacity (C/I leads to m)

Example

- Peak power of TDMA is 10logk higher then in FDMA (k … time slots)
- in practice TDMA have a 3-6 times better capacity

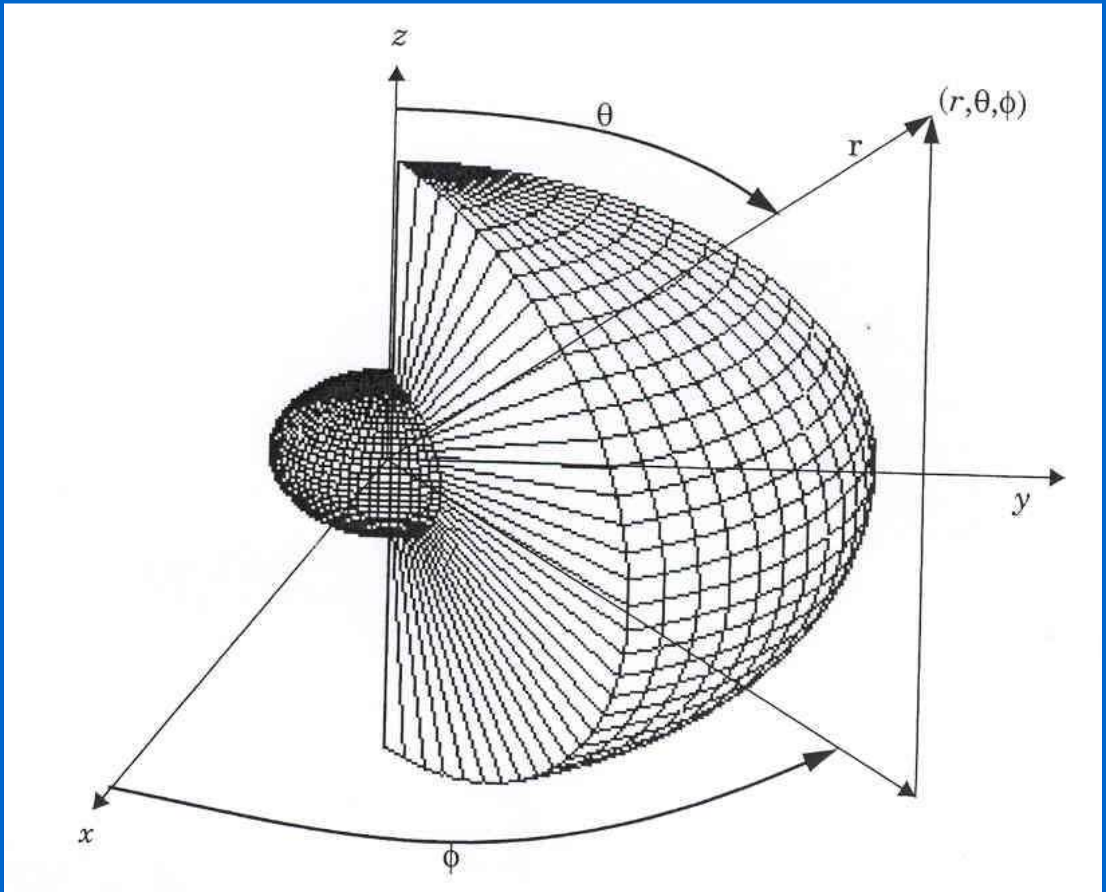
Capacity of SDMA systems

- one beam each user
- base station tracks each user as it moves
- adaptive antennas most powerful form
- beam pattern G(♐) has maximum gain in the direction of desired user
- beam is formed by N-element adaptive array antenna

Capacity of SDMA systems

- G(♐) steered in the horizontal ♐ -plane through 360°
- G(♐) has no variation in the elevation plane to account which are near to and far from the base station
- following picture shows a 60 degree beamwidth with a 6 dB sideslope level

Capacity of SDMA systems



Capacity of SDMA systems

- reverse link received signal power, from desired mobiles, is $P_{r;0}$
- interfering users i = 1,…,k-1 have received power $P_{r;I}$
- average total interference power I seen by a single desired user:

Capacity of SDMA

$$I = E\left\{\sum_{i=1}^{K-1} G(\theta_i)\, P_{r;I}\right\}$$

- θ_i … direction of the i-th user in the horizontal plane
- E … expectation operator

Capacity of SDMA systems

- in case of perfect power control (received power from each user is the same) :

$$P_{r;I} = P_c$$

- Average interference power seen by user 0:

$$I = P_c \, E\left\{ \sum_{i=1}^{K-1} G(\theta_i) \right\}$$

Capacity of SDMA systems

- users independently and identically distributed throughout the cell:

$$I = P_c *(k-1) * 1/D$$

- D … directivity of the antenna - given by max(G(☞))
- D typ. 3dB …10dB

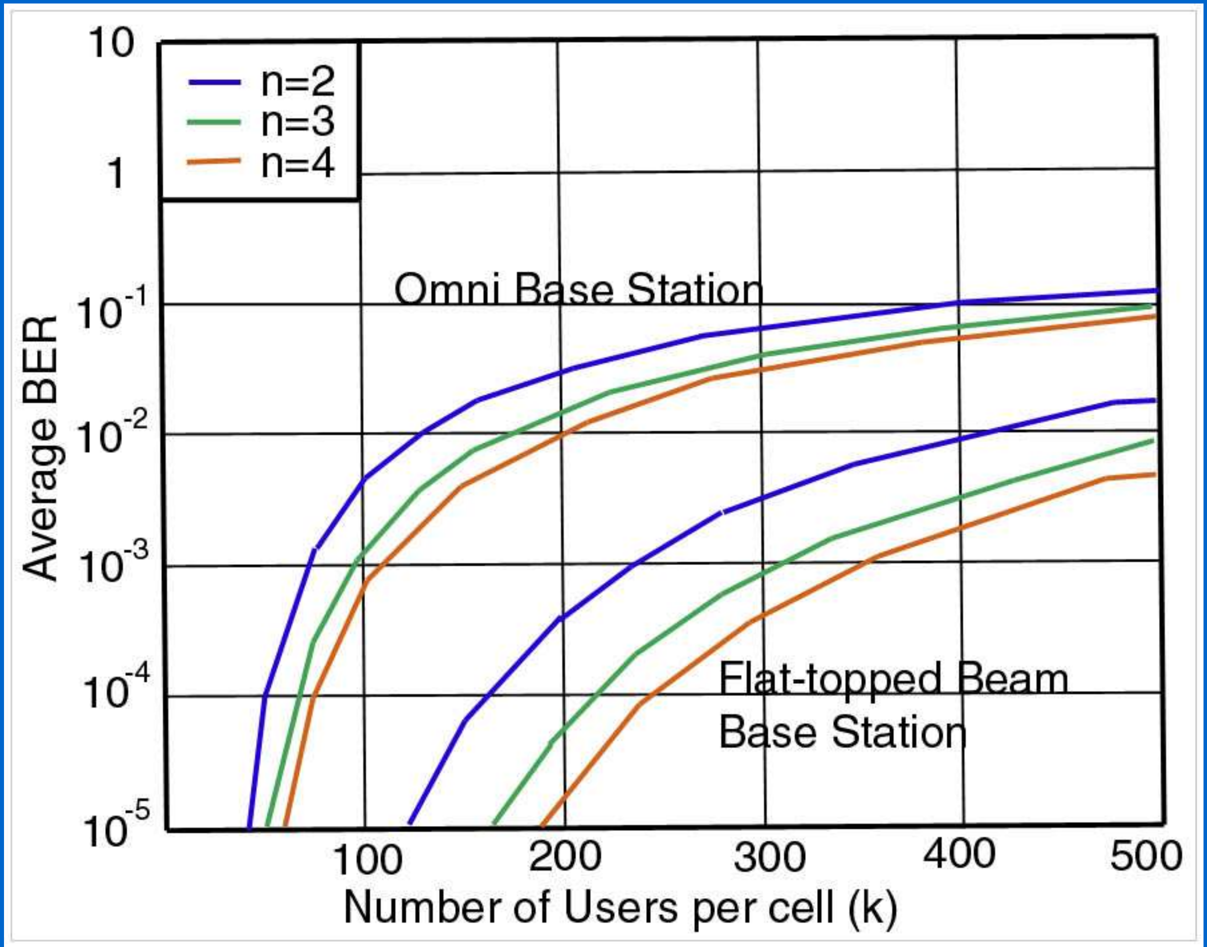
Capacity of SDMA systems

- Average bit error rate P_b for user 0:

$$P_b = Q\left(\sqrt{\frac{3DN}{K-1}}\right)$$

- D … directivity of the antenna
- Q(x) … standard Q-function
- N … spreading factor
- K … number of users in a cell

Capacity of SDMA systems



The Cellular Concept

3rd Chapter

Wireless Communication

By Theodore S. Rappaport

2nd Edition

Cellular Systems-Basic Concepts

Cellular system solves the problem of spectral congestion.

Offers high capacity in limited spectrum.

High capacity is achieved by limiting the coverage area of each BS to a small geographical area called cell.

Replaces high powered transmitter with several low power transmitters.

Each BS is allocated a portion of total channels and nearby cells are allocated completely different channels.

All available channels are allocated to small no of neighboring BS.

Interference between neighboring BSs is minimized by allocating different channels.

Cellular Systems-Basic Concepts

Same frequencies are reused by spatially separated BSs.

Interference between co-channels stations is kept below acceptable level.

Additional radio capacity is achieved.

Frequency Reuse-Fix no of channels serve an arbitrarily large no of subscribers

Frequency Reuse

used by service providers to improve the efficiency of a cellular network and to serve millions of subscribers using a **limited radio spectrum**

After covering a certain distance a radio wave gets attenuated and the signal falls below a point where it can no longer be used or cause any interference

A transmitter transmitting in a specific frequency range will have only a limited coverage area

Beyond this coverage area, that frequency can be reused by another transmitter.

The entire network coverage area is divided into cells based on the principle of frequency reuse

Frequency Reuse

A cell = basic geographical unit of a cellular network; is the area around an antenna where a specific frequency range is used.

when a subscriber moves to another cell, the antenna of the new cell takes over the signal transmission

a cluster is a group of adjacent cells, usually 7 cells; no frequency reuse is done within a cluster

the frequency spectrum is divided into sub-bands and each sub-band is used within one cell of the cluster

in heavy traffic zones cells are smaller, while in isolated zones cells are larger

Frequency Reuse

The design process of selecting and allocating channel groups for all of the cellular base stations within a system is called frequency reuse or frequency planning.

Cell labeled with same letter use the same set of frequencies.

Cell Shapes:

Circle, Square, Triangle and Hexagon.

Hexagonal cell shape is conceptual , in reality it is irregular in shape

Frequency Reuse

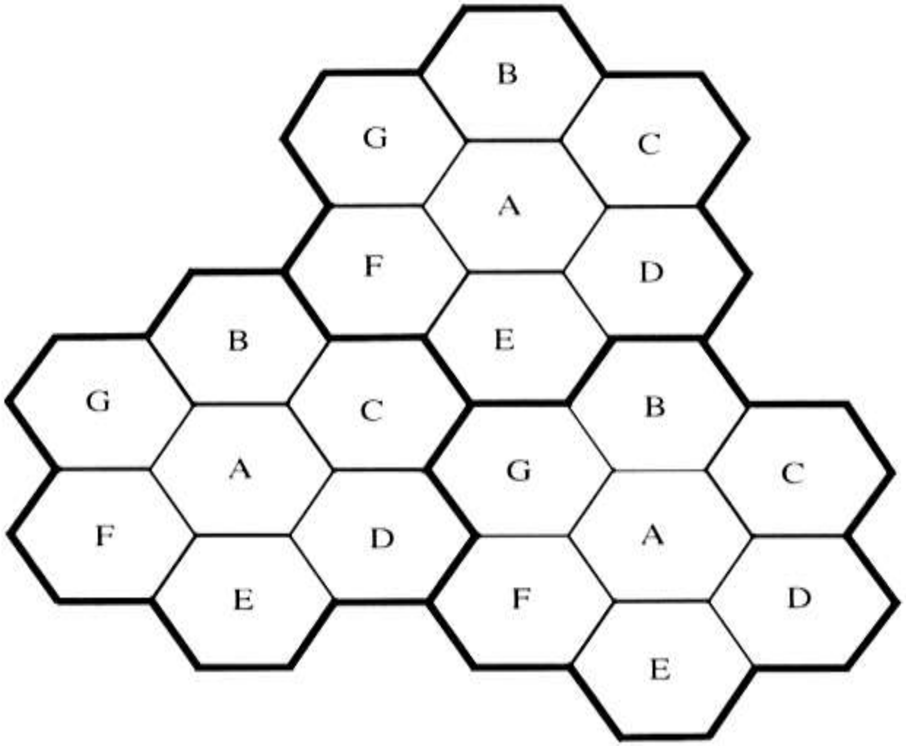


Figure 3.1 Illustration of the cellular frequency reuse concept. Cells with the same letter use the same set of frequencies. A cell cluster is outlined in bold and replicated over the coverage area. In this example, the cluster size, *N*, is equal to seven, and the frequency reuse factor is 1/7 since each cell contains one-seventh of the total number of available channels.

Frequency Reuse

In hexagonal cell model, BS transmitter can be in centre of cell or on its 3 vertices.

Centered excited cells use omni directional whereas edge excited cells use directional antennas.

A cellular system having ‘S’ duplex channels, each cell is allocated ‘k’ channels(k<S).

If S channels are allocated to N cells into unique and disjoint channels, the total no of available channel is S=kN.

Frequency Reuse

N cells collectively using all the channels is called a cluster, is a group of adjacent cells.

If cluster if repeated M times, the capacity C of system is given as

$$C=MkN=MS$$

Capacity of system is directly proportional to the no of times cluster is repeated.

Reducing the cluster size N while keeping the cell size constant, more clusters are required to cover the given area and hence more capacity.

Co-channel interference is dependent on cluster size, large cluster size less interference and vice versa.

Frequency Reuse

The Frequency Reuse factor is given as 1/N, each cell is assigned 1/N of total channels.

Lines joining a cell and each of its neighbor are separated by multiple of 60^0,certain cluster sizes and cell layout possible

Geometery of hexagon is such that no of cells per cluster i.e N, can only have values which satisfy the equation

$$N=i^2+ij+j^2$$

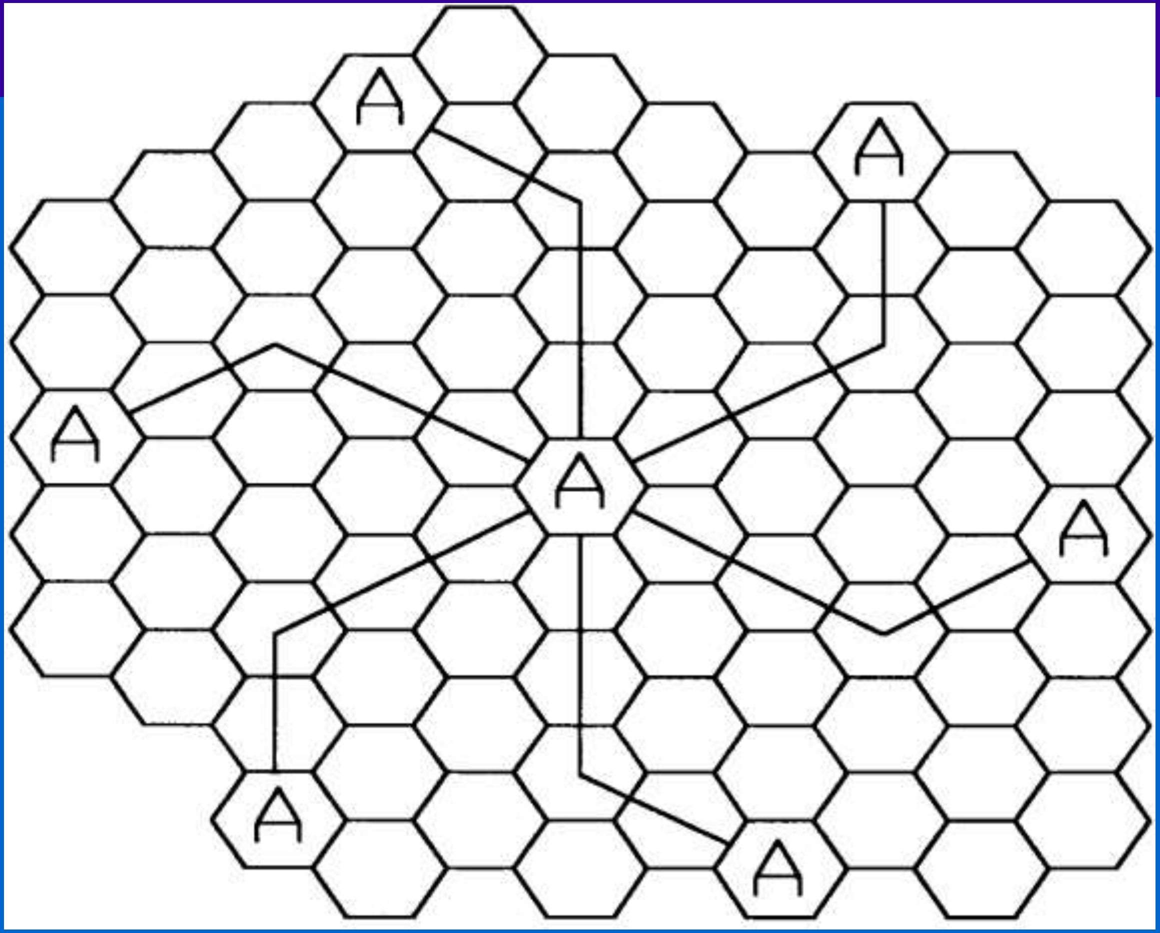
N, the cluster size is typically 4, 7 or 12.

In GSM normally N =7 is used.

i and j are integers, for i=3 and j=2 N=19.

Example from Book

Locating co-channel Cell



Channel Assignment Strategies

A scheme for increasing capacity and minimizing interference is required.

CAS can be classified as either fixed or dynamic

Choice of CAS impacts the performance of system.

In Fixed CA each cell is assigned a *predetermined* set of voice channels

Any call attempt within the cell can only be served by the *unused* channel in that particular cell

If all the channels in the cell are occupied, the call is *blocked*. The user does not get service.

In variation of FCA, a cell can *borrow channels* from its neighboring cell if its own channels are full.

Dynamic Channel Assignment

Voice channels are not allocated to different cells *permanently*.

Each time a call request is made, the *BS request* a channel from the MSC.

MSC allocates a channel to the requesting cell using an algorithm that takes into account

- likelihood of future blocking
- The reuse distance of the channel (should not cause interference)
- Other parameters like cost

To ensure min QoS, MSC only allocates a given frequency if that frequency is not currently in use in the cell or any other cell which falls within the *limiting reuse distance.*

DCA reduce the likelihood of blocking and increases capacity

Requires the MSC to collect realtime data on channel occupancy and traffic distribution on continous basis.

Hand-off

Mobile *moves into a different cell during* a conversation, MSC transfers the call to new channel belonging to new BS

Handoff operation involves *identifying the new BS* and *allocation of voice and control signal* to channels associated with new BS

Must be performed *successfully, infrequently* and *impercitble* to user

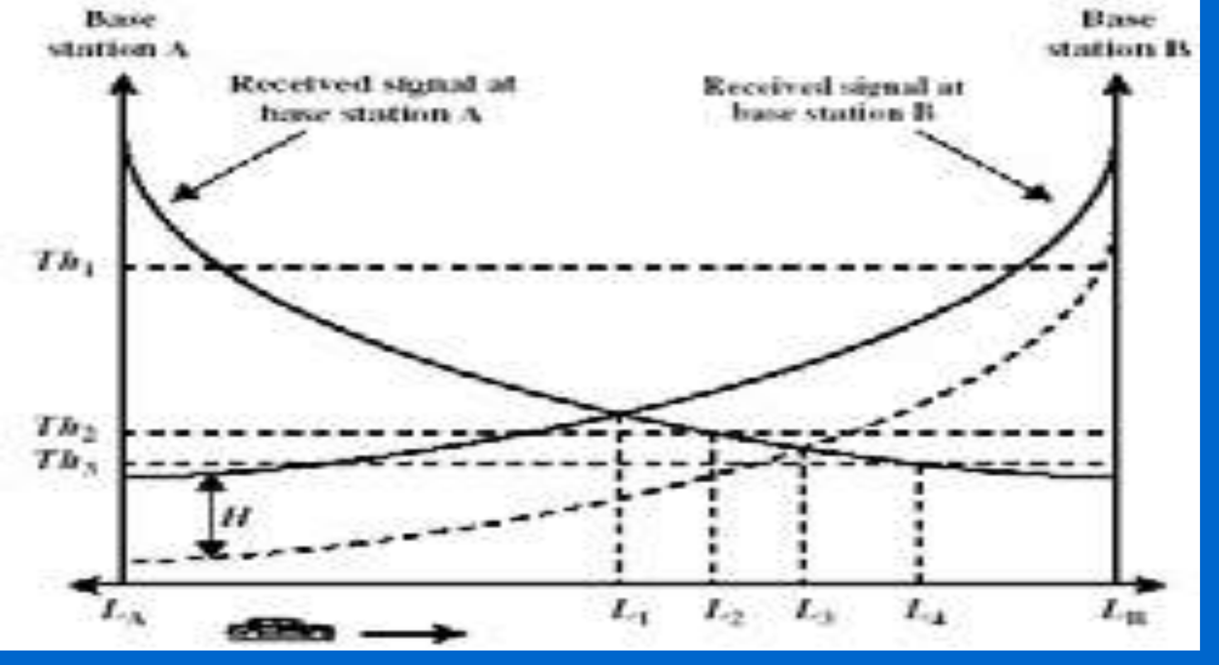
To meet these requirements an *optimum signal level* must be defined to initiate a handoff.

Min usuable signal for acceptable voice qualtiy -90 to -100 dBm

A slight higher value is used as *threshold*

Handoff

By looking at the variations of signal strength from either BS it is possible to decide on the optimum area where handoff can take place



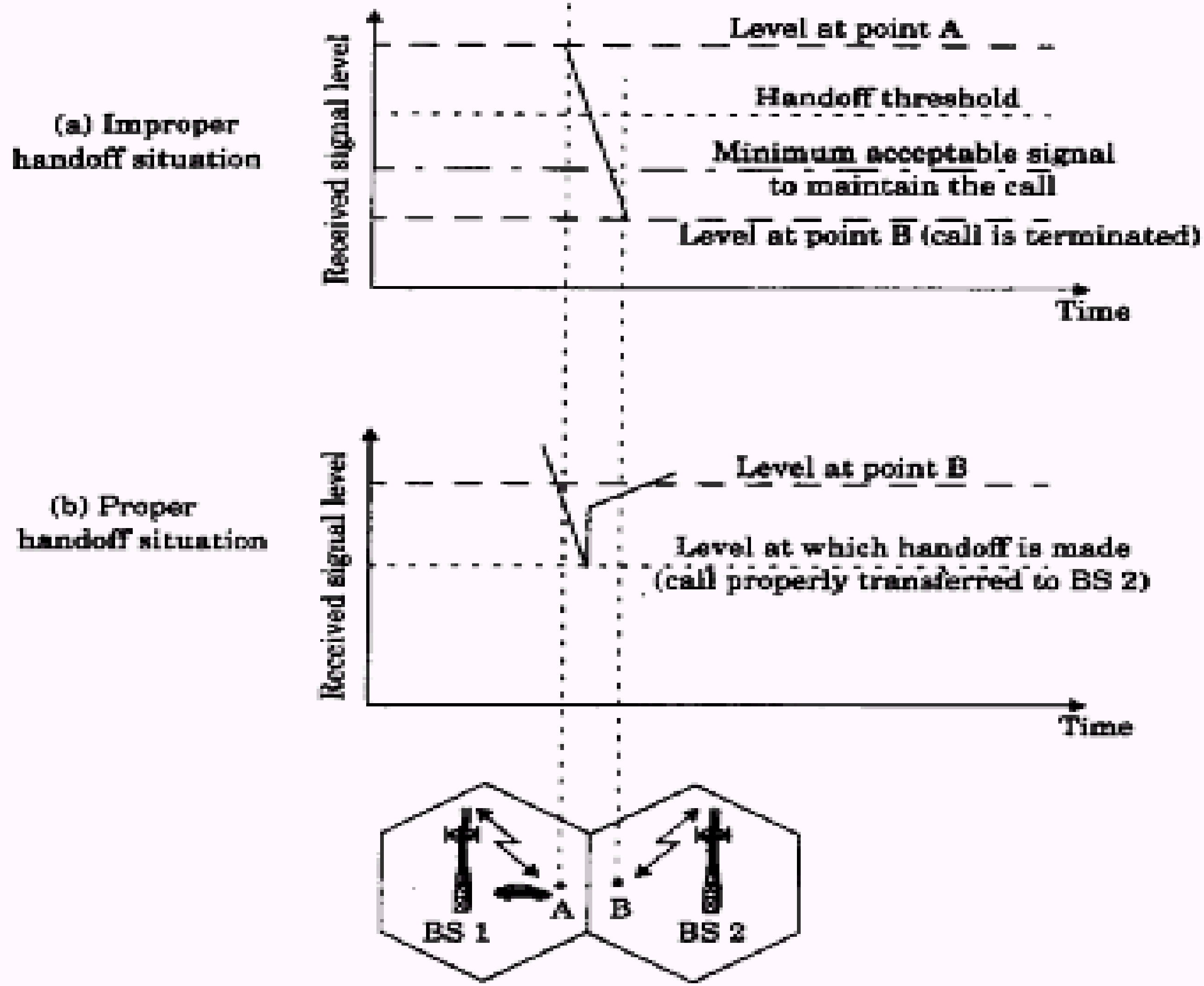


Figure 2.3
Illustration of a handoff scenario at cell boundary.

Hand-off strategies

Handoff is made when received signal at the BS *falls below* a certain threshold

During handoff: to avoid call termination, *safety margin should exist* and should not be too large or small

$$\Delta = Power_{handoff} - Power_{min\ usable}$$

Large Δ results in unecesarry handoff and for small Δ unsufficient time to complete handoff, so carefully chosen to meet the requirements.

Fig a, handoff not made and signal falls below min acceptable level to keep the channel active.

Can happen due to excessive delay by MSC in assigning handoff, or when threshold Δ is set to small.

Excessive delay may occur during high traffic conditions due to computional loading or non avialablilty of channels in nearby cells

Hand-off

In deciding when to handoff , it is important to ensure that the drop in signal level is not due to momentary fading.

In order to ensure this the BS monitors the signal for a certain period of time before initiating a handoff

The length of time needed to decide if handoff is necessary depends on the speed at which the mobile is moving

Hand-off strategies

In 1st generation analog cellular systems, the signal strength measurements are made by the BS and are supervised by the MSC.

A spare Rx in base station (locator Rx) monitors RSS of RVC's in neighboring cells

Tells Mobile Switching Center about these mobiles and their channels

Locator Rx can see if signal to this base station is significantly better than to the host base station

MSC monitors RSS from all base stations & decides on handoff

Hand-off strategies

In 2nd generation systems Mobile Assisted Handoffs (MAHO)are used

In MAHO, every MS measures the received power from the surrounding BS and continually reports these values to the corresponding BS.

Handoff is initiated if the signal strength of a neighboring BS exceeds that of current BS

MSC no longer monitors RSS of all channels

reduces computational load considerably

enables much more rapid and efficient handoffs

imperceptible to user

Soft Handoff

CDMA spread spectrum cellular systems provides a unique handoff capability

Unlike channelized wireless systems that assigns different radio channel during handoff (called hard handoff), the spread spectrum MS share the same channel in every cell

The term handoff here implies that a different BS handles the radio communication task

The ability to select between the instantaneous received signals from different BSs is called soft handoff

Inter system Handoff

If a mobile moves from one cellular system to a different system controlled by a different MSC, an inter-system handoff is necessary

MSC engages in intersystem handoff when signal becomes weak in a given cell and MSC cannot find another cell within its system to transfer the on-going call

Many issues must be resolved

- Local call may become long distance call
- Compatibility between the two MSCs

Prioritizing Handoffs

Issue: Perceived Grade of Service (GOS) – service quality as viewed by users

"quality" in terms of dropped or blocked calls (not voice quality)

assign higher priority to handoff vs. new call request

a dropped call is more aggravating than an occasional blocked call

Guard Channels

% of total available **cell** channels exclusively set aside for handoff requests

makes fewer channels available for new call requests

a good strategy is dynamic channel allocation (not fixed)

adjust number of guard channels as needed by demand

so channels are not wasted in cells with low traffic

Prioritizing Handoffs

Queuing of Handoff Requests

use time delay between handoff threshold and minimum useable signal level to place a blocked handoff request in queue

a handoff request can "*keep trying*" during that time period, instead of having a single block/no block decision

prioritize requests (based on mobile speed) and handoff as needed

calls will still be dropped if time period expires

Practical Handoff Considerations

Problems occur because of a *large range of mobile velocities*

pedestrian vs. vehicle user

Small cell sizes and/or micro-cells → *larger # handoffs*

MSC load is *heavy* when high speed users are passed between very small cells

Umbrella Cells

use *different antenna heights* and *Tx power levels* to provide large **and** small cell coverage

multiple antennas & Tx can be co-located at single location if necessary (saves on obtaining new tower licenses)

large cell → high speed traffic → fewer handoffs

small cell → low speed traffic

example areas: interstate highway passing through urban center, office park, or nearby shopping mall

Umbrella Cells

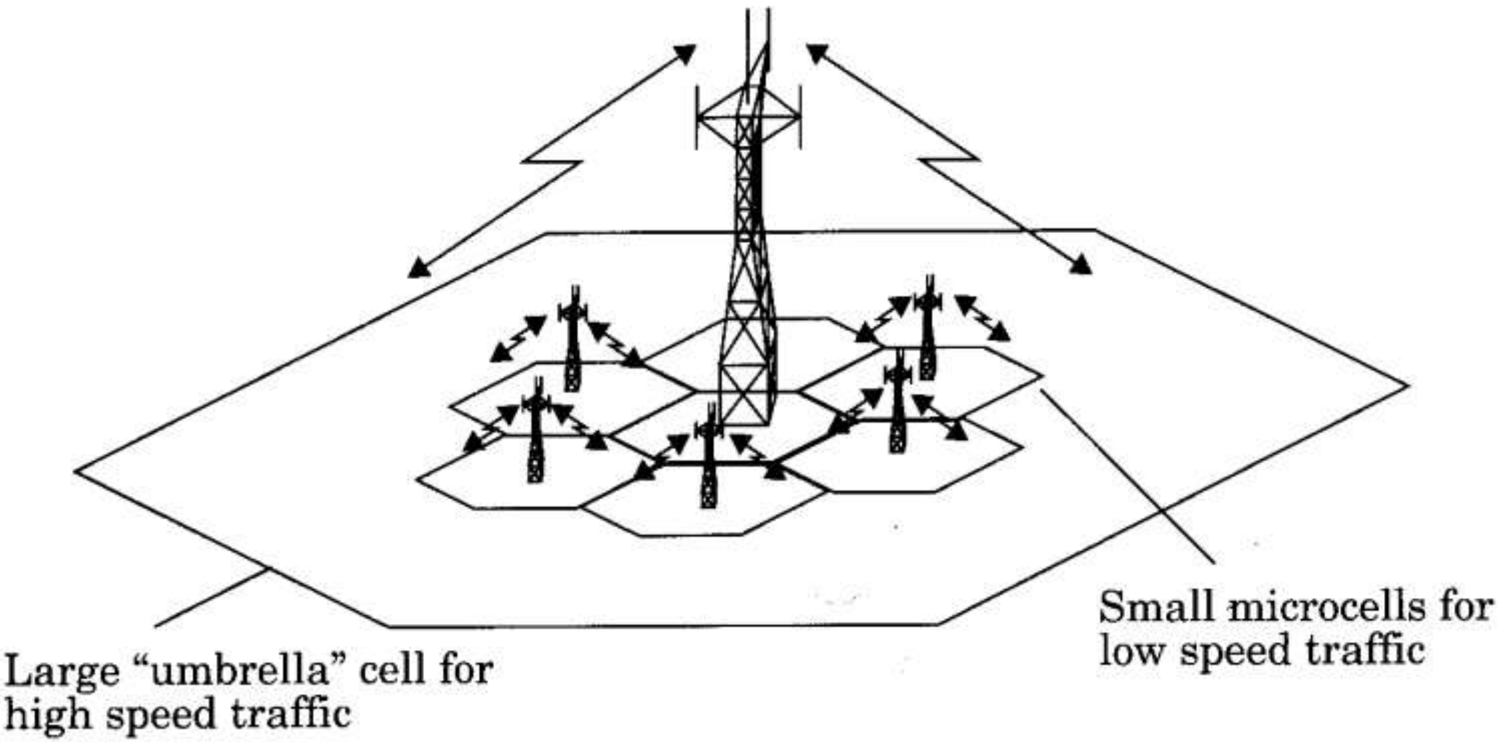


Figure 3.4 The umbrella cell approach.

Typical handoff parameters

Analog cellular (1st generation)

threshold margin $\Delta \approx$ 6 to 12 dB

total time to complete handoff $\approx$ 8 to 10 sec

Digital cellular (2nd generation)

total time to complete handoff $\approx$ 1 to 2 sec

lower necessary threshold margin $\Delta \approx$ 0 to 6 dB

enabled by mobile assisted handoff

Reuse Ratio:

For hexagonal cell reuse distance is given by $D=R(\sqrt{3N})$

Where R is cell size or cell radius and N is cluster size

D increases as we increase N

Reuse factor is given by $Q=D/R=(\sqrt{3N})$

Interference

Goals for this section

Co-Channel

Adjacent Channel

How to calculate signal to interference ratio

Interference

Interference is major limiting factor in the performance of cellular radio. It limits the capacity and increases the no of dropped calls.

Sources of interference include

Another mobile in same cell

A call in progress in a neighboring cell

Other BSs operating in the same frequency band

Effects of Interference

Interference in voice channels causes

- Crosstalk
- Noise in background

Interference in control channels causes

- Error in digital signaling, which causes
  - Missed calls
  - Blocked calls
  - Dropped calls

Interference

Two major types of Interferences

Co-channel Interference (CCI)

Adjacent channel Interference (ACI)

CCI is caused due to the cells that reuse the same frequency set. These cells using the same frequency set are called Co-channel cells

ACI is caused due to the signals that are adjacent in frequency

Co-channel Interference

Increase base station Tx power to improve radio signal reception?

will also increase interference into other co-channel cells by the same amount

no net improvement

Separate co-channel cells by some minimum distance to provide sufficient isolation from propagation of radio signals?

if all cell sizes, transmit powers, and coverage patterns ≈ same → co-channel interference is independent of Tx power

Co-channel Interference

co-channel interference depends on:

R : cell radius

D : distance to base station of nearest co-channel cell where D=R(√3N)

if *D/R* ↑ then spatial separation relative to cell coverage area ↑

improved isolation from co-channel RF energy

$Q = D / R$: co-channel reuse ratio

hexagonal cells → $Q = D/R = \sqrt{3N}$

Smaller value of Q provides larger capacity, but higher CCI

Hence there is tradeoff between capacity and interference.

small Q → small cluster size → more frequency reuse → larger system capacity

small Q → small cell separation → increased CCI

Signal to Interference ratio S/I

The Signal-to-Interference (S/I) for a mobile is

$$\text{Eq. (3.5)}: \quad \frac{S}{I} = \frac{S}{\sum_{i=1}^{i_o} I_i} \quad \text{where}$$

S is desired signal power ,I_i : interference power from i^{th} co-channel cell

The average received power at distance d is

$$P_r = P_o \, (d/d_o)^{-n}$$

The RSS decays as a power law of the distance of separation between transmitter and receiver

Where P_o is received power at reference distance d_o and n is the path loss exponent and ranges between 2-4

If D_i is the distance of i^{th} interferer, the received power is proportional to $(D_i)^{-n}$

Signal to Interference ratio S/I

The S/I for mobile is given by

$$\frac{S}{I} = \frac{\text{signal from intended base station when at edge of cell (R away)}}{\text{signals from other base stations (D away)}}$$

$$= \frac{R^{-n}}{\sum_{i=1}^{i_o}(D_i{}^{-n})}$$

$$\frac{S}{I} = \frac{R^{-n}}{\sum_{i=1}^{i_0=6}(D_i)^{-n}} = \frac{(D/R)^n}{6} = \frac{(\sqrt{3N})^n}{6} = \frac{Q^4}{6}$$

With only the first tier(layer of) equidistant interferers.

For a hexagonal cluster size, which always have 6 CC cell in first tier

Signal to Interference ratio S/I

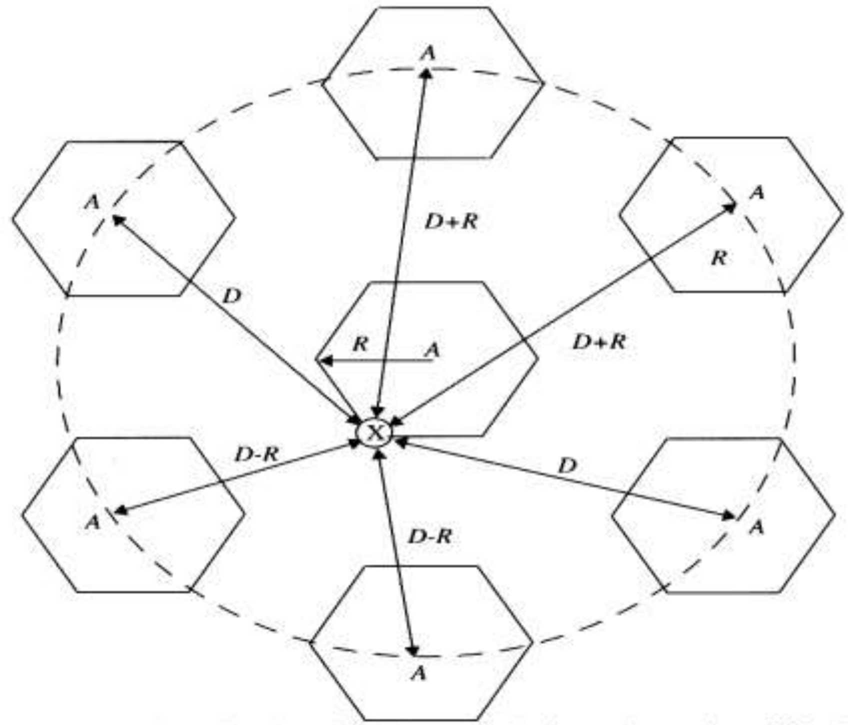


Figure 3.5 Illustration of the first tier of co-channel cells for a cluster size of $N = 7$. An approximation of the exact geometry is shown here, whereas the exact geometry is given in [Lee86]. When the mobile is at the cell boundary (point X), it experiences worst case co-channel interference on the forward channel. The marked distances between the mobile and different co-channel cells are based on approximations made for easy analysis.

$$\frac{S}{I} = \frac{R^{-4}}{2(D-R)^{-4} + 2(D+R)^{-4} + 2D^{-4}}$$

$$\frac{S}{I} = \frac{1}{2(Q-1)^{-4} + 2(Q+1)^{-4} + 2Q^{-4}}$$

Example S/I

Examples for Problem 2.3

TDMA can tolerate S/I = 15 dB

What is the optimal value of N for omni-directional antennas? Path loss = 4. **Co-channel Interference**

cluster size N =7 (choices 4,7,12)

path loss exponent (means)n=4

co-channel reuse ratio Q=sqrt(3N)=4.582576

Ratio of distance to radius Q=D/R=4.582576

number of neighboring cells i_o=6 # of sides of hexagon

signal to interference ratio S/I= $(D/R)^{\wedge n} / i_o$ =73.5

convert to dB, S/I=10log(S/I) =18.66287dB

S/I is greater than required, it will work.

Example S/I

cluster size N=4 (choices 4,7,12)

path loss exponent (means) n=4

co-channel reuse ratio Q= sqrt(3N)=3.464102

Ratio of distance to radius Q=D/R=3.464102

number of neighboring cells io=6 ,# of sides of hexagon

signal to interference ratio S/I= $(D/R)^{\wedge n}$ / i_o =24

convert to dB, S/I= 10log(S/I)=13.80211dB

S/I is less than required, it will not work!

cluster size N=7

path loss exponent n=3

Q=sqrt(3N)=4.582576

number of neighboring cells io= 6, # of sides of hexagon

signal to interference ratio=S/I= $(D/R)^{\wedge n}$ / i_o =16.03901

convert to dB, S/I=10log(S/I)=12.05178dB

S/I is less than required, it will not work!

Adjacent Channel Interference

Results from imperfect receiver filters, allowing nearby frequencies to leak into pass-band.

Can be minimized by careful filtering and channel assignments.

Channels are assigned such that frequency separations between channels are maximized.

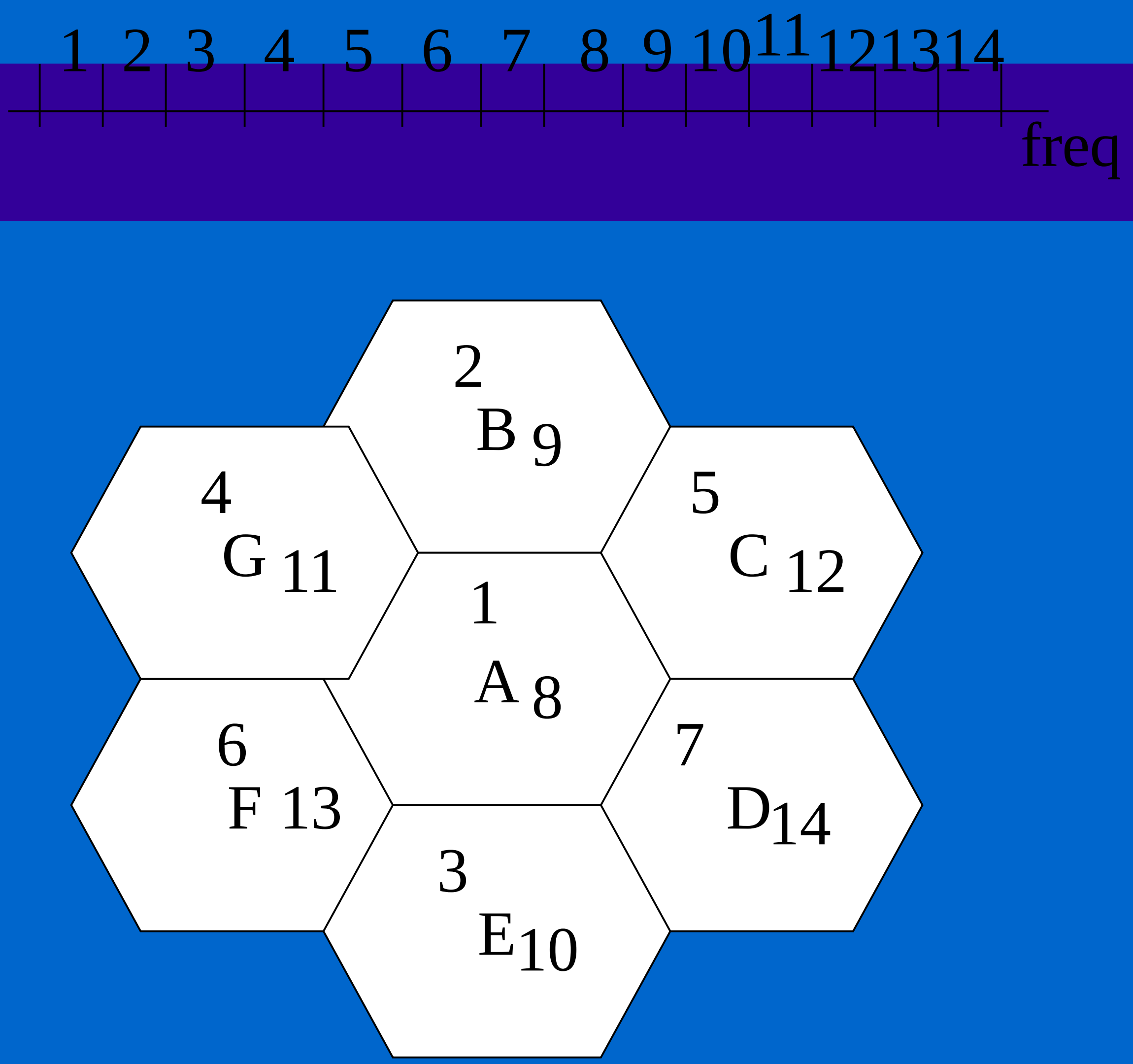
For example, by sequentially assigning adjacent bands to different cells

Total 832 channels, divided into two groups with 416 channels each.

Out of 416, 395 are voice and 21 are control channels.

395 channels are divided into 21 subsets, each containing almost 19 channels, with closet channel 21 channels away

If N=7 is used, each cell uses 3 subsets, assigned in such a way that each channel in a cell is 7 channels away.



Frequency Planning/Channel Assignment

Table 3.2 AMPS Channel Allocation for A and B Side Carriers

1A	2A	3A	4A	5A	6A	7A	1B	2B	3B	4B	5B	6B	7B	1C	2C	3C	4C	5C	6C	7C	
1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	
22	23	24	25	26	27	28	29	30	31	32	33	34	35	36	37	38	39	40	41	42	
43	44	45	46	47	48	49	50	51	52	53	54	55	56	57	58	59	60	61	62	63	
64	65	66	67	68	69	70	71	72	73	74	75	76	77	78	79	80	81	82	83	84	
85	86	87	88	89	90	91	92	93	94	95	96	97	98	99	100	101	102	103	104	105	
106	107	108	109	110	111	112	113	114	115	116	117	118	119	120	121	122	123	124	125	126	
127	128	129	130	131	132	133	134	135	136	137	138	139	140	141	142	143	144	145	146	147	
148	149	150	151	152	153	154	155	156	157	158	159	160	161	162	163	164	165	166	167	168	
169	170	171	172	173	174	175	176	177	178	179	180	181	182	183	184	185	186	187	188	189	
190	191	192	193	194	195	196	197	198	199	20	201	202	203	204	205	206	207	208	209	210	A
211	212	213	214	215	216	217	218	219	220	221	222	223	224	225	226	227	228	229	230	231	SIDE
232	233	234	235	236	237	238	239	240	241	242	243	244	245	246	247	248	249	250	251	252	
253	254	255	256	257	258	259	260	261	262	263	264	265	266	267	268	269	270	271	272	273	
274	275	276	277	278	279	280	281	282	283	284	285	286	287	288	289	290	291	292	293	294	
295	296	297	298	299	300	301	302	303	304	305	306	307	308	309	310	311	312	-	-	-	
313	314	315	316	317	318	319	320	321	322	323	324	325	326	327	328	329	330	331	332	333	
-	-	-	-	-	-	-	-	-	-	-	-	-	-	-	-	-	-	667	668	669	
670	671	672	673	674	675	676	677	678	679	680	681	682	683	684	685	686	687	688	689	690	
691	692	693	694	695	696	697	698	699	700	701	702	703	704	705	706	707	708	709	710	711	
712	713	714	715	716	-	-	-	-	991	992	993	994	995	996	997	998	999	1000	1001	1002	
1003	1004	1005	1006	1007	1008	1009	1010	1011	1012	1013	1014	1015	1016	1017	1018	1019	1020	1021	1022	1023	
334	335	336	337	338	339	340	341	342	343	344	345	346	347	348	349	350	351	352	353	354	
355	356	357	358	359	360	361	362	363	364	365	366	367	368	369	370	371	372	373	374	375	
376	377	378	379	380	381	382	383	384	385	386	387	388	389	390	391	392	393	394	395	396	
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418	419	420	421	422	423	424	425	426	427	428	429	430	431	432	433	434	435	436	437	438	
439	440	441	442	443	444	445	446	447	448	449	450	451	452	453	454	455	456	457	458	459	
460	461	462	463	464	465	466	467	468	469	470	471	472	473	474	475	476	477	478	479	480	
481	482	483	484	485	486	487	488	489	490	491	492	493	494	495	496	497	498	499	500	501	
502	503	504	505	506	507	508	509	510	511	512	513	514	515	516	517	518	519	520	521	522	
523	524	525	526	527	528	529	530	531	532	533	534	535	536	537	538	539	540	541	542	543	B
544	545	546	547	548	549	550	551	552	553	554	555	556	557	558	559	560	561	562	563	564	SIDE
565	566	567	568	569	570	571	572	573	574	575	576	577	578	579	580	581	582	583	584	585	
586	587	588	589	590	591	592	593	594	595	596	597	598	599	600	601	602	603	604	605	606	
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628	629	630	631	632	633	634	635	636	637	638	639	640	641	642	643	644	645	646	647	648	
649	650	651	652	653	654	655	656	657	658	659	660	661	662	663	664	665	666	-	-	-	
-	-	-	-	-	717	718	719	720	721	722	723	724	725	726	727	728	729	730	731	732	
733	734	735	736	737	738	739	740	741	742	743	744	745	746	747	748	749	750	751	752	753	
754	755	756	757	758	759	760	761	762	763	764	765	766	767	768	769	770	771	772	773	774	
775	776	777	778	779	780	781	782	783	784	785	786	787	788	789	790	791	792	793	794	795	
796	797	798	799																		

Learning Objectives

Concept of Trunking

Key definitions in Trunking /Traffic Theory

Erlang-(unit of traffic)

Grade of Service

Two Types of Trunked Systems

Trunking Efficiency

Trunking & Grade of Service

Cellular radio systems rely on *trunking to accommodate a large number of users in a limited radio spectrum.*

Trunking allows a large no of users to share a relatively small number of channels in a cell by providing access to each user, on demand, from a pool of available channels.

In a trunked radio system (TRS) each user is allocated a channel on a per call basis, upon termination of the call, the previously occupied channel is immediately returned to the pool of available channels.

Key Definitions

Setup Time: Time required to allocate a radio channel to a requesting user

Blocked Call: Call which cannot be completed at the time of request, due to congestion(*lost call)*

Holding Time: Average duration of a typical call. Denoted by *H(in seconds)*

Request Rate: The average number of calls requests per unit time(λ)

Traffic Intensity: Measure of channel time utilization or the average channel occupancy measured in Erlangs. Dimensionless quantity. Denoted by A

Load: Traffic intensity across the entire TRS (Erlangs)

Erlang-a unit of traffic

The fundamentals of trunking theory were developed by Erlang, a Danish mathematician, the unit bears his name.

An Erlang is a unit of telecommunications traffic measurement.

Erlang represents the continuous use of one voice path.

It is used to describe the total traffic volume of one hour

A channel kept busy for one hour is defined as having a load of one Erlang

For example, a radio channel that is occupied for thirty minutes during an hour carries 0.5 Erlangs of traffic

For 1 channel

Min load=0 Erlang (0% time utilization)

Max load=1 Erlang (100% time utilization)

Erlang-a unit of traffic

For example, if a group of 100 users made 30 calls in one hour, and each call had an average call duration(holding time) of 5 minutes, then the number of Erlangs this represents is worked out as follows:

Minutes of traffic in the hour = number of calls x duration

Minutes of traffic in the hour = 30 x 5 = 150

Hours of traffic in the hour = 150 / 60 = 2.5

Traffic Intensity= 2.5 Erlangs

Traffic Concepts

Traffic Intensity offered by each user(Au): Equals average call arrival rate multiplied by the holding time(service time)

Au=λ*H(Erlangs)*

Total Offered Traffic Intensity for a system of U users (A):

A =U*Au(Erlangs)

Traffic Intensity per channel, in a C channel trunked system

Ac=U*Au/C(Erlangs)

Trunking & Grade of Service

In a TRS, when a particular user requests service and all the available radio channels are already in use , the user is *blocked or denied access to the system. In some systems a queue may be used to hold the requesting users until a channel becomes available.*

Trunking systems must be designed carefully in order to ensure that there is a low likelihood that a user will be blocked or denied access.

The likelihood that a call is blocked, or the likelihood that a call experiences a delay greater than a certain queuing time is called **"Grade of Service" (GOS)''.**

Trunking & Grade of Service

Grade of Service (GOS): Measure of ability of a user to access a trunked system during the busiest hour. Measure of the congestion which is specified as a probability.

The probability of a call being blocked

Blocked calls cleared(BCC) or Lost Call Cleared(LCC) or Erlang B systems

The probability of a call being delayed beyond a certain amount of time before being granted access

Blocked call delayed or Lost Call Delayed(LCD) or Erlang C systems

Blocked Call Cleared Systems

When a user requests service, there is a minimal call set-up time and the user is given immediate access to a channel if one is available

If channels are already in use and no new channels are available, call is blocked without access to the system

The user does not receive service, but is free to try again later

All blocked calls are instantly returned to the user pool

Modeling of BCC Systems

The Erlang B model is based on following assumptions :

Calls are assumed to arrive with a Poisson distribution

There are nearly an infinite number of users

Call requests are memory less ,implying that all users, including blocked users, may request a channel at any time

All free channels are fully available for servicing calls until all channels are occupied

The probability of a user occupying a channel(called service time) is exponentially distributed. Longer calls are less likely to happen

There are a finite number of channels available in the trunking pool.

Inter-arrival times of call requests are independent of each other

Modeling of BCC Systems

Erlang B formula is given by

$$\Pr[\text{blocking}] = \frac{A^C / C!}{\sum_{k=0}^{C} \frac{A^k}{k!}}$$

where ***C*** *is the number of trunked channels offered by a trunked radio system and* ***A*** *is the total offered traffic.*

Erlang B

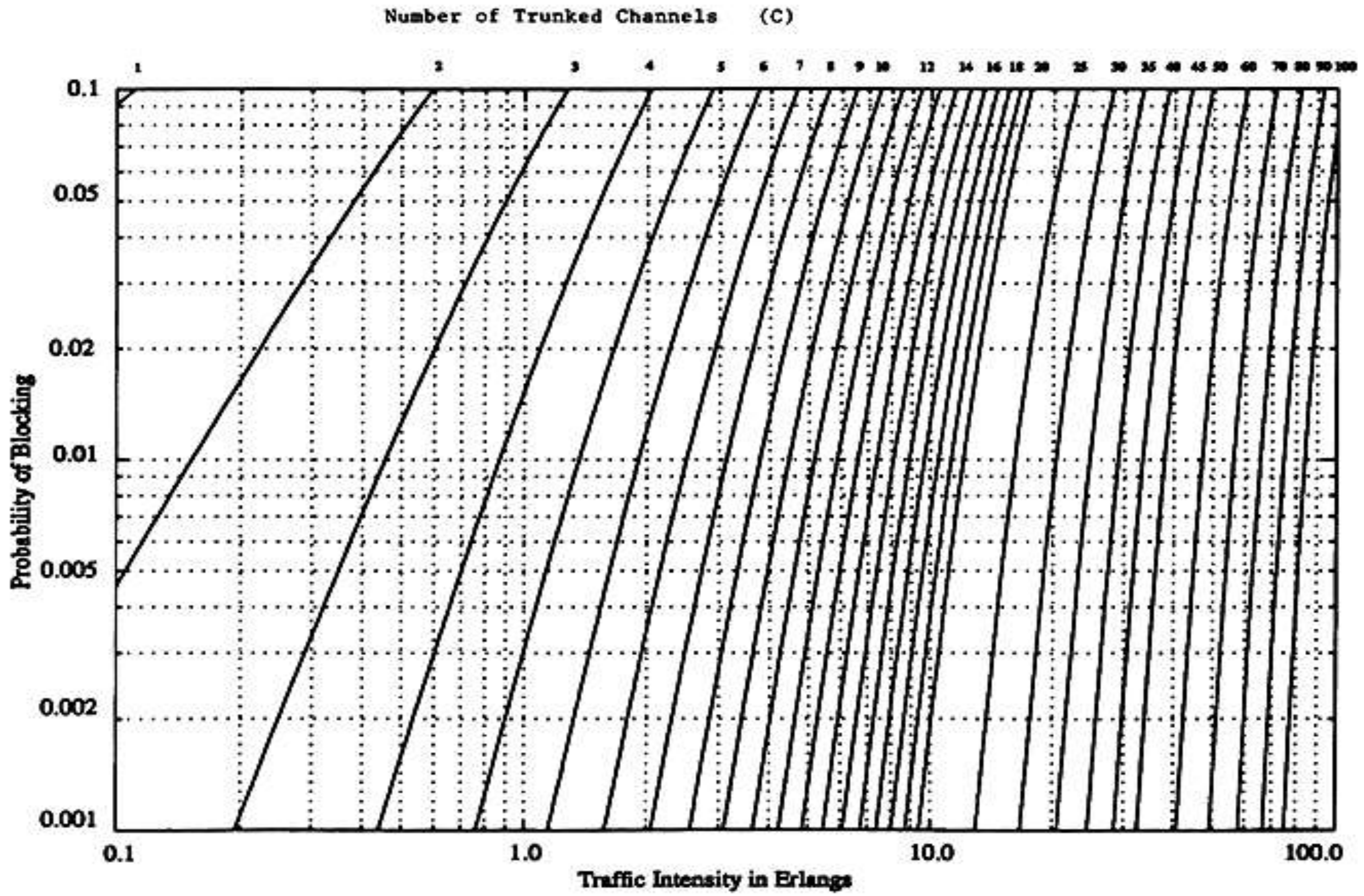


Figure 3.6 The Erlang B chart showing the probability of blocking as functions of the number of channels and traffic intensity in Erlangs.

Example 3.4

How many users can be supported for 0.5% blocking probability for the following number of trunked channels in a BCC system? (a) 5, (b) 10,(c)=20. Assumed that each user generates 0.1 Erlangs of traffic.

Solution

Given C=5, GOS=0.005, Au=0.1,

From graph/Table using C=5 and GOS=0.005,A=1.13

Total Number of users U=A/Au=1.13/0.1=11 users

Given C=10, GOS=0.005, Au=0.1,

From graph/Table using C=5 and GOS=0.005,A=3.96

Total Number of users U=A/Au=3.96/0.1=39 users

Given C=20, GOS=0.005, Au=0.1,

From graph/Table using C=20 and GOS=0.005,A=11.10

Total Number of users U=A/Au=11.10/0.1=110 users

Erlang B Trunking GOS

Table 3.4 Capacity of an Erlang B System

Number of Channels *C*	Capacity (Erlangs) for GOS = 0.01	= 0.005	= 0.002	= 0.001
2	0.153	0.105	0.065	0.046
4	0.869	0.701	0.535	0.439
5	1.36	1.13	0.900	0.762
10	4.46	3.96	3.43	3.09
20	12.0	11.1	10.1	9.41
24	15.3	14.2	13.0	12.2
40	29.0	27.3	25.7	24.5
70	56.1	53.7	51.0	49.2
100	84.1	80.9	77.4	75.2

BCC System Example

Assuming that each user in a system generates a traffic intensity of 0.2 Erlangs, how many users can be supported for 0.1% probability of blocking in an Erlang B system for a number of trunked channels equal to 60.

Solution 1:

System is an Erlang B

A_u = 0.2 Erlangs

Pr [Blocking] = 0.001

C = 60 Channels

From the Erlang B figure, we see that

$A \approx 40$ *Erlangs*

Therefore U=A/Au=40/0.02=2000users.

Blocked Call Delayed(BCD) Systems

Queues are used to hold call requests that are initially blocked

When a user attempts a call and a channel is not immediately available, the call request may be delayed until a channel becomes available

Mathematical modeling of such systems is done by Erlang C formula

The Erlang C model is based on following assumptions :

Similar to those of Erlang B

Additionally, if offered call cannot be assigned a channel, it is placed in a queue of infinite length

Each call is then serviced in the order of its arrival

Blocked Call Delayed Systems

Erlang C formula which gives likelihood of a call not having immediate access to a channel (all channels are already in use)

$$\Pr(delay > 0) = \frac{A^C}{A^C + C!\left(1 - \frac{A}{C}\right)\sum_{k=0}^{C-1}\frac{A^k}{k!}}$$

Erlang C

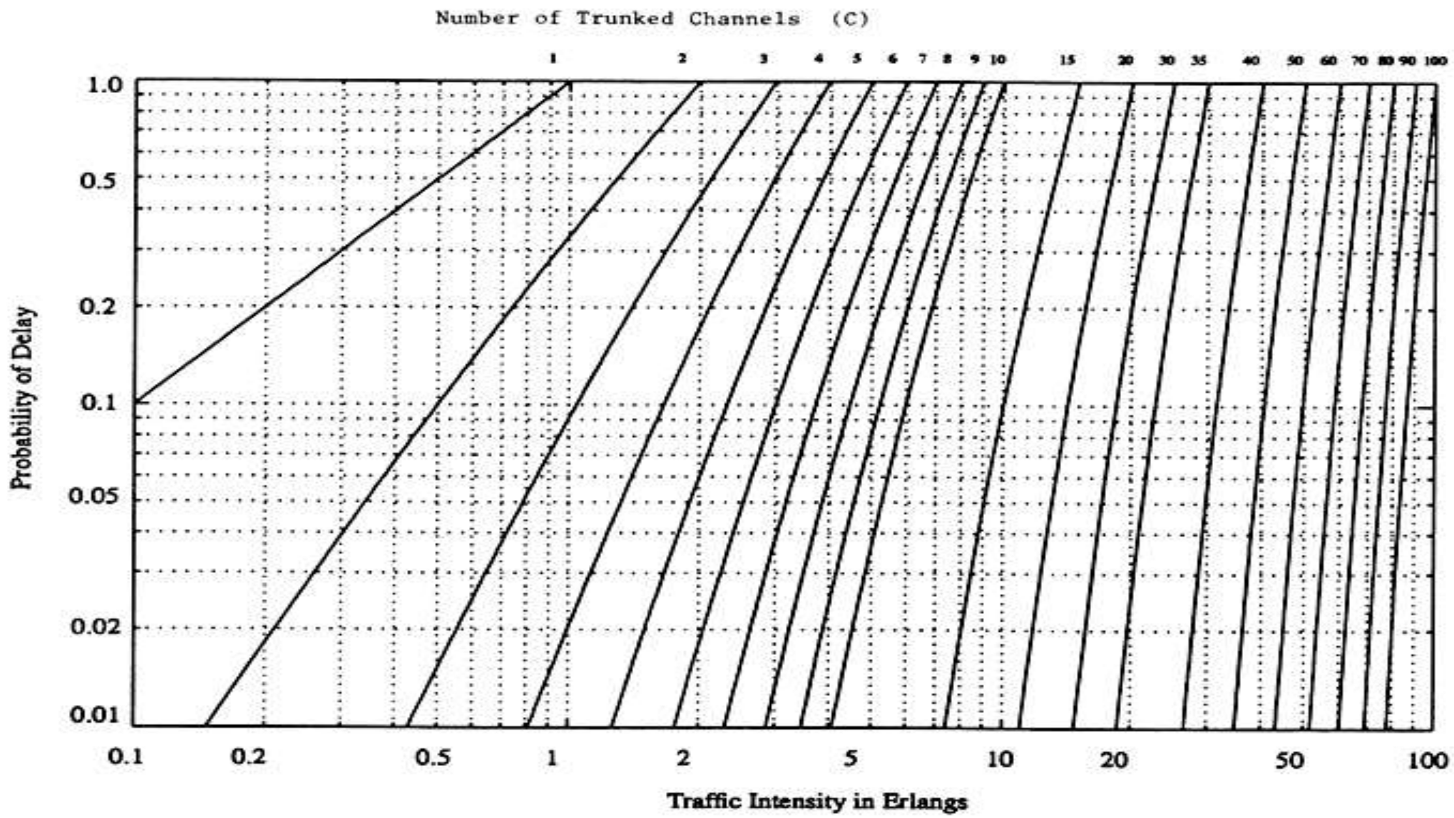


Figure 3.7 The Erlang C chart showing the probability of a call being delayed as a function of the number of channels and traffic intensity in Erlangs.

Modeling of BCD Systems

Probability that any caller is delayed in queue for a wait time greater than **t seconds is given as GOS of a BCD System**

The probability of a call getting delayed for any period of time greater than zero is

P[delayed call is forced to wait > t sec]=P[delayed] x Conditional P[delay is >t sec]

Mathematically;

Pr[delay>t] = Pr [delay>0] Pr [delay>t| delay>0]

Where P[delay>t| delay>0]= $e^{(-(C-A)t/H)}$

Pr[delay>t] = Pr [delay>0] e(-(C-A)t/H)

where C = total number of channels, *t =delay time of interest, H=average duration of call*

Trunking Efficiency

Trunking efficiency is a measure of the number of users which can be offered a particular GOS with a particular configuration of fixed channels.

The way in which channels are grouped can substantially alter the number of users handled by a trunked system.

Example:

10 trunked channels at a GOS of 0.01 can support 4.46 Erlangs, where as two groups of 5 trunked channels can support 2x1.36=2.72 Erlangs of traffic

10 trunked channels can offer 60% more traffic at a specific GOS than two 5 channel trunks.

Therefore, if in a certain situation we sub-divide the total channels in a cell into smaller channel groups then the total carried traffic will reduce with increasing number of groups

Erlang B Trunking GOS

Table 3.4 Capacity of an Erlang B System

Number of Channels *C*	Capacity (Erlangs) for GOS = 0.01	= 0.005	= 0.002	= 0.001
2	0.153	0.105	0.065	0.046
4	0.869	0.701	0.535	0.439
5	1.36	1.13	0.900	0.762
10	4.46	3.96	3.43	3.09
20	12.0	11.1	10.1	9.41
24	15.3	14.2	13.0	12.2
40	29.0	27.3	25.7	24.5
70	56.1	53.7	51.0	49.2
100	84.1	80.9	77.4	75.2

Improving Capacity

As demand for service increases, system designers have to provide more channel per unit coverage area

Common Techniques are: Cell Splitting, Sectoring and Microcell Zoning

Cell Splitting increases the number of BS deployed and allows an orderly growth of the cellular system

Sectoring uses directional antennas to further control interference

Micro cell Zoning distributes the coverage of cell and extends the cell boundary to hard-to-reach areas

Cell Splitting

Cell splitting is the process of subdividing a congested cell into smaller cells with

their own BS

a corresponding reduction in antenna height

a corresponding reduction in transmit power

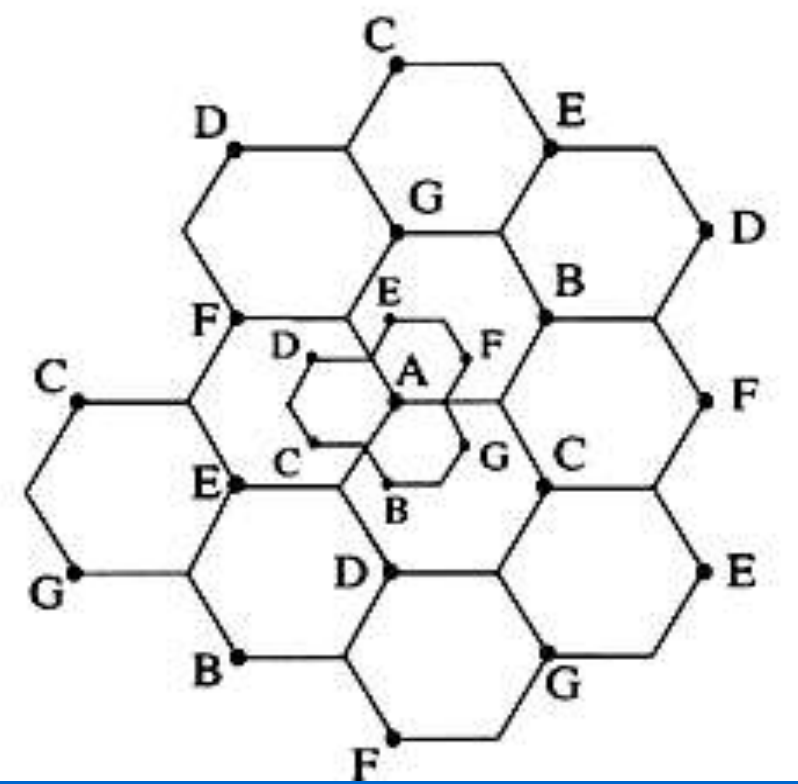
Splitting the cell reduces the cell size and thus more number of cells have to be used

For the new cells to be smaller in size the transmit power of these cells must be reduced.

Idea is to keep $Q=D/R$ constant while decreasing R

More number of cells ▸ more number of clusters ▸ more channels ▸high capacity

Cells are split to add channels with no new spectrum usage



Example S/I

cluster size N=4 (choices 4,7,12)

path loss exponent (means) n=4

co-channel reuse ratio Q= sqrt(3N)=3.464102

Ratio of distance to radius Q=D/R=3.464102

number of neighboring cells io=6 ,# of sides of hexagon

signal to interference ratio S/I= $(D/R)^{\wedge n}$ / i_o =24

convert to dB, S/I= 10log(S/I)=13.80211dB

S/I is less than required, it will not work!

cluster size N=7

path loss exponent n=3

Q=sqrt(3N)=4.582576

number of neighboring cells io= 6, # of sides of hexagon

signal to interference ratio=S/I= $(D/R)^{\wedge n}$ / i_o =73.334

convert to dB, S/I=10log(S/I)=18.65dB

S/I is less than required, it will work!

Cell Splitting-Power Issues

Suppose the cell radius of new cells is reduced by half

What is the required transmit power for these new cells??

Pr[at old cell boundary]=Pt1R^{-n}

Pr[at new cell boundary]= Pt2(R/2)$^{-n}$

where Pt1and Pt2are the transmit powers of the larger and smaller cell base stations respectively, and n is the path loss exponent.

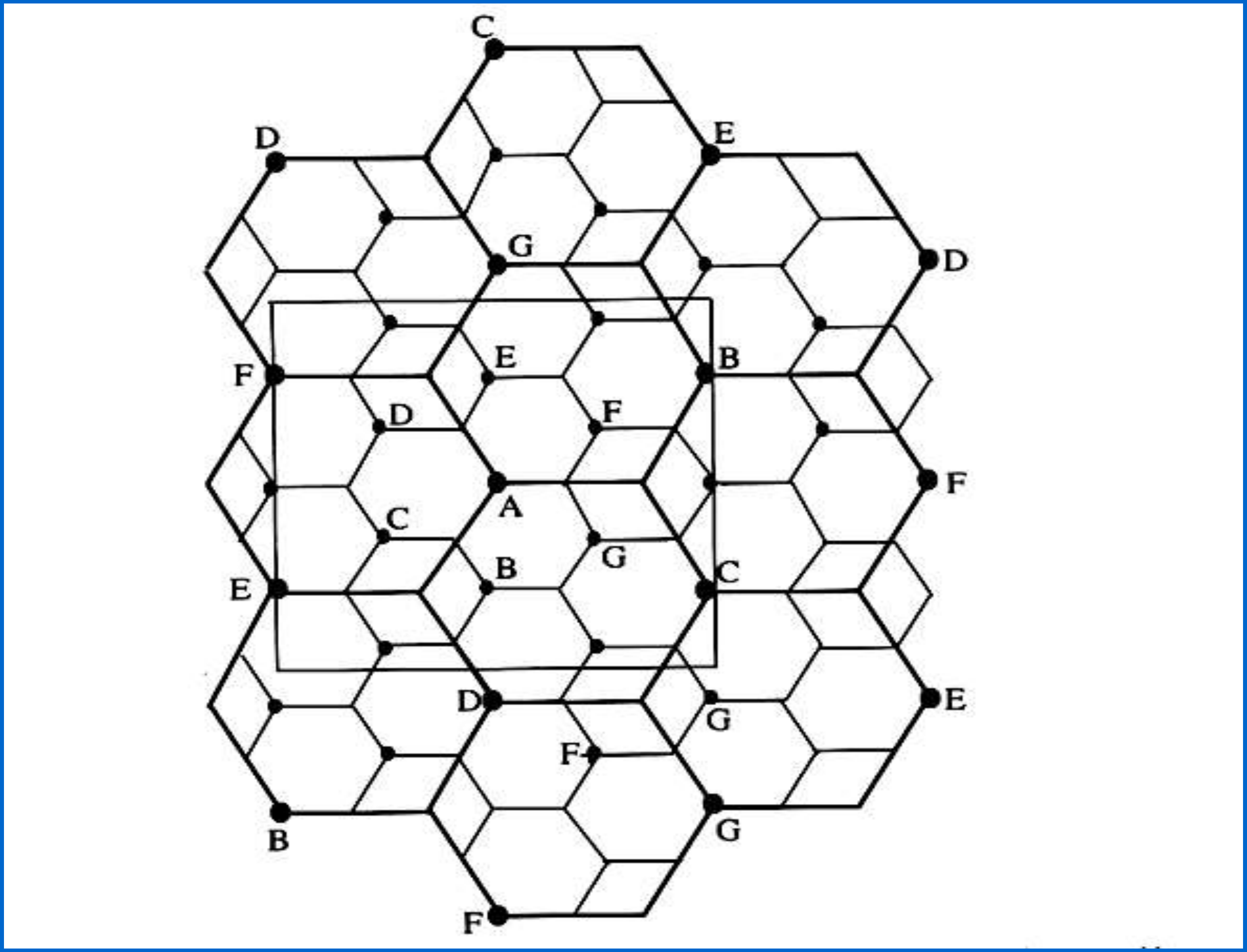
So, **Pt2= Pt1/2^{n}**

If we take *n=3 and the received powers equal to each other, then*

Pt2=Pt1/8

In other words, the transmit power must be reduced by 9dB in order to fill in the original coverage area while maintaining the S/I requirement

Illustration of cell splitting in 3x3 square centered around base station A



Cell Splitting

In practice not all the cells are split at the same time hence different size cells will exist simultaneously.

In such situations, special care needs to be taken to keep the distance between co-channel cells at the required minimum, and hence channel assignments become more complicated.

To overcome handoff problem:

Channels in the old cell must be broken down into two channel groups, one for smaller cell and other for larger cell

The larger cell is usually dedicated to high speed traffic so that handoffs occur less frequently

At start small power group has less channels and large power group has large no of channels, at maturity of the system large power group does not have any channel

Umbrella Cells

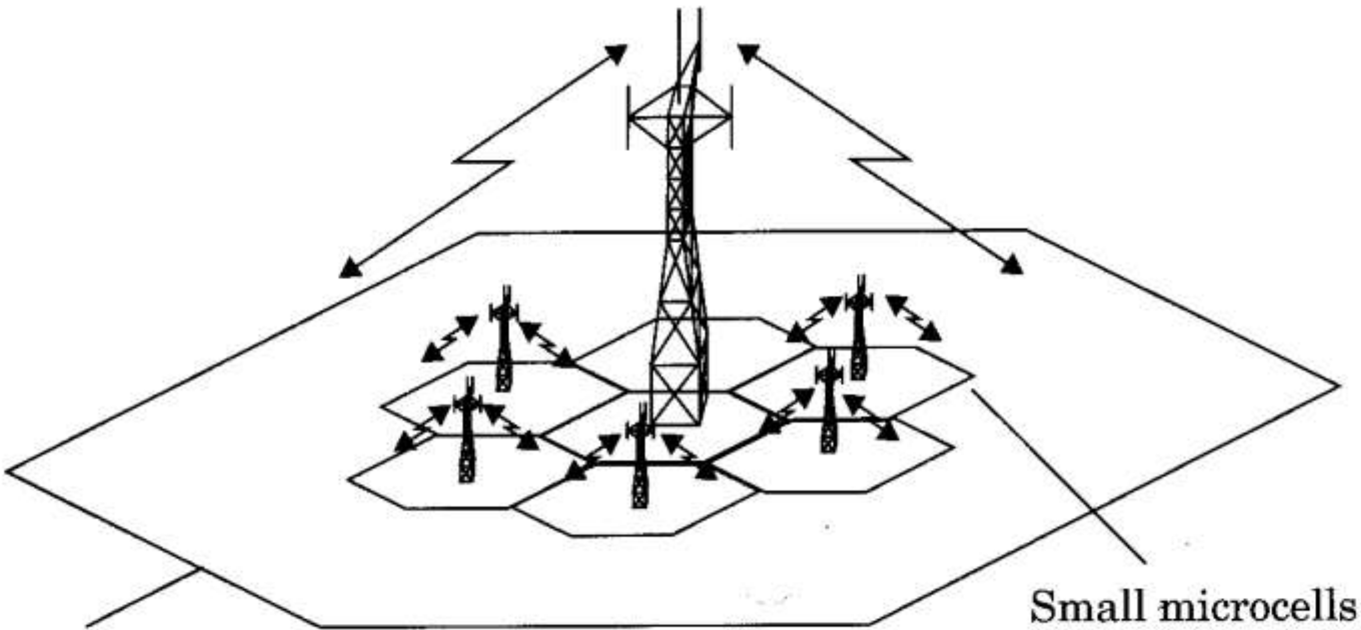
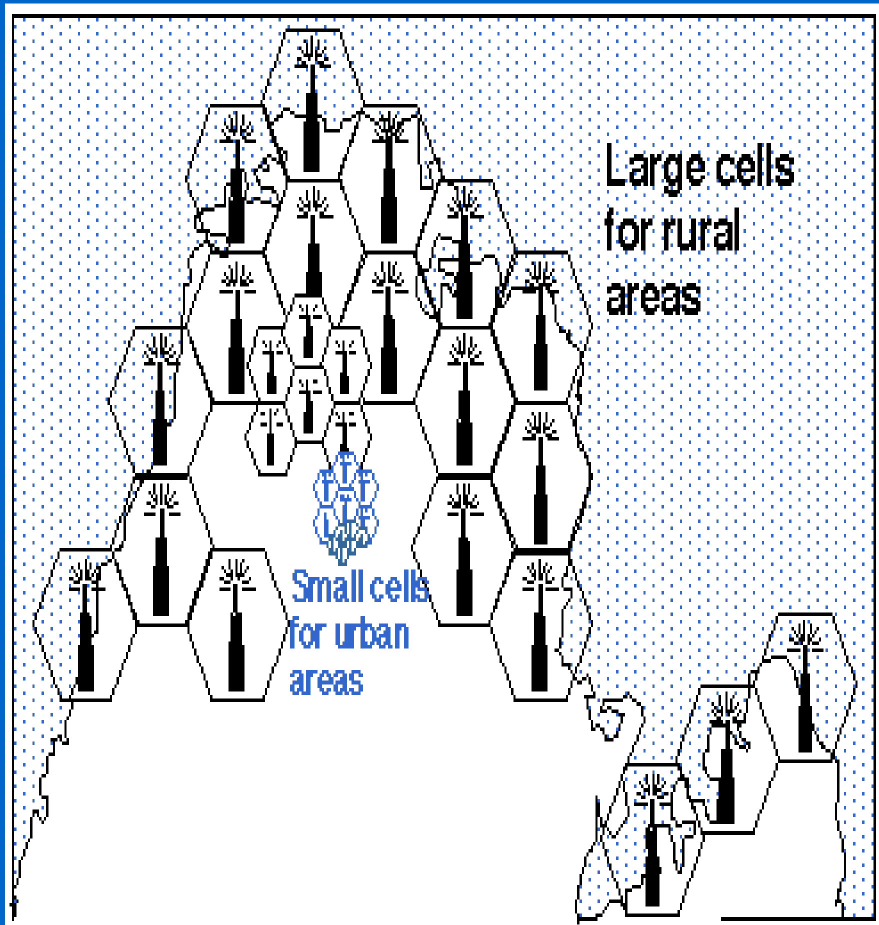
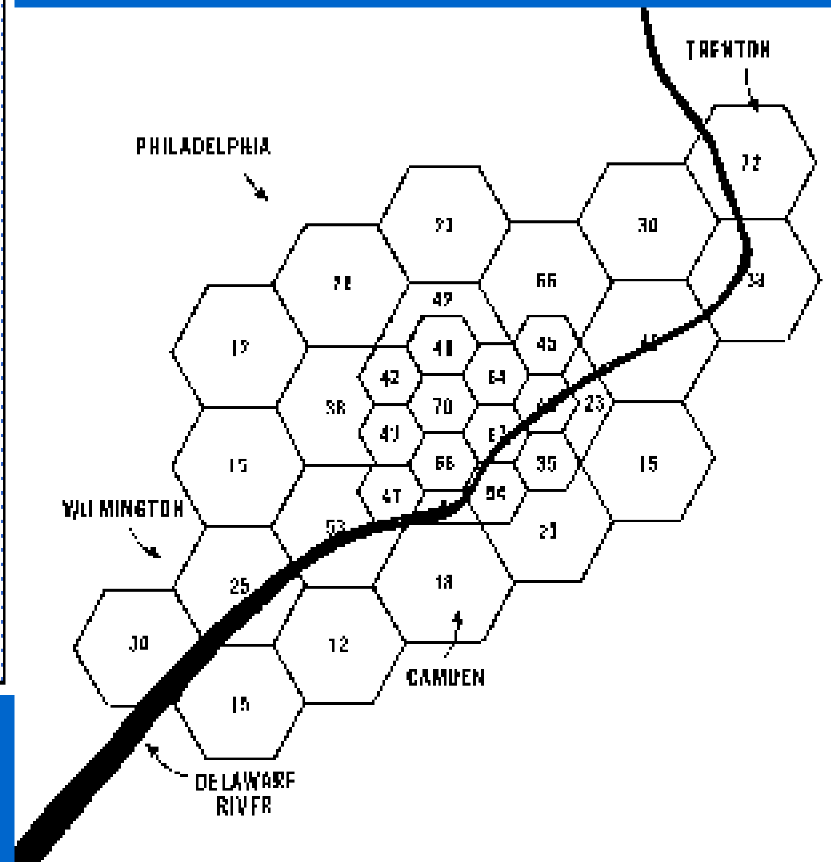


Figure 3.4 The umbrella cell approach.


Large cells for rural areas
Small cells for urban areas


TRENTON
PHILADELPHIA
WILMINGTON
CAMDEN
DELAWARE RIVER

Sectoring

In this approach

first SIR is improved using directional antennas,

capacity improvement is achieved by reducing the number of cells in a cluster thus increasing frequency reuse

The CCI decreased by replacing the single omni-directional antenna by several directional antennas, each radiating within a specified sector

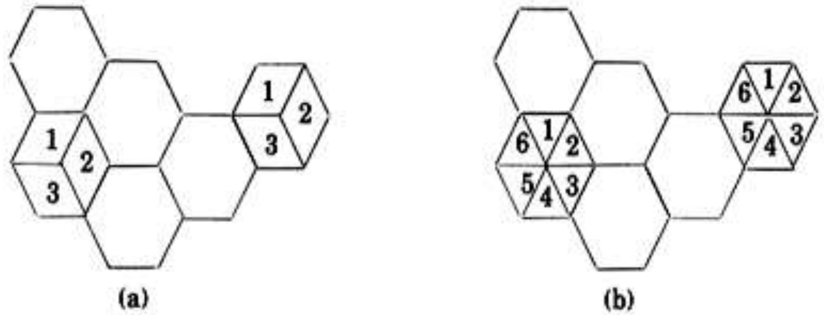


Figure 3.10 (a) 120° sectoring; (b) 60° sectoring.

Sectoring

A directional antenna transmits to and receives from only a fraction of total of the co-channel cells. Thus CCI is reduced

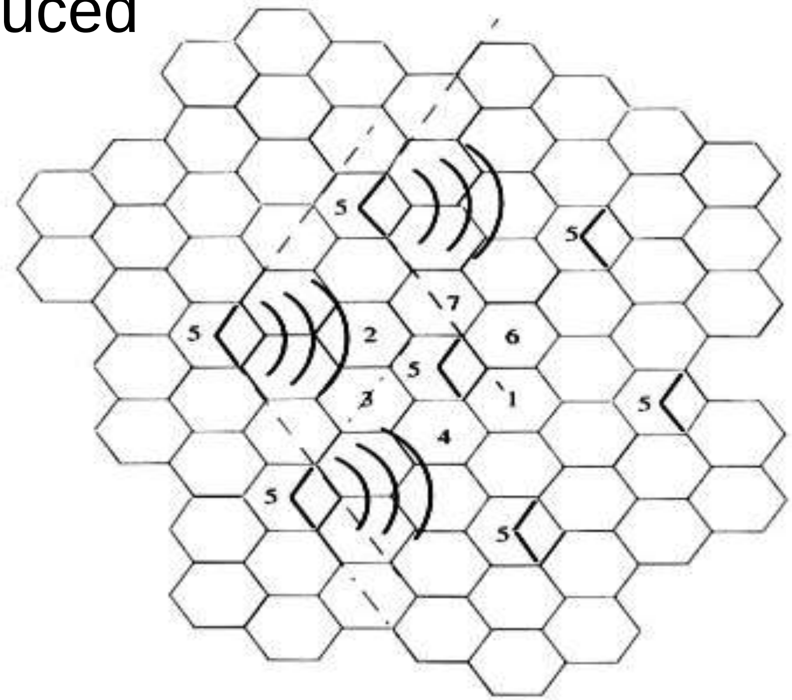


Figure 3.11 Illustration of how 120° sectoring reduces interference from co-channel cells. Out of the 6 co-channel cells in the first tier, only two of them interfere with the center cell. If omnidirectional antennas were used at each base station, all six co-channel cells would interfere with the center cell.

Problems with Sectoring

Increases the number of antennas at each BS

Decrease in trunking efficiency due to sectoring(dividing the bigger pool of channels into smaller groups)

Increase number of handoffs(sector-to sector)

Good news:Many modern BS support sectoring and related handoff without help of MSC

Microcell Zone Concept

The Problems of sectoring can be addressed by Microcell Zone Concept

A cell is conceptually divided into microcells or zones

Each microcell(zone) is connected to the same base station(fiber/microwave link)

Doing something in middle of cell splitting and sectoring by extracting good points of both

Each zone uses a directional antenna

Each zone radiates power into the cell.

MS is served by strongest zone

As mobile travels from one zone to another, it retains the same channel, i.e. no hand off

The BS simply switches the channel to the next zone site

Micro Zone Cell Concept

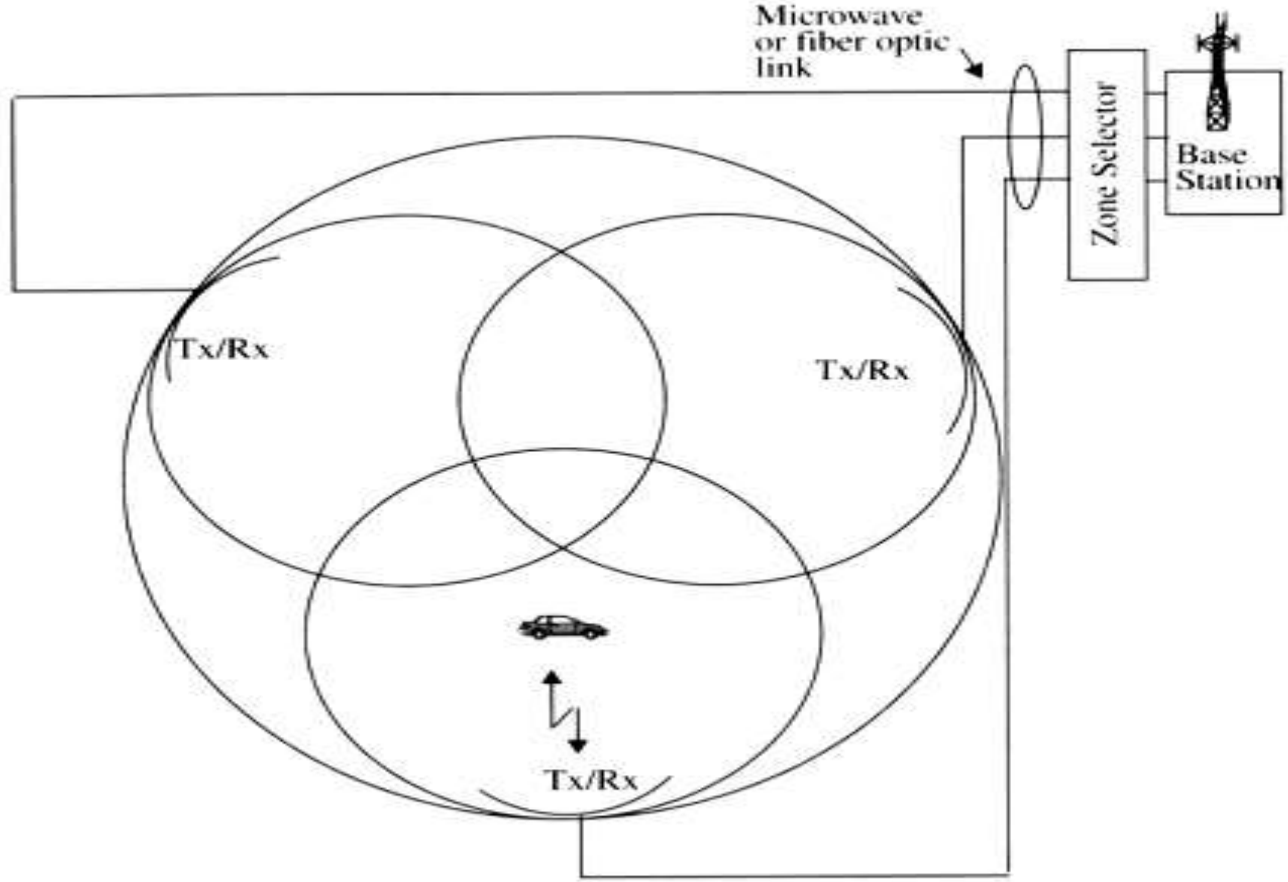


Figure 3.13 The microcell concept [adapted from [Lee91b] © IEEE].

Microcell Zone Concept

Reduced Interference (Zone radius is small so small and directional antennas are used).

Decrease in CCI improves the signal quality and capacity.

No loss in trunking efficiency (all channels are used by all cells).

No extra handoffs.

Increase in capacity (since smaller cluster size can be used).

Repeaters for Range Extension

Useful for hard to reach areas

Buildings

Tunnels

Valleys

Radio transmitters called Repeaters can be used to provide coverage in these area

Repeaters are **bi-directional**

Rx signals from BS

Amplify the signals

Re-radiate the signals

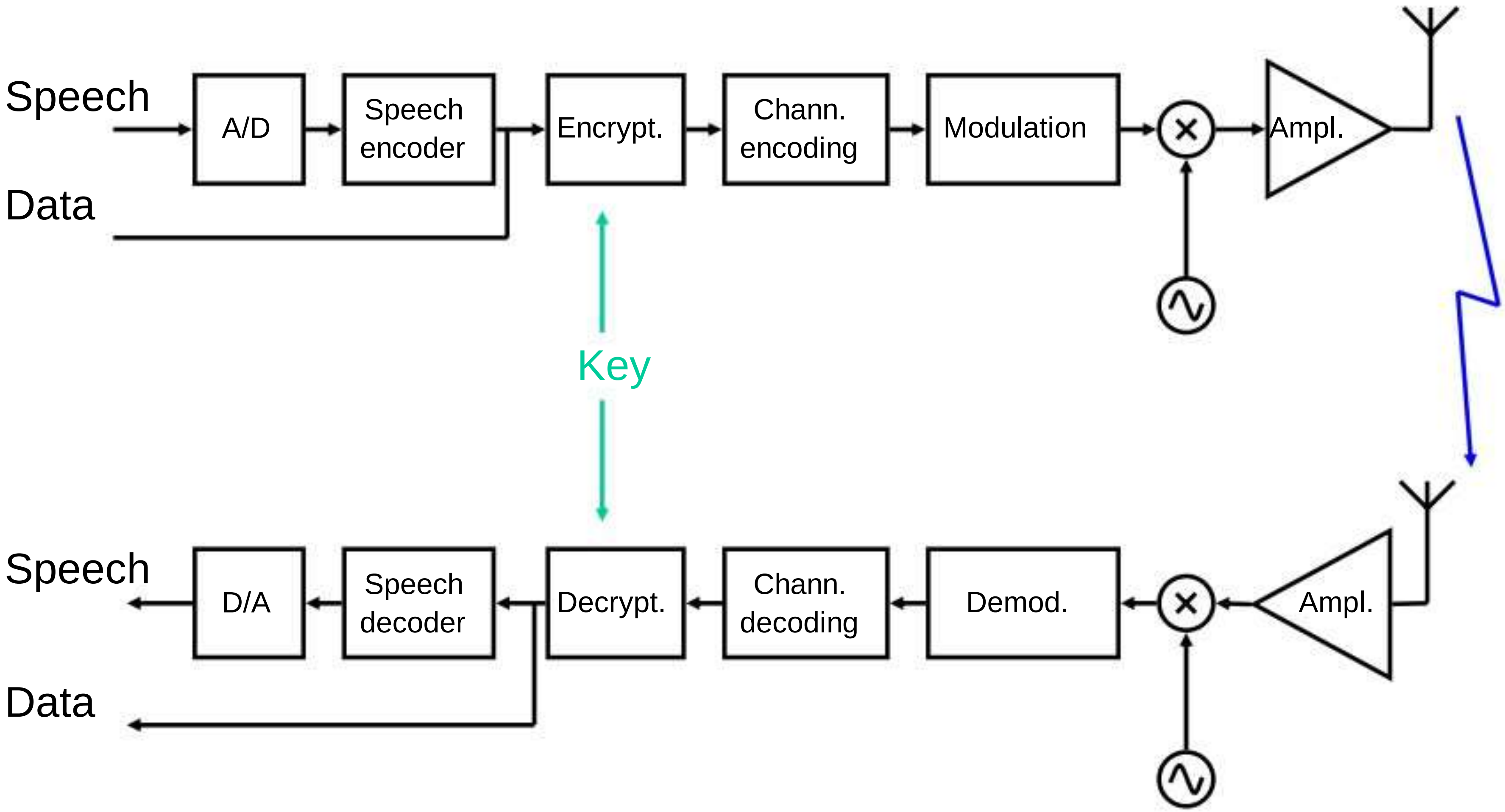
Received noise and interference is also re-radiated

UNIT - III WIRELESS TRANSCEIVERS

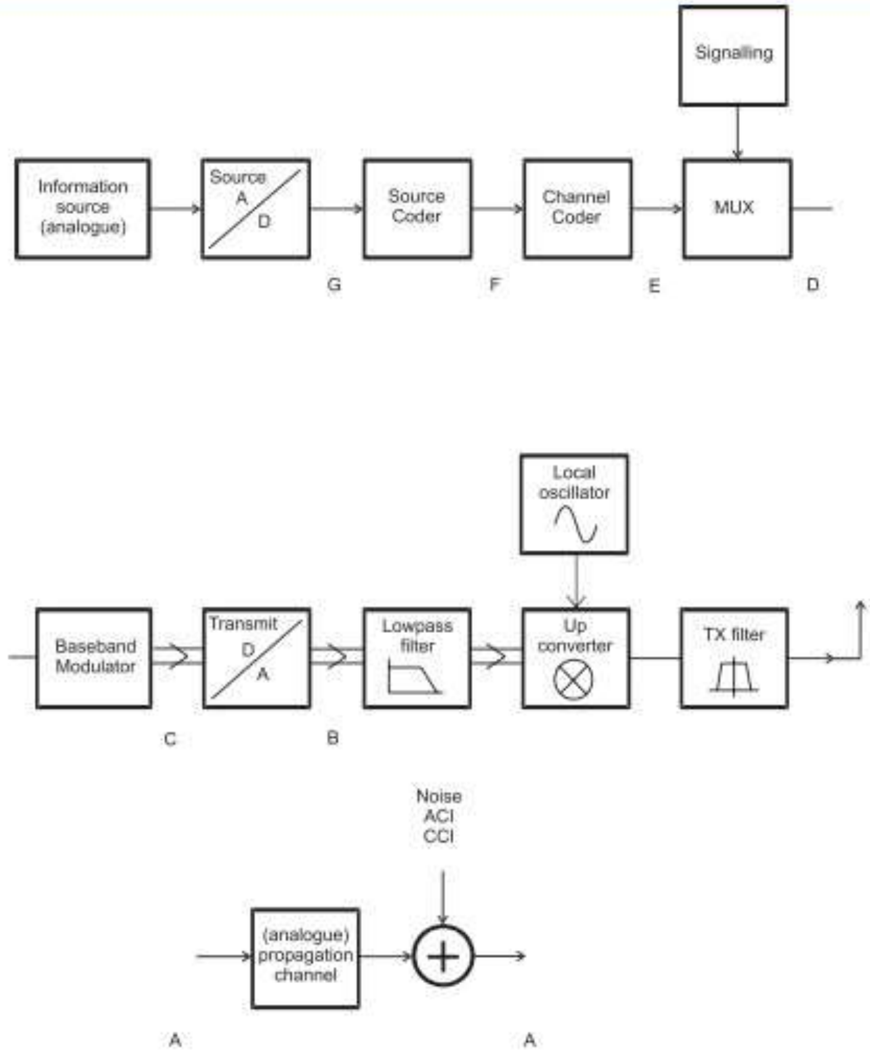
- Unit Syllabus
  - Structure of a Wireless Communication Link
  - Modulation
    - QPSK
    - $\pi/4$ - DQPSK
    - OQPSK
    - BFSK
    - MSK
    - GMSK
  - Demodulation
    - Error Probability in AWGN
    - Error Probability in Flat - Fading Channels
    - Error Probability in Delay and Frequency Dispersive Fading Channels

Structure of a wireless communications link

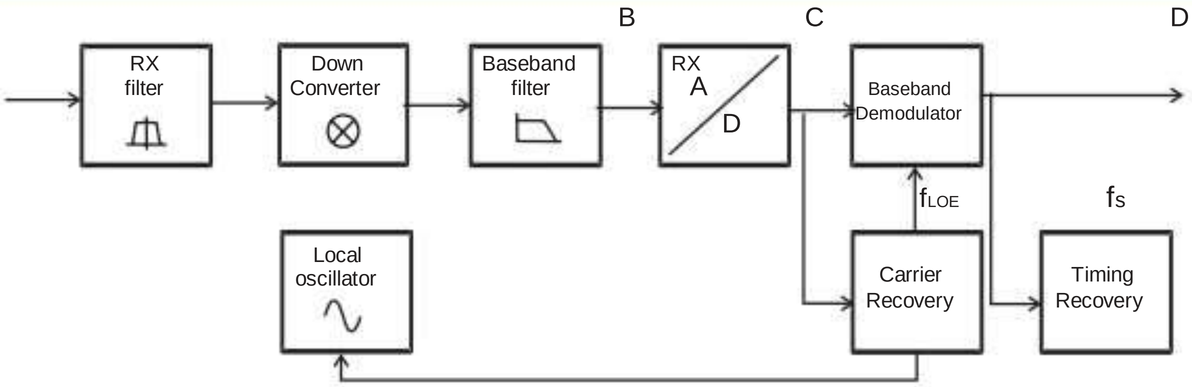
Block diagram

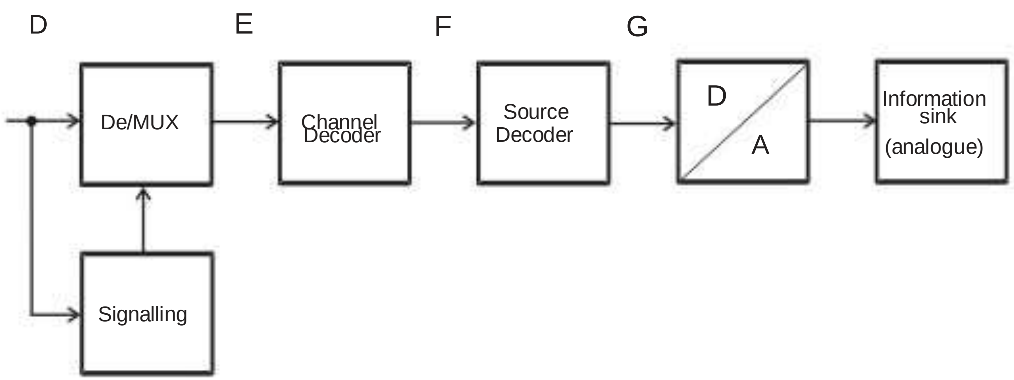


Block diagram transmitter



Block diagram receiver





Modulation

RADIO SIGNALS AND COMPLEX NOTATION

Simple model of a radio signal

- A transmitted radio signal can be written

$$s(t) = A\cos(2\pi f t + \phi)$$

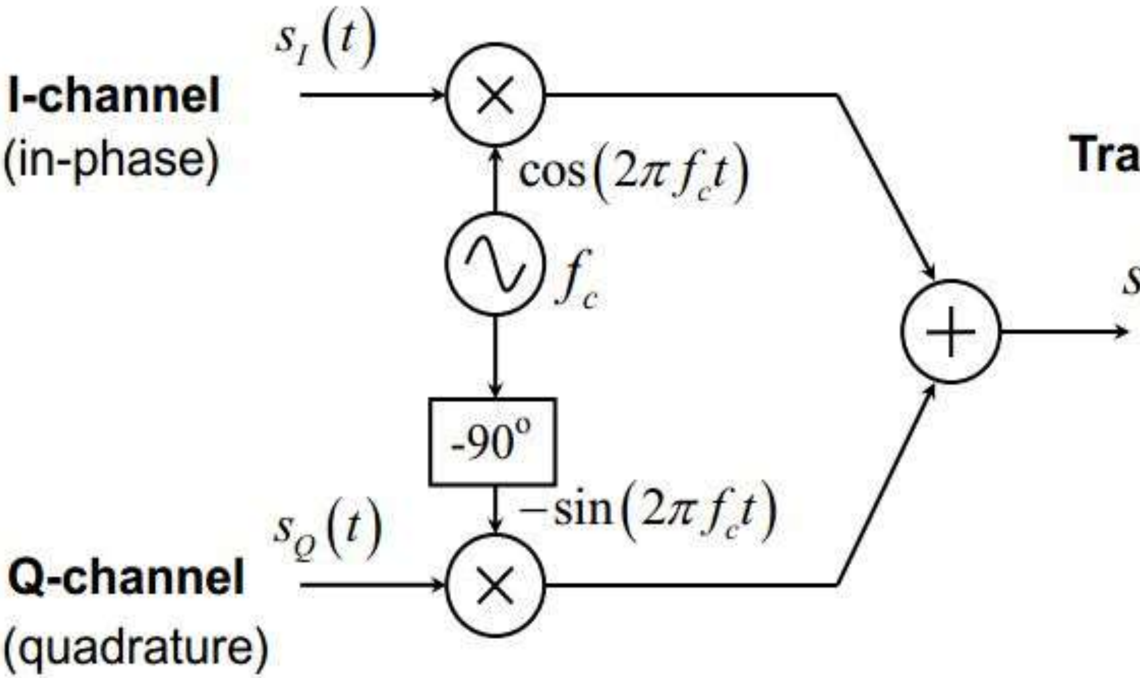
Amplitude Frequency Phase

- By letting the transmitted information change the amplitude, the frequency, or the phase, we get the tree basic types of digital modulation techniques

  - ASK (Amplitude Shift Keying)
  - FSK (Frequency Shift Keying)
  - PSK (Phase Shift Keying)

Constant amplitude

The IQ modulator



Transmited radio signal

$$s(t) = s_I(t)\cos(2\pi f_c t) - s_Q(t)\sin(2\pi f_c t)$$

Take a step into the complex domain:

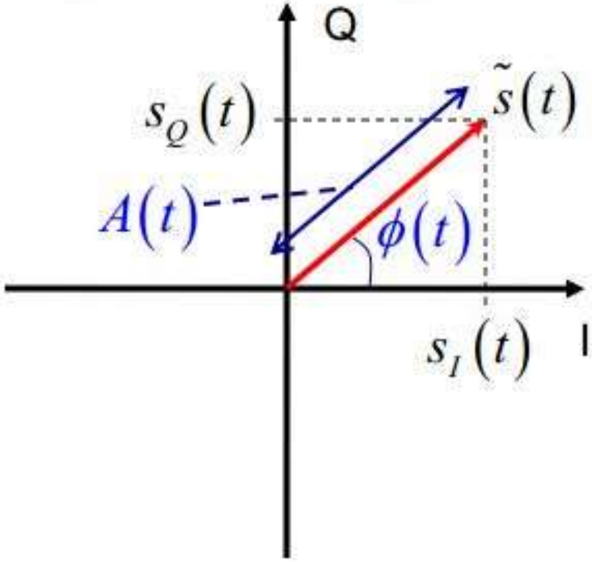
Complex envelope $\tilde{s}(t) = s_I(t) + j s_Q(t)$

Carrier factor $e^{j2\pi f_c t}$

$$\Rightarrow \quad s(t) = \mathrm{Re}\left\{\tilde{s}(t)\, e^{j2\pi f_c t}\right\}$$

Interpreting the complex notation

Complex envelope (phasor)



Polar coordinates:

$$\tilde{s}(t) = s_I(t) + js_Q(t) = A(t)e^{j\phi(t)}$$

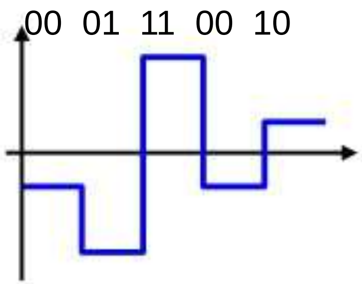
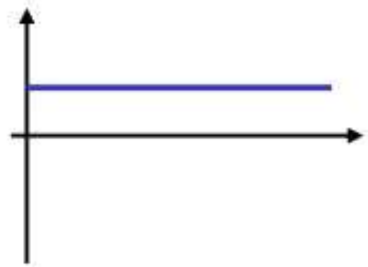
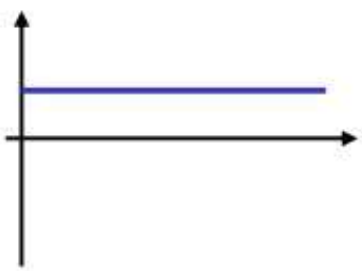
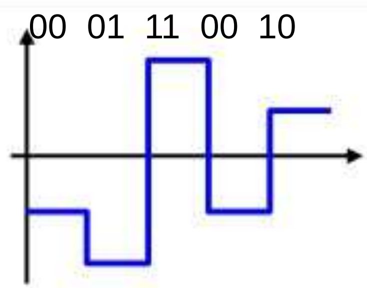
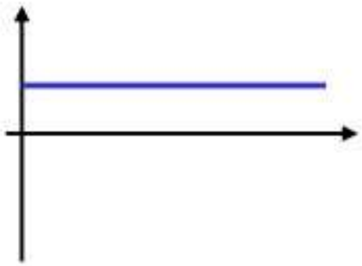
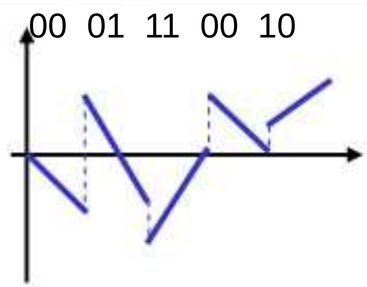
Transmitted radio signal

$$\begin{aligned} s(t) &= \mathrm{Re}\left\{\tilde{s}(t)e^{j2\pi f_c t}\right\} \\ &= \mathrm{Re}\left\{A(t)e^{j\phi(t)}e^{j2\pi f_c t}\right\} \\ &= \mathrm{Re}\left\{A(t)e^{j(2\pi f_c t + \phi(t))}\right\} \\ &= A(t)\cos\left(2\pi f_c t + \phi(t)\right) \end{aligned}$$

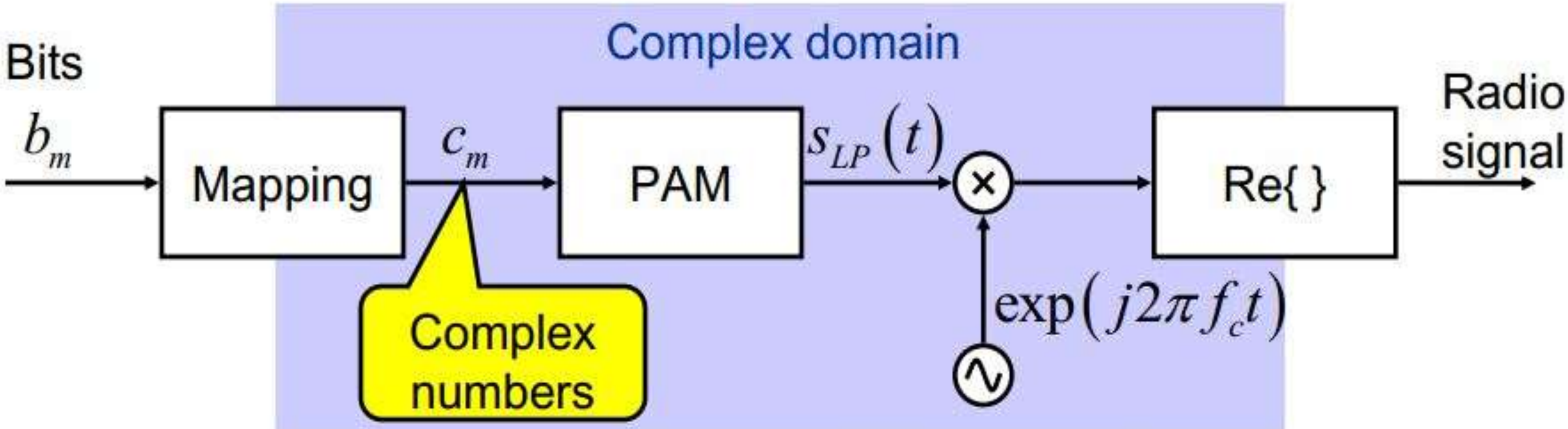
By manipulating the amplitude A(t) and the phase ϕ (t) of the complex envelope (phasor), we can create any type of modulation/radio signal.

Example: Amplitude, phase and frequency modulation

$$s(t) = A(t)\cos\left(2\pi f_c t + \phi(t)\right)$$

	$A(t)$	$\phi(t)$	Comment:
4ASK			- Amplitude carries information - Phase constant (arbitrary)
4PSK			- Amplitude constant (arbitrary) - Phase carries information
4FSK			- Amplitude constant (arbitrary) - Phase slope (frequency) carries information

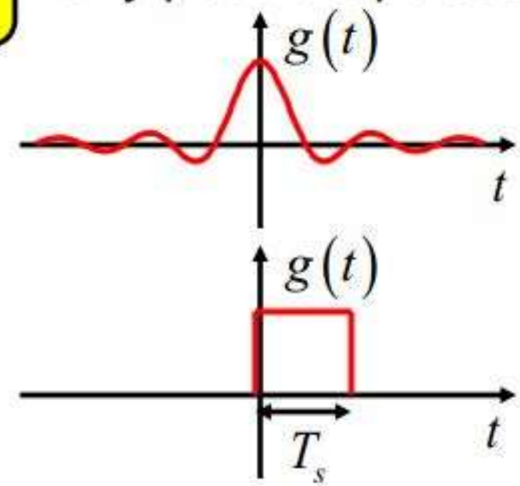
Pulse amplitude modulation (PAM)
The modulation process



PAM: $s_{LP}(t) = \sum_{m=-\infty}^{\infty} c_m g(t - mT_s)$

Symbol time

Many possible pulses



"Standard" basis pulse criteria

$\int_{-\infty}^{\infty} |g(t)|^2 \, dt = 1 \text{ or } = T_s$ (energy norm.)

$\int_{-\infty}^{\infty} g(t) g^*(t - mT_s) \, dt = 0, \; m \neq 0$ (orthogonality)

Pulse amplitude modulation (PAM)
Basis pulses and spectrum

Assuming that the complex numbers c_m representing the data are independent, then the power spectral density of the base band PAM signal becomes:

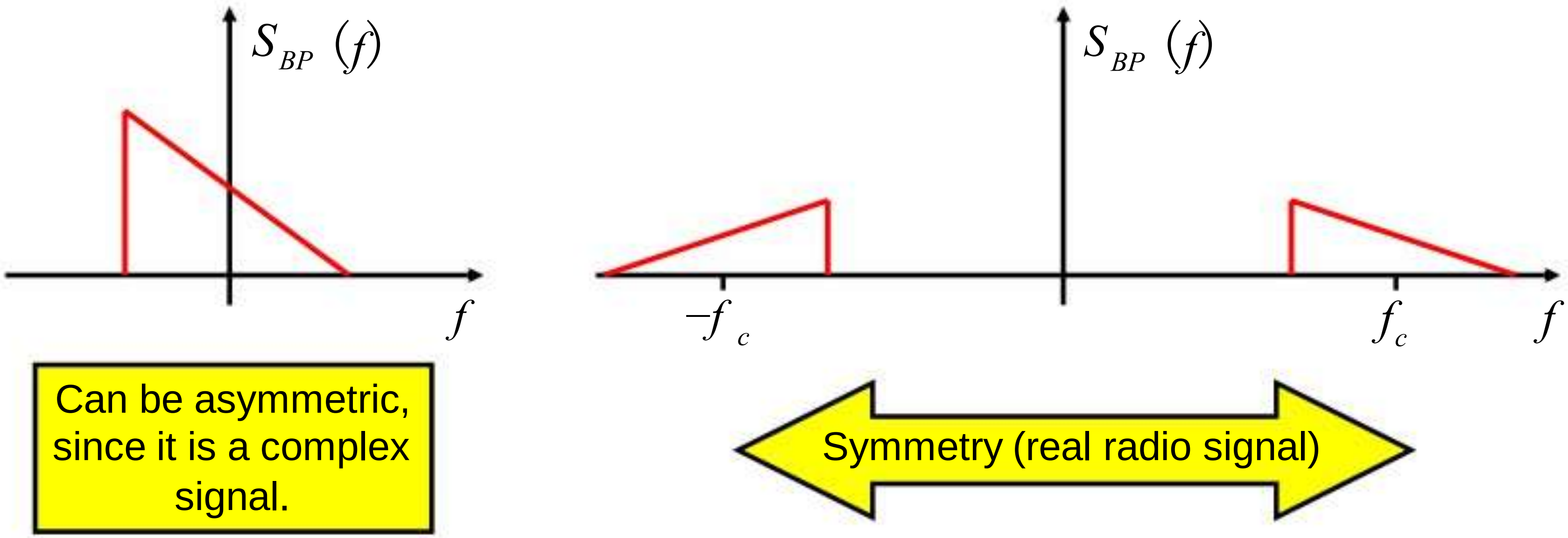
$$S_{LP}(f) \propto \left| \int_{-\infty}^{\infty} g(t)\, e^{-j2\pi ft}\, dt \right|^2$$

which translates into a radio signal (band pass) with

$$S_{BP}(f) = \frac{1}{2}\left(S_{LP}(f - f_c) + S_{LP}(-f - f_c) \right)$$

Pulse amplitude modulation (PAM)
Basis pulses and spectrum

Illustration of power spectral density of the (complex) base-band signal, $S_{LP}(f)$, and the (real) radio signal, $S_{BP}(f)$.



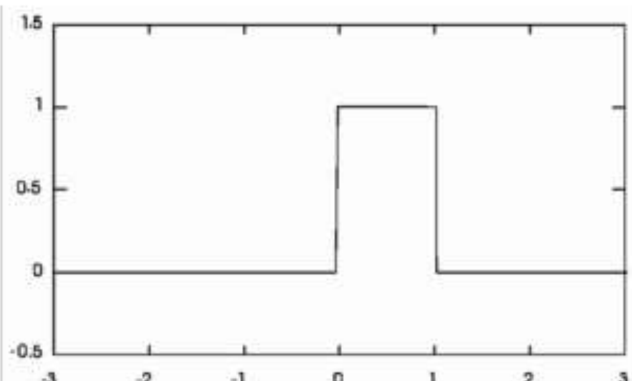
What we need are basis pulses g(t) with nice properties like:

- Narrow spectrum (low side-lobes)
- Relatively short in time (low delay)

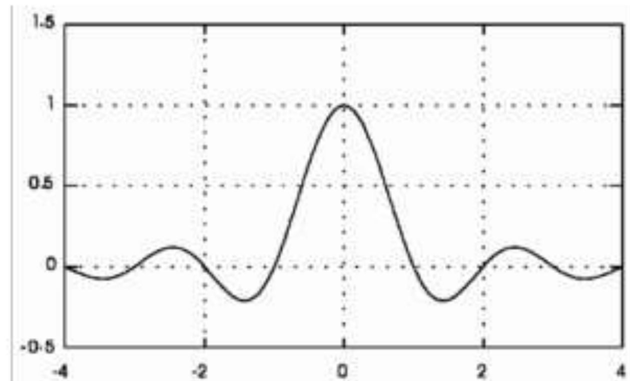
Pulse amplitude modulation (PAM)
Basis pulses

TIME DOMAIN | FREQ. DOMAIN

Rectangular [in time]

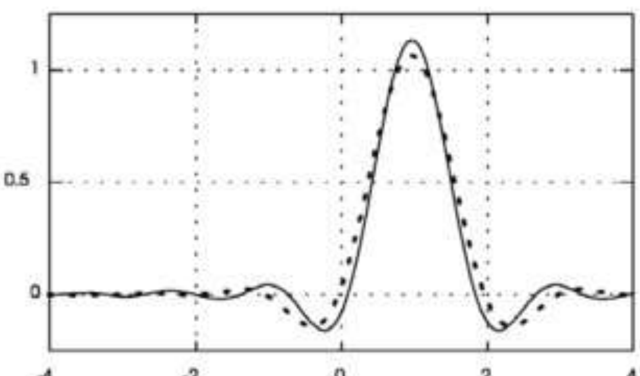


Normalized time t/T_s

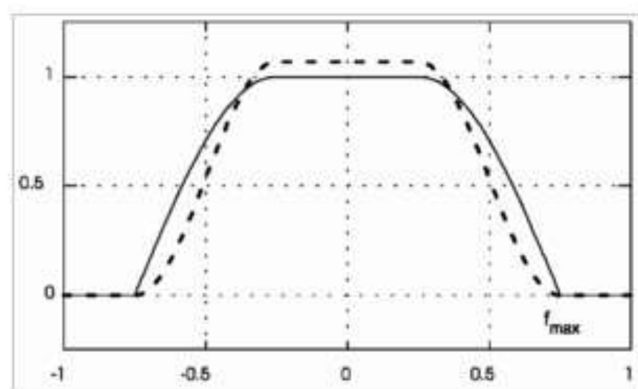


Normalized freq. f/T_s

(Root-) Raised-cosine [in freq.]



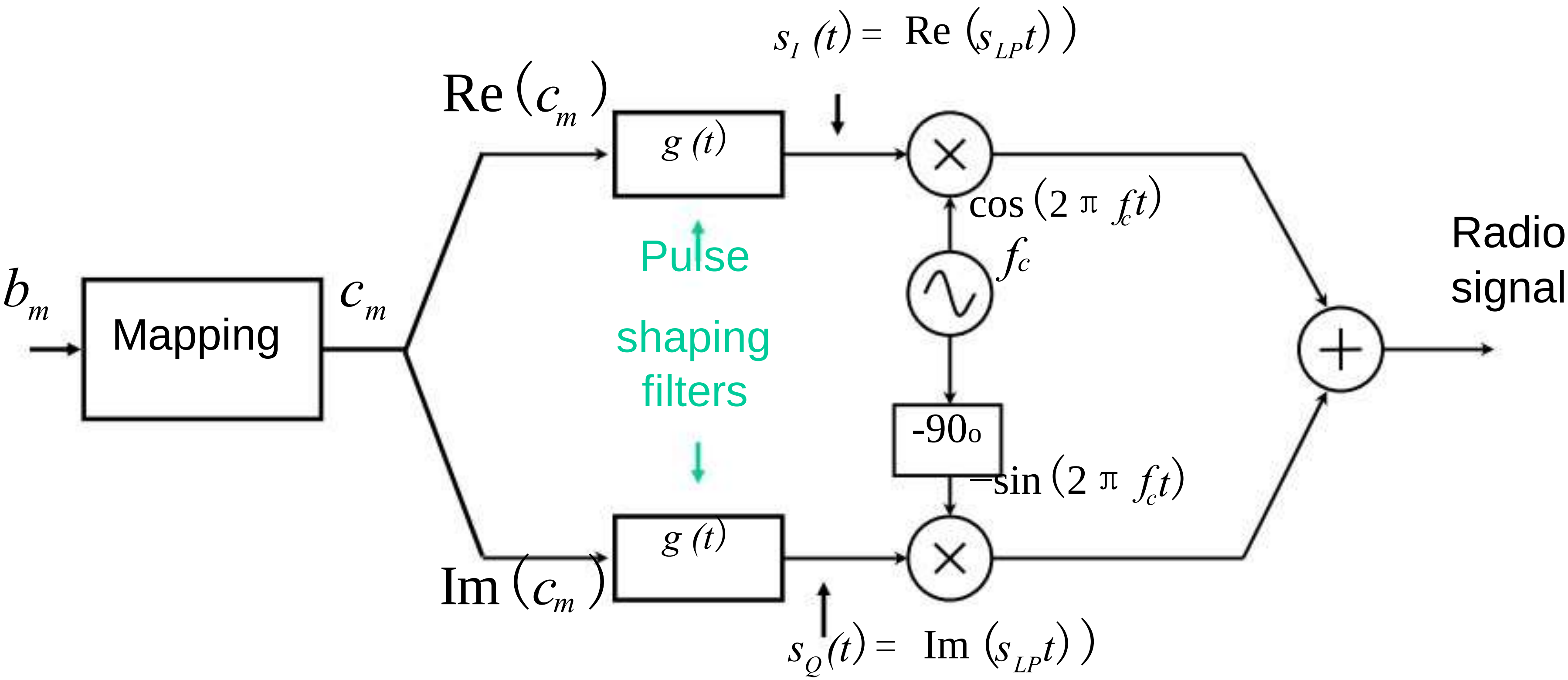
Normalized time t/T_s



Normalized freq. f/T_s

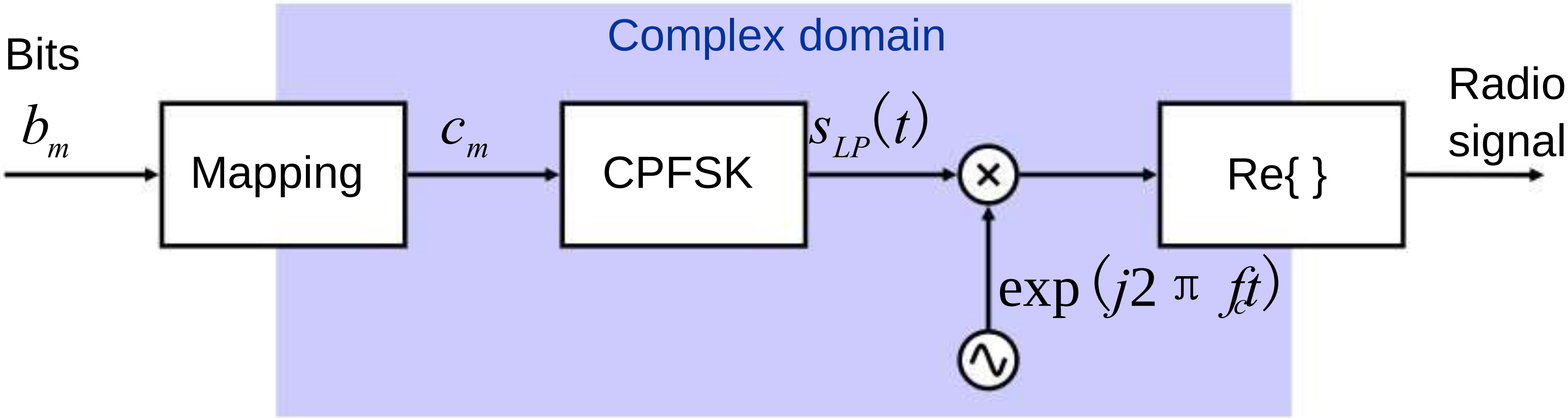
Pulse amplitude modulation (PAM) Interpretation as IQ-modulator

For real valued basis functions g(t) we can view PAM as:



(Both the rectangular and the (root-) raised-cosine pulses are real valued.)

Continuous-phase FSK (CPFSK)
The modulation process



CPFSK: $$s_{LP}(t) = A\exp\left(j\Phi_{CPFSK}(t)\right)$$

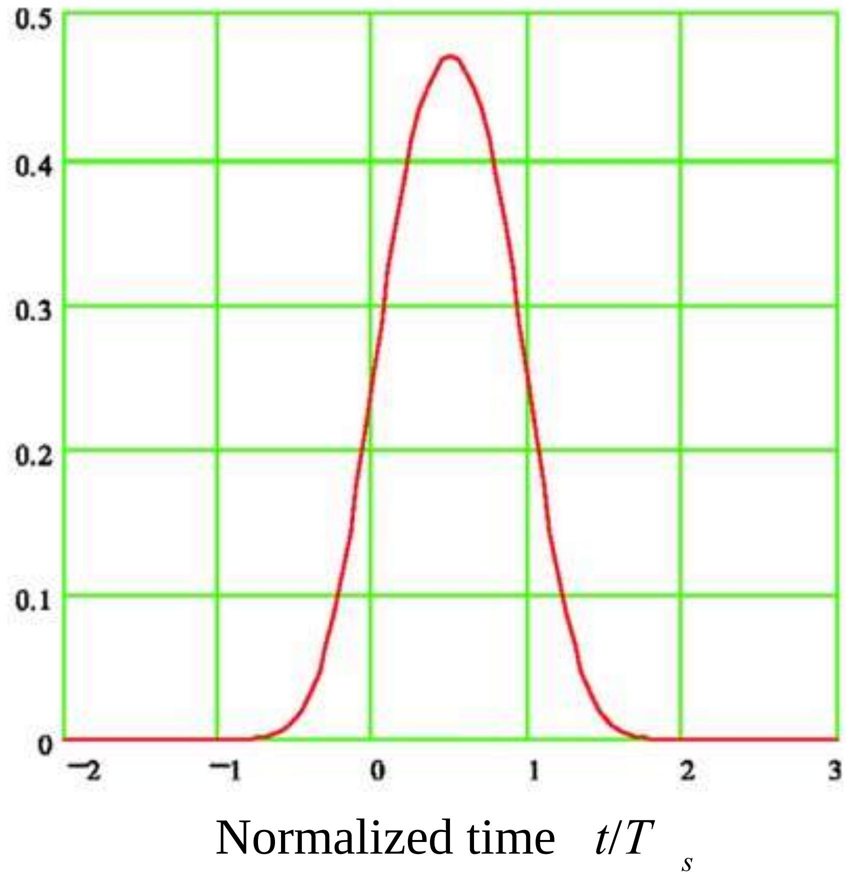
where the amplitude A is constant and the phase is

$$\Phi_{CPFSK}(t) = 2\pi h_{\mathrm{mod}} \sum_{m=-\infty}^{\infty} c_m \int_{-\infty}^{t} g(u-mT)\,du$$

Phase basis pulse

where h_{mod} is the modulation index.

Continuous-phase FSK (CPFSK)
The Gaussian phase basis pulse



$BT_s=0.5$

SIGNAL SPACE DIAGRAM

Principle of signal-space diagram (1)

- Represent a continuous signal by a discrete vector
- Choice of expansion functions:
  - In passband, usually

$$\varphi_{BP,1}(t) = \sqrt{\frac{2}{T_S}} \cos(2\pi f_c t)$$

$$\varphi_{BP,2}(t) = \sqrt{\frac{2}{T_S}} \sin(2\pi f_c t) \quad .$$

  - In baseband, usually

$$\varphi_1(t) = \sqrt{\frac{1}{T_S}} \cdot 1$$

$$\varphi_2(t) = \sqrt{\frac{1}{T_S}} \cdot j.$$

Principle of signal-space diagram (2)

- Signal vector for m-th signal

$$s_{m,n} = \int_0^{T_S} s_m(t)\varphi_n^*(t)dt$$

- Energy contained in signal

$$E_{S,m} = \int_0^{T_S} s_{BP,m}{}^2(t)dt = \|\mathbf{s}_{BP,m}\|^2$$

$$E_{S,m} \approx \frac{1}{2}\int_0^{T_S} \|s_{LP,m}(t)\|^2 dt = \frac{1}{2}\|\mathbf{s}_{LP,m}\|^2$$

- Correlation coefficients between signals *k* and *m*

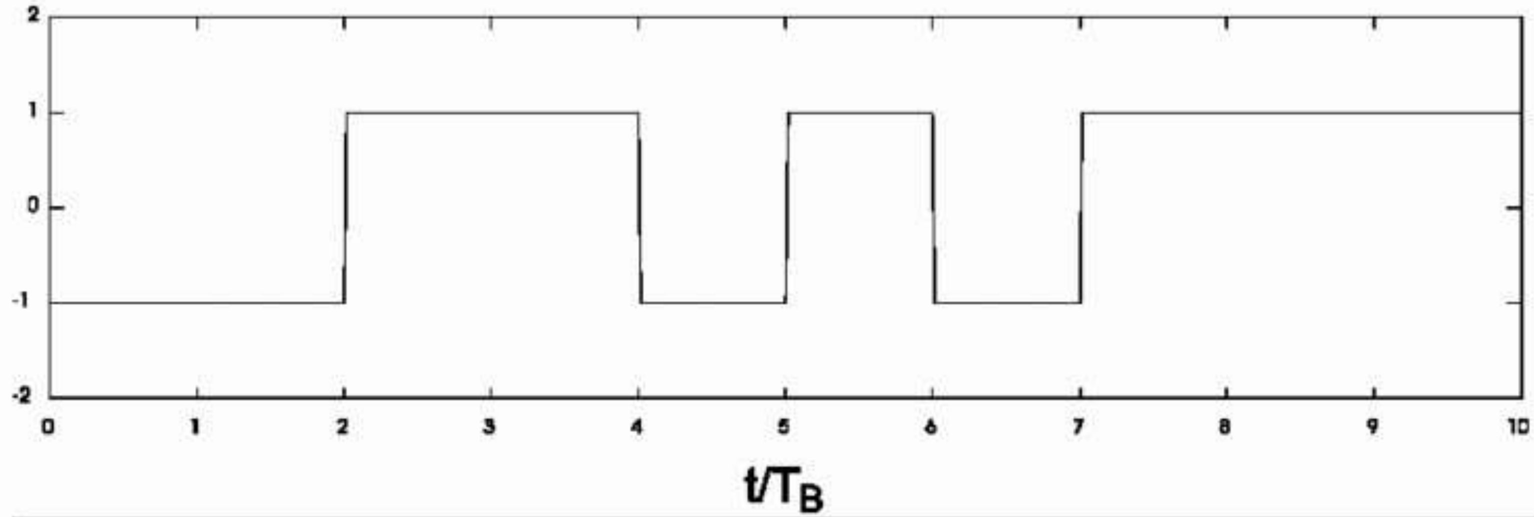
$$\mathrm{Re}\{\rho_{k,m}\} = \frac{\mathbf{s}_{BP,m}\mathbf{s}_{BP,k}}{\|\mathbf{s}_{BP,m}\|\,\|\mathbf{s}_{BP,k}\|}$$

- *Take care about normalization BP vs. LP*

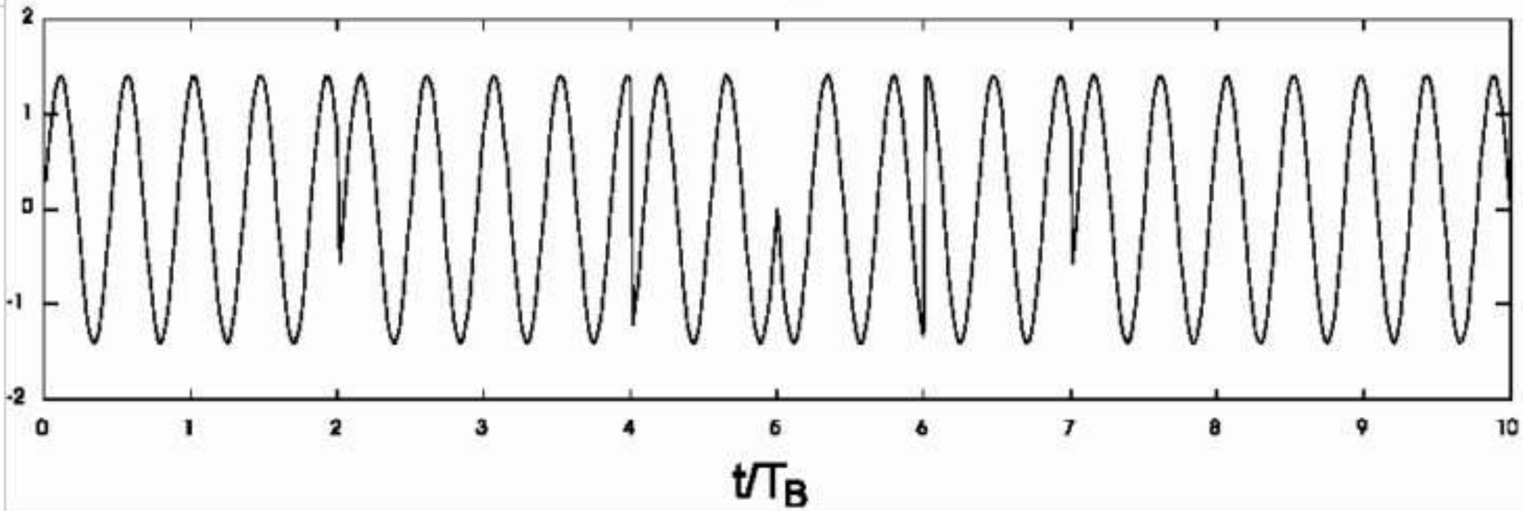
IMPORTANT MODULATION FORMATS

Binary phase-shift keying (BPSK)
Rectangular pulses

Base-band

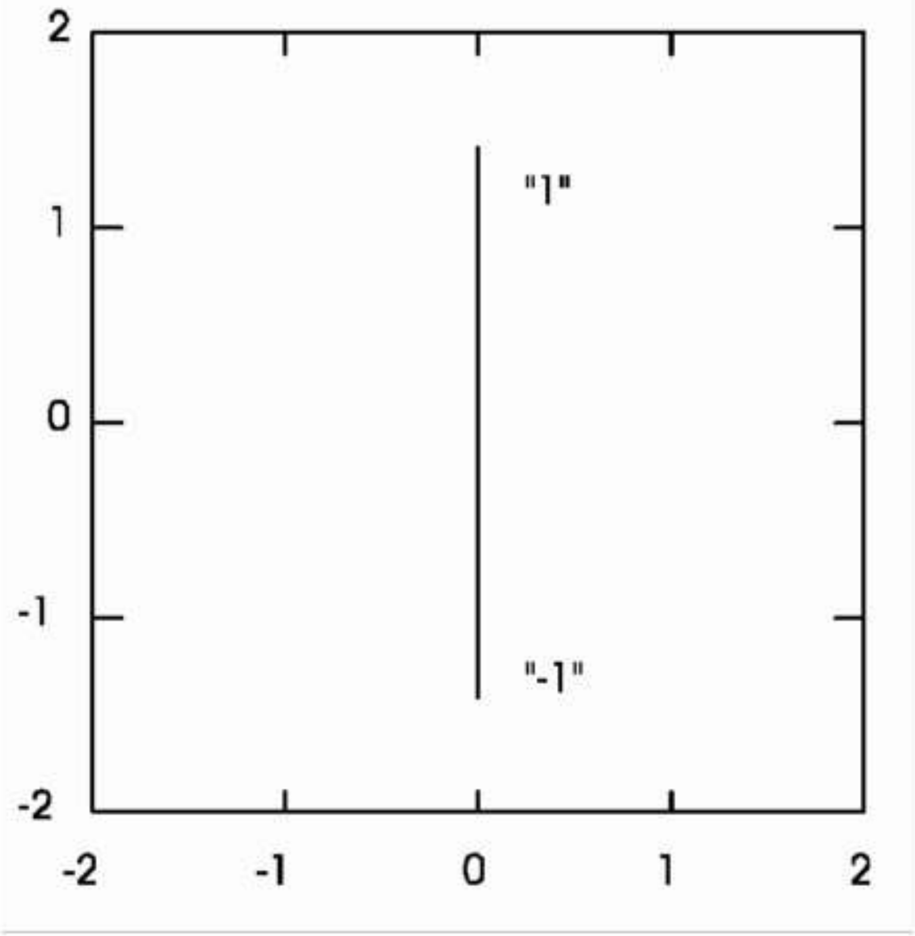


Radio signal

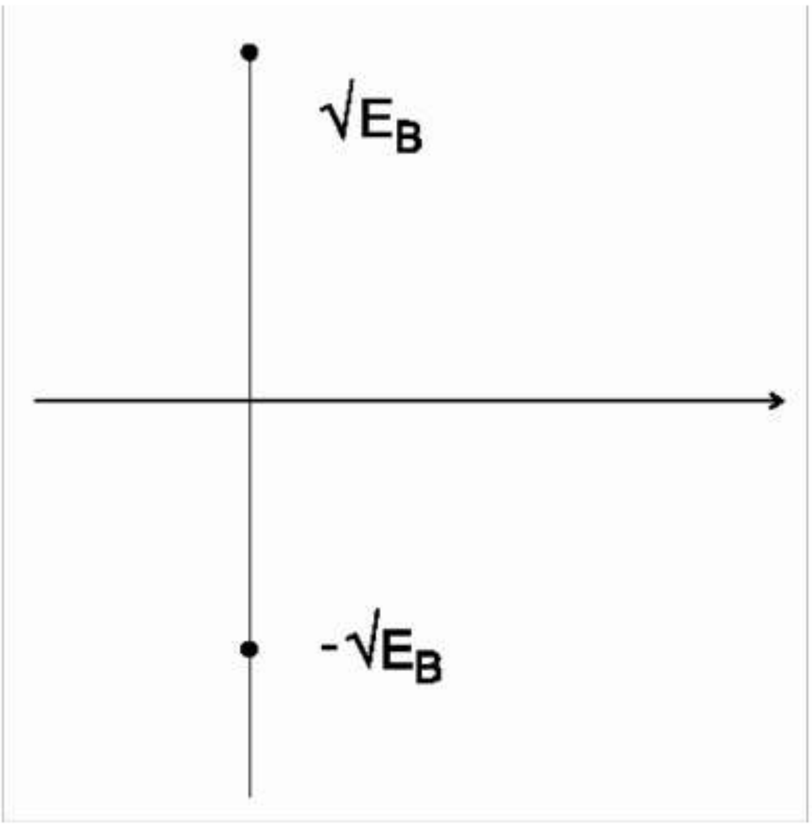


Binary phase-shift keying (BPSK)
Rectangular pulses

Complex representation

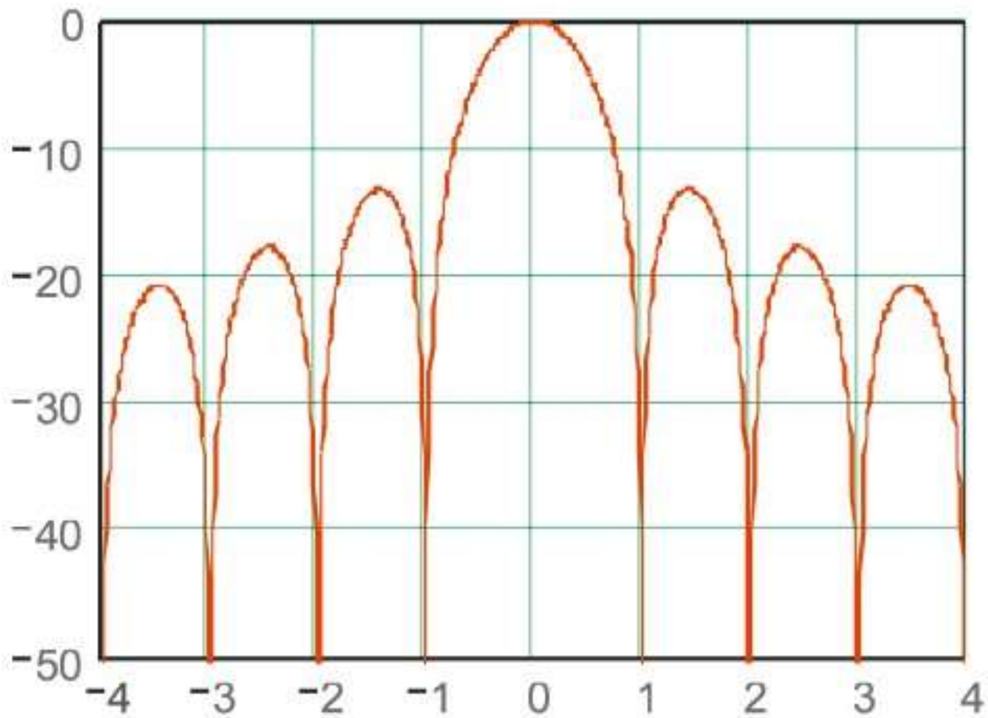


Signal space diagram



Binary phase-shift keying (BPSK) Rectangular pulses

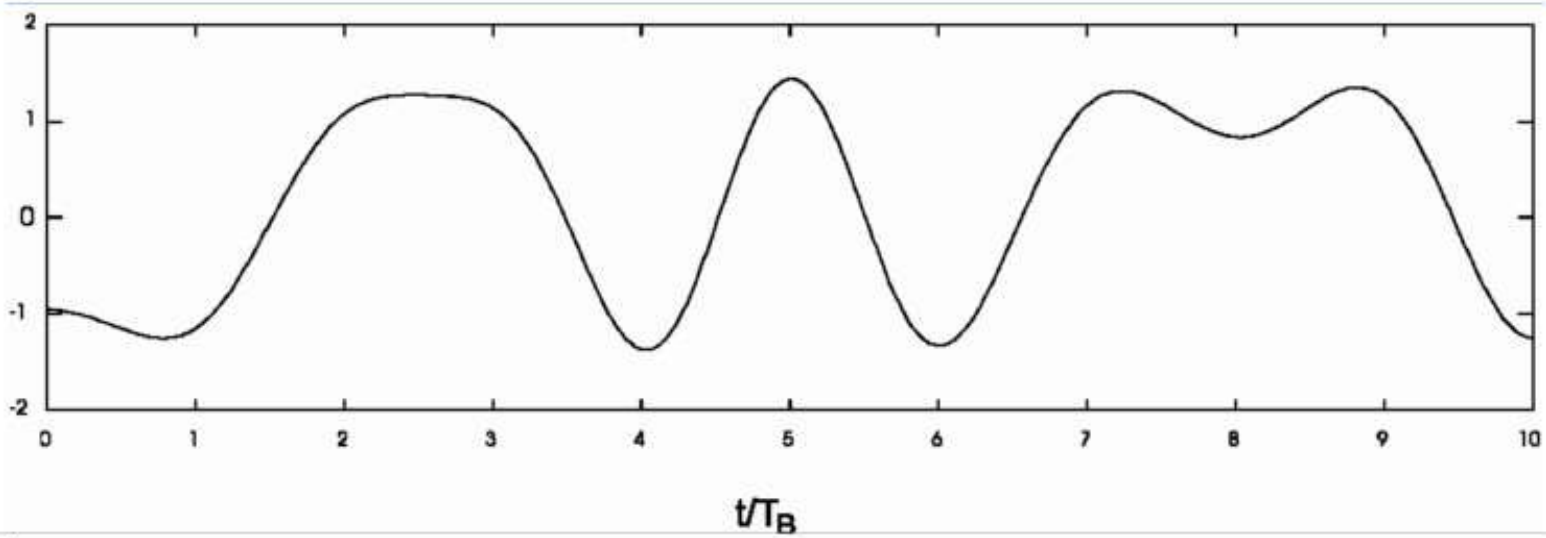
Power spectral density for BPSK



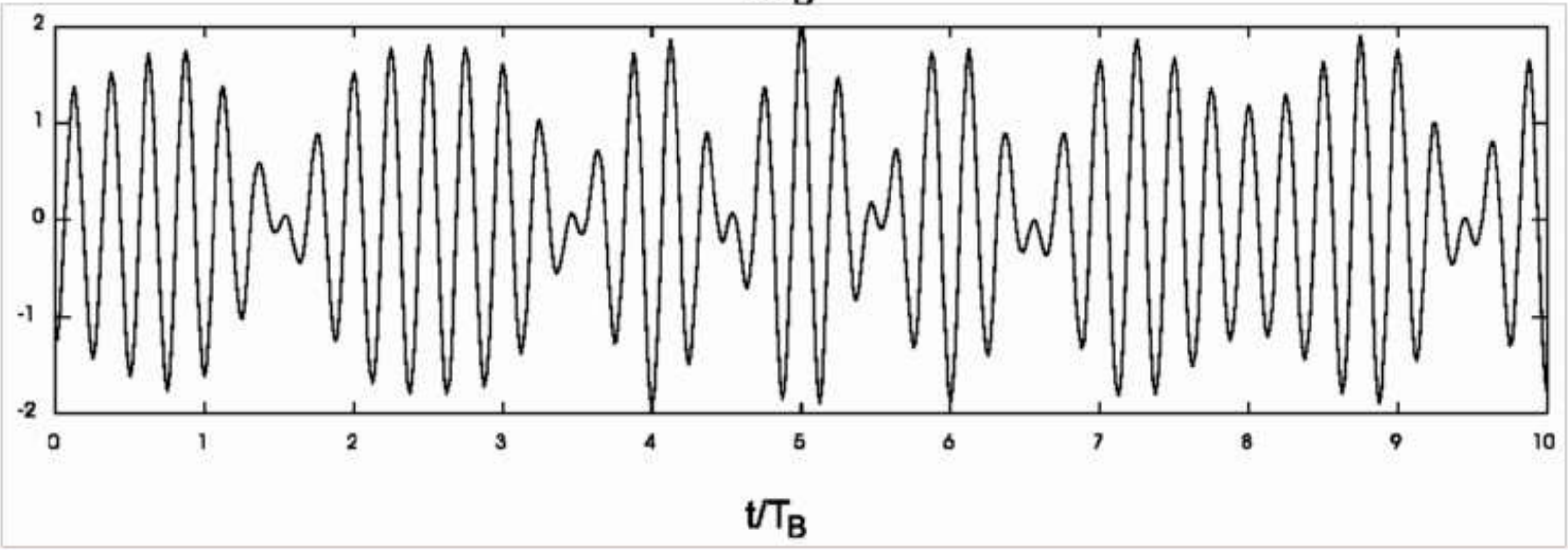
Contained percentage of total energy	spectral efficiency
90%	$0.59 Bit/s/Hz$
99%	$0.05 Bit/s/Hz$

Binary amplitude modulation (BAM)
Raised-cosine pulses (roll-off 0.5)

Base-band

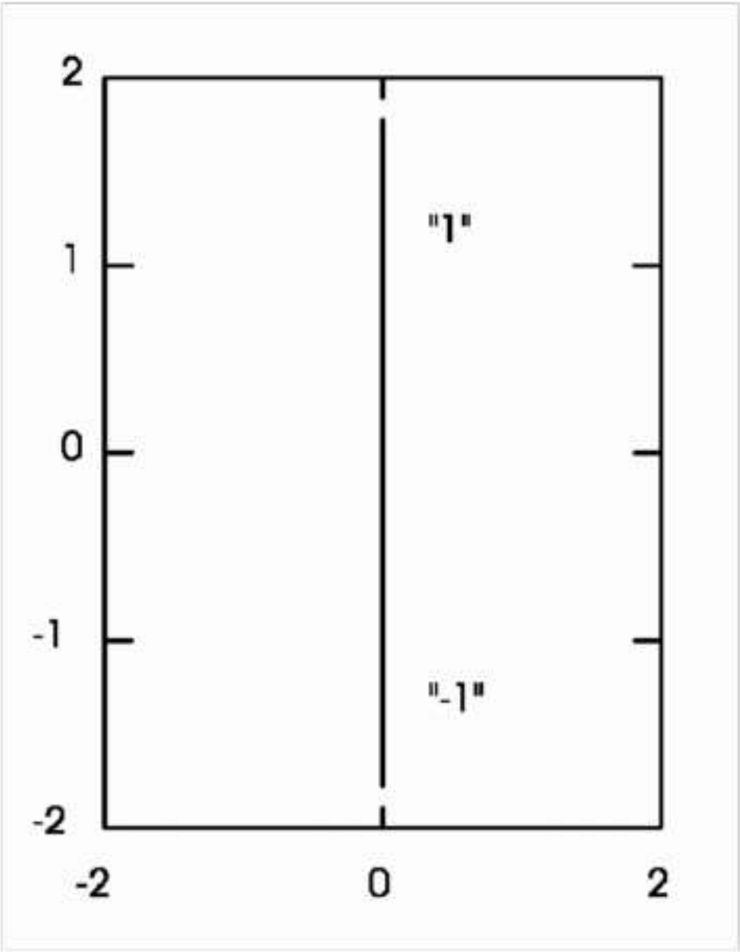


Radio signal

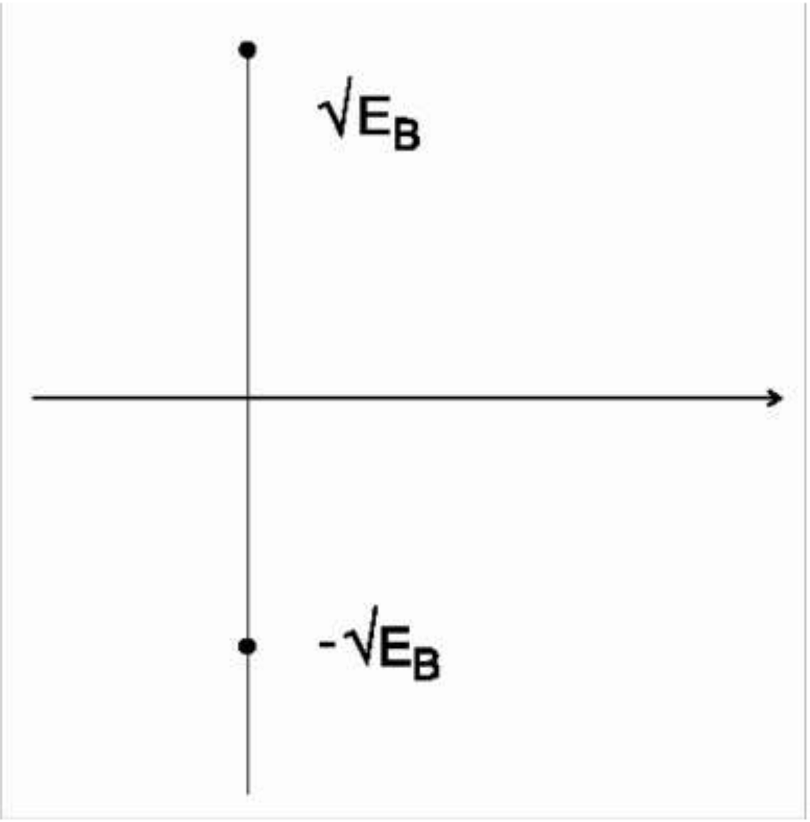


Binary amplitude modulation (BAM)
Raised-cosine pulses (roll-off 0.5)

Complex representation

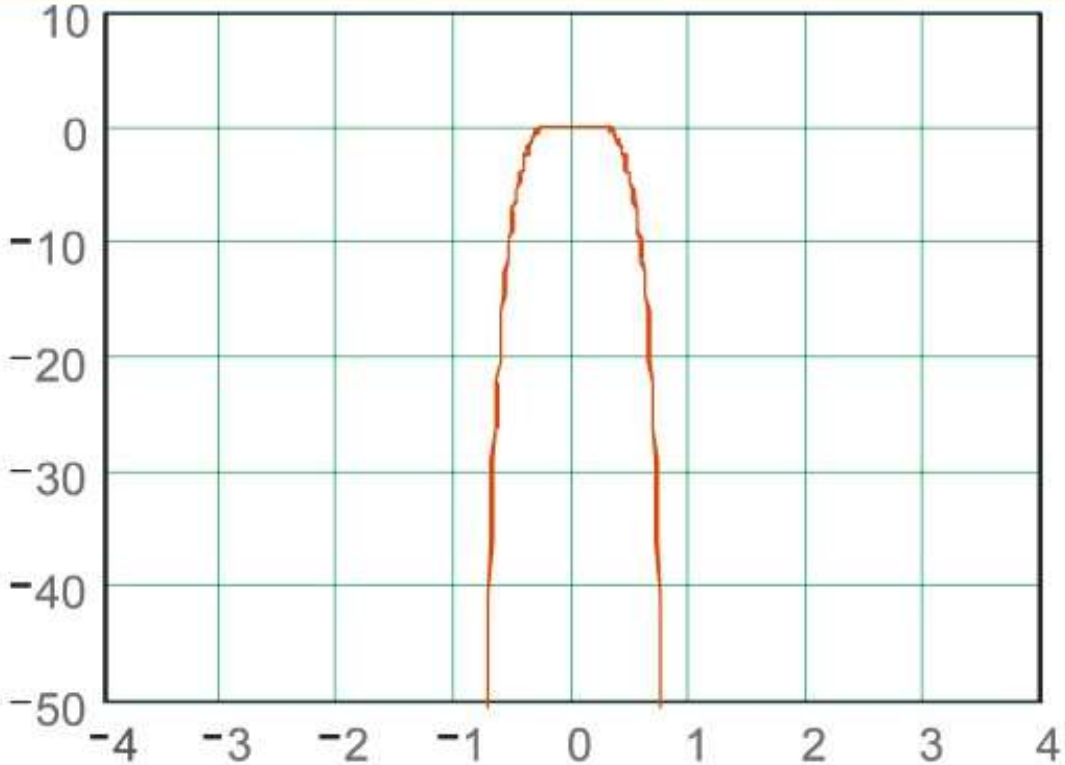


Signal space diagram



Binary amplitude modulation (BAM)
Raised-cosine pulses (roll-off 0.5)

Power spectral density for BAM

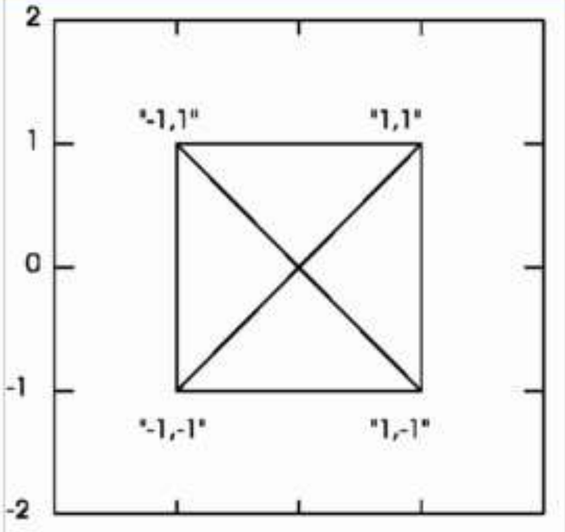


Normalized freq. $f \cdot T_b$

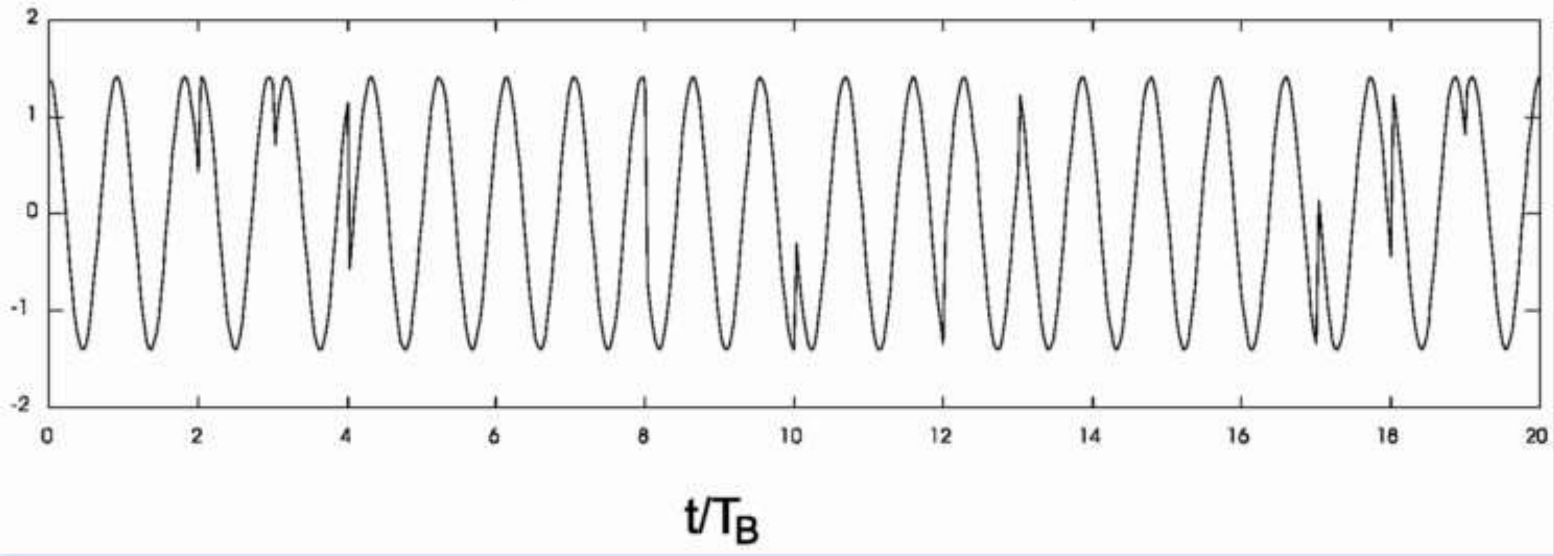
Contained percentage of total energy	spectral efficiency
90%	$1.02 Bit/s/Hz$
99%	$0.79 Bit/s/Hz$

Quaternary PSK (QPSK or 4-PSK) Rectangular pulses

Complex representation

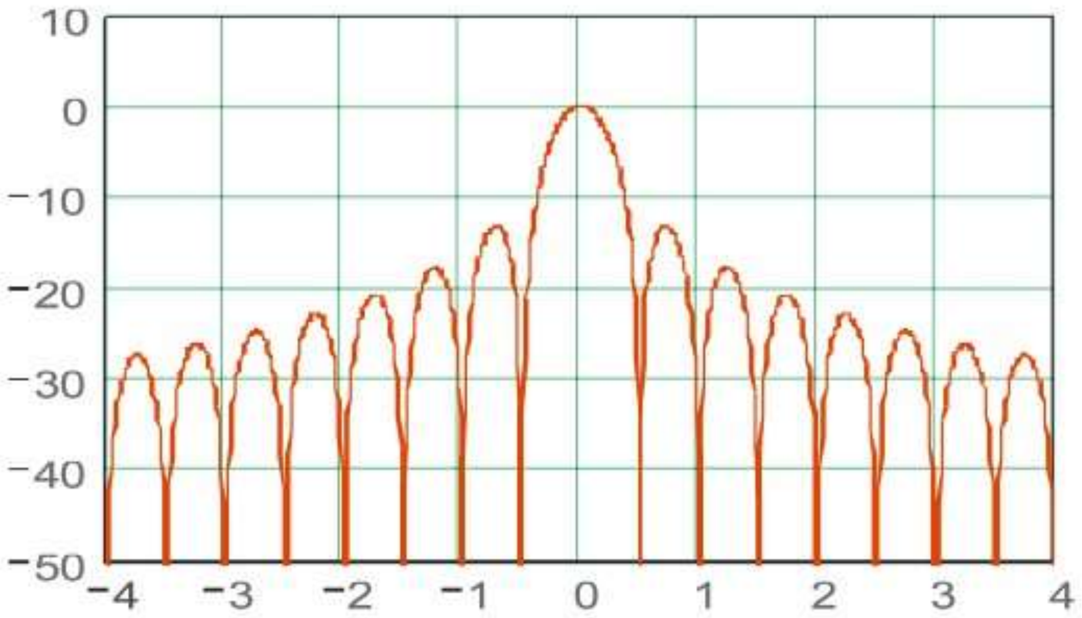


Radio signal



Quaternary PSK (QPSK or 4-PSK) Rectangular pulses

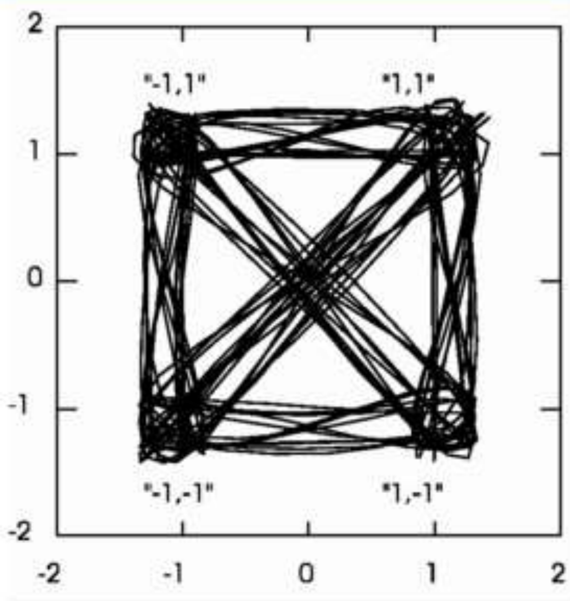
Power spectral density for QPSK



Contained percentage of total energy	spectral efficiency
90%	$1,18 Bit/s/Hz$
99%	$0.10 Bit/s/Hz$

Quadrature ampl.-modulation (QAM)
Root raised-cos pulses (roll-off 0.5)

Complex representation

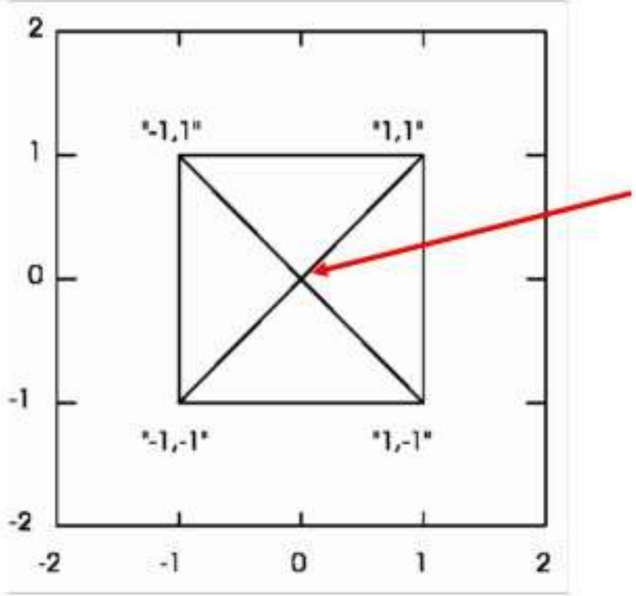


Contained percentage of total energy	spectral efficiency
90%	$2.04 Bit/s/Hz$
99%	$1.58 Bit/s/Hz$

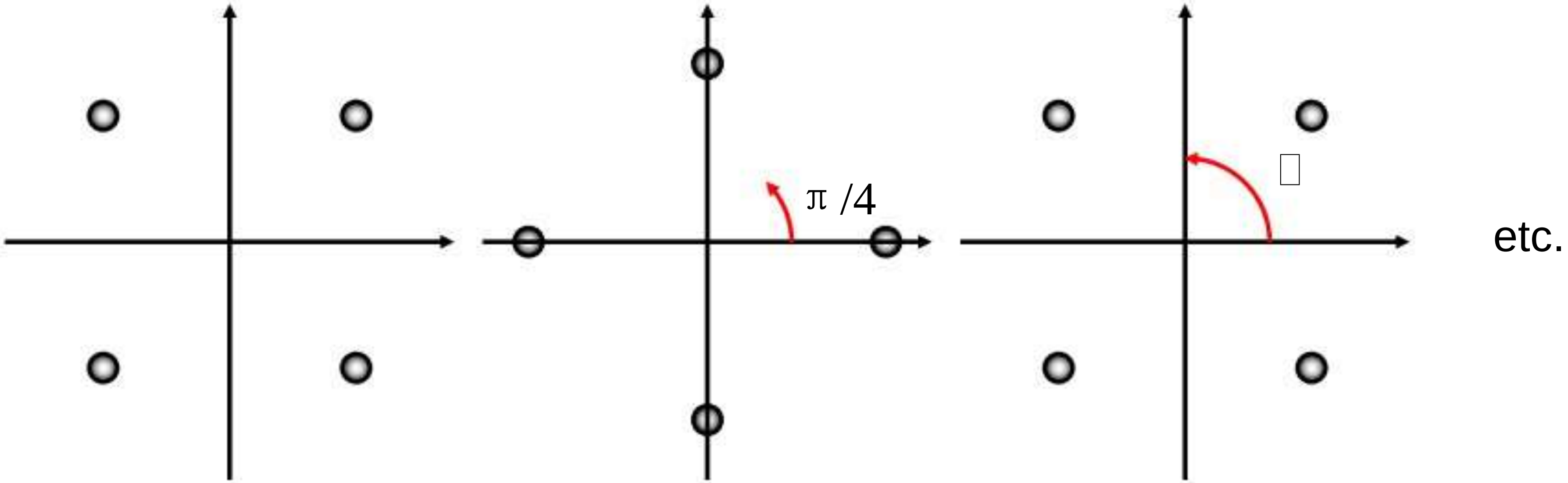
Amplitude variations
The problem

Signals with high amplitude variations leads to less efficient amplifiers.

Complex representation of QPSK



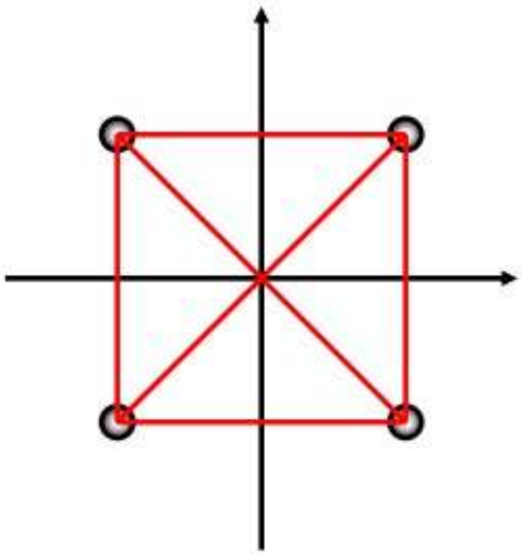
Amplitude variations
A solution



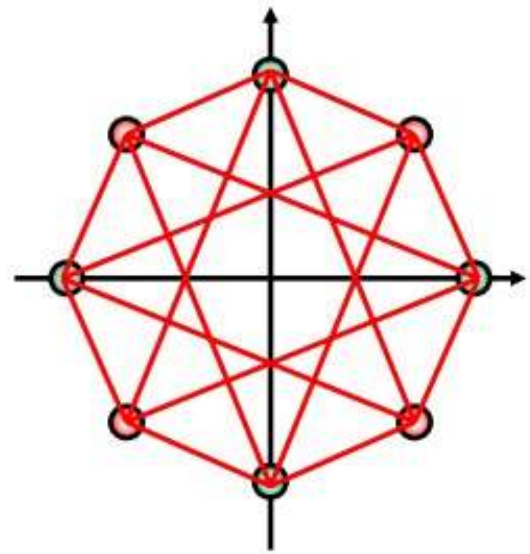
Amplitude variations
A solution

Looking at the complex representation ...

QPSK without rotation



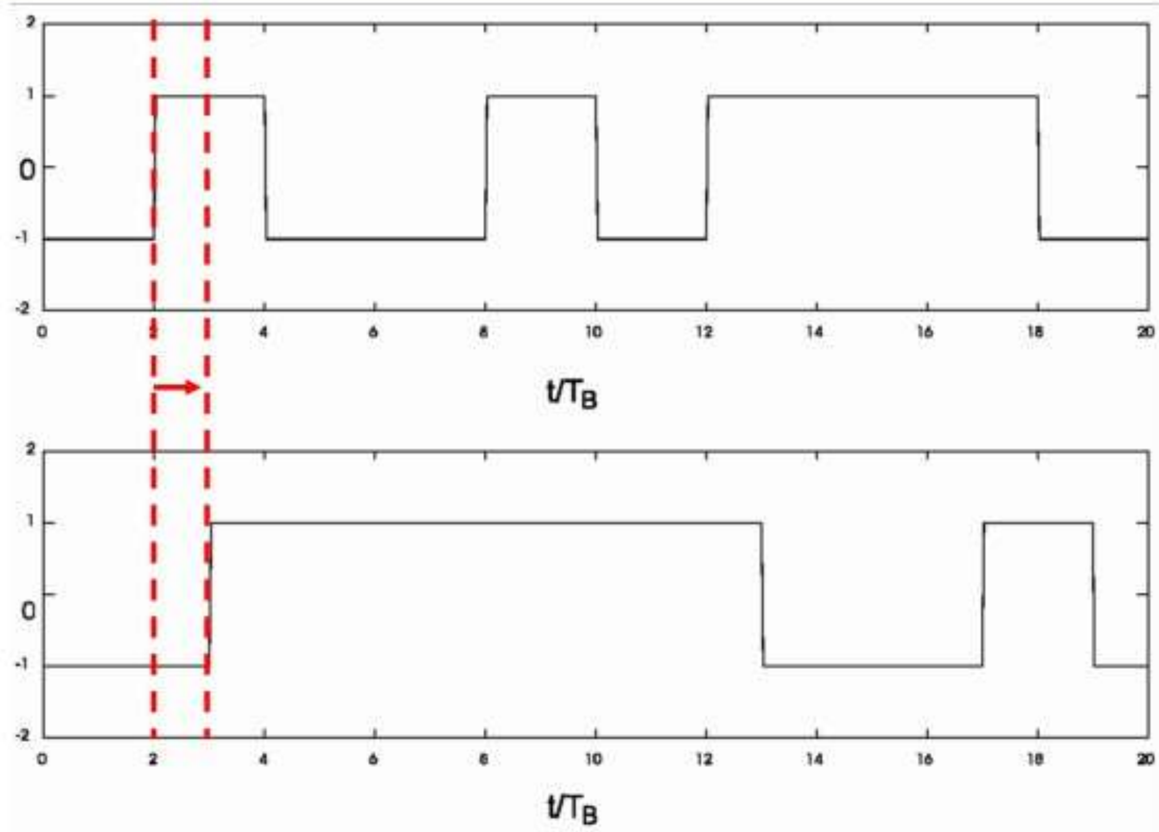
QPSK with rotation



Offset QPSK (OQPSK)
Rectangular pulses

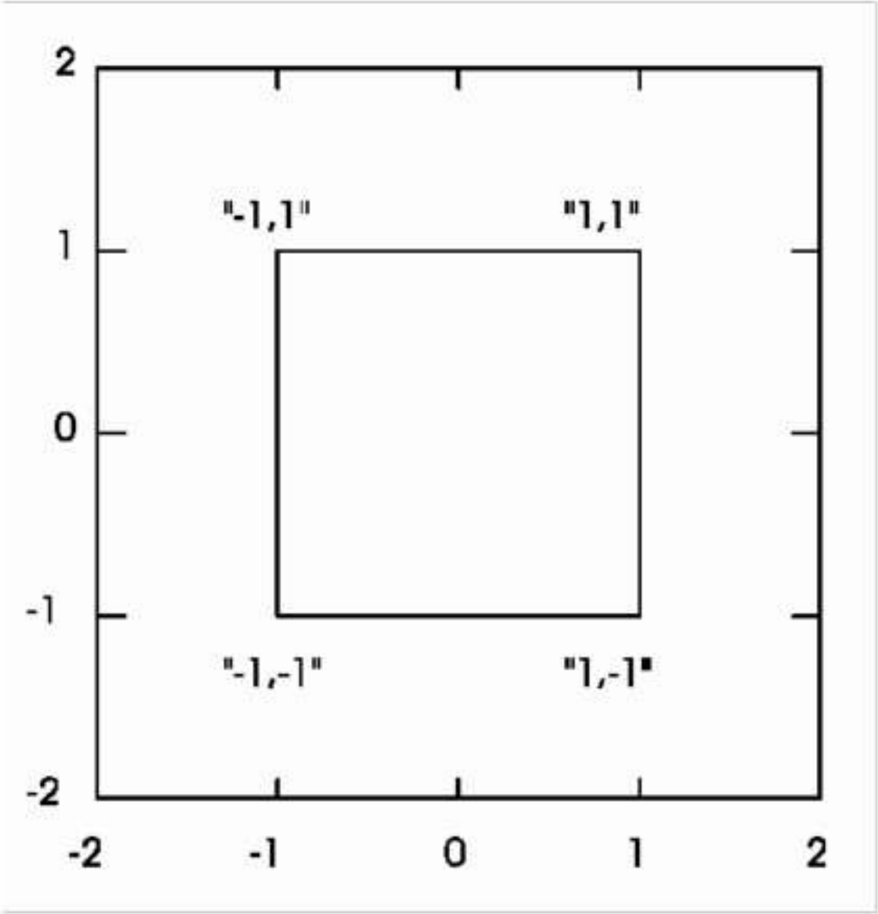
In-phase signal

Quadrature signal



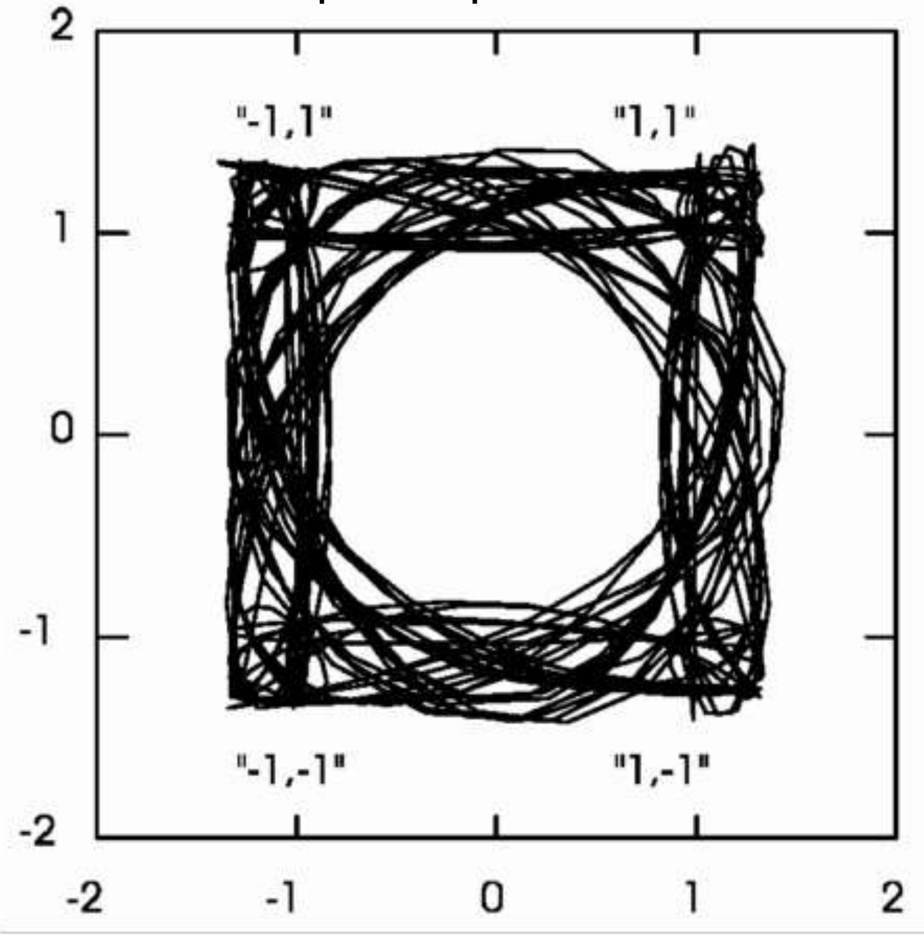
Offset QPSK
Rectangular pulses

Complex representation



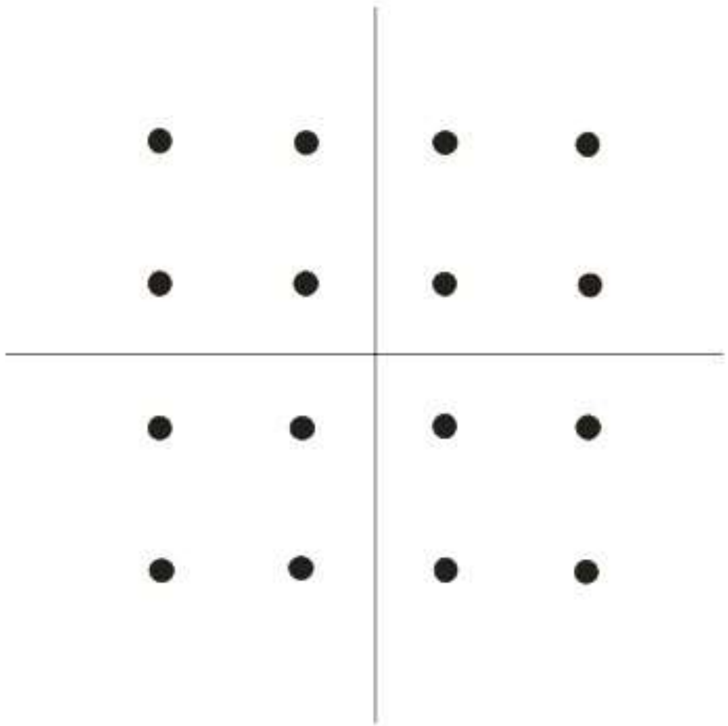
Offset QAM (OQAM)
Raised-cosine pulses

Complex representation



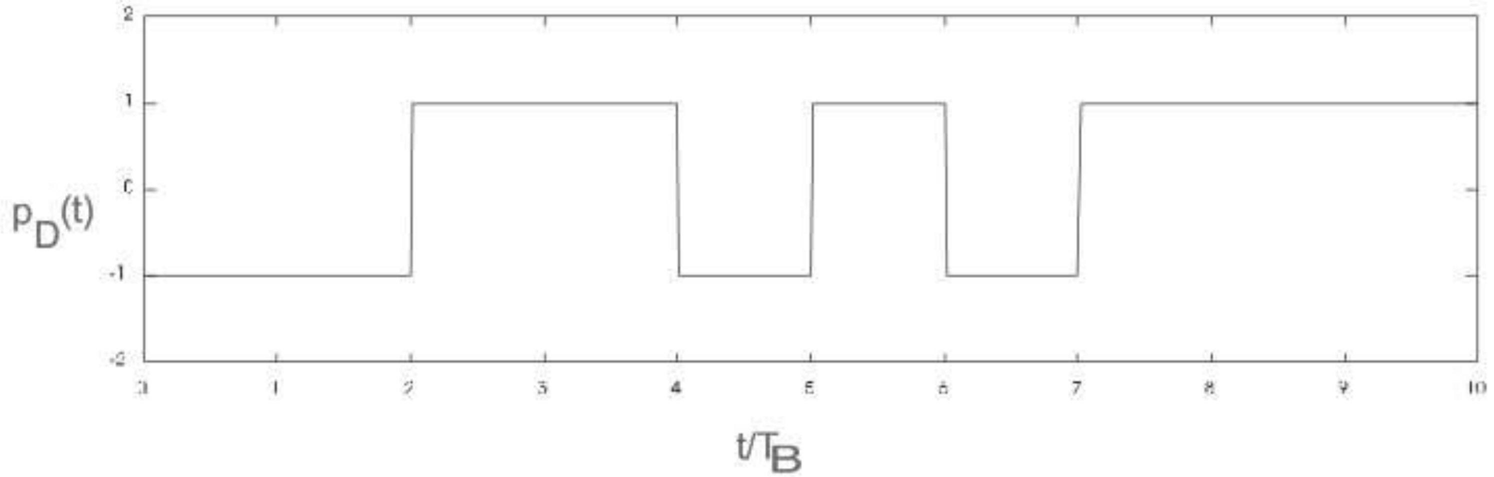
Higher-order modulation

16-QAM signal space diagram

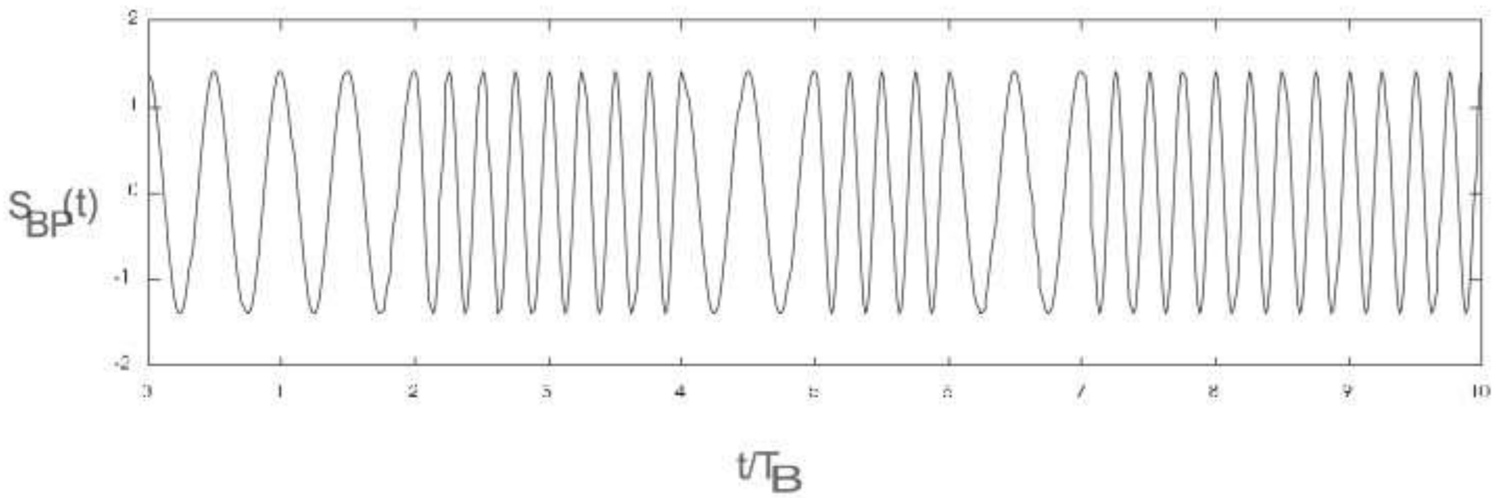


Binary frequency-shift keying (BFSK)
Rectangular pulses

Base-band

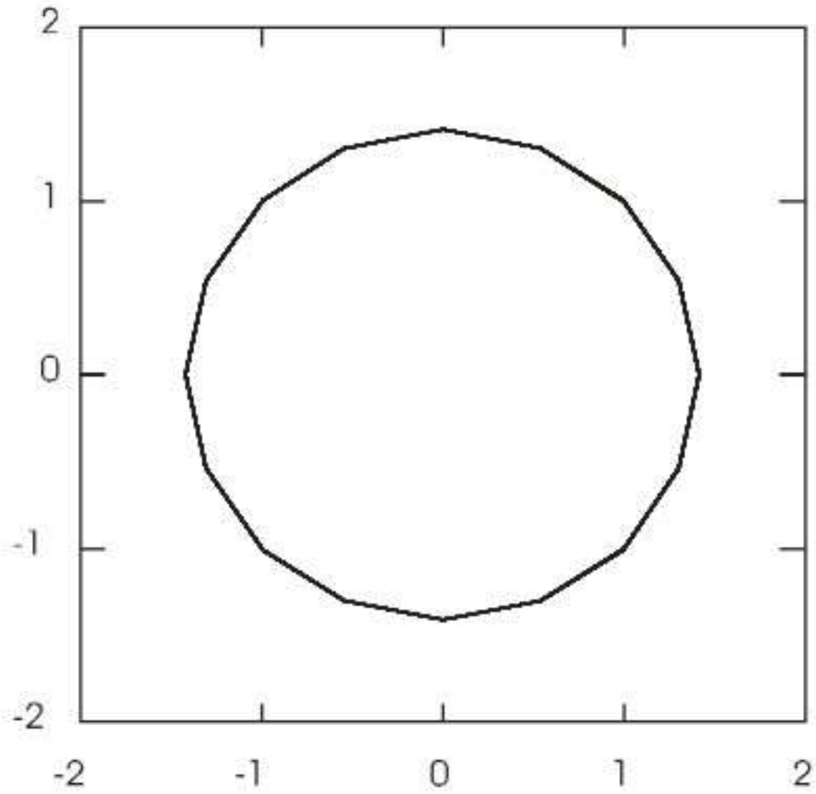


Radio signal

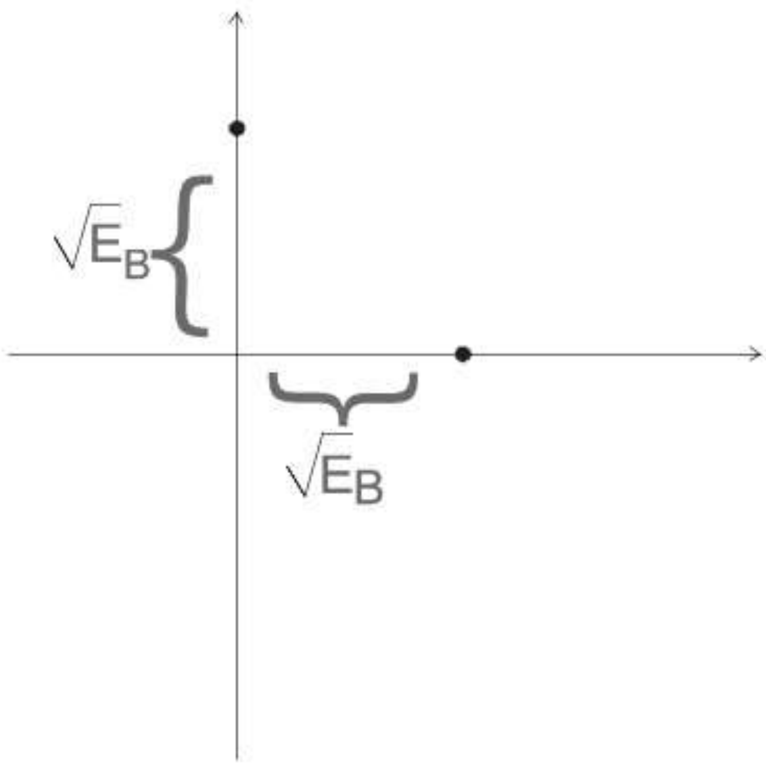


Binary frequency-shift keying (BFSK)
Rectangular pulses

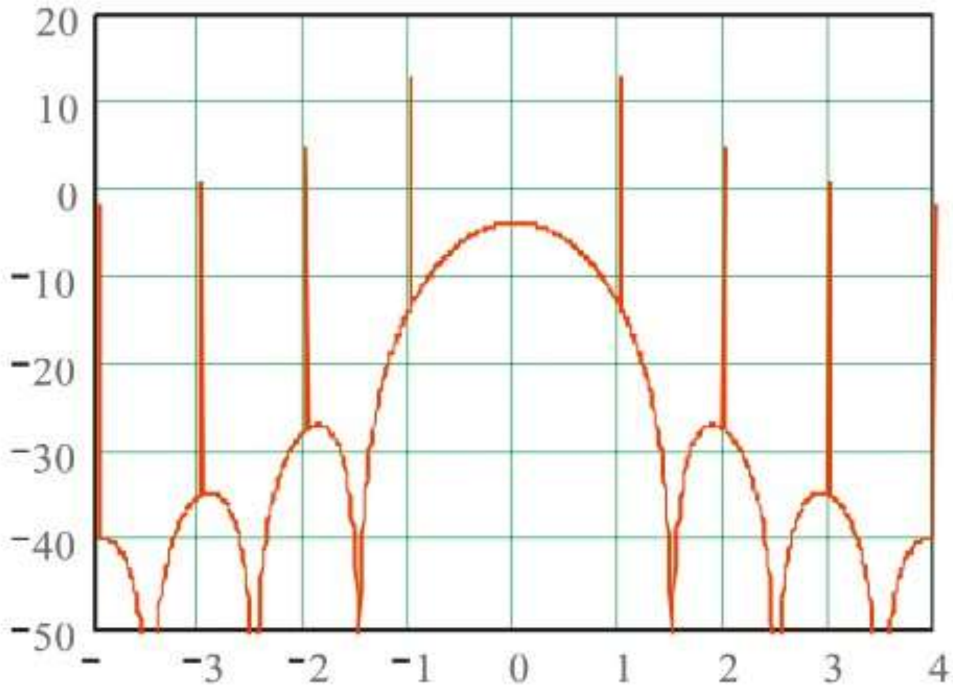
Complex representation



Signal space diagram



Binary frequency-shift keying (BFSK)
Rectangular pulses



Contained percentage of total energy	spectral efficiency
90%	$0.59 Bit/s/Hz$
99%	$0.05 Bit/s/Hz$

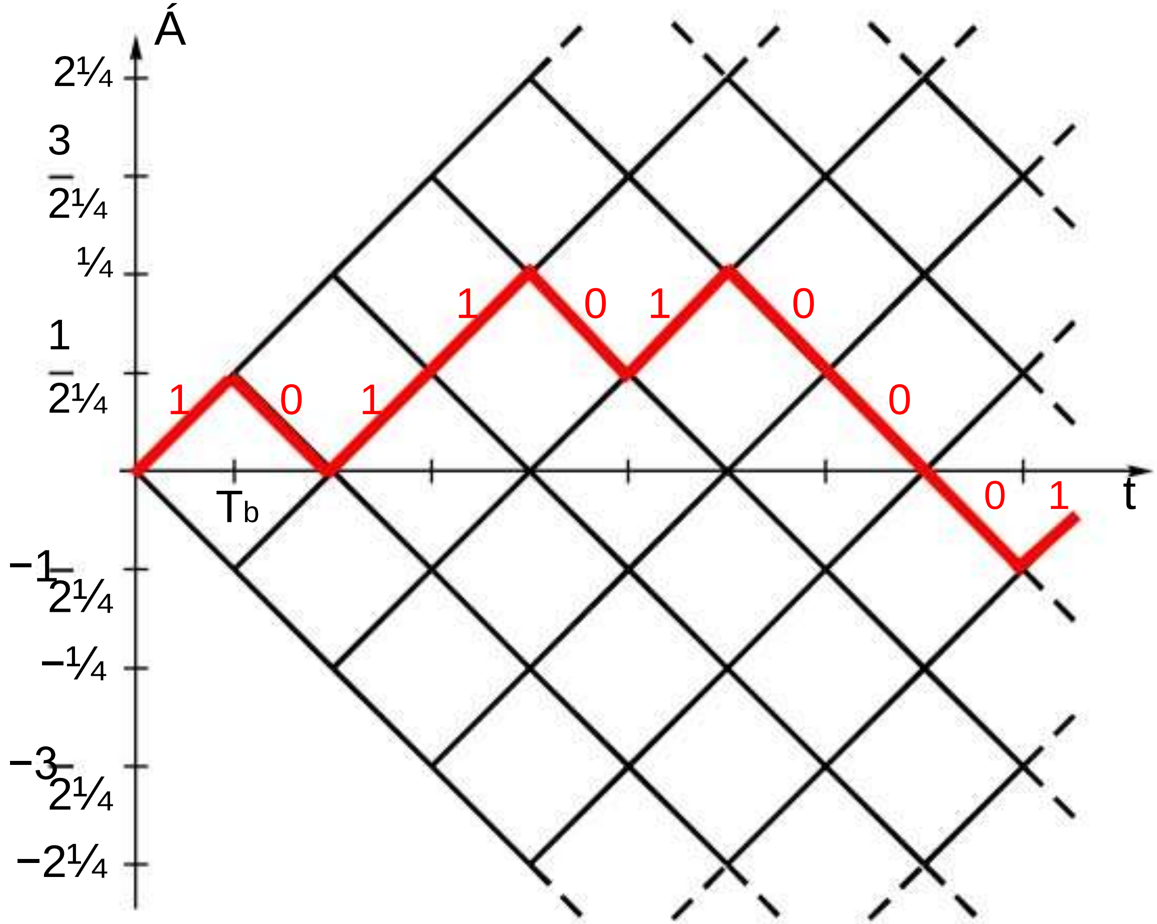
Continuous-phase modulation

Basic idea:

- Keep **amplitude constant**

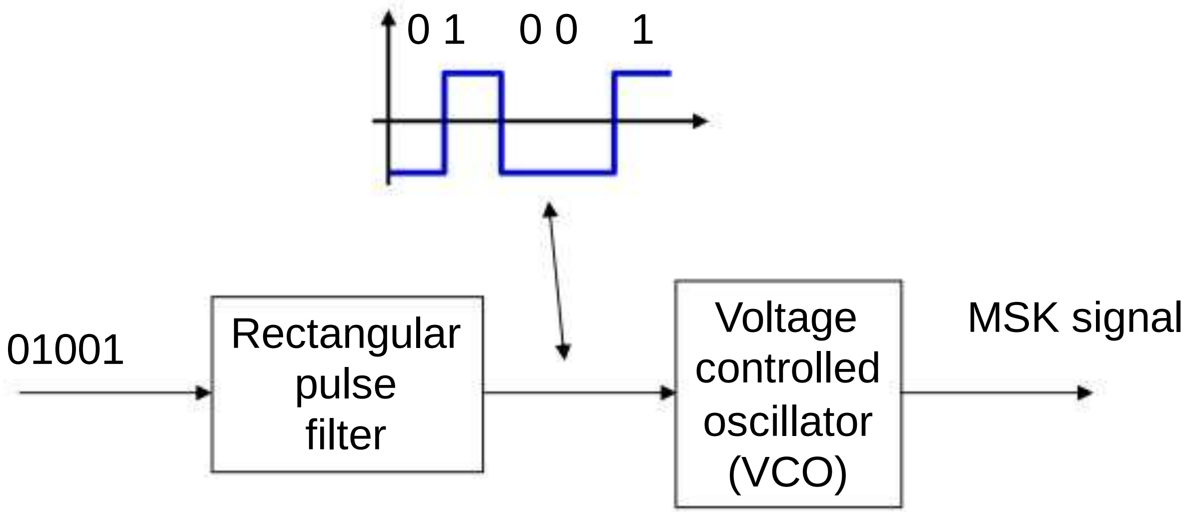
- Change phase continuously

MSK/FFSK



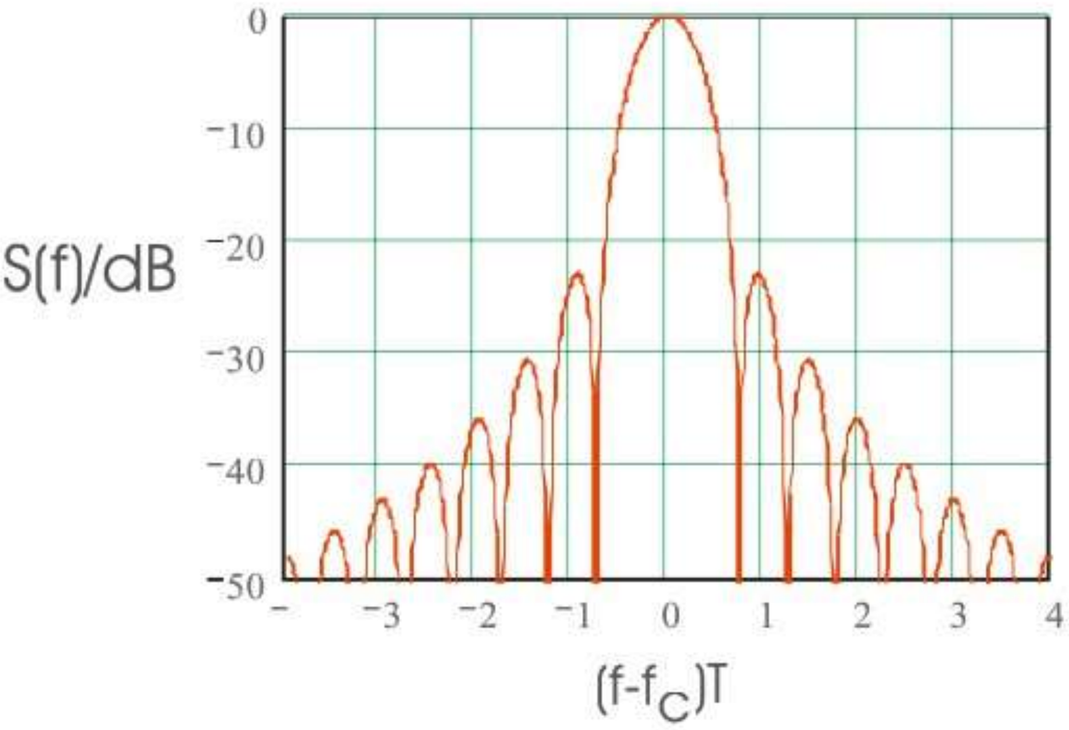
Minimum shift keying (MSK)

Simple MSK implementation



Minimum shift keying (MSK)

Power spectral density of MSK

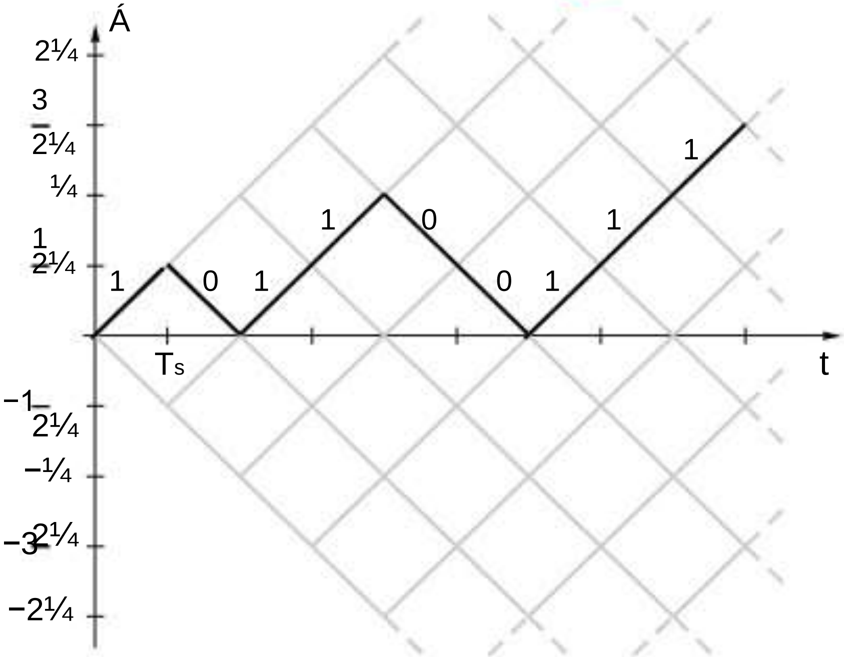


Contained percentage of total energy	spectral efficiency
90 %	1,29 Bit / s / Hz
99 %	0,85 Bit / s / Hz

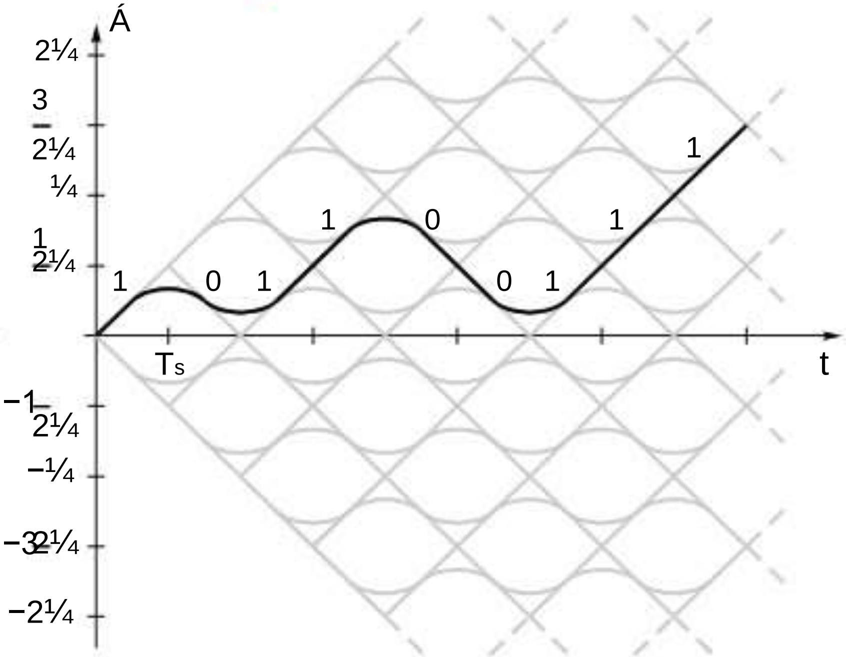
Gaussian filtered MSK (GMSK)

Further improvement of the phase: Remove ’corners’

(Simplified figure)

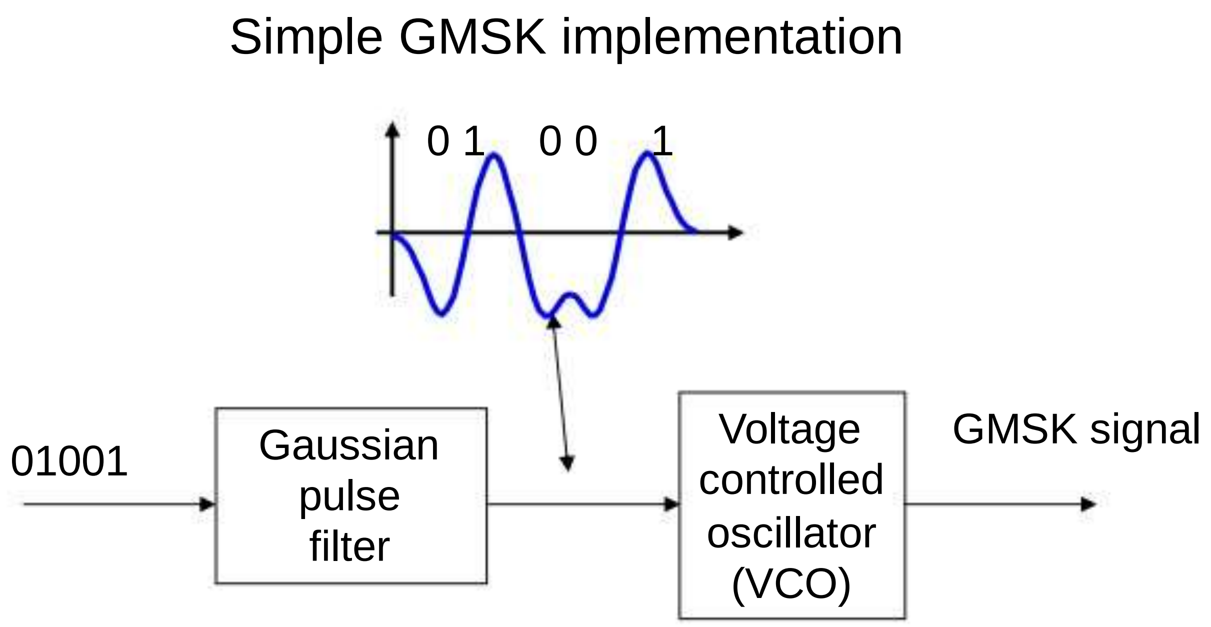


MSK
(Rectangular pulse filter)



Gaussian filtered MSK - GMSK
(Gaussian pulse filter)

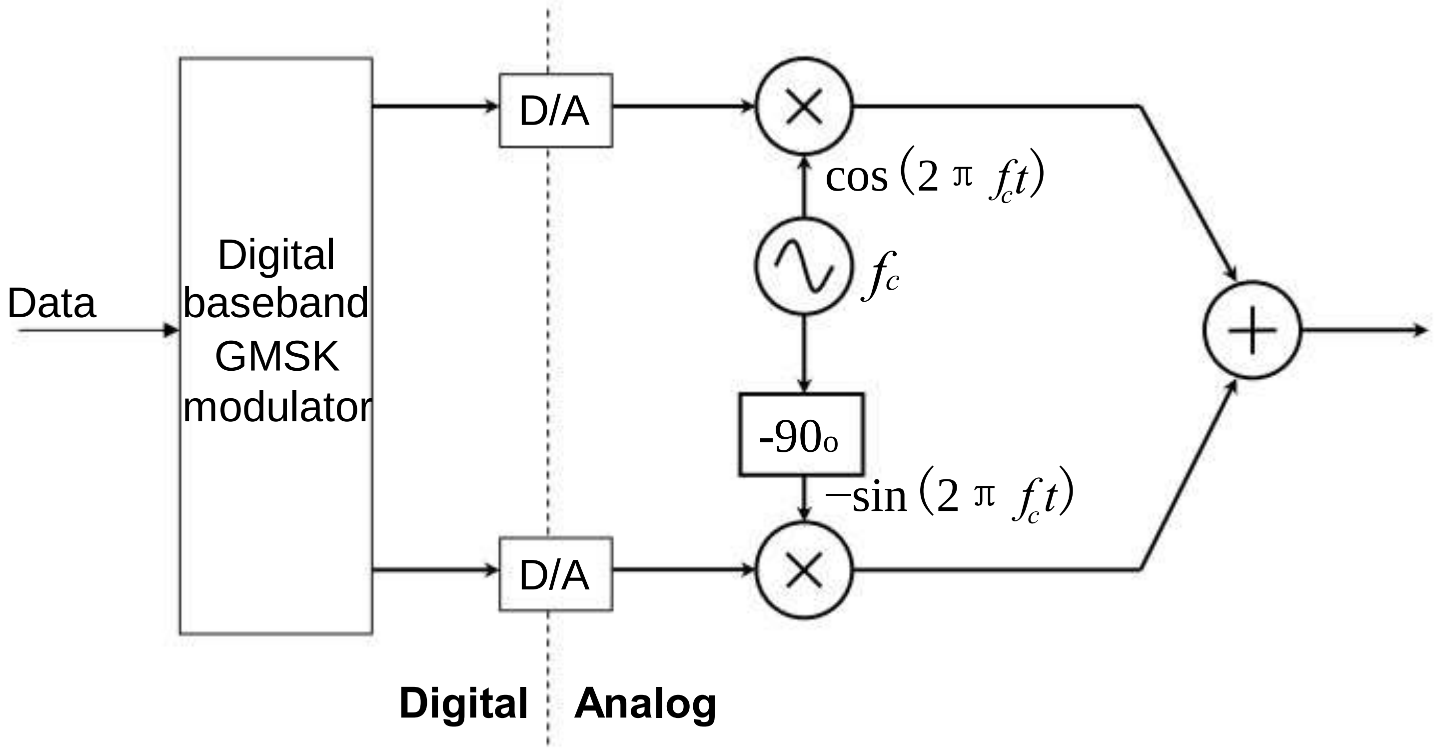
Gaussian filtered MSK (GMSK)



GSFK is used in e.g. Bluetooth.

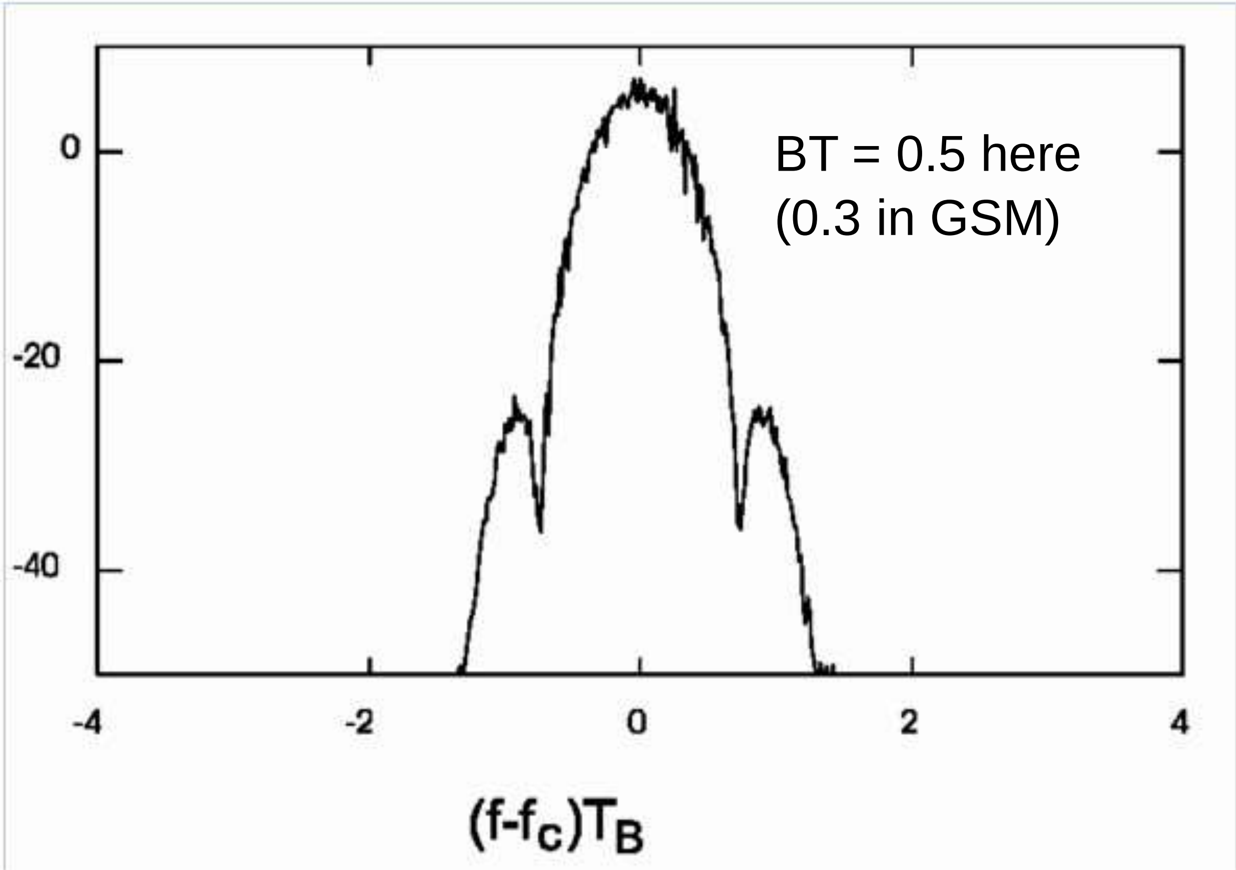
Gaussian filtered MSK (GMSK)

Digital GMSK implementation



Gaussian filtered MSK (GMSK)

Power spectral density of GMSK.



Contained percentage of total energy	spectral efficiency
90 %	1,45 Bit / s / Hz
99 %	0,97 Bit / s / Hz

How do we use all these spectral efficiencies?

Example: Assume that we want to use MSK to transmit 50 kbit/sec, and want to know the required transmission bandwidth.

Take a look at the spectral efficiency table:

Contained percentage of total energy	spectral efficiency
90 %	1,29 Bit / s / Hz
99 %	0,85 Bit / s / Hz

The 90% and 99% bandwidths become:

$$B_{90\%} = 50000 / 1.29 = 38.8 \text{ kHz}$$

$$B_{99\%} = 50000 / 0.85 = 58.8 \text{ kHz}$$

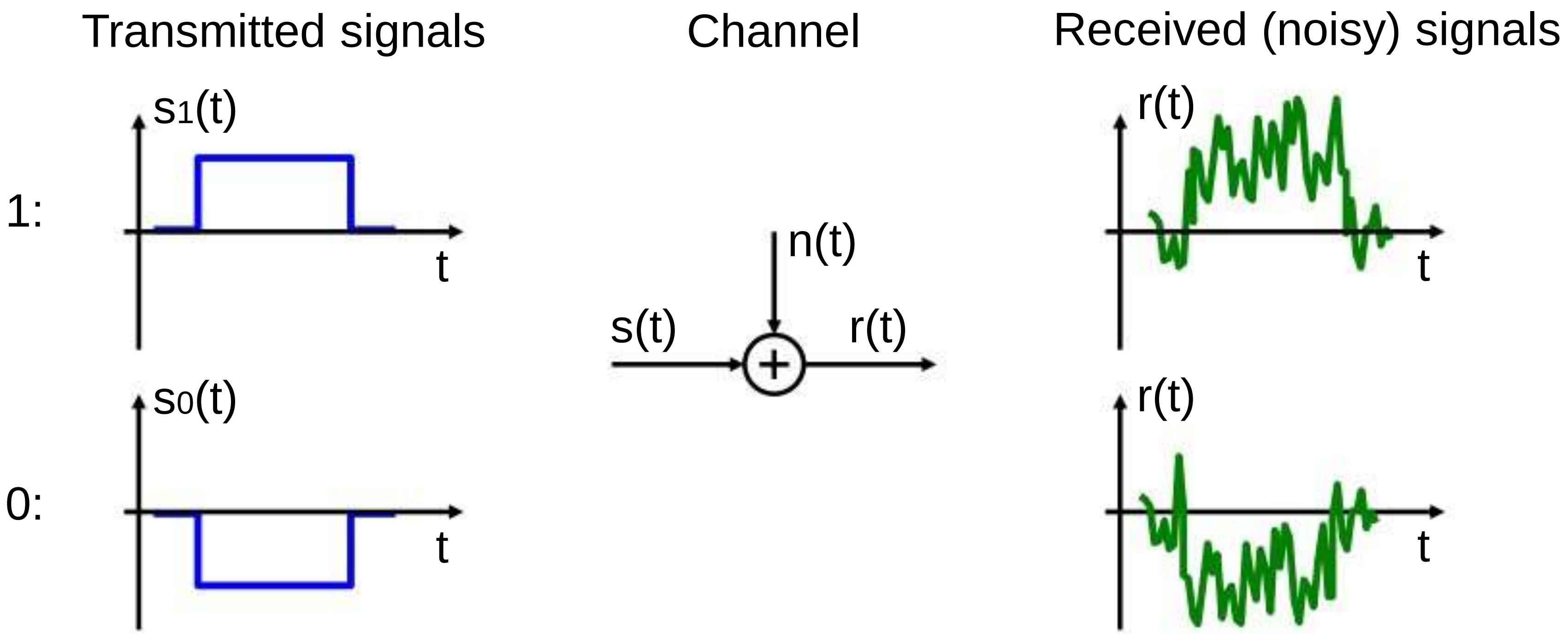
Summary

Modulation method	spectral efficiency for 90 % of total energy Bit / s / Hz	spectral efficiency for 99 % of total energy Bit / s / Hz	envelope variations w ratio of maximum and minimum amplitude
BPSK	0,59	0,05	1
BAM (α=0.5)	1,02	0,79	∞
QPSK, OQPSK, $\pi/4$-QPSK	1,18	0,10	1
MSK	1,29	0,85	1
GMSK (B_G T= 0.5)	1,45	0,97	1
QAM ($\alpha = 0.5$)	2,04	1,58	∞
OQAM ($\alpha = 0.5$)	2,04	1,58	2.6
FSK		$< 1/(2 f_D T_B)$	1

Demodulation and BER computation

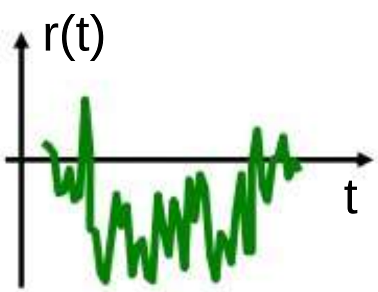
OPTIMAL RECEIVER
AND
BIT ERROR PROBABILITY
IN AWGN CHANNELS

Optimal receiver
Transmitted and received signal



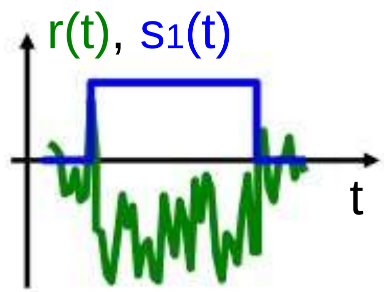
Optimal receiver
A first “intuitive” approach

Assume that the following signal is received:

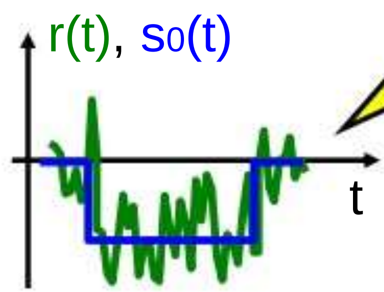


Comparing it to the two possible noise free received signals:

1:



0:



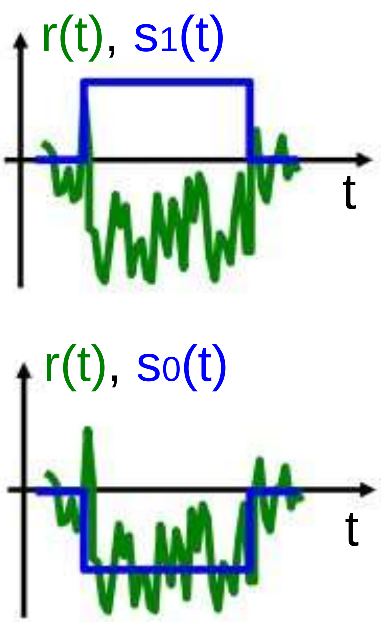
This seems to be the best “fit”. We assume that “0” was the transmitted bit.

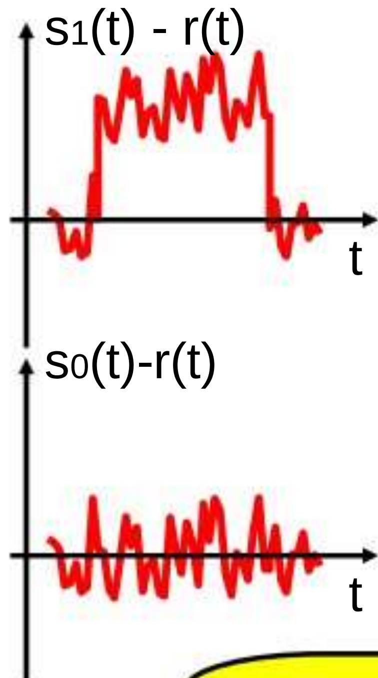
Optimal receiver
Let’s make it more measurable

To be able to better measure the “fit” we look at the energy of the residual (difference) between received and the possible noise free signals:

1:

0:





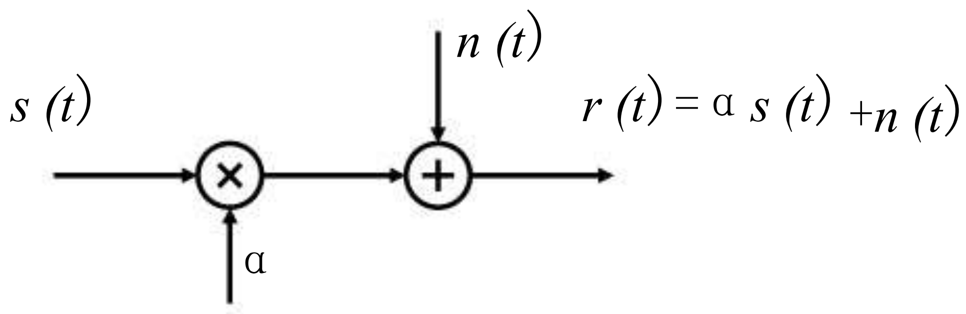
$$\int \left| s_1(t) - r(t) \right|^2 dt$$

$$\int \left| s_0(t) - r(t) \right|^2 dt$$

This residual energy is much smaller. We assume that “0” was transmitted.

Optimal receiver
The AWGN channel

The additive white Gaussian noise (AWGN) channel



$s(t)$ - transmitted signal

α - channel attenuation

$n(t)$ - white Gaussian noise

$r(t)$ - received signal

In our digital transmission system, the transmitted signal s(t) would be one of, let's say M, different alternatives $s_0(t)$, $s_1(t)$, ... , $s_{M-1}(t)$.

Optimal receiver
The AWGN channel, cont.

For a received $r(t)$, the residual energy e_i for each possible transmitted alternative $s_i(t)$ is calculated as

$$e_i = \int \left| r(t) - \alpha s_i(t) \right|^2 dt = \int \left(r(t) - \alpha s_i(t) \right) \left(r(t) - \alpha s_i(t) \right)^* dt$$

$$= \underline{\int \left| r(t) \right|^2 dt} - \underbrace{2\,\mathrm{Re}\left\{ \alpha^* \int r(t) s_i^*(t)\, dt \right\}} + \underline{|\alpha|^2 \int \left| s_i(t) \right|^2 dt}$$

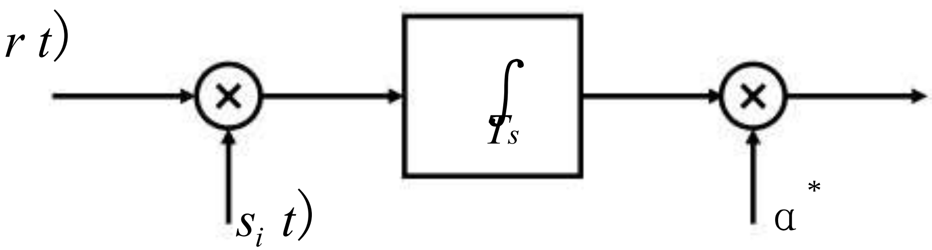
Same for all *i*

The residual energy is minimized by **maximizing** this part of the expression.

Same for all *i*, if the transmitted signals are of equal energy.

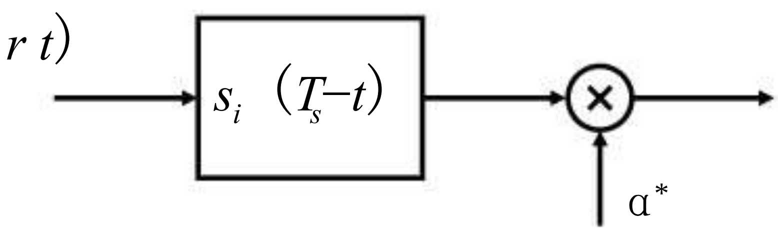
Optimal receiver
The AWGN channel, cont.

The central part of the comparison of different signal alternatives is a correlation, that can be implemented as a correlator:



The real part of the output from either of these is sampled at t = T_s

or a matched filter



where T_s is the symbol time (duration).

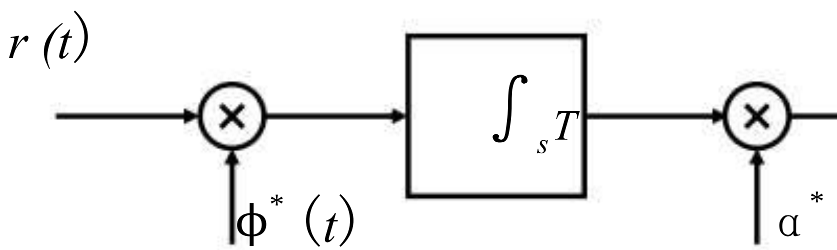
Optimal receiver
Antipodal signals

In antipodal signaling, the alternatives (for “0” and “1”) are

$$s_0(t) = \phi(t)$$

$$s_1(t) = -\phi(t)$$

This means that we only need ONE correlation in the receiver for simplicity:

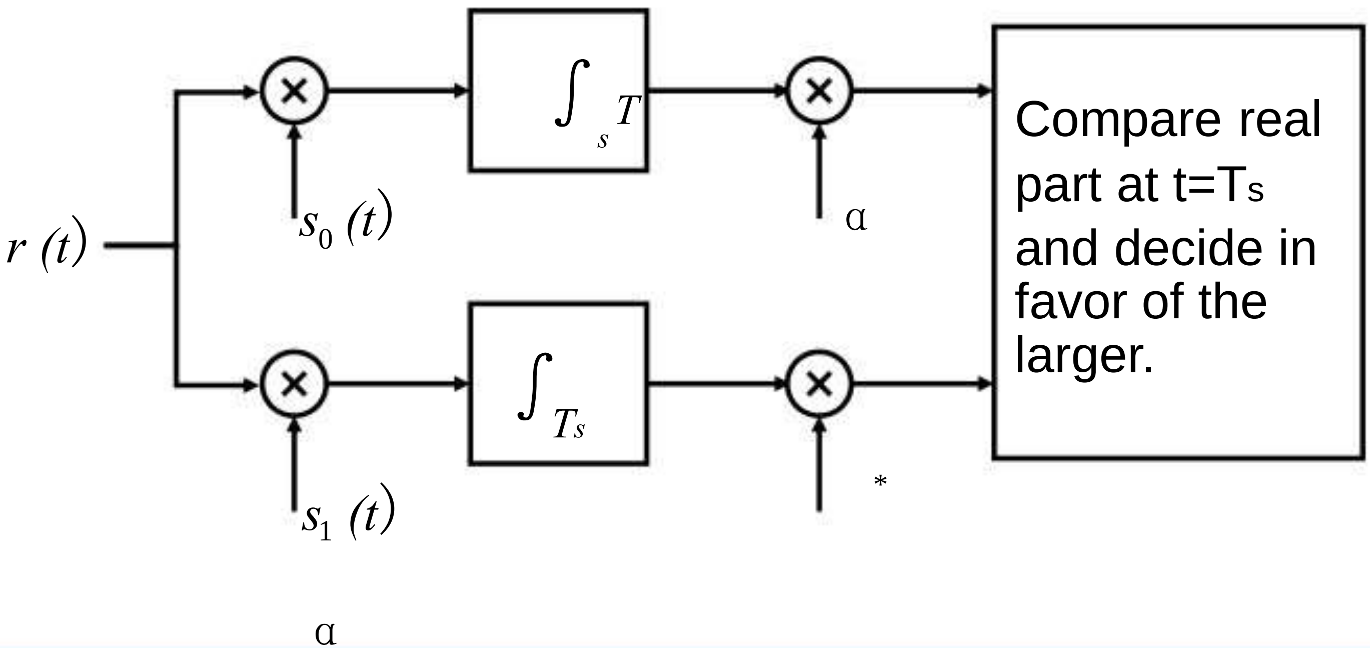


If the real part at T=Ts is
>0 decide “0”
<0 decide “1”

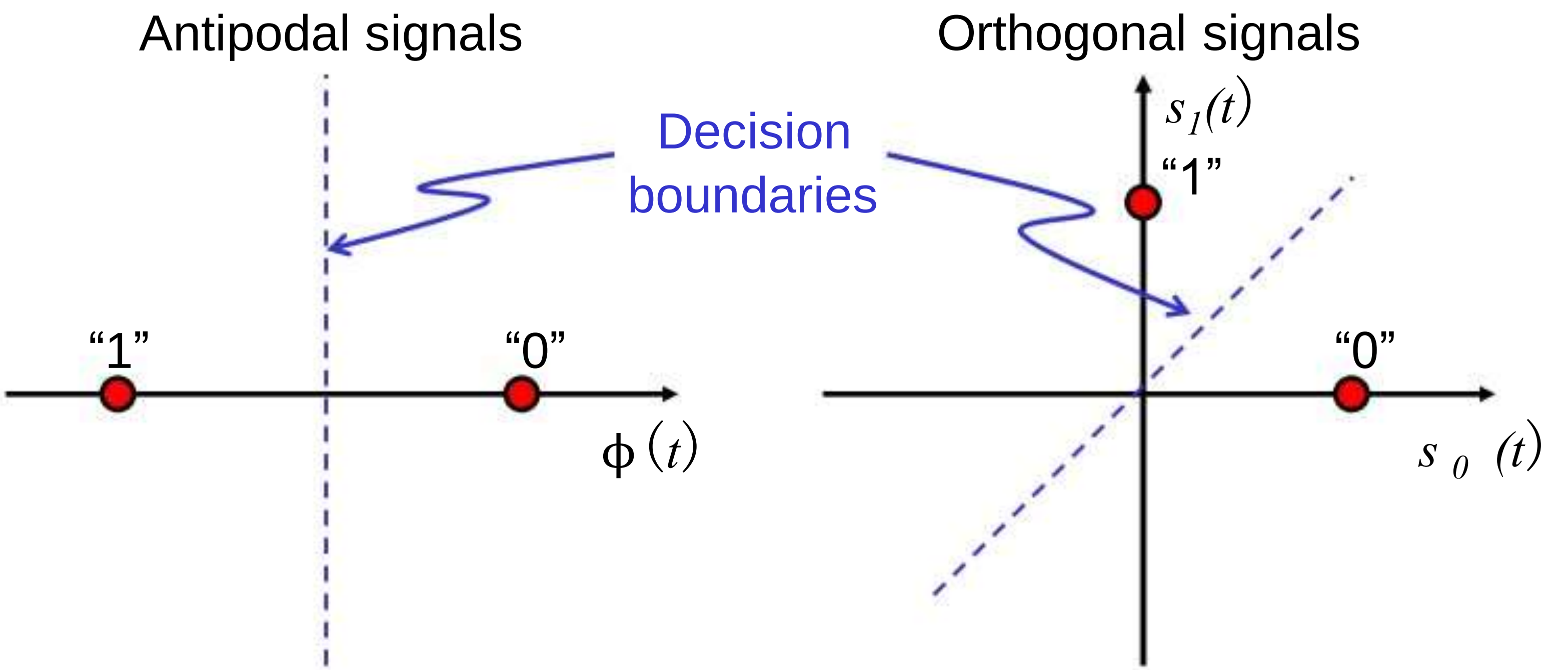
Optimal receiver
Orthogonal signals

In binary orthogonal signaling, with equal energy alternatives $s_0(t)$ and $s_1(t)$ (for “0” and “1”) we require the property:

$$\langle s_0(t), s_1(t) \rangle = \int_0 s_0(t)\, s_1^*(t)\, dt = 0$$

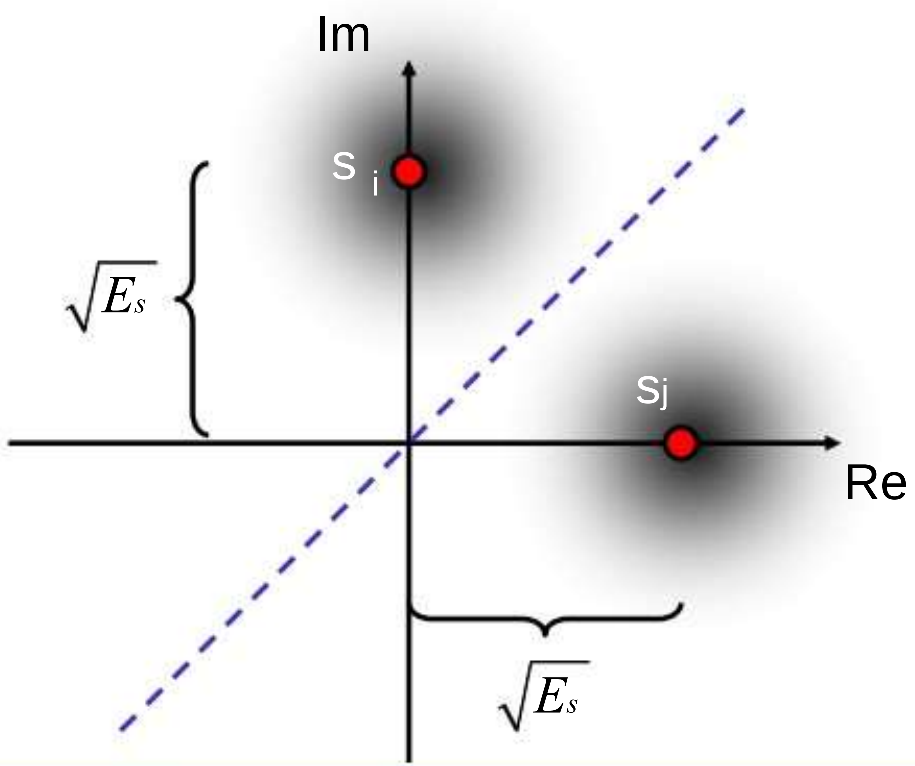


Optimal receiver
Interpretation in signal space



Optimal receiver
The noise contribution

Assume a 2-dimensional signal space, here viewed as the complex plane



This normalization of axes implies that the noise centered around each alternative is complex Gaussian

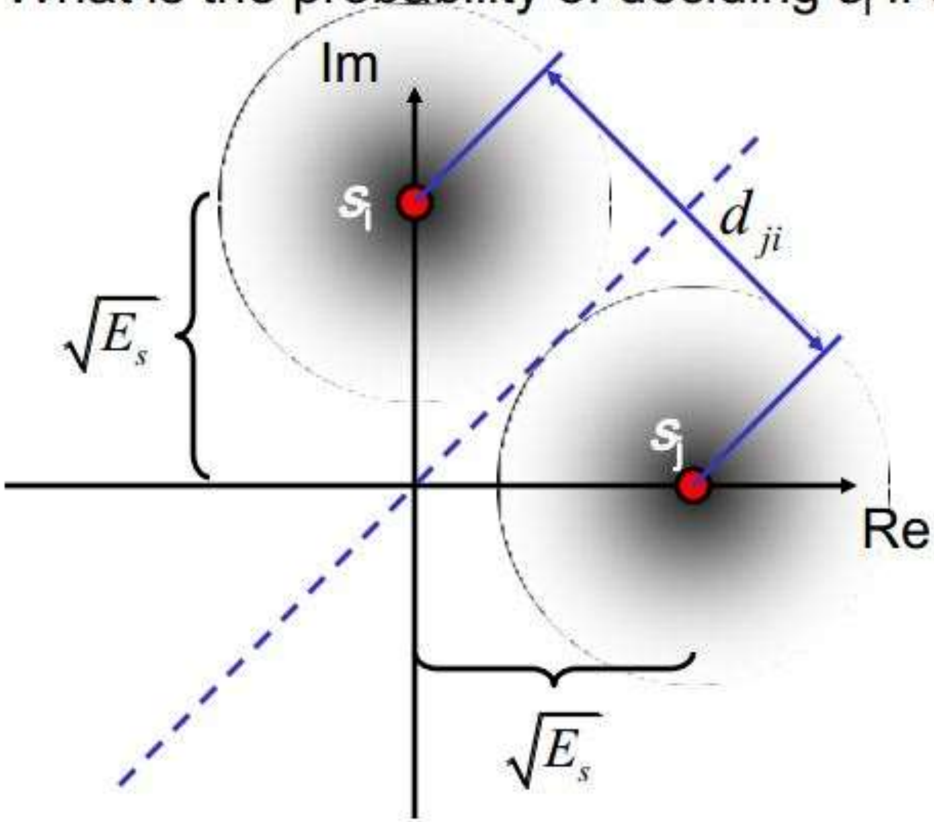
$$N\left(0,\sigma^2\right)+jN\left(0,\sigma^2\right)$$

with variance $\sigma^2 = N_0/2$ in each direction.

Fundamental question: What is the probability that we end up on the wrong side of the decision boundary?

Optimal receiver
Pair-wise symbol error probability

What is the probability of deciding s_i if s_j was transmitted?



We need the distance between the two symbols. In this orthogonal case:

$$d_{ji} = \sqrt{\sqrt{E_s}^2 + \sqrt{E_s}^2} = \sqrt{2E_s}$$

The probability of the noise pushing us across the boundary at distance $d_{ji}/2$ is

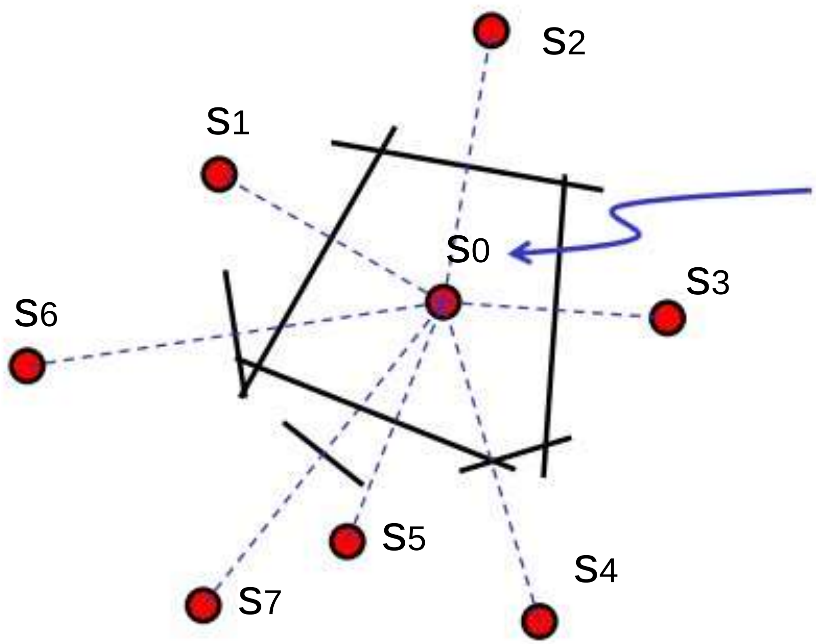
$$P(s_j \to s_i) = Q\left(\frac{d_{ji}/2}{\sqrt{N_0/2}}\right) = Q\left(\sqrt{\frac{E_s}{N_0}}\right)$$

$$= \frac{1}{2}\operatorname{erfc}\left(\sqrt{\frac{E_s}{2N_0}}\right)$$

Optimal receiver
The union bound

Calculation of symbol error probability is simple for two signals!

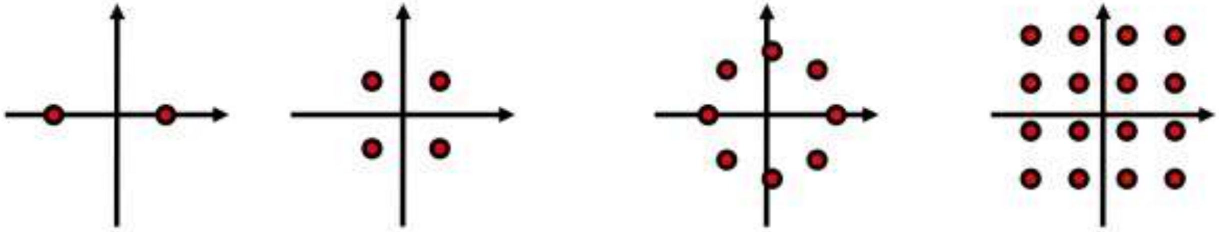
When we have many signal alternatives, it may be impossible to calculate an exact symbol error rate.



When s_0 is the transmitted signal, an error occurs when the received signal is outside this polygon.

Optimal receiver
Bit-error rates (BER)

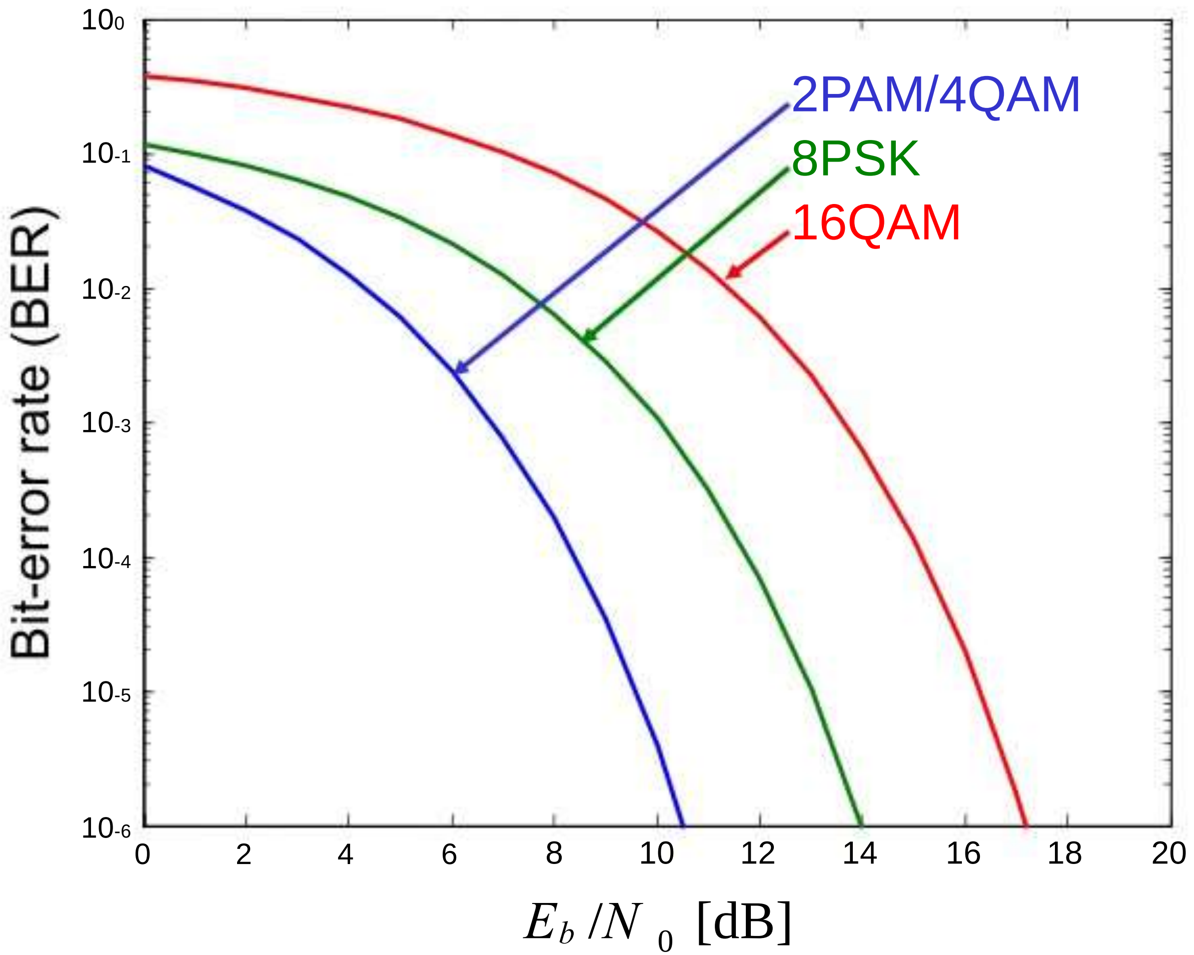
EXAMPLES:	2PAM	4QAM	8PSK	16QAM



Bits/symbol	1	2	3	4
Symbol energy	E_b	$2E_b$	$3E_b$	$4E_b$
BER	$Q\sqrt{\frac{2E_b}{N_0}}$	$Q\sqrt{\frac{2E_b}{N_0}}$	$\approx \frac{2}{3}Q\sqrt{0.87\frac{E_b}{N_0}}$	$\approx \frac{3}{2}Q\sqrt{\frac{E_{b,\max}}{2.25N_0}}$

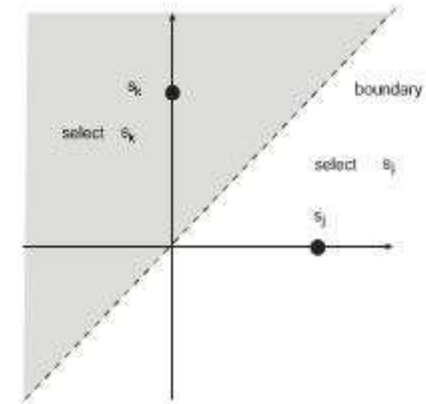
Gray coding is used when calculating these BER.

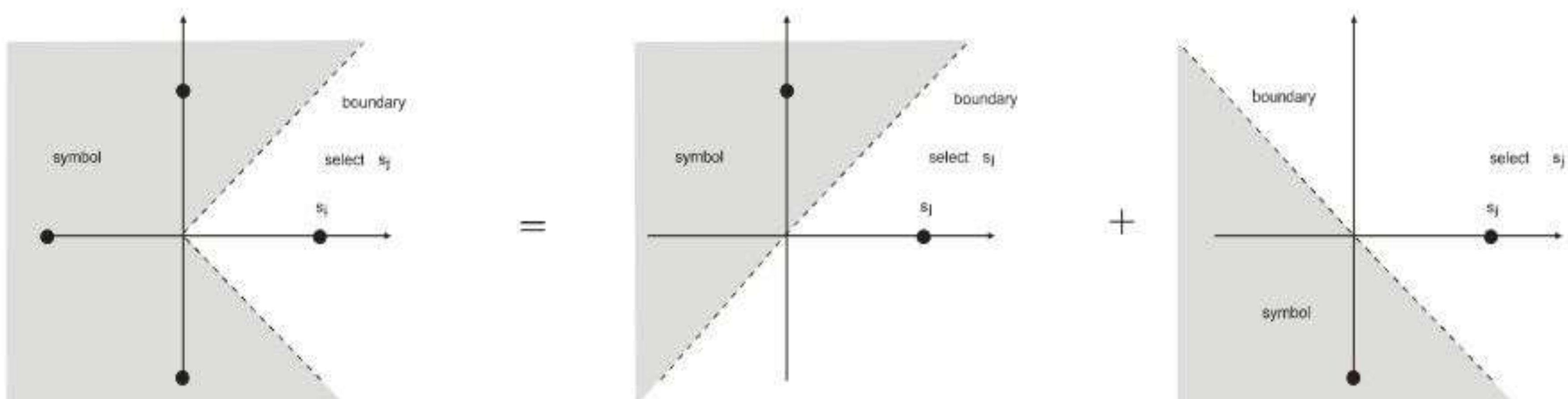
Optimal receiver
Bit-error rates (BER), cont.



Optimal receiver – BER of QPSK

- Compute via union bound
- Pairwise error probability $Q\left(\sqrt{2\gamma_{\mathrm{B}}}\right)$
- Symbol error probability $SER \approx 2Q(\sqrt{2\gamma_{\mathrm{B}}})$
- Bit error probability $BER = Q(\sqrt{2\gamma_{\mathrm{B}}})$





Optimal receiver
Where do we get E_b and N_0?

Where do those magic numbers E_b and N_0 come from?

The noise power spectral density N_0 is calculated according to

$$N_0 = kT_0F_0 \Leftrightarrow N_{0|dB} = -204 + F_{0|dB}$$

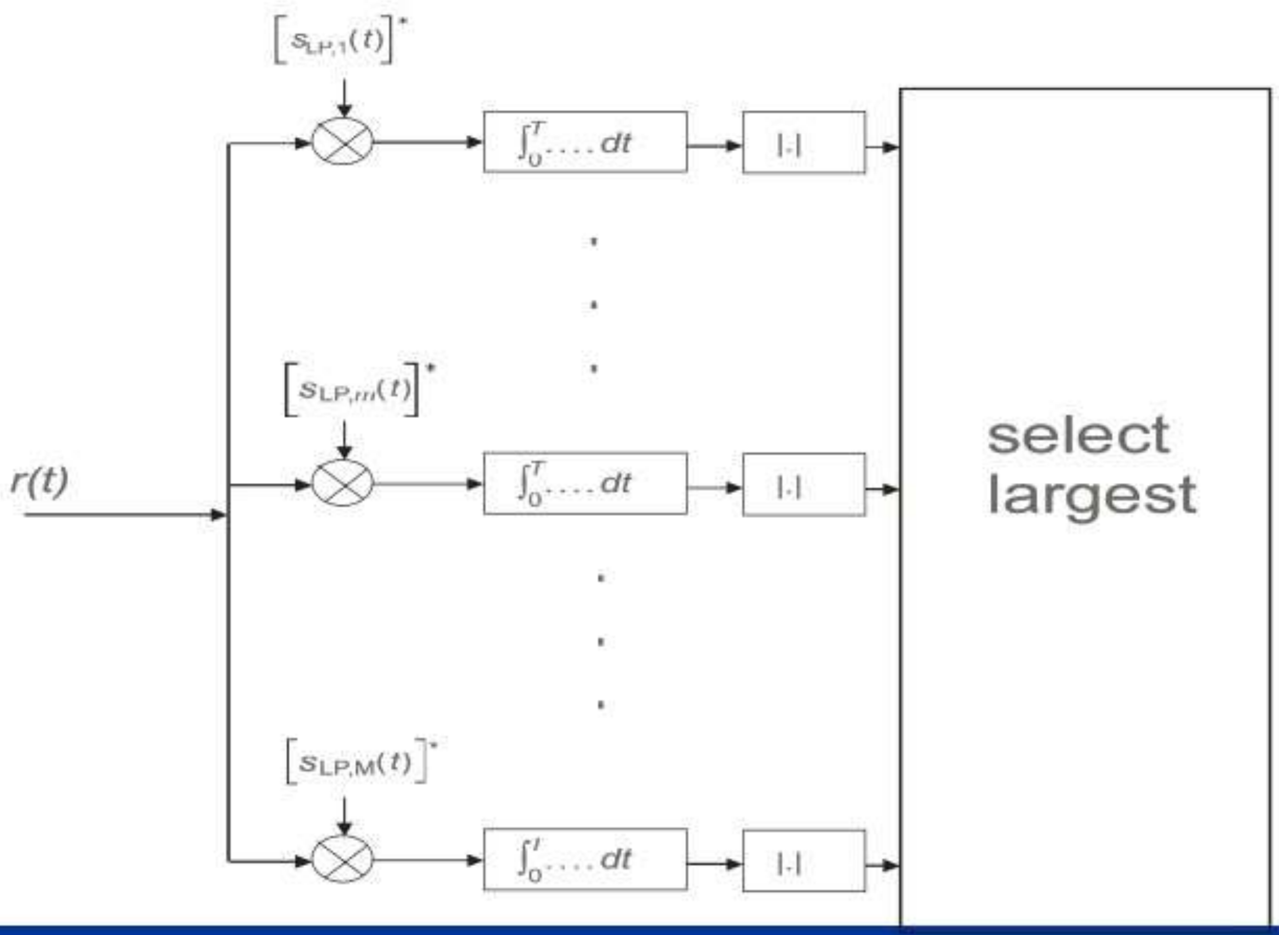
where F_0 is the noise factor of the "equivalent" receiver noise source.

The bit energy E_b can be calculated from the received power C (at the same reference point as N_0). Given a certain data-rate d_b [bits per second], we have the relation

$$E_b = C / d_b \Leftrightarrow E_{b|dB} = C_{|dB} - d_{b|dB}$$

THESE ARE THE EQUATIONS THAT RELATE DETECTOR PERFORMANCE ANALYSIS TO LINK BUDGET CALCULATIONS!

Noncoherent detection (1)



BER for differential receiver

- Differential BPSK

$$\Phi_i = \Phi_{i-1} + \begin{cases} +\frac{\pi}{2} & b_i = +1 \\ -\frac{\pi}{2} & b_i = -1 \end{cases}$$

- BER for differentially detected BPSK:

$$BER = \frac{1}{2}\exp(-\gamma_b)\ .$$

Noncoherent detection (2)

- Error probability for noncoherent detection

$$BER = Q_{\mathrm{M}}(a,b) - \tfrac{1}{2} I_0(ab)\exp\left(-\tfrac{1}{2}(a^2 + b^2)\right)$$

$$a = \sqrt{\frac{\gamma_{\mathrm{B}}}{2}\left(1 - \sqrt{1 - |\rho|^2}\right)} \qquad b = \sqrt{\frac{\gamma_{\mathrm{B}}}{2}\left(1 + \sqrt{1 - |\rho|^2}\right)} .$$

- For phase modulation, |ρ|=1, therefore SNR=0

BER IN FADING CHANNELS AND DISPERSION-INDUCED ERRORS

BER in fading channels (1)

We have (or can calculate) BER expressions for non-fading AWGN channels.

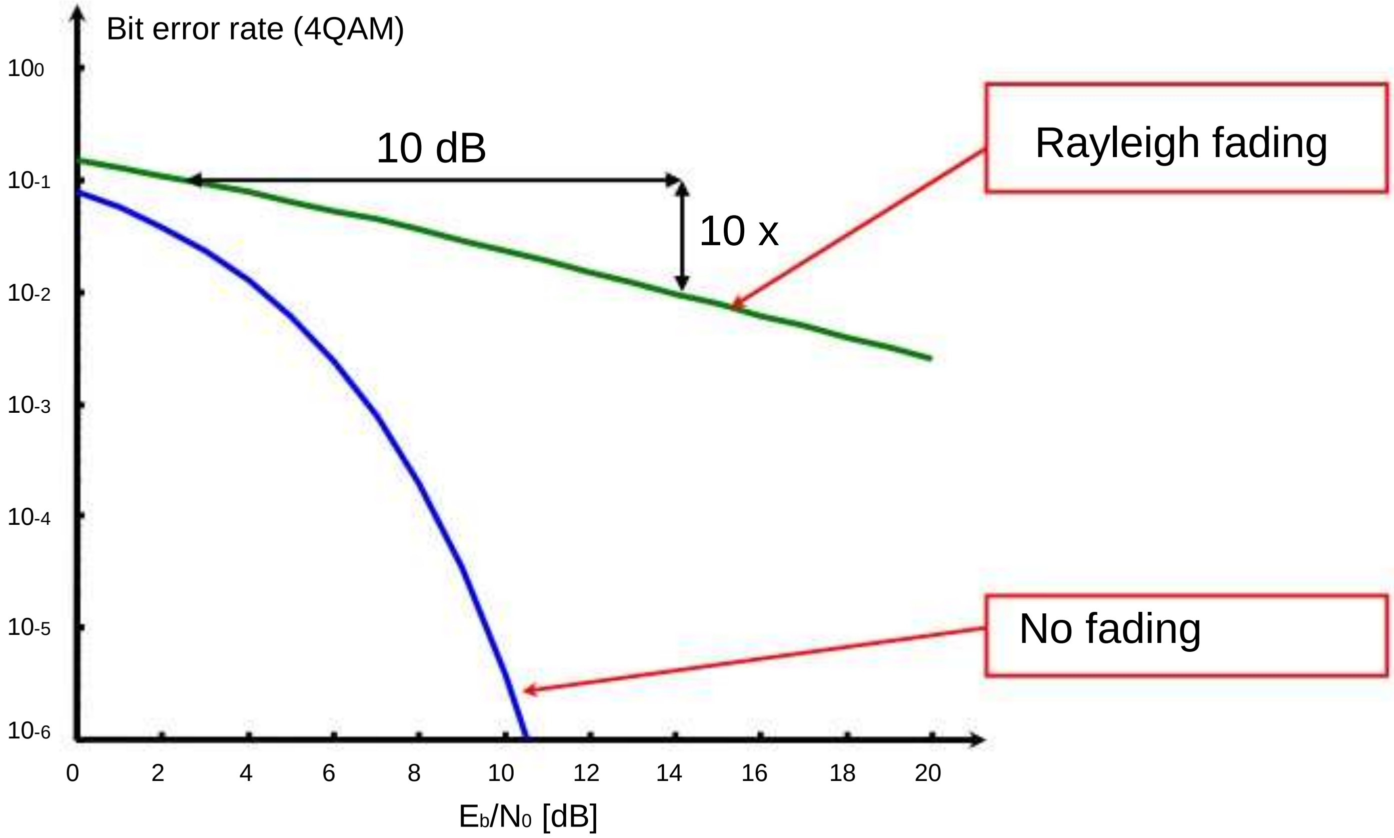
$$pdf\left(\gamma_b\right) = \frac{1}{\overline{\gamma_b}} e^{-\gamma_b / \overline{\gamma_b}}$$

γ_b	-- E_b/N_0
$\overline{\gamma_b}$	-- average E_b/N_0

$$BER_{Rayleigh}\left(\overline{\gamma_b}\right) = \int_0^\infty BER_{AWGN}\left(\gamma_b\right) \times pdf\left(\gamma_b\right) d\gamma_b$$

BER in fading channels (2)

THIS IS A SERIOUS PROBLEM!



BER in fading channels (3)

- Coherent detection of antipodal signals

$$\overline{BER} = \frac{1}{2}\left[1 - \sqrt{\frac{\overline{\gamma_B}}{1+\overline{\gamma_B}}}\right] \approx \frac{1}{4\overline{\gamma_B}}$$

- Coherent detection of orthogonal signals

$$\overline{BER} = \frac{1}{2}\left[1 - \sqrt{\frac{\overline{\gamma_B}}{2+\overline{\gamma_B}}}\right] \approx \frac{1}{2\overline{\gamma_B}}$$

- Differential detection of antipodal signals

$$\overline{BER} = \frac{1}{2+\overline{\gamma_B}} \approx \frac{1}{\overline{\gamma_B}}$$

- Differential detection of orthogonal signals

$$\overline{BER} = \frac{1}{2(1+\overline{\gamma_B})} \approx \frac{1}{2\overline{\gamma_B}}$$

Alternative computation of BER

- Alternative representation of Q-function

$$Q(x) = \frac{1}{\pi}\int_0^{\pi/2} \exp\left(-\frac{x^2}{2\sin^2\theta}\right) d\theta$$

- Example: SER of M-ary PSK in AWGN channel:

$$SER = \frac{1}{\pi}\int_0^{(M-1)\pi/M} \exp\left(-\frac{\gamma_S}{\sin^2\theta}\sin^2(\pi/M)\right) d\theta$$

- Averaged SER:

$$\overline{SER} = \frac{1}{\pi}\int_0^{(M-1)\pi/M} \int_0^{\infty} pdf_\gamma(\gamma) \exp\left(-\frac{\gamma_S}{\sin^2\theta}\sin^2(\pi/M)\right) d\theta$$

- This can be expressed in terms of the characteristic function of the fading distribution $M_\gamma(s)$

$$\overline{SER} = \frac{1}{\pi}\int_0^{(M-1)\pi/M} .M_\gamma\left(\frac{\gamma_S}{\sin^2\theta}\sin^2(\pi/M))\right) d\theta$$

Doppler-induced errors

- Distortion on the channel causes irreducible errors (cannot be eliminated by increasing transmit power)
- Frequency dispersion:
  - Due to Doppler effect
  - Instantaneous frequency can be computed as

$$f_{\text{inst}}(t) = \frac{\mathrm{Im}\left(r^*(t)\frac{dr(t)}{dt}\right)}{|r(t)|^2}$$

  - Large frequency shift in fading dips
  - Resulting BER (for MSK)

$$\overline{BER}_{\text{Doppler}} = \frac{1}{2}\pi^2(\nu_{\max}T_{\text{B}})^2 \quad .$$

  - Mostly relevant for low datarates

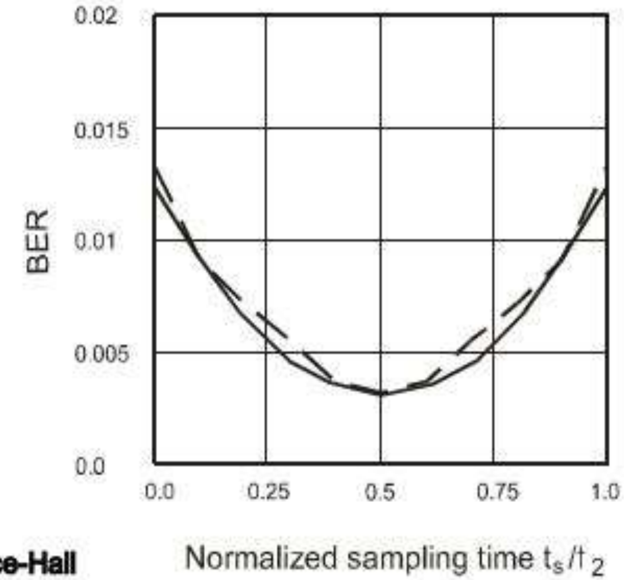
Errors induced by delay dispersion

- Delay dispersion causes intersymbol interference

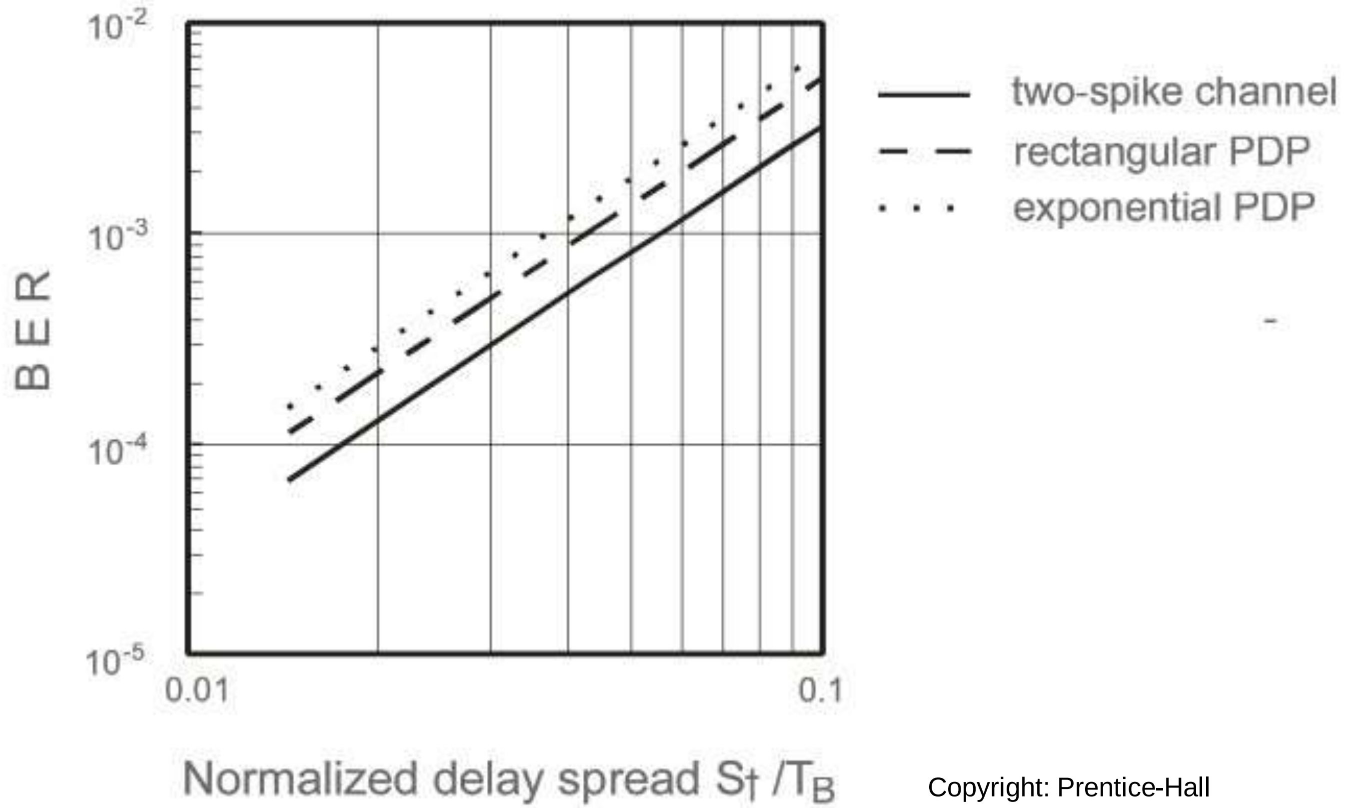
- Average BER

$$\overline{BER} = K(\frac{S_\tau}{T_B})^2$$

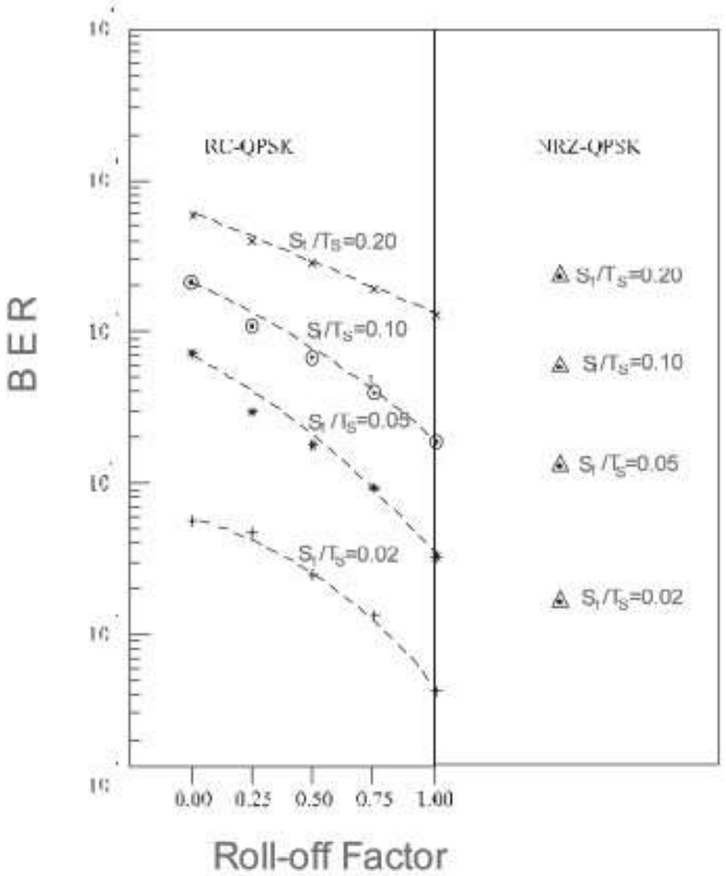
- Influenced by sampling instant



Errors induced by delay dispersion (2)



Impact of filtering



Computation methods (1)

- Group delay method: distortion of signal phase is related to group delay

$$\Phi_c(\omega) = \Phi_c(0) + \omega \frac{\partial \Phi_c}{\partial \omega}\Big|_{\omega=0} + \frac{1}{2}\omega^2 \frac{\partial^2 \Phi_c}{\partial \omega^2}\Big|_{\omega=0} + \ldots..$$

$$\approx \Phi_c(0) - \omega T_g$$

- Statistics of group delay

$$pdf_{T_g}(T_g) = \frac{1}{2S_\tau} \frac{1}{\left[1+\left(T_g/S_\tau\right)^2\right]^{3/2}}$$

- ->

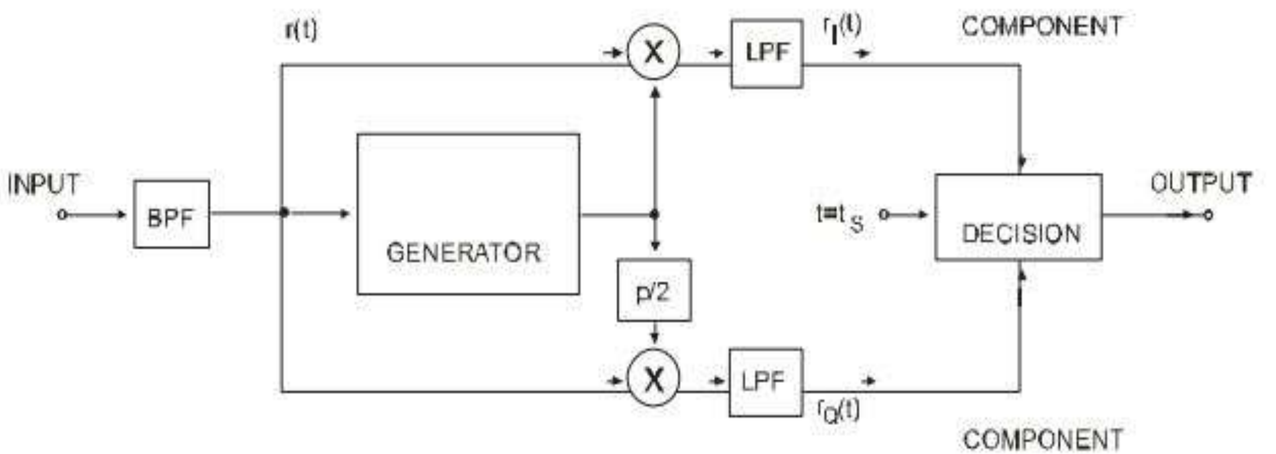
$$BER = \frac{4}{9}\left(\frac{S_\tau}{T_B}\right)^2 \approx \frac{1}{2}\left(\frac{S_\tau}{T_B}\right)^2.$$

Computation method (2)

- Quadratic form of Gaussian variables
- Formulate error event as

$$D = A|X|^2 + B|Y|^2 + CXY^* + C^*X^*Y < 0$$

- Canonical receiver



Computation method (3)

- Differentially-detected MSK

$$X = r(t_s) \; Y = r(t_s - T)$$

- Error condition is

$$\mathrm{Re}\{b_0 X Y^* \exp(-j\pi/2)\} \;\; < 0$$

- BER can be computed as

$$\overline{BER} = \frac{1}{2} - \frac{1}{2}\frac{b_0 \,\mathrm{Im}\{\rho_{XY}\}}{\sqrt{\mathrm{Im}\{\rho_{XY}\}^2 + (1 - |\rho_{XY}|^2)}}$$

MULTI PATH MITIGATION TECHNIQUES

Wireless Communication

Introduction

- The mobile radio channel places fundamental limitations on the performance of a wireless communication system
- The wireless transmission path may be
  - Line of Sight (LOS)
  - Non line of Sight (NLOS)
- Radio channels are random and time varying
- Modeling radio channels have been one of the difficult parts of mobile radio design and is done in statistical manner
- When electrons move, they create EM waves that can propagate through space.
- By using antennas we can transmit and receive these EM wave
- Microwave ,Infrared visible light and radio waves can be used.

Properties of Radio Waves

- Are easy to generate
- Can travel long distances
- Can penetrate buildings
- May be used for both indoor and outdoor coverage
- Are omni-directional-can travel in all directions
- Can be narrowly focused at high frequencies(>100MHz) using parabolic antenna

Properties of Radio Waves

- Frequency dependence
  - Behave more like light at high frequencies
    - Difficulty in passing obstacle
    - Follow direct paths
    - Absorbed by rain
  - Behave more like radio at lower frequencies
    - Can pass obstacles
    - Power falls off sharply with distance from source
- Subject to interference from other radio waves

Propagation Models

- The statistical modeling is usually done based on data measurements made specifically for
  - the intended communication system
  - the intended spectrum

- They are tools used for:
  - Predicting the average signal strength at a given distance from the transmitter
  - Estimating the variability of the signal strength in close spatial proximity to a particular locations

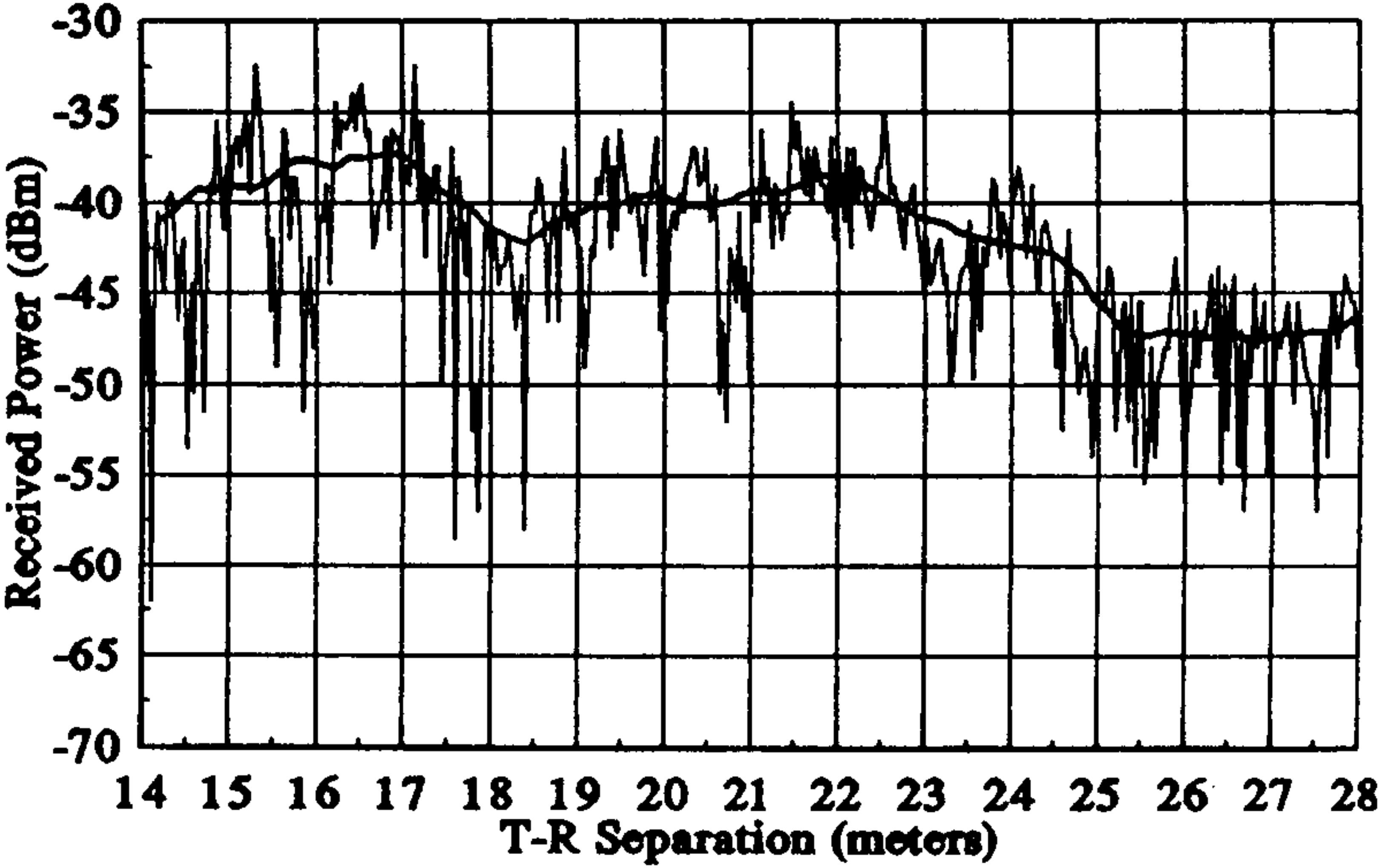
Propagation Models

- Large Scale Propagation Model:
  - Predict the mean signal strength for an arbitrary transmitter-receiver(T-R) separation
  - Estimate radio coverage of a transmitter
  - Characterize signal strength over large T-R separation distances(several 100's to 1000's meters)

Propagation Models

- Small Scale or Fading Models:
  - Characterize rapid fluctuations of received signal strength over
    - Very short travel distances(a few wavelengths)
    - Short time durations(on the order of seconds)

Small-scale and large-scale fading



Free Space Propagation Model

- For clear LOS between T-R
  - Ex: satellite & microwave communications

- Assumes that received power decays as a function of T-R distance separation raised to some power.

- Given by Friis free space eqn:

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

'L' is the system loss factor

L >1 indicates loss due to transmission line attenuation, filter losses & antenna losses

L = 1 indicates no loss in the system hardware

- Gain of antenna is related to its effective aperture A_e by

$$G = 4\pi A_e / \lambda^2$$

Free Space Propagation Model

- Effective Aperture A_e is related to physical size of antenna.

  $\lambda = c/f$.
- c is speed of light,
- P_t and P_r must be in same units
- G_t ad G_r are dimensionless

- An isotropic radiator, an ideal radiator which radiates power with unit gain uniformly in all directions, and is often used as reference

- Effective Isotropic Radiated Power (EIRP) is defined as

  EIRP= Pt Gt
- Represents the max radiated power available from a transmitter in direction of maximum antenna gain, as compared to an isotropic radiator

Free Space Propagation Model

- In practice Effective Radiated Power (ERP) is used instead of (EIRP)

- Effective Radiated Power (ERP) is radiated power compared to half wave dipole antennas

- Since dipole antenna has gain of 1.64(2.15 dB)
  ERP=EIRP-2.15(dB)

- the ERP will be 2.15dB smaller than the EIRP for same Transmission medium

Free Space Propagation Model

- Path Loss (PL) represents signal attenuation and is defined as difference between the effective transmitted power and received power

Path loss PL(dB) = 10 log [Pt/Pr]

= -10 log {GtGr λ^2/(4π)^2d^2}

- Without antenna gains (with unit antenna gains)

PL = - 10 log { λ^2/(4π)^2d^2}

- Friis free space model is valid predictor for P_r for values of d which are in the far-field of transmitting antenna

Free Space Propagation Model

- The far field or Fraunhofer region that is beyond far field distance d_f given as : $d_f = 2D^2/\lambda$
- D is the largest physical linear dimension of the transmitter antenna
- Additionally, $d_f >> D$ and $d_f >> \lambda$
- The Friis free space equation does not hold for d=0
- Large Scale Propagation models use a close-in distance, d_o, as received power reference point, chosen such that $d_o >= d_f$
- Received power in free space at a distance greater then do

$$Pr(d) = Pr(do)(do/d)^2 \quad d > d_o > d_f$$

Pr with reference to 1 mW is represented as

$$Pr(d) = 10\log(Pr(do)/0.001W) + 20\log(do/d)$$

Electrostatic,inductive and radiated fields are launched, due to flow of current from anntena.

Regions far away from transmitter electrostatic and inductive fields become negligible and only radiated field components are considered.

Example

- What will be the far-field distance for a Base station antenna with
- Largest dimension D=0.5m
- Frequency of operation fc=900MHz,1800MHz

- For 900MHz
- *λ =3*10^8/900*10^6)=0.33m*
- *df= 2D^2/ λ =2(0.5)^2/0.33=1.5m*

Example

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

solution

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).

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

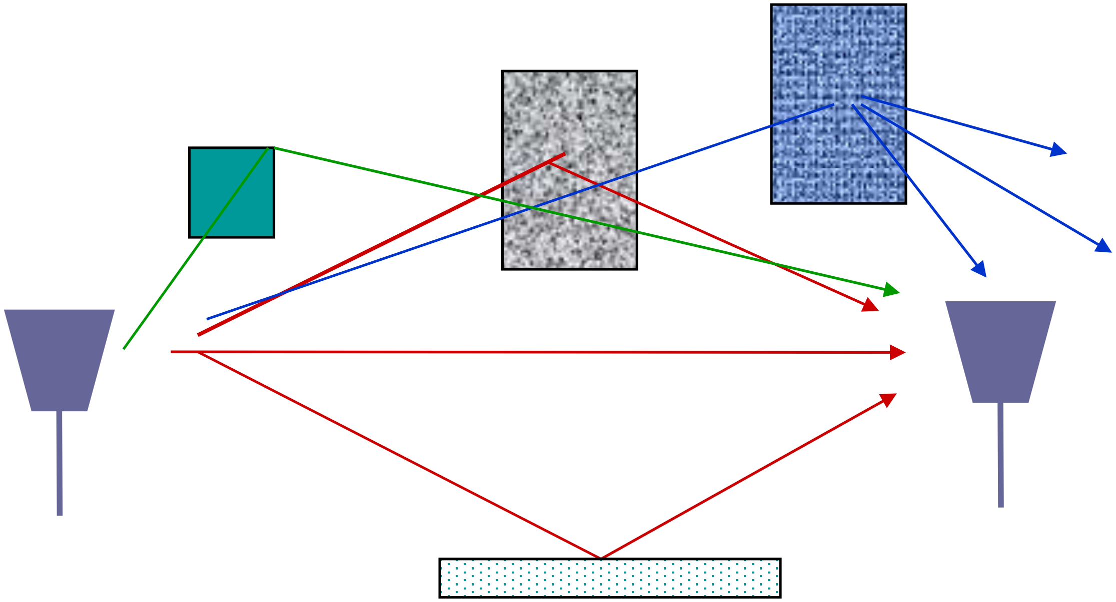
$$P_r(\text{dBm}) = 10\log P_r(\text{mW}) = 10\log\left(3.5 \times 10^{-3} \text{ mW}\right) = -24.5 \text{ dBm}.$$

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

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

Propagation Mechanisms

- Three basic propagation mechanism which impact propagation in mobile radio communication system are:
  - Reflection
  - Diffraction
  - Scattering



Propagation Mechanisms

- Reflection occurs when a propagating electromagnetic wave impinges on an object which has very large dimensions as compared to wavelength e.g. surface of earth , buildings, walls

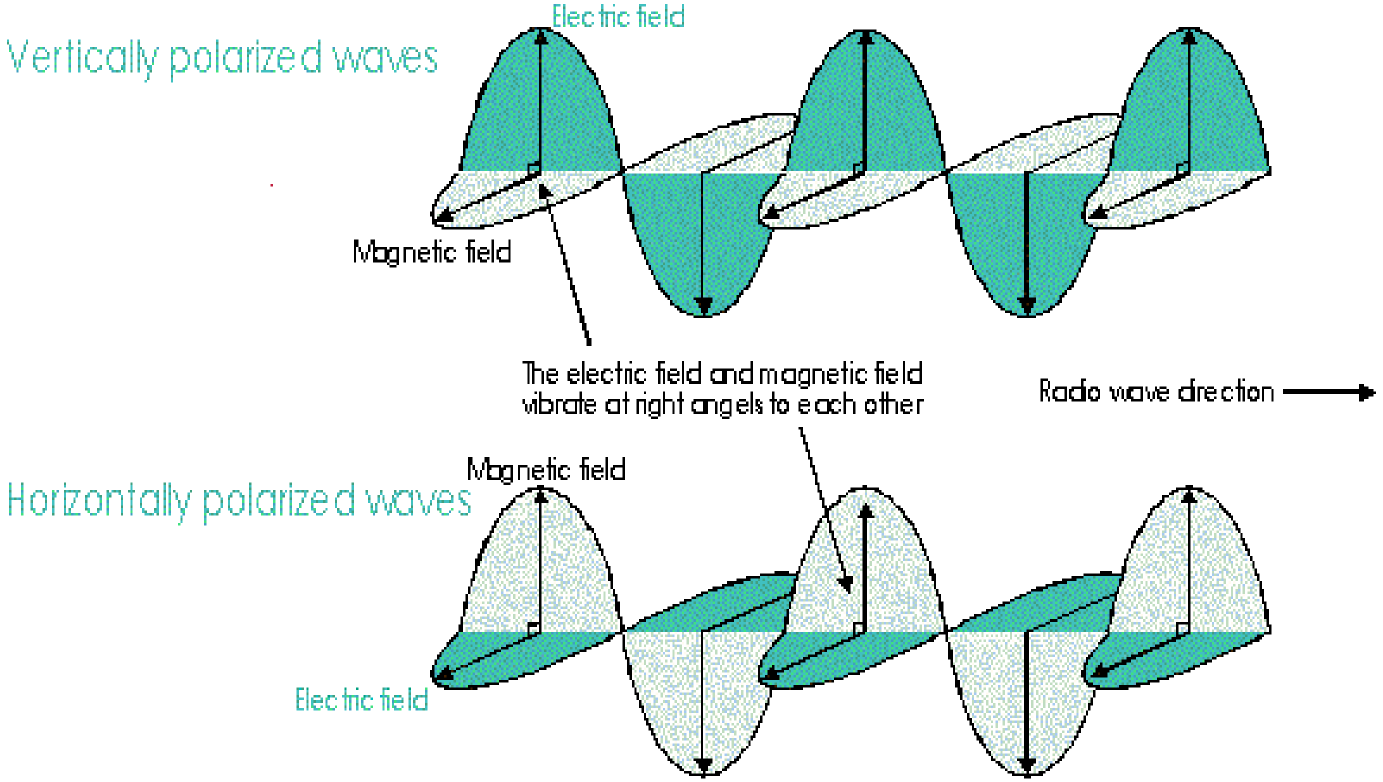
- Diffraction occurs when the radio path between the transmitter and receiver is obstructed by a surface that has sharp irregularities(edges)
  - Explains how radio signals can travel urban and rural environments without a line of sight path

- Scattering occurs when medium has objects that are smaller or comparable to the wavelength (small objects, irregularities on channel, foliage, street signs etc)

Reflection

- Occurs when a radio wave propagating in one medium impinges upon another medium having different electrical properties
- If radio wave is incident on a perfect dielectric
  - Part of energy is reflected back
  - Part of energy is transmitted
- In addition to the change of direction, the interaction between the wave and boundary causes the energy to be split between reflected and transmitted waves
- The amplitudes of the reflected and transmitted waves are given relative to the incident wave amplitude by Fresnel reflection coefficients

Vertical and Horizontal polarization



Reflection- Dielectrics

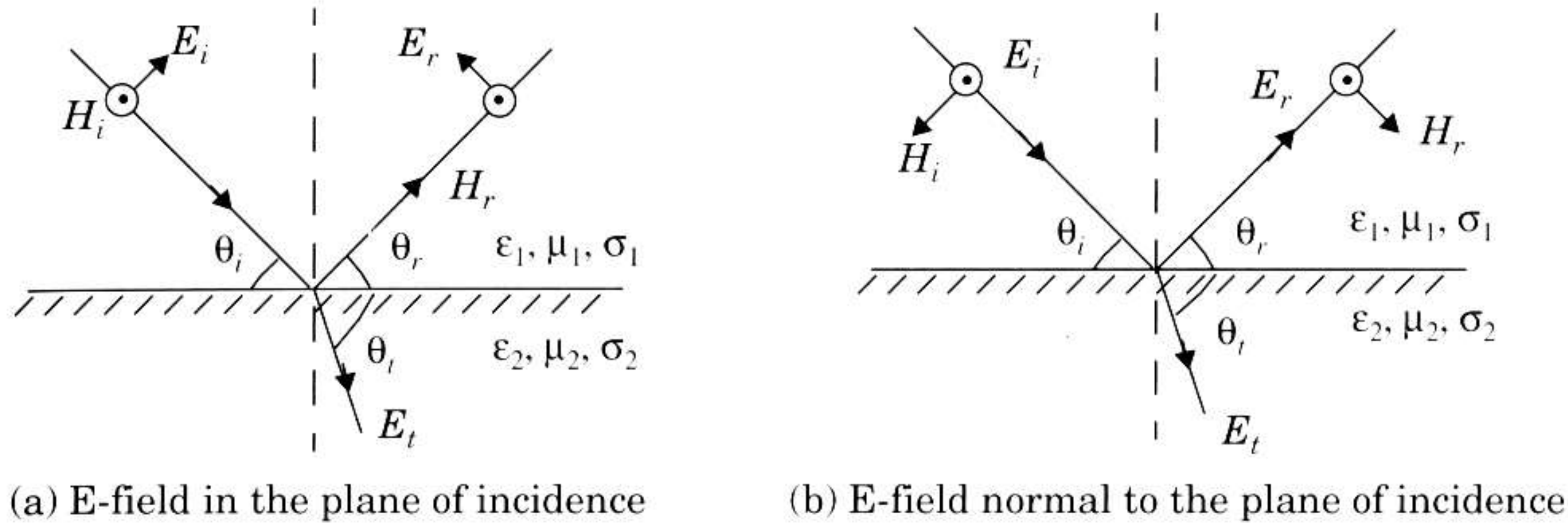


Figure 4.4 Geometry for calculating the reflection coefficients between two dielectrics.

Reflection

- $\Gamma(\parallel) = \frac{E_r}{E_i} = \frac{\eta_2 \sin\theta_t - \eta_1 \sin\theta_i}{\eta_1 \sin\theta_t + \eta_1 \sin\theta_i}$ (Paralell E-field polarization)
- $\Gamma(\perp) = \frac{E_r}{E_i} = \frac{\eta_2 \sin\theta_i - \eta_1 \sin\theta_t}{\eta_1 \sin\theta_i + \eta_1 \sin\theta_t}$ (Perpendicular E-field polarization)
- These expressions express ratio of reflected electric fields to the incident electric field and depend on impedance of media and on angles
- η is the intrinsic impedance given by $\sqrt{(\mu/\varepsilon)}$
- μ=permeability,ε=permittivity

- Γ(∥)= / = (Paralell E-field polarization)
- Γ(⊥)= = (Perpendicular E-field polarization)
- These expressions express ratio of reflected electric fields to the incident electric field and depend on impedance of media and on angles
- η is the intrinsic impedance given by
- μ=permeability,ε=permittivity

Reflection-Perfect Conductor

- If incident on a perfect conductor the entire EM energy is reflected back
- Here we have $\theta_r = \theta_i$
- $E_i = E_r$ (E-field in plane of incidence)
- $E_i = -E_r$ (E field normal to plane of incidence)
- Γ(parallel)= 1
- Γ(perpendicular)= -1

Reflection - Brewster Angle

- It is the angle at which no reflection occur in the medium of origin. It occurs when the incident angle θ_B is such that the reflection coefficient Γ(parallel) is equal to zero.
- It is given in terms of θ_B as given below

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

- When first medium is a free space and second medium has an relative permittivity of $\varepsilon_r\ then$

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

- Brewster angle only occur for parallel polarization

- It is the angle at which no reflection occur in the medium of origin. It occurs when the incident angle is such that the reflection coefficient Γ(parallel) is equal to zero.
- It is given in terms of as given below
- When first medium is a free space and second medium has an relative permittivity of
- Brewster angle only occur for parallel polarization

Ground Reflection(Two Ray) Model

- In mobile radio channel, single direct path between base station and mobile and is seldom only physical means for propagation
- Free space model as a stand alone is inaccurate
- Two ray ground reflection model is useful
  - Based on geometric optics
  - Considers both direct and ground reflected path
- Reasonably accurate for predicting large scale signal strength over several kms that use tall tower height
- Assumption: The height of Transmitter >50 meters

Ground Reflection(Two Ray) Model

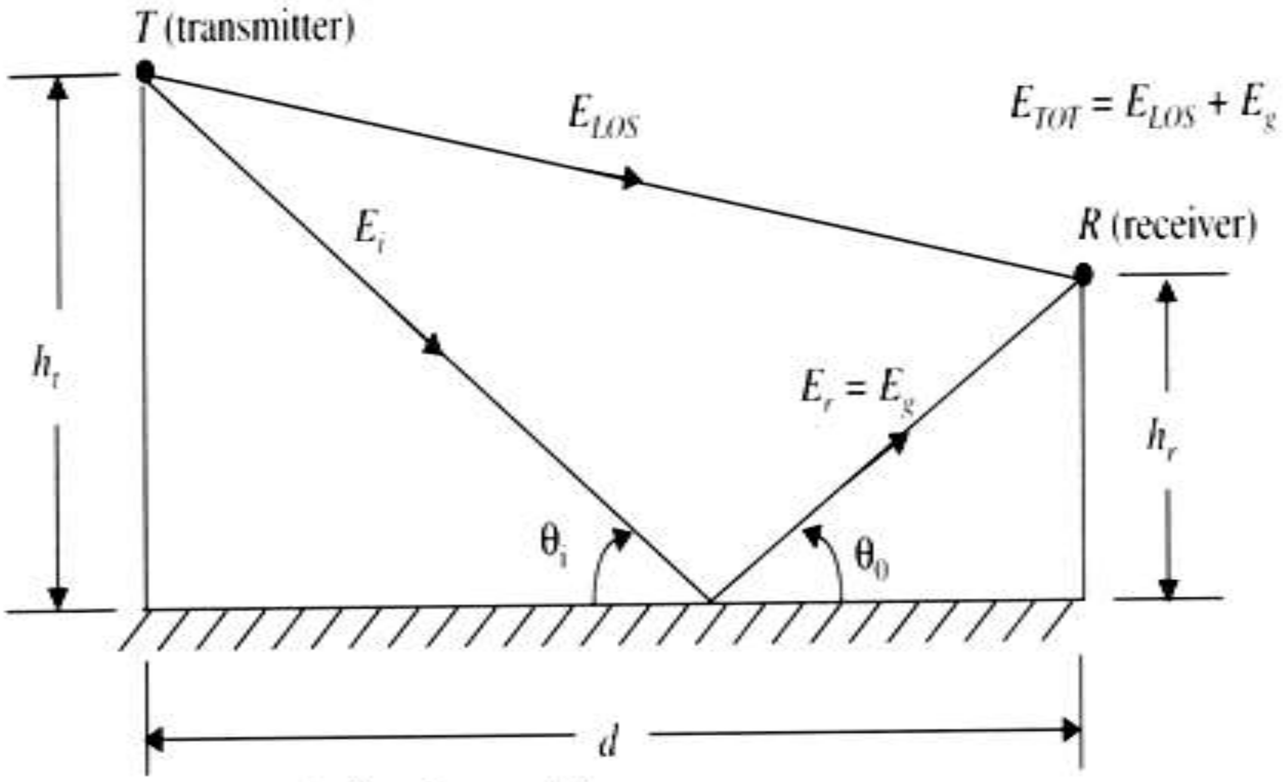


Figure 4.7 Two-ray ground reflection model.

Ground Reflection(Two Ray) Model

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

let E_o be $|\ \vec{E}\ |$ at reference point d_o then

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

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

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

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

Ground Reflection(Two Ray) Model

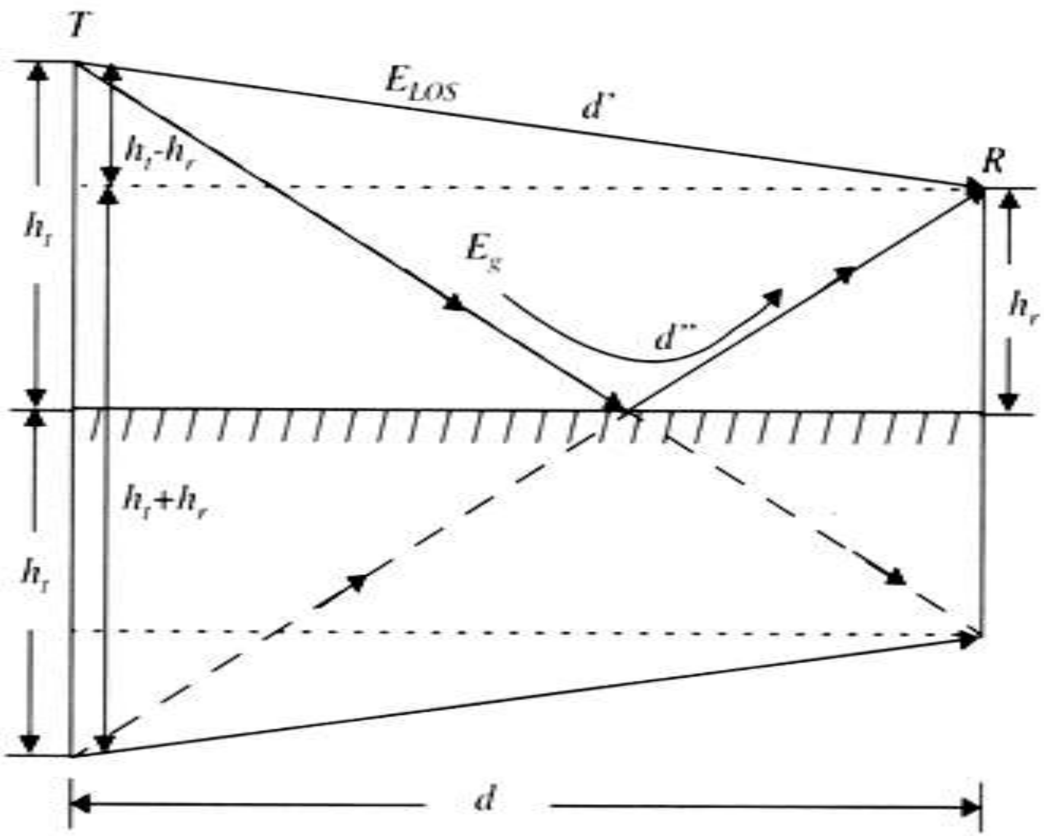


Figure 4.8 The method of images is used to find the path difference between the line-of-sight and the ground reflected paths.

Path Difference

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

$$= d\sqrt{\left(\left(\frac{h_t + h_r}{d}\right)^2 + 1\right)} - d\sqrt{\left(\left(\frac{h_t - h_r}{d}\right)^2 + 1\right)}$$

$$\approx d\left(1 + \frac{1}{2}\left(\frac{h_t + h_r}{d}\right)^2\right) - d\left(1 + \frac{1}{2}\left(\frac{h_t - h_r}{d}\right)^2\right)$$

$$\approx \frac{1}{2d}\left((h_t + h_r)^2 - (h_t - h_r)^2\right)$$

$$\approx \frac{1}{2d}\left(\left(h_t^2 + 2h_t h_r + h_r^2\right) - \left(h_t^2 - 2h_t h_r + h_r^2\right)\right)$$

$$\approx \frac{2h_t h_r}{d}$$

Phase difference

$$\theta_\Delta \text{ radians} = \frac{2\pi\Delta}{\lambda} = \frac{2\pi\Delta}{\left(c/f_c \right)} = \frac{\omega_c \Delta}{c}$$

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

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

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

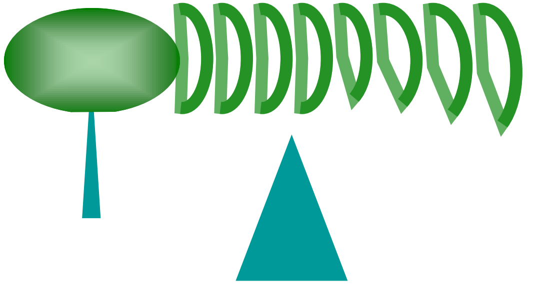
$$\overline{P}_r = P_t G_t G_r \frac{h_t^2 h_r^2}{d^4}$$

Diffraction

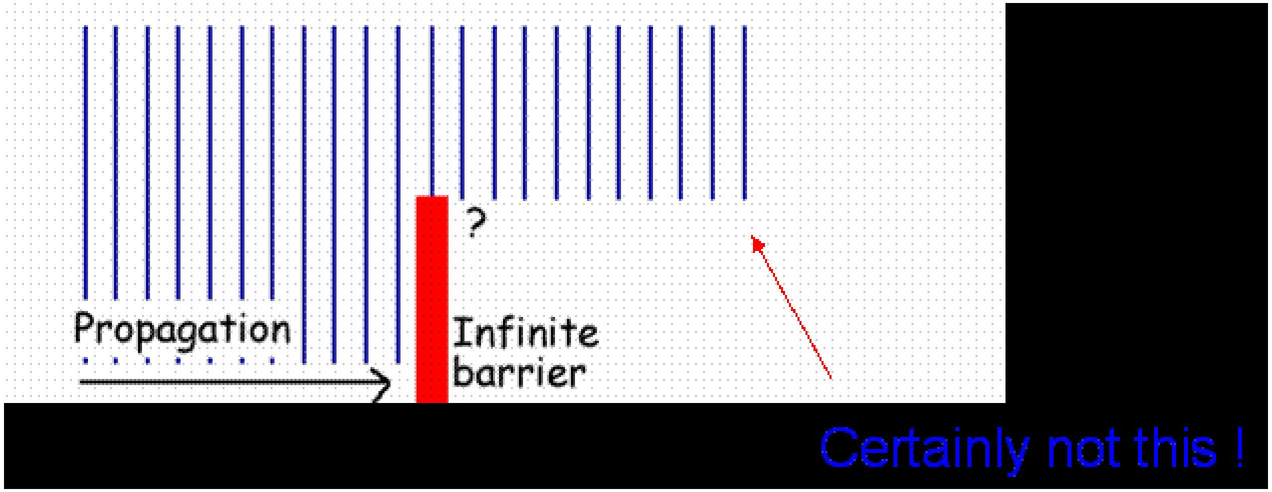
- Diffraction is the bending of wave fronts around obstacles.

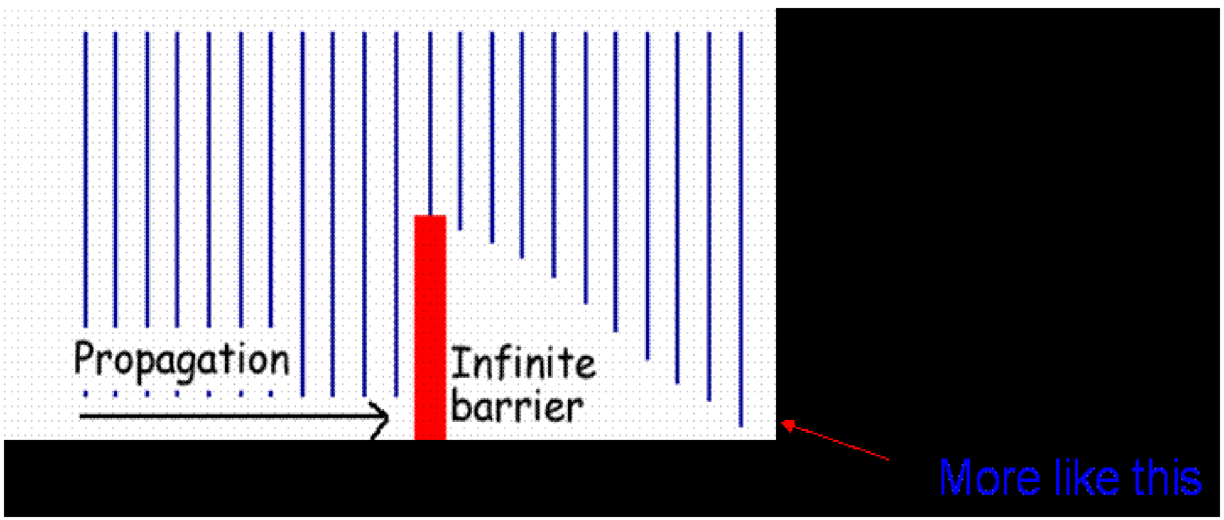
- Diffraction allows radio signals to propagate behind obstructions and is thus one of the factors why we receive signals at locations where there is no line-of-sight from base stations

- Although the received field strength decreases rapidly as a receiver moves deeper into an obstructed (shadowed) region, the diffraction field still exists and often has sufficient signal strength to produce a useful signal.



Diffraction





Knife-edge Diffraction Model

- Estimating the signal attenuation caused by diffraction of radio waves over hills and buildings is essential in predicting the field strength in a given service area.
- As a starting point, the limiting case of propagation over a knife edge gives good in sight into the order of magnitude diffraction loss.
- When shadowing is caused by a single object such as a building, the attenuation caused by diffraction can be estimated by treating the obstruction as a diffracting knife edge

Knife-edge Diffraction Model

Consider a receiver at point *R located in the shadowed region. The field strength at point R is a vector sum of the fields due to all of the secondary Huygens sources in the plane above the knife edge.*

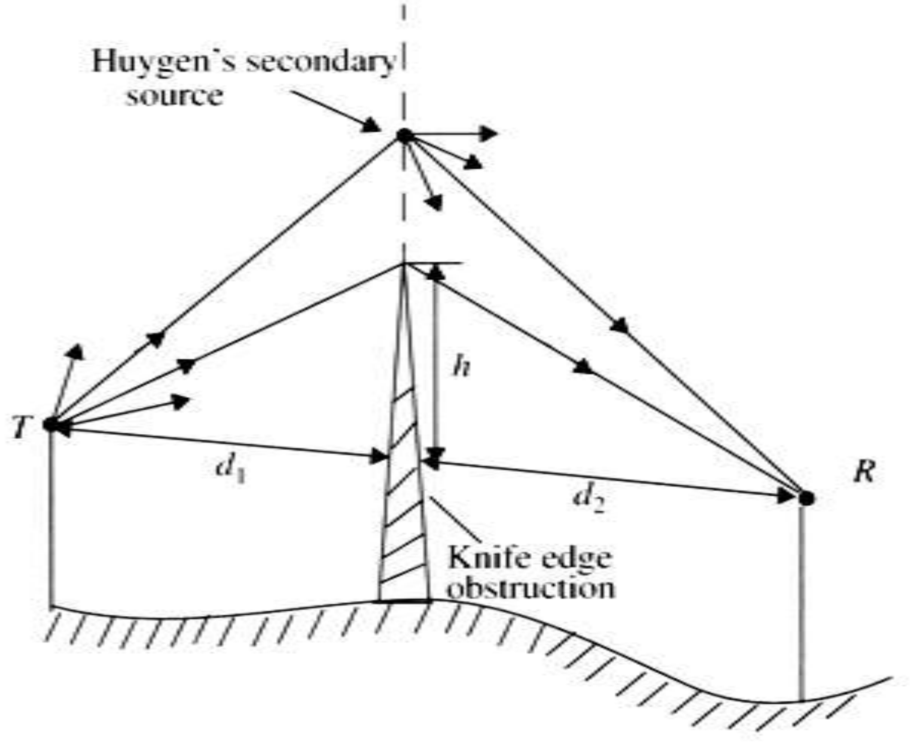


Figure 4.13 Illustration of knife-edge diffraction geometry. The receiver *R* is located in the shadow region.

Knife-edge Diffraction Model

- The difference between the direct path and diffracted path, call *excess path length*

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

- The corresponding phase difference

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

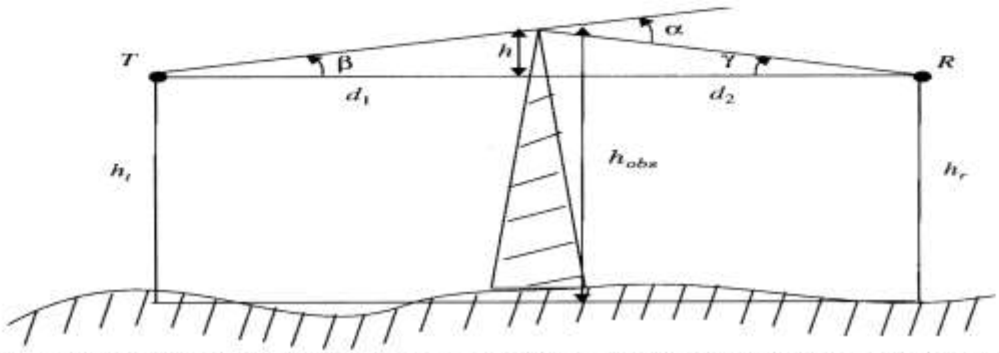
- *Fresnel-Kirchoff* diffraction parameter is used to normalize the phased term and given as

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

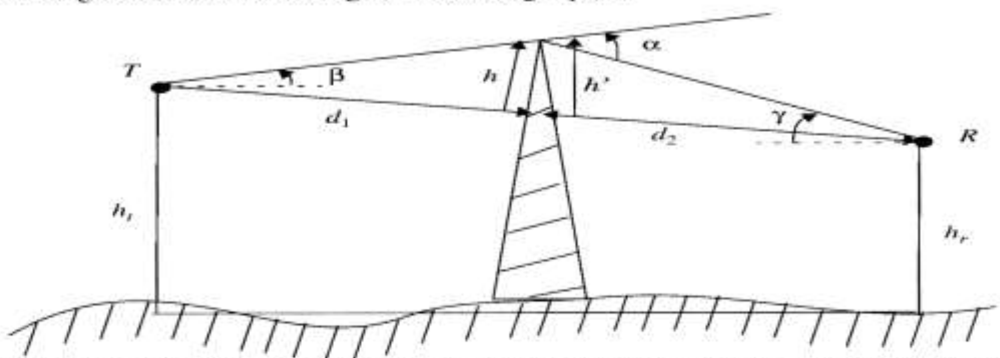
Which gives $\phi = \frac{\pi}{2}v^2$

- where $\alpha = h\left(\frac{d_1+d_2}{d_1 d_2}\right)$

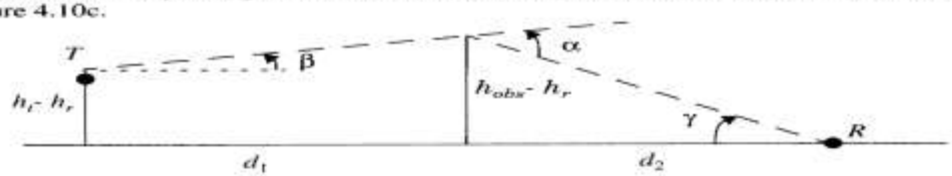
Knife-edge Diffraction Model



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



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



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

Figure 4.10 Diagrams of knife-edge geometry.

Fresnel zones

- Fresnel zones represent successive regions where secondary waves have a path length from the TX to the RX which are $n\lambda/2$ greater in path length than of the LOS path. The plane below illustrates successive Fresnel zones.

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

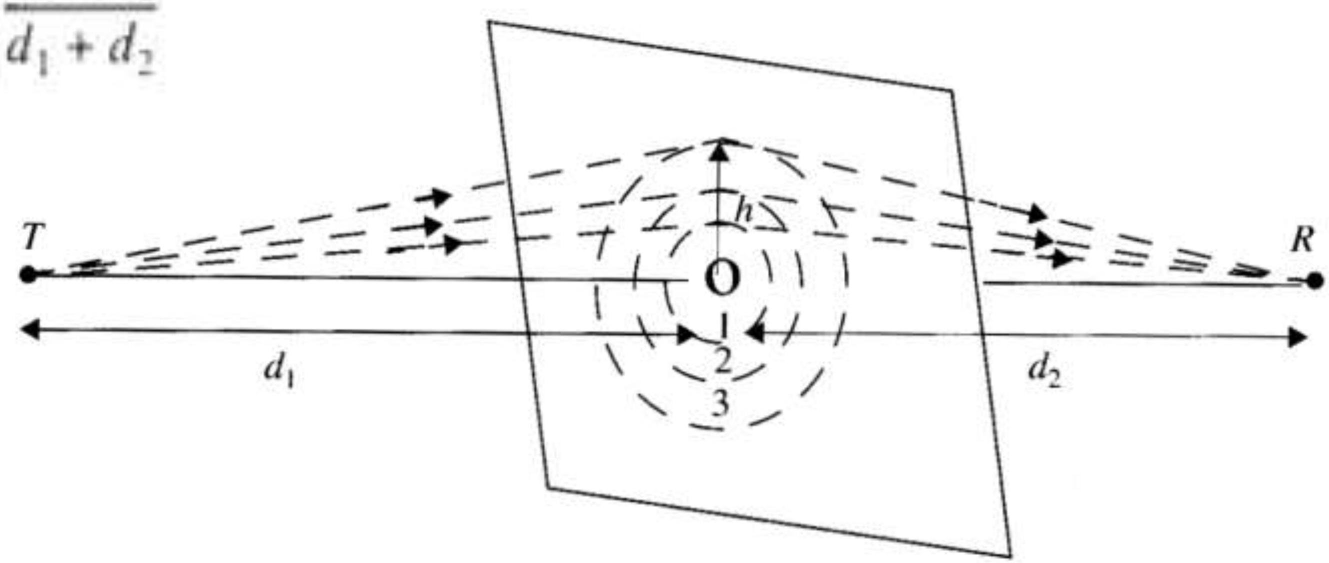
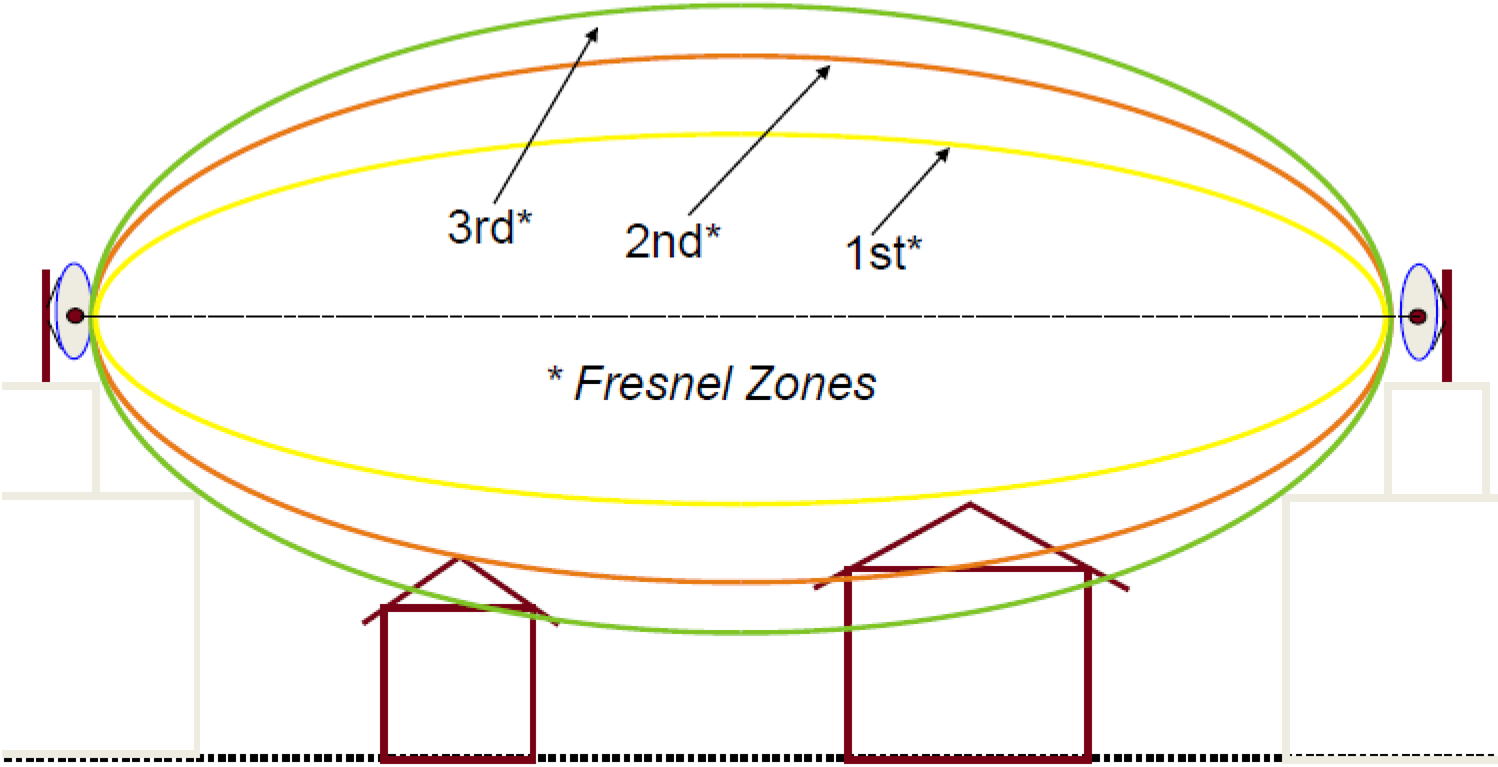


Figure 4.11 Concentric circles which define the boundaries of successive Fresnel zones.

Fresnel zones



Diffraction gain

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

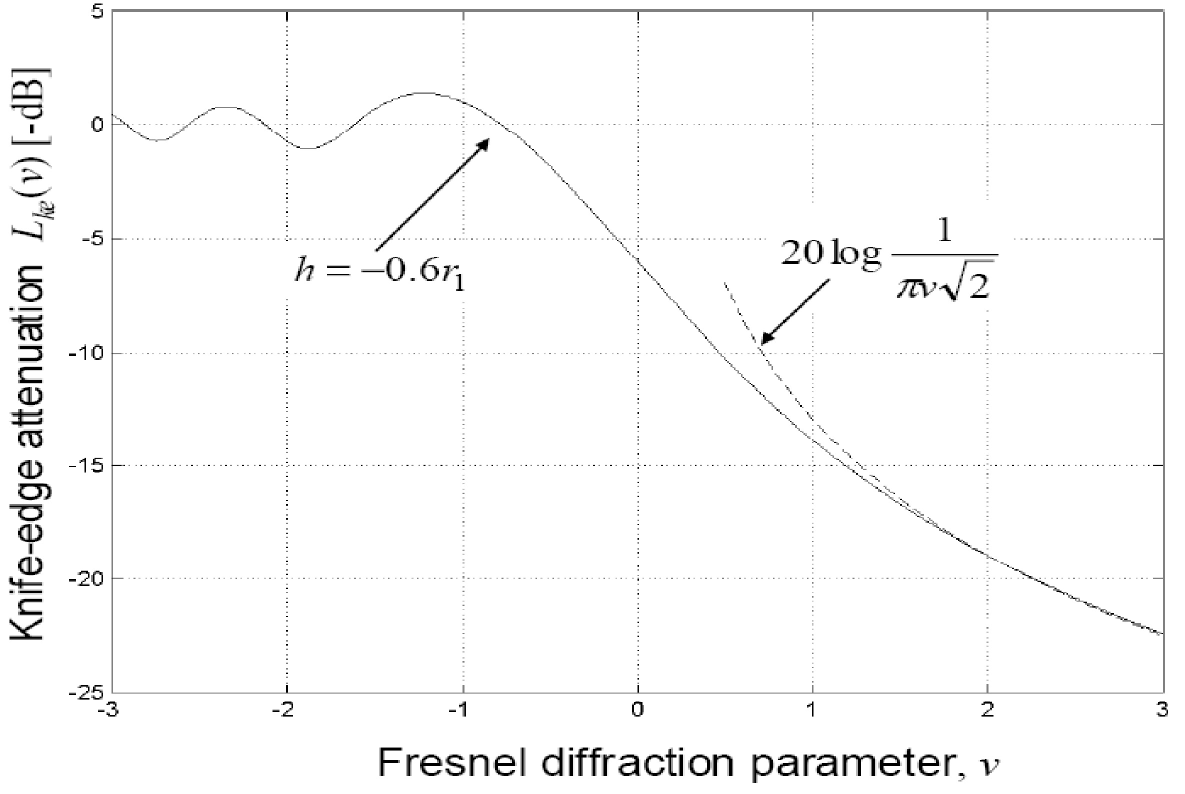
$$G_d(\text{dB}) = 20\log|F(v)|$$

- The electric field strength, *Ed, of a knife edge diffracted wave is given by*

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

- *Eo : is the free space field strength in the absence of both the ground and the knife edge.*
- *F(v): is the complex fresnel integral.*
- *v: is the Fresnel-Kirchoff diffraction parameter*

Graphical Calculation of diffraction attenuation



Numerical solution

- An approximate numerical solution for equation

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

- Can be found using set of equations given below for different values of v

$G_d(dB)$	v
0	≤ -1
$20 \log(0.5-0.62v)$	[-1,0]
$20 \log(0.5\, e^{-0.95v})$	[0,1]
$20 \log(0.4-(0.1184-(0.38-0.1v)^2)^{1/2})$	[1, 2.4]
$20 \log(0.225/v)$	> 2.4

Example

Example 4.7

Compute the diffraction loss for the three cases shown in Figure 4.12. Assume $\lambda = 1/3$ m, $d_1 = 1$km, $d_2 = 1$ km, and (a) h = 25 m, (b) $h = 0$, (c) h = –25 m. Compare your answers using values from Figure 4.14, as well as the approximate solution given by Equation (4.61.a)–(4.61.e). For each of these cases, identify the Fresnel zone within which the tip of the obstruction lies.

Given:

$\lambda = 1/3$ m

$d_1 = 1$ km

$d_2 = 1$ km

(a)h = 25 m

Using Equation (4.56), the Fresnel diffraction parameter is obtained as

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

From Figure 4.14, the diffraction loss is obtained as 22 dB.

Using the numerical approximation in Equation (4.61.e), the diffraction loss is equal to 21.7 dB.

The path length difference between the direct and diffracted rays is given by Equation (4.54) as

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

To find the Fresnel zone in which the tip of the obstruction lies, we need to compute n which satisfies the relation $\Delta = n\lambda/2$. For $\lambda = 1/3$ m, and $\Delta = 0/625$ m, we obtain

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

Therefore, the tip of the obstruction completely blocks the first three Fresnel zones.

(b) $h = 0$ m

Therefore, the Fresnel diffraction parameter $v = 0$.

From Figure 4.14, the diffraction loss is obtained as 6 dB.

Using the numerical approximation in Equation (4.61.b), the diffraction loss is equal to 6 dB.

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

(c) $h = -25$ m

Using Equation (4.56), the Fresnel diffraction parameter is obtained as -2.74.

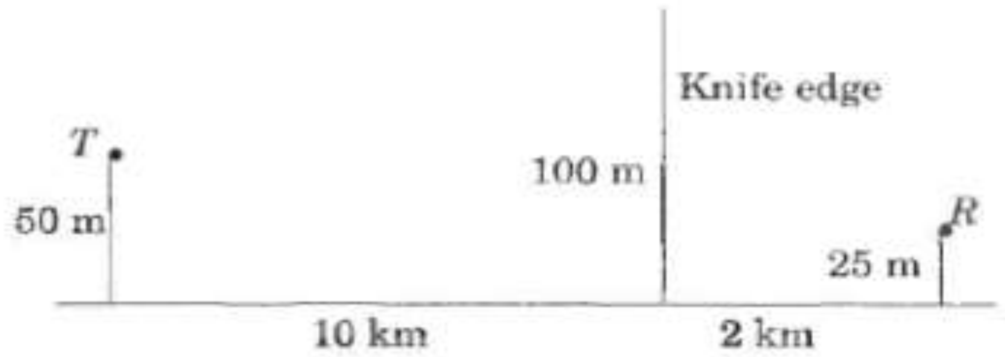
From Figure 4.14, the diffraction loss is approximately equal to 1 dB.

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

Example

Example 4.8

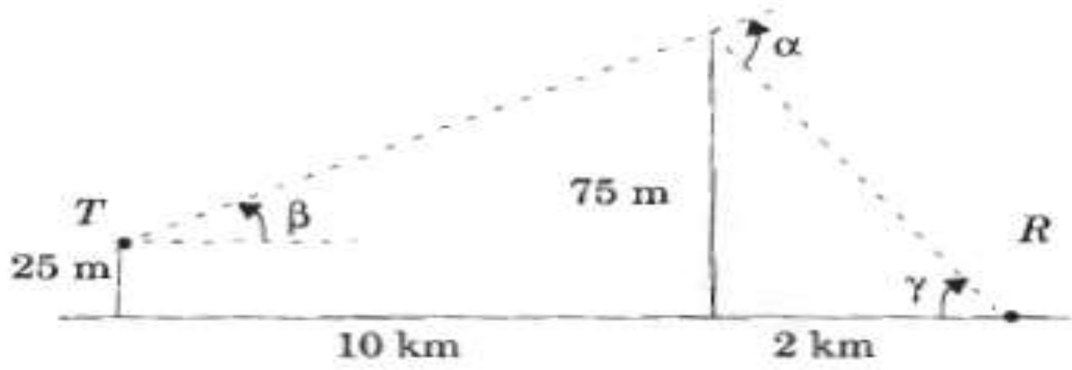
Given the following geometry, determine (a) the loss due to knife-edge diffraction, and (b) the height of the obstacle required to induce 6 dB diffraction loss. Assume $f = 900$ MHz.



Solution

(a) The wavelength $\lambda = \frac{c}{f} = \frac{3 \times 10^8}{900 \times 10^6} = \frac{1}{3}$ m.

Redraw the geometry by subtracting the height of the smallest structure.



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

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

and

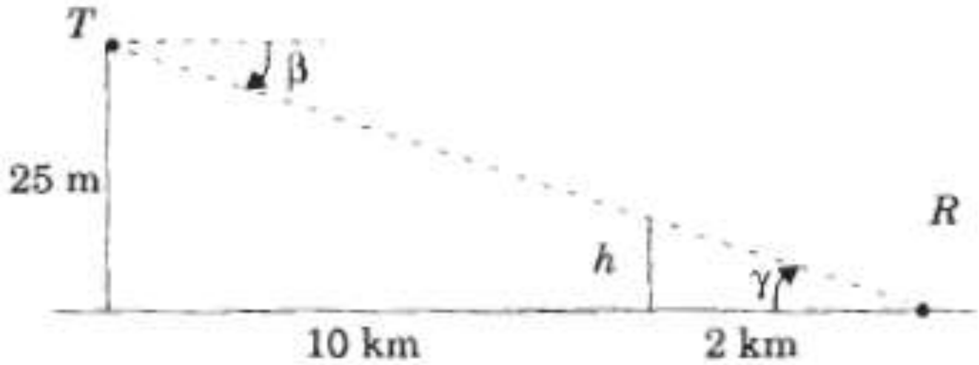
$$\alpha = \beta + \gamma = 2.434^\circ = 0.0424 \text{ rad}$$

Then using Equation (4.56)

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

From Figure 4.14 or (4.61.e), the diffraction loss is 25.5 dB.

(b) For 6 dB diffraction loss, $v = 0$. The obstruction height h may be found using similar triangles ($\beta = \gamma$), as shown below.



It follows that $\frac{h}{2000} = \frac{25}{12000}$, thus $h = 4.16$ m.

Multiple Knife Edge Diffraction

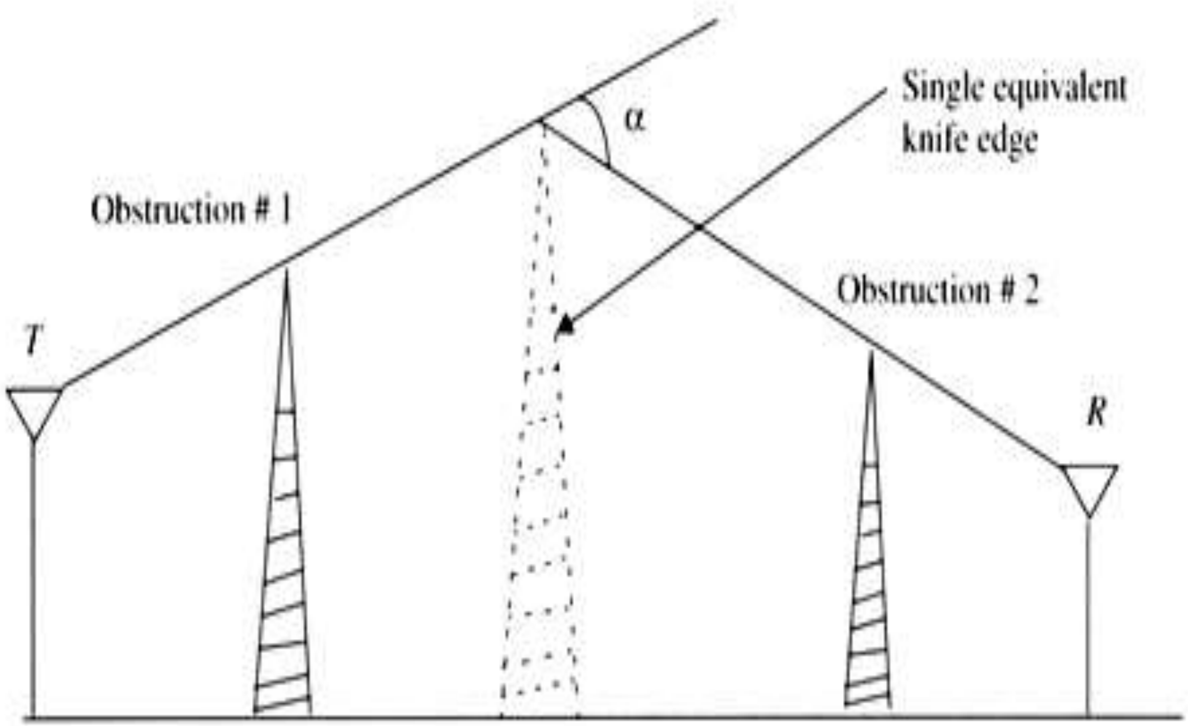


Figure 4.15 Bullington's construction of an equivalent knife edge [from [Bul47] © IEEE].

Scattering

- Scattering occurs when the medium through which the wave travels consists of objects with dimensions that are small compared to the wavelength, and where the number of obstacles per unit volume is large.

- Scattered waves are produced by
  - rough surfaces,
  - small objects,
  - or by other irregularities in the channel.

- Scattering is caused by trees, lamp posts, towers, etc.

Scattering

- Received signal strength is often stronger than that predicted by reflection/diffraction models alone

- The EM wave incident upon a rough or complex surface is **scattered** in **many** directions and provides more energy at a receiver
  - energy that would have been absorbed is instead reflected to the Rx.

- flat surface → EM reflection (one direction)
- rough surface → EM scattering (many directions)

Scattering

- Rayleigh criterion: used for testing surface roughness
- A surface is considered smooth if its min to max protuberance (bumps) h is less than critical height hc

$h_c = \lambda/8 \sin\Theta_i$

- Scattering path loss factor ρ_s is given by

$$\rho_s = \exp[-8[(\pi*\sigma_h*\sin\Theta_i)/\lambda]^2]$$

Where h is surface height and σ_h is standard deviation of surface height about mean surface height.

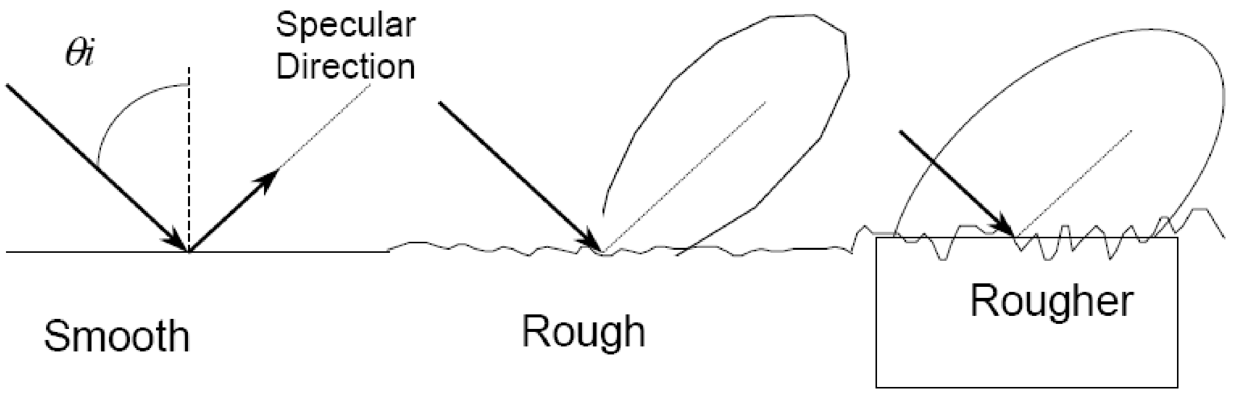
- For rough surface, the flat surface reflection coefficient is multiplied by scattering loss factor ρ_s to account for diminished electric field

- Reflected E-fields for $h > h_c$ for rough surface can be calculated as

$\Gamma_{rough} = \rho_s \Gamma$

Scattering

Rough Surface Scattering



Roughness depends on :

- Surface height range
- Angle of incidence
- Wavelength

14

Outdoor propagation Environment

- Based on the coverage area, the Outdoor propagation environment may be divided into three categories

  1. Propagation in Macro cells
  2. Propagation in Micro cells
  3. Propagation in street Micro cells

Outdoor propagation Environment

Macrocells versus Microcells

	Macrocell	Microcell
Cell Radius	1 to 20 km	0.1 to 1 km
Tx Power	1 to 10 W	0.1 to 1 W
Fading	Rayleigh	Nakgami-Rice
RMS Delay Spread	0.1 to 10 µs	10 to 100ns
Max. Bit Rate	0.3 Mbps	1 Mbps

Outdoor propagation Models

- Outdoor radio transmission takes place over an irregular terrain.
- The terrain profile must be taken into consideration for estimating the path loss

  e.g. trees buildings and hills must be taken into consideration

- Some common models used are
  - Longley Rice Model
  - Okumura Model
  - Hatta model

Longley Rice Model

- Longley Rice Model is applicable to point to point communication.
- It covers 40MHz to 300 GHz
- It can be used in wide range of terrain
- Path geometry of terrain and the refractivity of troposphere is used for transmission path loss calculations
- Geometrical optics is also used along with the two ray model for the calculation of signal strength.
- Two modes
  - Point to point mode prediction
  - Area mode prediction

Longley Rice Model

- Longley Rice Model is normally available as a computer program which takes inputs as
  - Transmission frequency
  - Path length
  - Polarization
  - Antenna heights
  - Surface reflectivity
  - Ground conductivity and dialectic constants
  - Climate factors
  - A problem with Longley rice is that It doesn't take into account the buildings and multipath.

Okumura Model

- In 1968 Okumura did a lot of measurements and produce a new model.
- The new model was used for signal prediction in Urban areas.
- Okumura introduced a graphical method to predict the median attenuation relative to free-space for a quasi-smooth terrain
- The model consists of a set of curves developed from measurements and is valid for a particular set of system parameters in terms of carrier frequency, antenna height, etc.

Okumura Model

- First of all the model determined the free space path loss of link.
- After the free-space path loss has been computed, the median attenuation, as given by Okumura's curves has to be taken to account
- The model was designed for use in the frequency range 200 up to 1920 MHz and mostly in an urban propagation environment.
- Okumura's model assumes that the path loss between the TX and RX in the terrestrial propagation environment can be expressed as:

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

Okumura Model

- Estimating path loss using Okumura Model

1. Determine free space loss and $A_{mu}(f,d)$, between points of interest
2. Add $Amu(f,d)$ and correction factors to account for terrain

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

L_{50} = 50% value of propagation path loss (median)

L_F = free space propagation loss

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

$G(h_{te})$ = base station antenna height gain factor

$G(h_{re})$ = mobile antenna height gain factor

G_{AREA} = gain due to environment

Okumura Model

- $A_{mu}(f,d)$ & G_{AREA} have been plotted for wide range of frequencies
- Antenna gain varies at rate of 20dB or 10dB per decade

$G(h_{te}) = 20\log\frac{h_{te}}{200}$ $\quad$ $10m < h_{te} < 1000m$

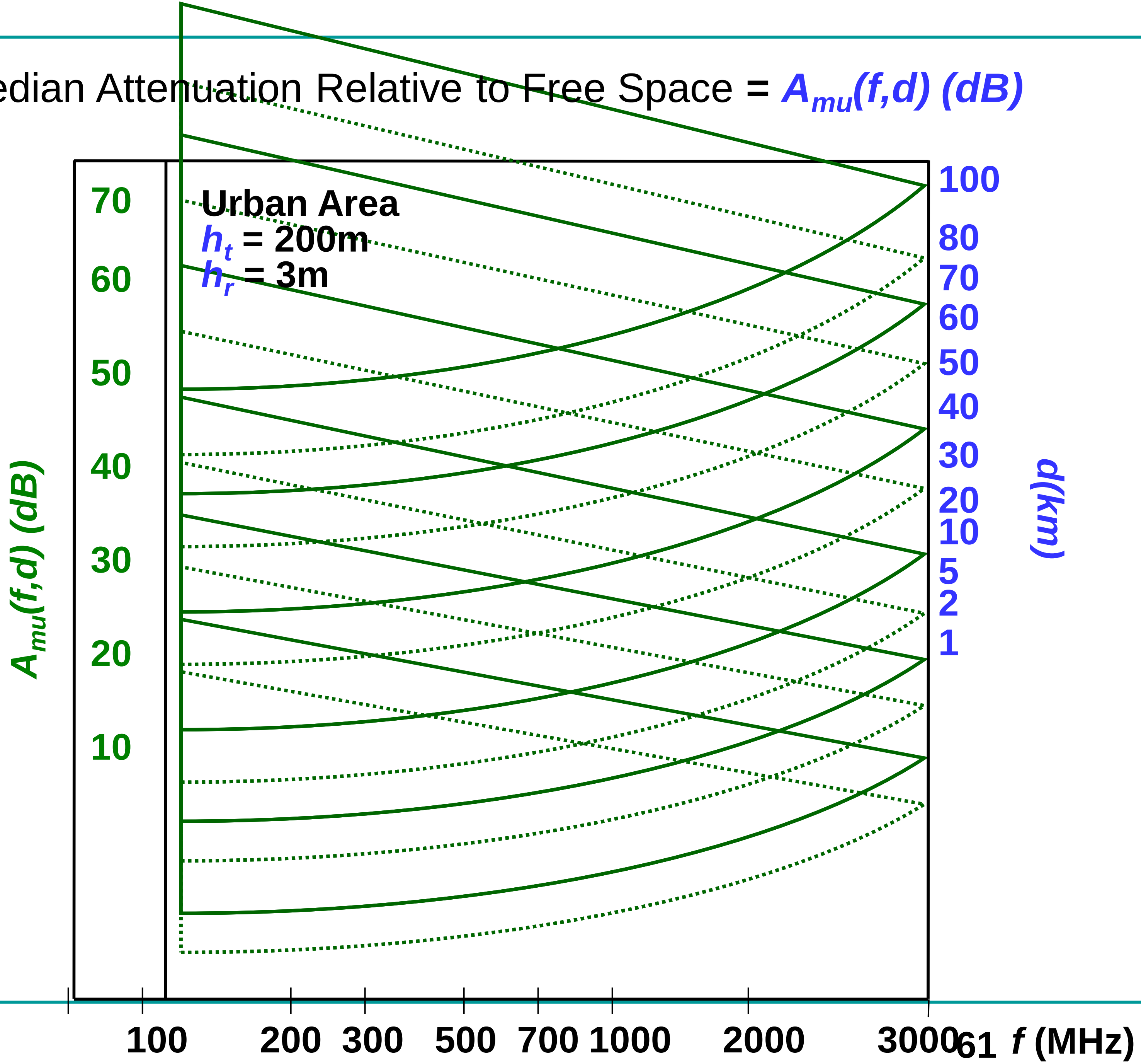
$G(h_{re}) = 10\log\frac{h_{re}}{3}$ $\quad$ $h_{re} \leq 3m$

$G(h_{re}) = 20\log\frac{h_{re}}{3}$ $\quad$ $3m < h_{re} < 10m$

- **model corrected** for

Δh = terrain undulation height, isolated ridge height average terrain slope and mixed land/sea parameter

Median Attenuation Relative to Free Space = $A_{mu}(f,d)$ (dB)



Correction Factor G_{AREA}

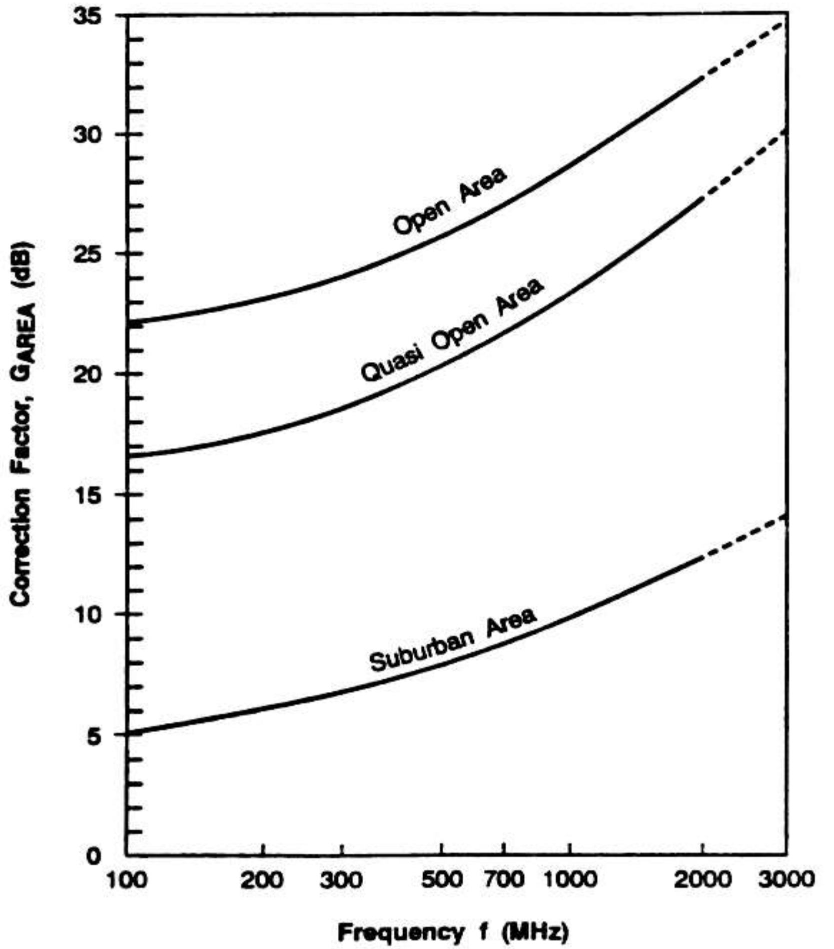


Figure 4.24 Correction factor, G_{AREA}, for different types of terrain [from [Oku68] © IEEE].

Example

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

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

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

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

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

- Most widely used model in Radio frequency.
- Predicting the behavior of cellular communication in built up areas.
- Applicable to the transmission inside cities.
- Suited for point to point and broadcast transmission.
- 150 MHz to 1.5 GHz, Transmission height up to 200m and link distance less than 20 Km.

Hata Model

- Hata transformed Okumura's graphical model into an analytical framework.

- The Hata model for urban areas is given by the empirical formula:

$$L50,_{urban} = 69.55\text{ dB} + 26.16\log(f_c) - 3.82\log(h_t) - a(h_r) + (44.9 - 6.55\log(h_t))\log(d)$$

- Where $L50,_{urban}$ is the median path loss in dB.

- The formula is valid for

$150\text{ MHz} \le f_c \le 1.5\text{GHz}$,

$1\text{ m} \le h_r \le 10\text{m}$, $30\text{m} \le h_t \le 200\text{m}$,

$1\text{km} < d < 20\text{km}$

Hata Model

- The correction factor $a(h_r)$ for mobile antenna height hr for a small or medium-sized city is given by:

$a(h_r) = (1.1 \log f_c - 0.7)h_r - (1.56 \log(f_c) - 0.8)$ dB

- For a large city it is given by

$a(h_r) = 8.29[\log(1.54h_r)]^2 - 1.10$ dB for $f_c <= 300$ MHz

$3.20[\log (11.75h_r)]^2 - 4.97$ dB for $f_c >= 300$ MHz

- To obtain path loss for suburban area the standard Hata model is modified as

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

- For rural areas

$L_{50} = L_{50}(urban) - 4.78\log(f_c)^2 - 18.33\log f_c - 40.98$

Indoor Models

- Indoor Channels are different from traditional channels in two ways

  1. The distances covered are much smaller

  2. The variability of environment is much greater for a much small range of Tx and Rx separation.

- Propagation inside a building is influenced by:

  - Layout of the building
  - Construction materials
  - Building Type: office , Home or factory

Indoor Models

- Indoor models are dominated by the same mechanism as out door models:

  - Reflection, Diffraction and scattering

- Conditions are much more variable

  - Doors/Windows open or not
  - Antenna mounting : desk ceiling etc
  - The levels of floor

- Indoor models are classifies as

  - Line of sight (LOS)
  - Obstructed (OBS) with varying degree of clutter

Indoor Models

- Portable receiver usually experience

  - Rayleigh fading for OBS propagation paths

  - Ricean fading for LOS propagation path

- Indoors models are effected by type of building e.g. Residential buildings, offices, stores and sports area etc.

- Multipath delay spread

  - Building with small amount of metal and hard partition have small delay spread 30 to 60ns

  - Building with large amount of metal and open isles have delay spread up to 300ns

Partition losses (same floor)

- Two types of partitions
  1. hard partitions: Walls of room
  2. Soft partitions : Moveable partitions that donot span to ceiling

- Partitions vary widely in their Physical and electrical properties.

- Path loss depend upon the types of partitions

Partition losses (same floor)

Partition Losses (Same Floor)

Material Type	Loss (dB)	Frequency
All metal partition	26	815 MHz
Concrete Block wall	13	1300 MHz
Empty Cardboard boxes	3 – 6 dB	1300 MHz
Dry Plywood (0.75 inches)	1 dB	9.6 GHz
Dry Plywood (0.75 inches)	4 dB	28.8 GHz

Partitions losses (between floors)

- Partition losses between the two floors depend on

1. External dimension and material used for buildings
2. Types of construction used to create floors
3. External surroundings
4. No of windows used
5. Tinting on the windows

- Floor Attenuation Factor (FAF) increases as we increase the no of floors

Partitions losses (between floors)

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

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

Log distance path loss model

- Path loss can be given as

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

where n is path loss exponent and σ is standard deviation

- n and σ depend on the building type.
- Smaller value of σ indicates better accuracy of path loss model

Log distance path loss model

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

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

Ericsson Multiple Break Point Model

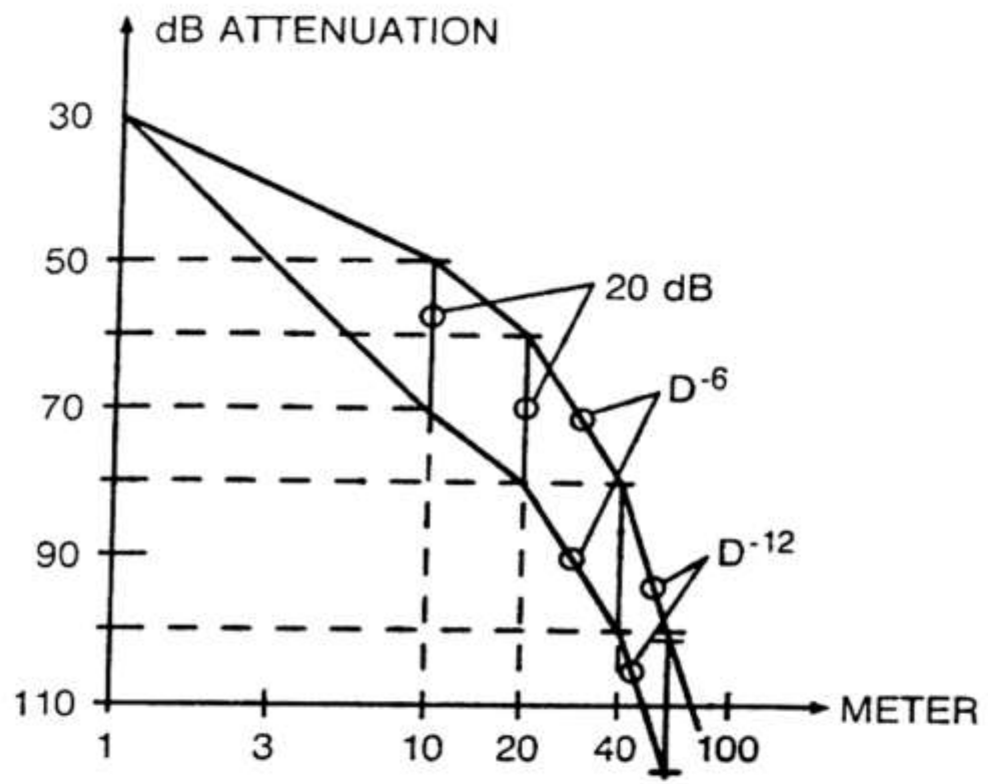


Figure 4.27 Ericsson in-building path loss model [from [Ake88] © IEEE].

Attenuation factor model

- Obtained by measurement in multiple floors building

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

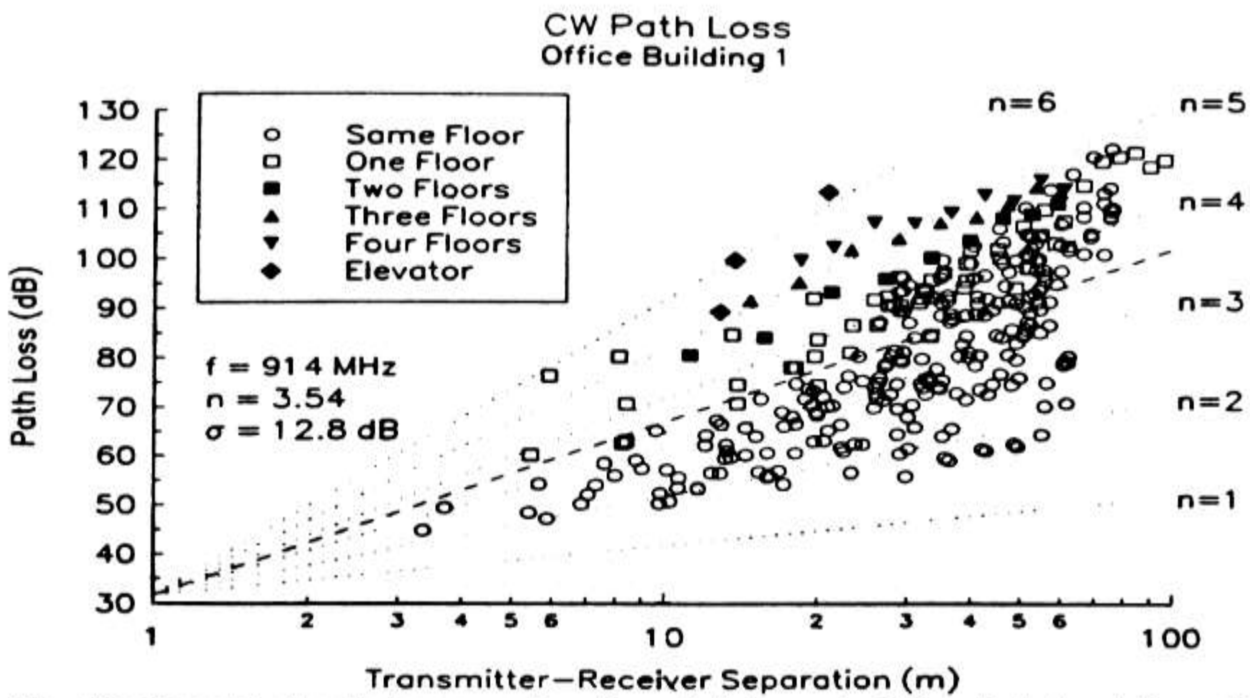


Figure 4.28 Scatter plot of path loss as a function of distance in Office Building 1 [from [Sei92b] © IEEE].

Attenuation factor model

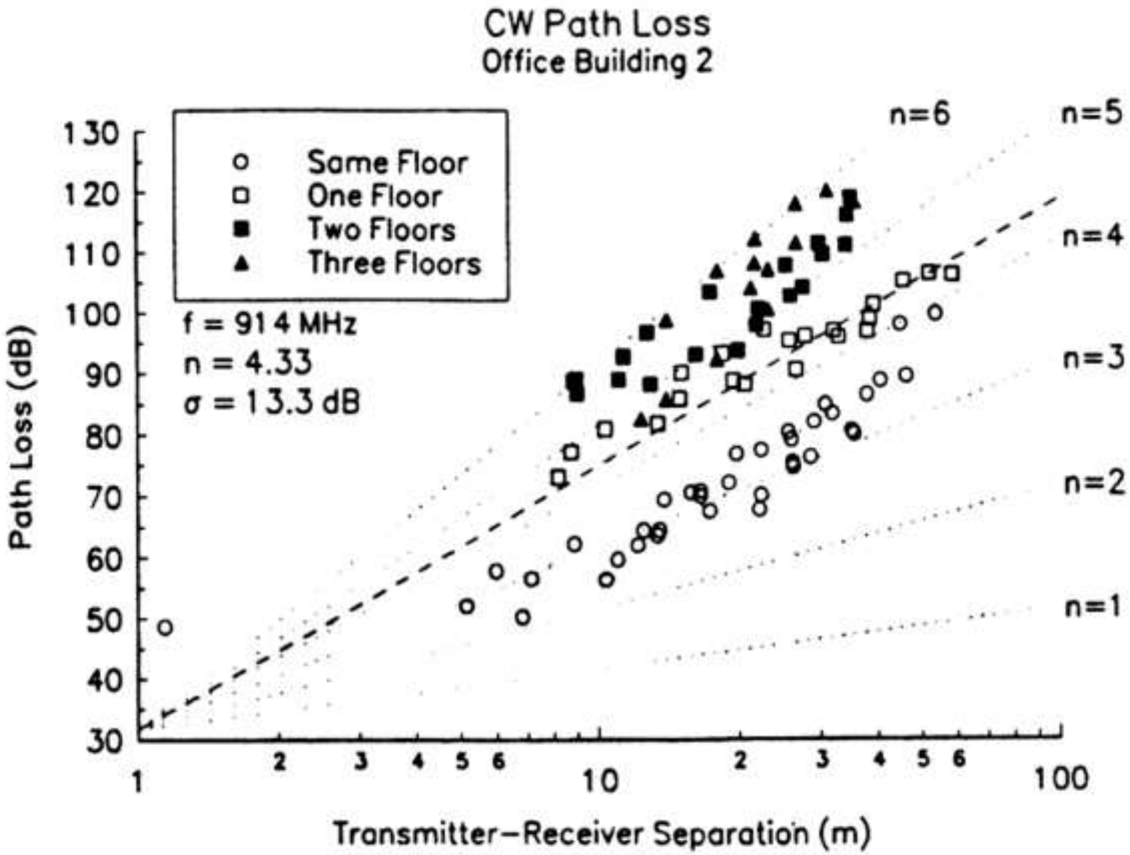


Figure 4.29 Scatter plot of path loss as a function of distance in Office Building 2 [from [Sei92b] © IEEE].

Signal penetration into building

- <u>Effect of frequency</u>

  - Penetration loss decreases with increasing frequency

- <u>Effect of Height</u>

- Penetration loss decreases with the height of building up to some certain height.

  - At lower heights the Urban clutter induces greater attenuation
  - Up to some height attenuation decreases but then again increase after a few floors
  - Increase in attenuation at higher floors is due to the Shadowing effects of adjacent buildings

Large, medium and small scale fading

- Large Scale Fading: Average signal power attenuation/path loss due to motion over large areas.
- Medium scale fading: Local variation in the average signal power around mean average power due to shadowing by local obstructions
- Small scale fading: large variation in the signal power due to small changes in the distance between transmitter and receiver (Also called Rayleigh fading when no LOS available). It is called Rayleigh fading due to the fact that various multipaths at the receiver with random amplitude & delay add up together to render rayleigh PDF for total signal.

Cause of Multipath Fading

- Fading : Fluctuation in the received signal power due to
  - Variations in the received singal amplitude (Different objects present on radio signal path produce attenuation of it's power as they can scatter or absorb part of the signal power, thus producing a variation of the amplitude
  - Variations in the signal phase
  - Variations in the received signal angle of arrival (different paths travelling different distances may have different phases & angle of arrival)

Causes of Multipath fading Cont..

- Reflections and diffraction from object create many different EM waves which are received in mobile antenna. These waves usually come from many different directions and delay varies.
- In the receiver, the waves are added either constructively or destructively and create a Rx signal which may very rapidly in phase and amplitude depending on the local objects and how mobile moves

Practical examples of small scale multipath fading

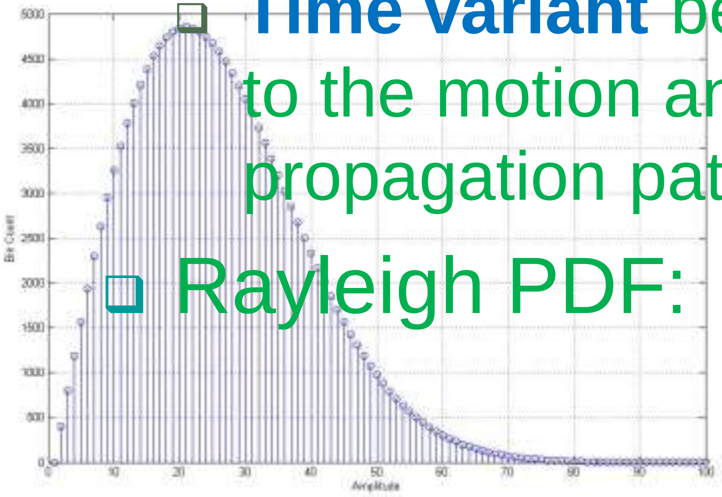
- Common examples of multipath fading are
  - temporary failure of communication due to a severe drop in the channel signal to noise ratio (You may have also experienced this. And you moved a steps away & noted that reception is better. It is due to small scale fading effects. ☺)
  - FM radio transmission experiencing intermittent loss of broadcast when away from station.

Multipath Fading- Most difficult

- Fades of 40 dB or more below local average level are frequent, with successive nulls occurring every half wavelength or so
- Referred to as Rayleigh Fading

Rayleigh Fading Mechanism

- Rayleigh fading manifests in two mechanism
  - **Time spreading** due to multipath (time dispersion)
  - **Time variant** behaviour of the channel due to the motion and subsequent changes in propagation paths
- Rayleigh PDF:



Rayleigh Fading

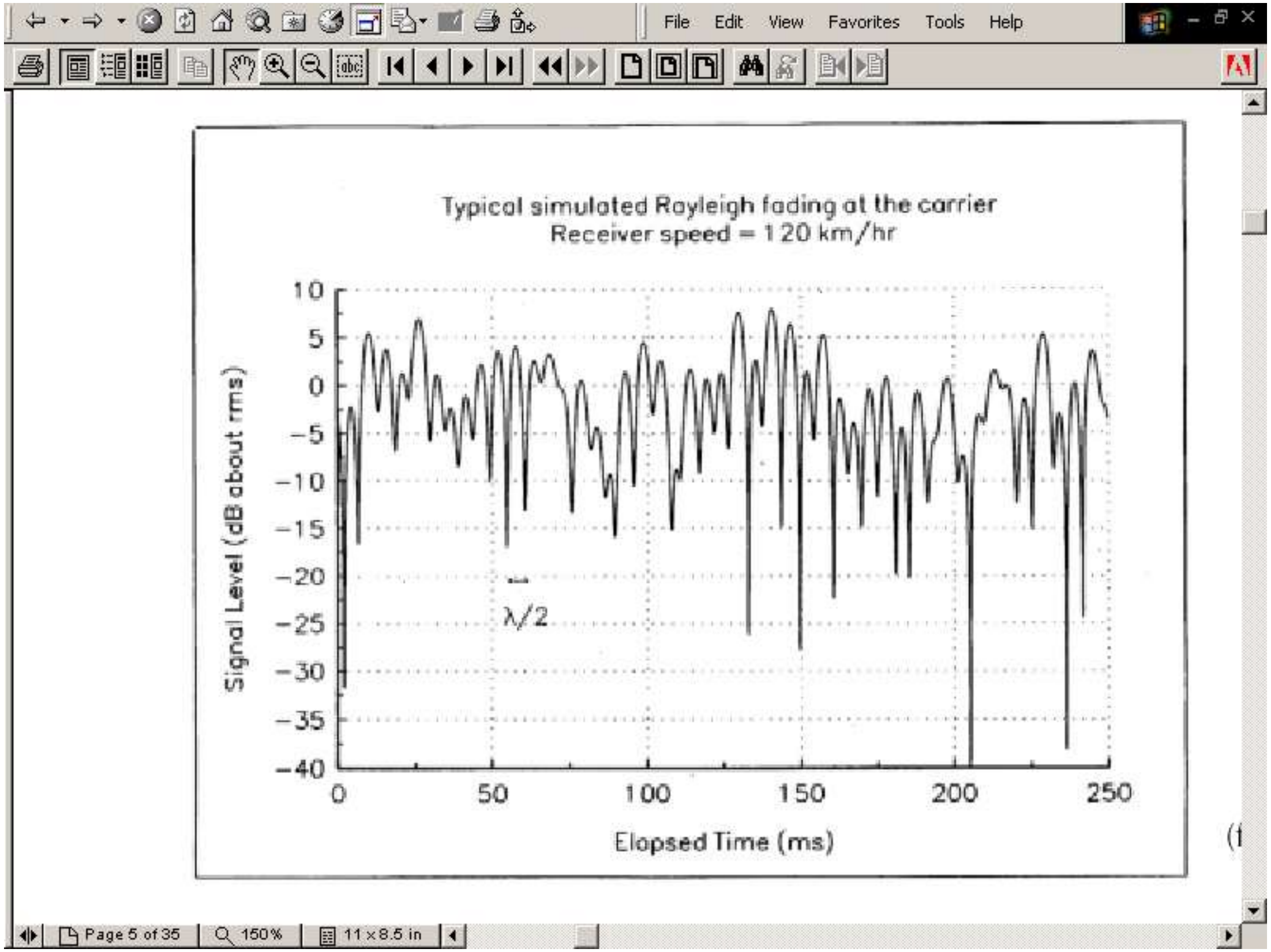
- The Rayleigh pdf is

$$p(r) = \begin{cases} \frac{r}{\sigma^2} \exp\left(-\frac{r^2}{2\sigma^2}\right) & for\ r > 0 \\ 0 & otherwise \end{cases}$$

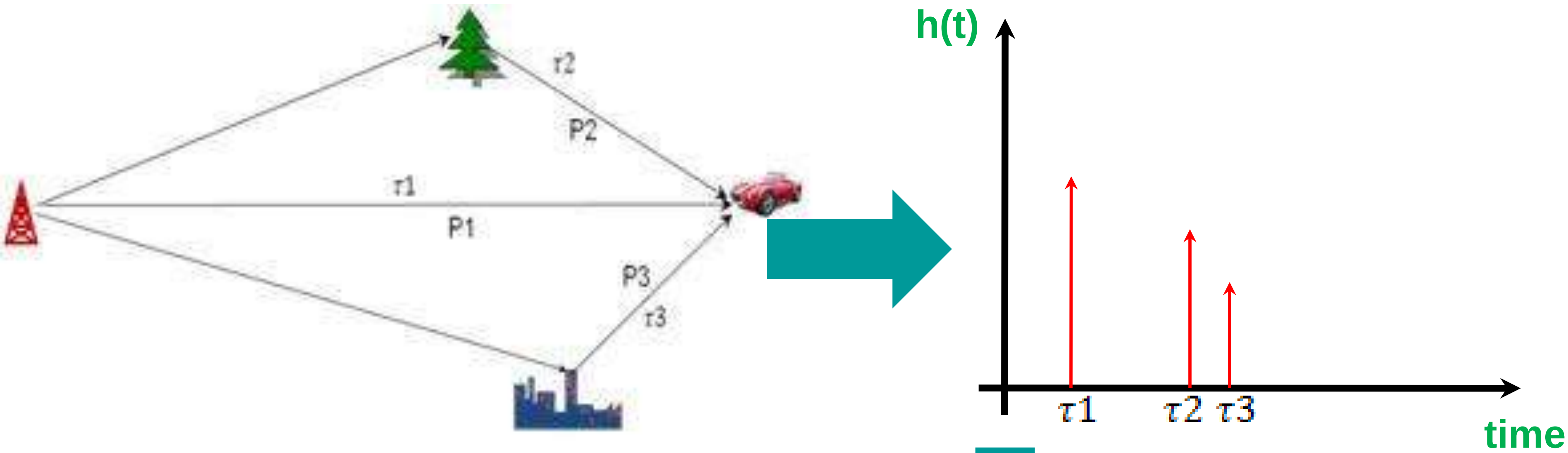
$Where\ r\ is\ the\ envelope\ amplitude\ of\ Rx\ signal\ \&$
$2\sigma^2\ is\ the\ power\ of\ the\ signal$

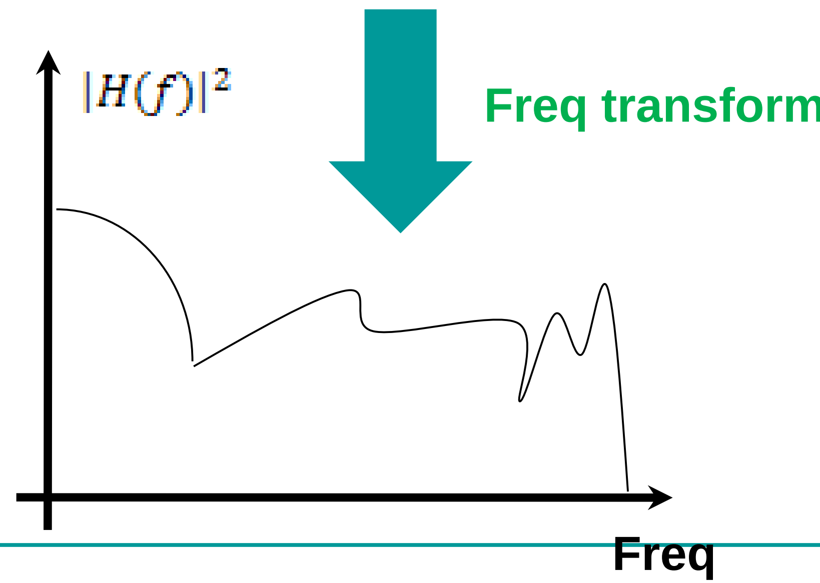
With Rayleigh Fading

Rayleigh Fading waveform envelope



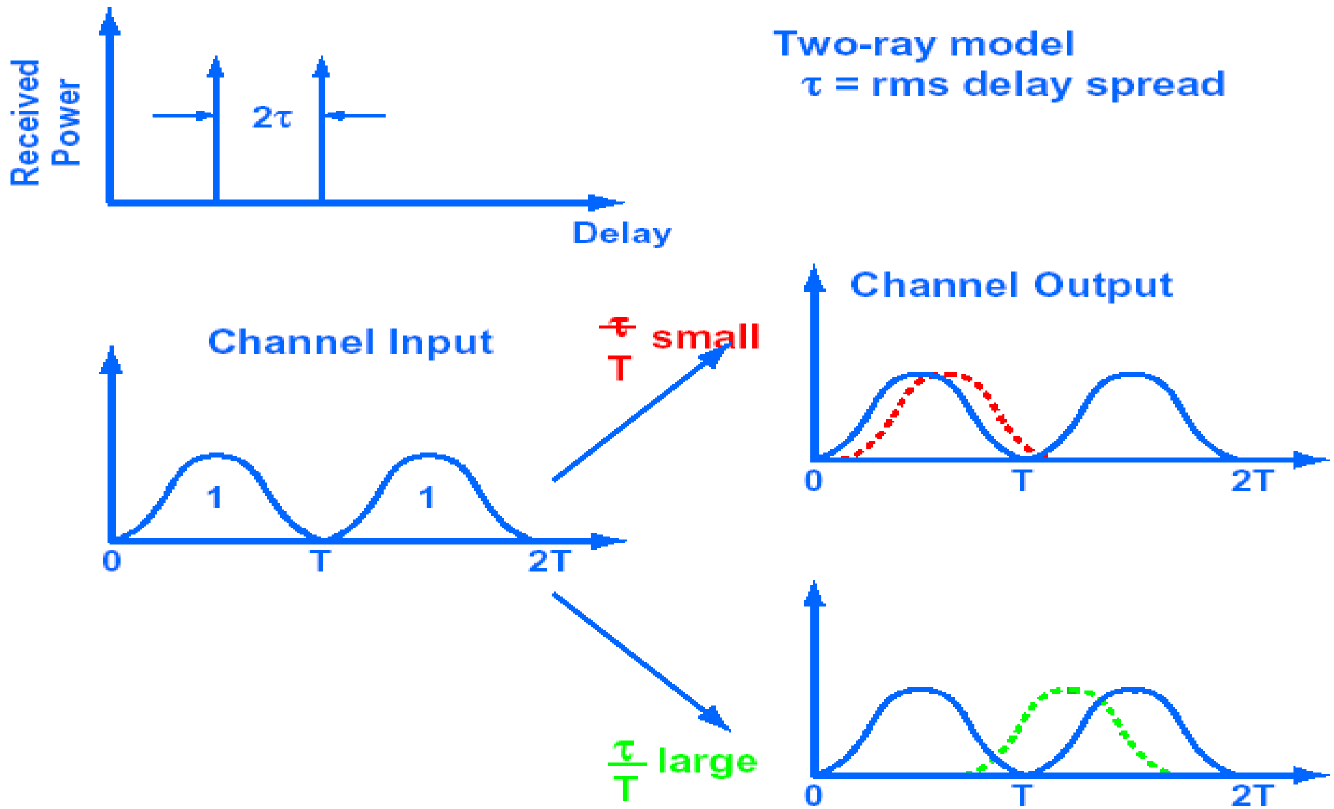
Time Dispersion phenomenon





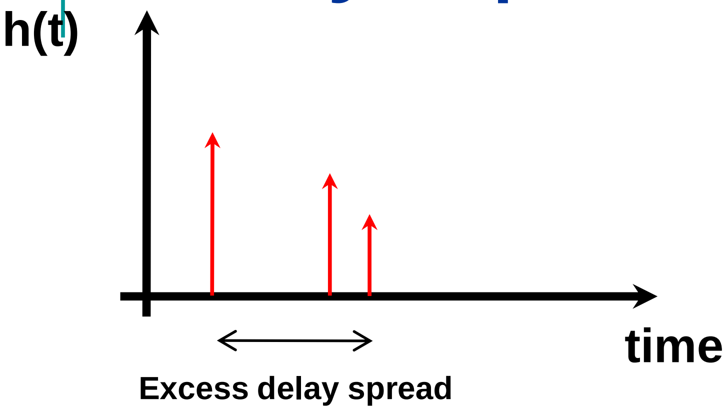
Different frequencies suffer different attenuation

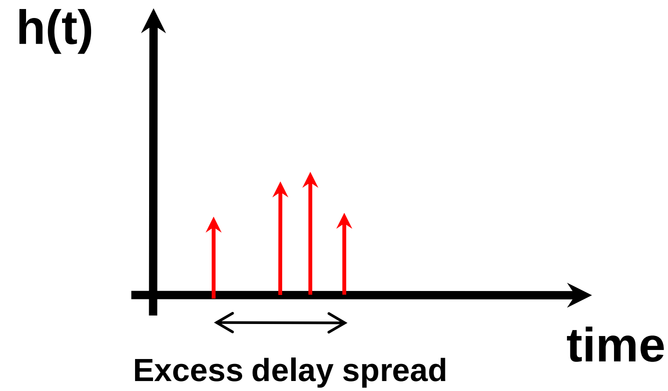
Delay Spread –Time Domain interpretation



- $\frac{\tau}{T}$ small ⇒ negligible intersymbol interference
- $\frac{\tau}{T}$ large ⇒ significant intersymbol interference, which causes an irreducible error floor

Delay Spread





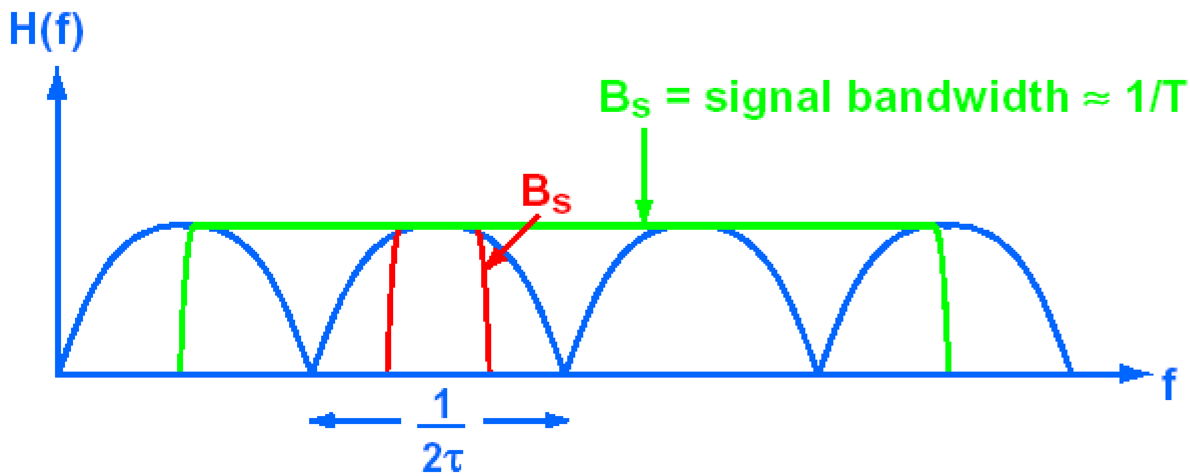
- Multiple impulses of varying power correspond to various multipaths. This time dispersion is also referred to as multipath delay spread.
- Delay between first significant path & last significant paths is loosely termed as channel excess delay spread.
- Two totally different channels can have same excess delay spread.
- A better measure of delay spread is rms delay spread σ_τ
  - L is the number of paths & β_i is the amplitude of the path i arriving at time τ_i

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

$$\overline{\tau^2} = \frac{\sum_{i=1}^{L} \tau_i^2 \beta_i^2}{\sum_{i=1}^{L} \beta_i^2}$$

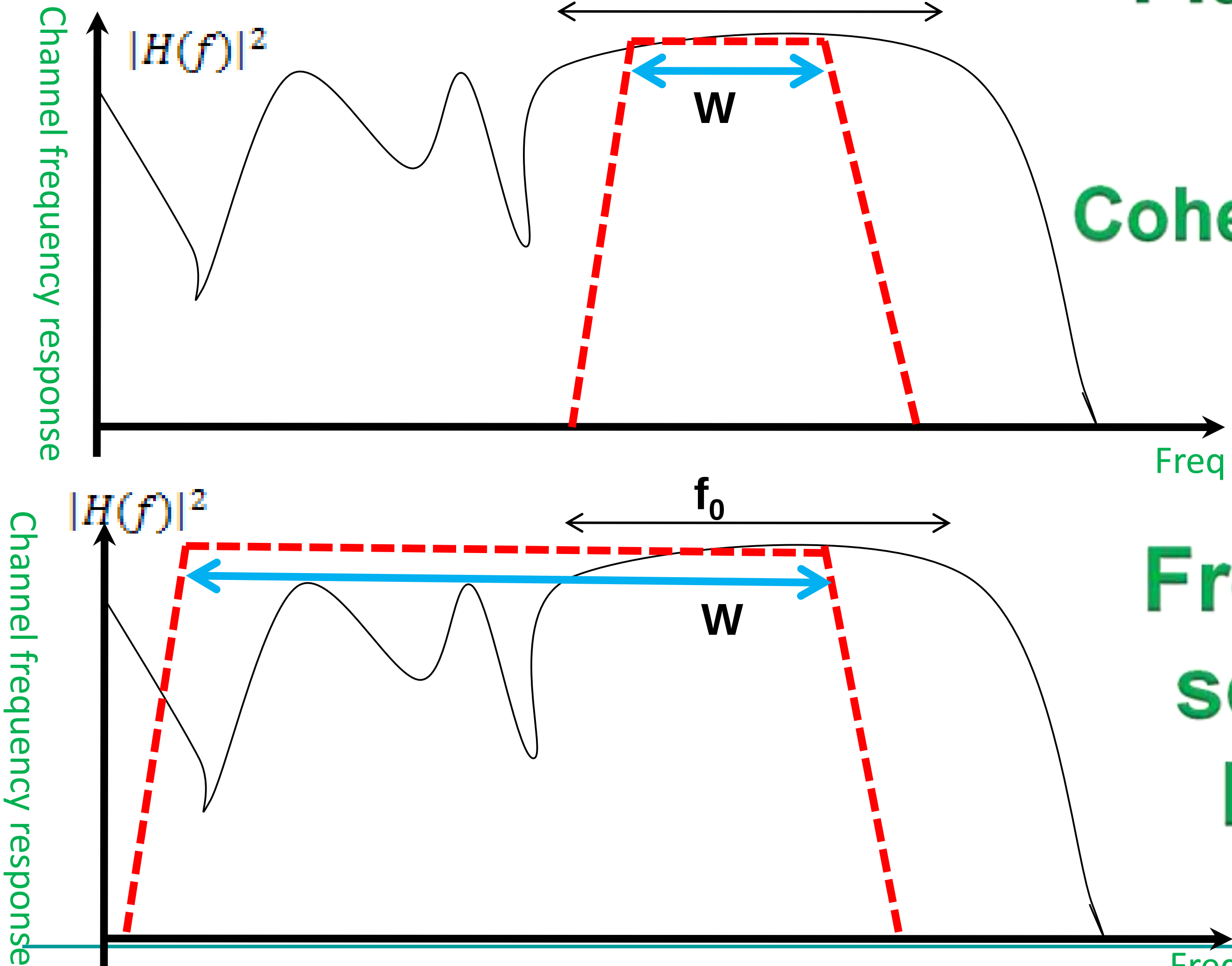
$\overline{\tau^2}$ is the second moment

Delay Spread- Freq. Domain Interpretation



- $\frac{\tau}{T}$ small ➡ flat fading
- $\frac{\tau}{T}$ large ➡ frequency-selective fading

Time spreading : Coherence Bandwidth



Flat Fading

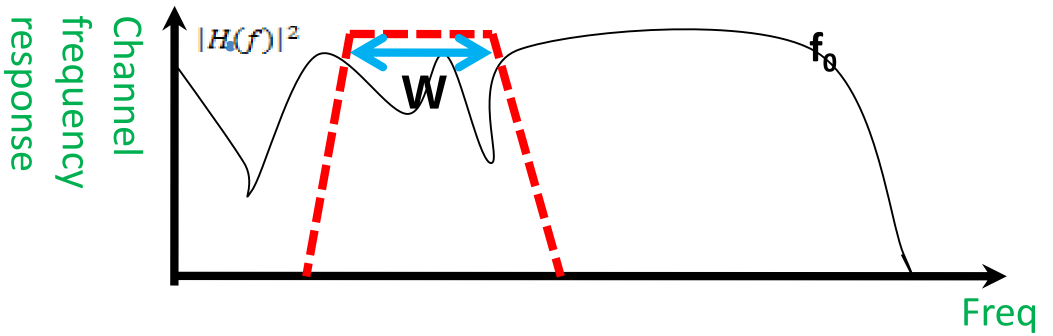
$W < f_0$

Coherence BW = f_0

Frequency selective Fading

$W > f_0$

More on flat fading



Condition f0 > W does not guarantee flat fading. As shown above, frequency nulls (frequency selective fading) may be there occasionally even though f0 > W.

Similarly, frequency selective fading channel may also show flat fading sometimes.

Bit Rate Limitations by Delay Spread

- QPSK modulation
- Bit error rate is 10^{-4}

	τ	Maximum Bit Rate
Mobile (rural)	25 μs	8 kbps
Mobile (city)	2.5 μs	80 kbps
Microcells	500 ns	400 kbps
Large Building	100 ns	2 Mbps

Coherence Bandwidth and delay spread

- There is no exact relationship between Coherence bandwidth and delay spread. For at least 0.9 correlation for channel's complex frequency transfer function, Coherence bandwidth f_0 is approximated by following relation: $$f_0 \approx \frac{1}{50\sigma_\tau}$$ **Where σ_τ is r.m.s. delay spread**

- For dense scatterer model which is useful for urban surroundings, coherence bandwidth is defined as assuming at least 0.5 correlation: $$f_0 \approx \frac{0.276}{\sigma_\tau}$$

- Another popular approximation assuming at least 0.5 correlation: $$f_0 \approx \frac{1}{5\sigma_\tau}$$

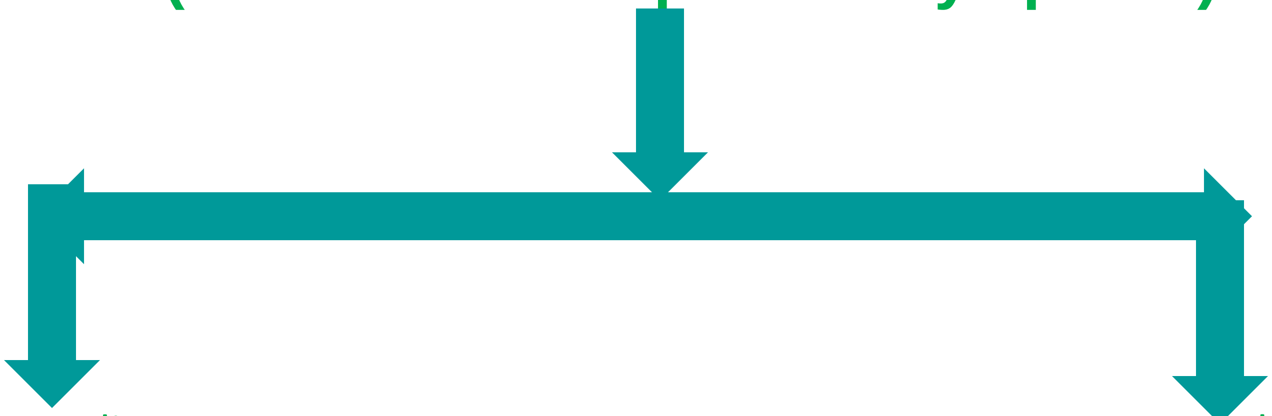
Effects of Flat & frequency selective fading

- Flat fading
  - Reduces SNR forcing various mitigation techniques to handle that. Not such a bad thing.
- Frequency selecting fading
  - ISI distortion (need equalizer in receiver)
  - Pulse mutilation
  - Irreducible BER

Summary of Time dispersion

Small scale fading

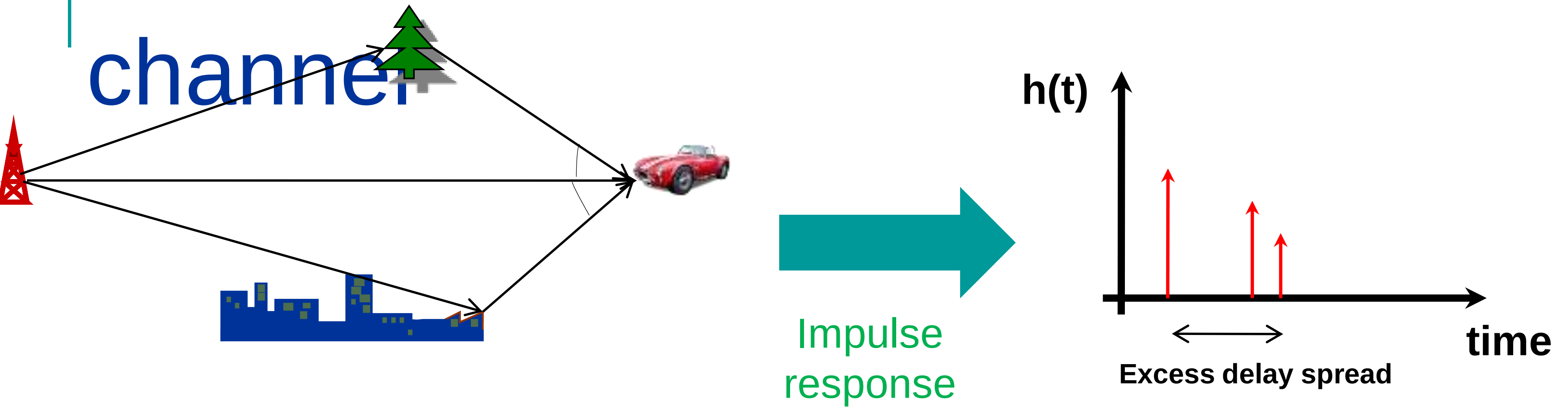
(based on multipath delay spread)

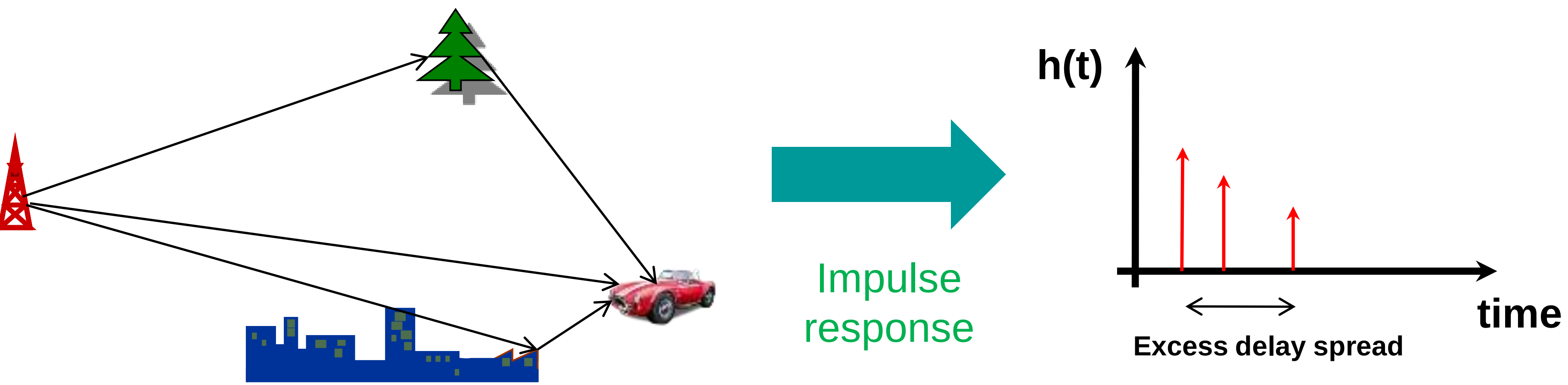


Flat Fading
BW of signal < BW of channel
Or
Delay Spread < Symbol period

Frequency selective Fading
BW of signal > BW of channel
Or
Delay Spread > Symbol period

Time variant behavior of the channel





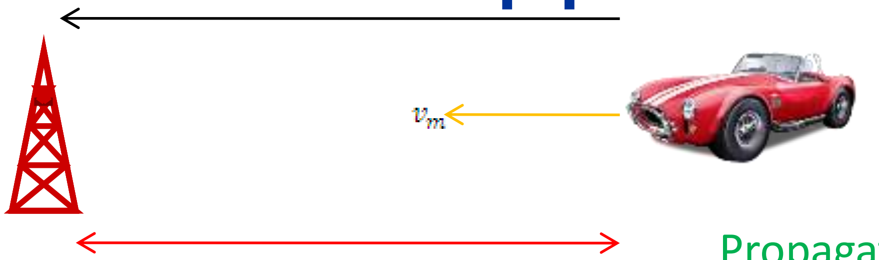
Relative movement between transmitter and receiver or objects between those causes variation in channel's characteristics over time. This happens due to propagation path change over time. Relative movement also creates frequency spreading due to Doppler effect

Time Variance

- Variance in channel conditions over time is an important factor when designing a mobile communication system.
- If fast variations happen, it can lead to severe pulse distortion and loss of SNR subsequently causing irreducible BER.

t)

Basic Doppler effect



$$\tau(t) = \frac{d(t)}{c} = \frac{d0 - v_m.t}{c} = \tau 0 - \frac{v_m.t}{c}$$

c is the light velocity and v_m is the car speed

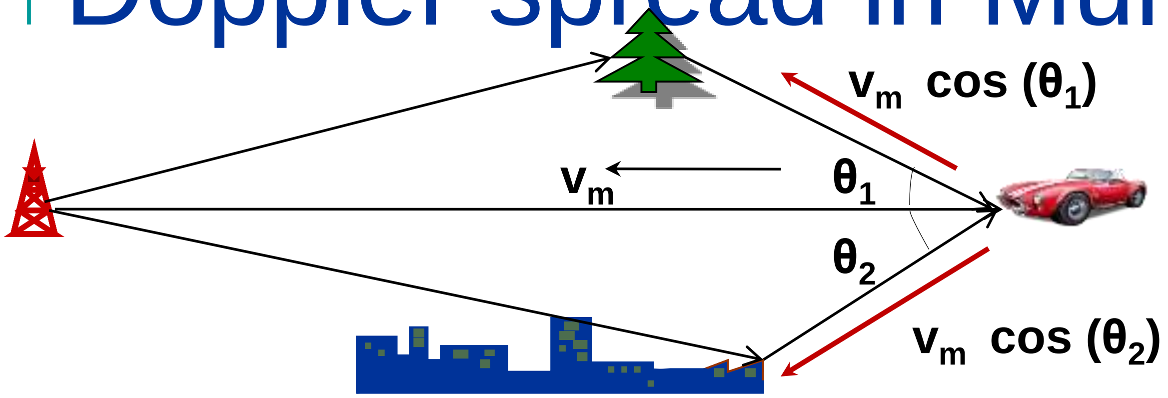
Propagation time is a function of time due to mobile car.

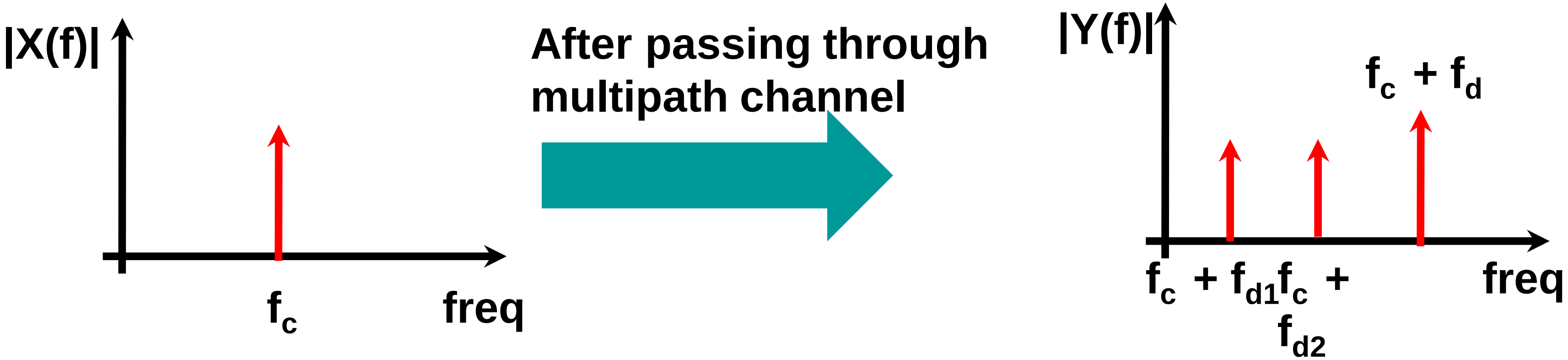
$$transmitted\ signal: \cos(2\pi f_c t)$$

$$Received\ signal: \cos[2\pi f_c(t - \tau(t))]$$
$$= \cos\left[2\pi f_c\left(t - \tau 0 + \frac{v_m.t}{c}\right)\right]$$
$$= \cos\,[2\pi(f_c + f_d)t - \emptyset]$$

$$Doppler\ frequency\ shift\ f_d = \frac{v_m}{c}.f_c$$

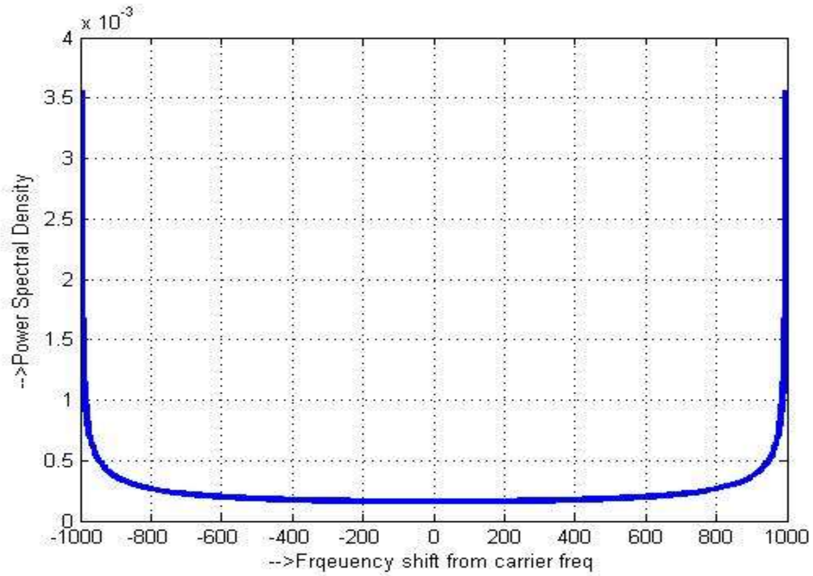
Doppler spread in Multipath





Due to multipaths, a single sinusoid by base station is perceived as summation of 3 sinusoids f_c+f_{d1}, f_c+f_{d2} and f_c+f_d , where f_d is maximum doppler frequency = $f_c*(v_m/d)$. Due to different arrivals of angle due to multipaths, perceived velocity is different for multipaths.

Doppler Spectrum



Imagine now multiple paths with different angles of arrival causing amagamalation of various frequencies between $f_c + f_d$ & $f_c - f_d$.

A popular model assumes that distribution of angle of arrival is distributed uniformly between 0 & 2π which leads to following spectrum

$$D(f) = \frac{1}{2\pi f_d \sqrt{1 - (\frac{f}{f_d})^2}}$$

This is called classical Doppler spectrum & shows how a single sinusoid ends up having a broad spectrum due to multipath & relative motion between Tx and Rx.

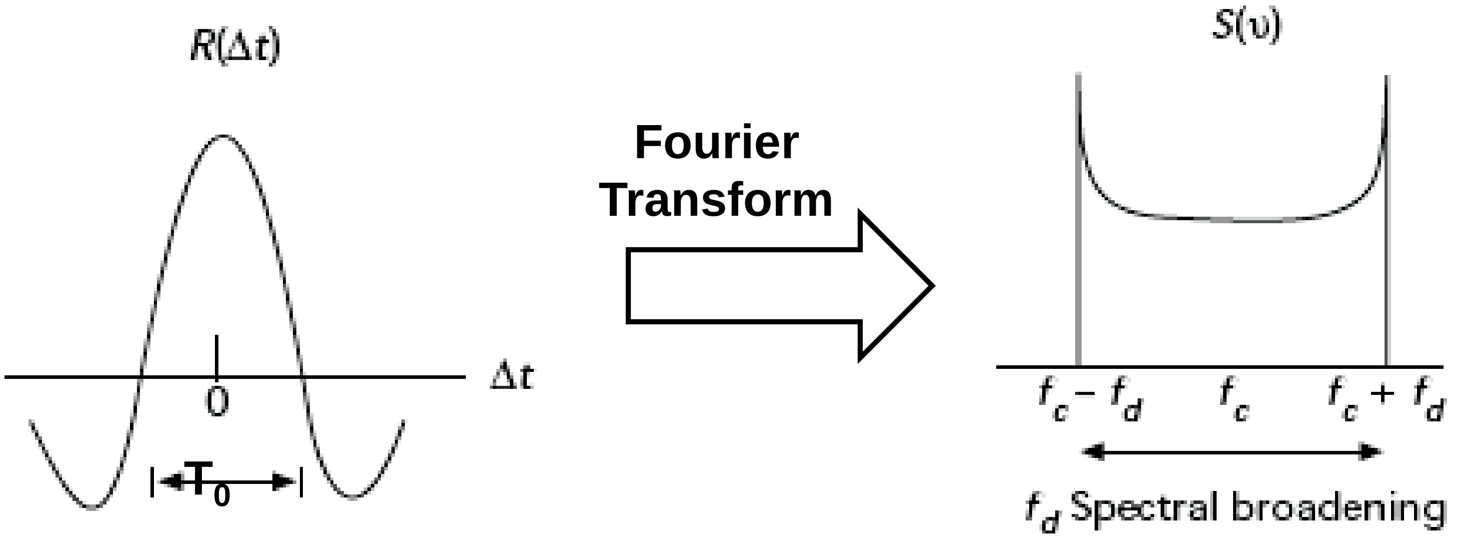
Time variant Channel: Coherence Time

- **Maximum doppler frequency is an important measure of time variance of channel characteristics. It depends on relative speed of any movement between Tx & Rx and the carrier frequency**
- **Coherence time: Approximate time duration over which the channel's response remains invariant**

$$T_0 \approx \frac{1}{f_d}$$

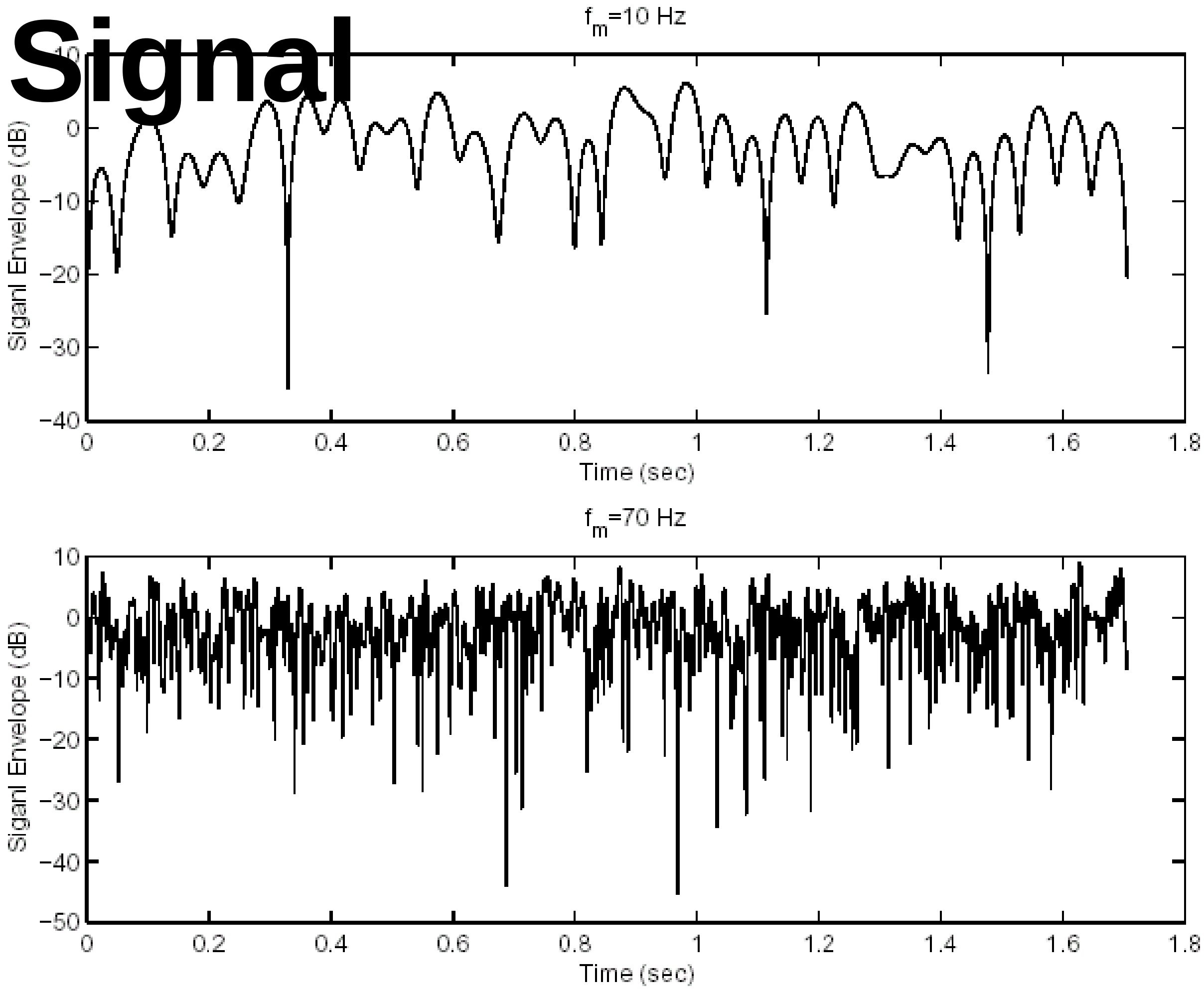
- **Where is Maximum Doppler Frequency**

Frequency Dual



Function $R(\Delta t)$ denotes space time correlation for the channel response to a sinusoid . So this indicates the amount of correlation between two sinusoids sent at different times t_1 & t_2 .

Waveform of Rayleigh Fading Signal



Rayleigh fading envelopes of signals with different maximum Doppler frequencies at carrier frequency of 800 MHz

Time Variance : Fast Fading

Fast Fading :

$T_0 < T_s$ Where T_s : Transmitted Symbol time

Or

$f_d > W$ Where W: Transmitted bandwidth

Above relationship means that channel changes drastically many times while a symbol is propagating;

Only highly mobile systems (~500 Km/Hr) will have fd ~1 kHz so systems having signalling rate of that order will be fast fading.

Impact of fast fading:

- Severe distortion of baseband pulse leading to detection problems
- Loss in SNR
- Synchronization problems (e.g. Failure of PLL)

Time variance: Slow Fading

Slow Fading :

$T_0 > T_s$ where T_s : Transmitted Symbol time

Or

$f_d < W$ where W: Transmitted bandwidth

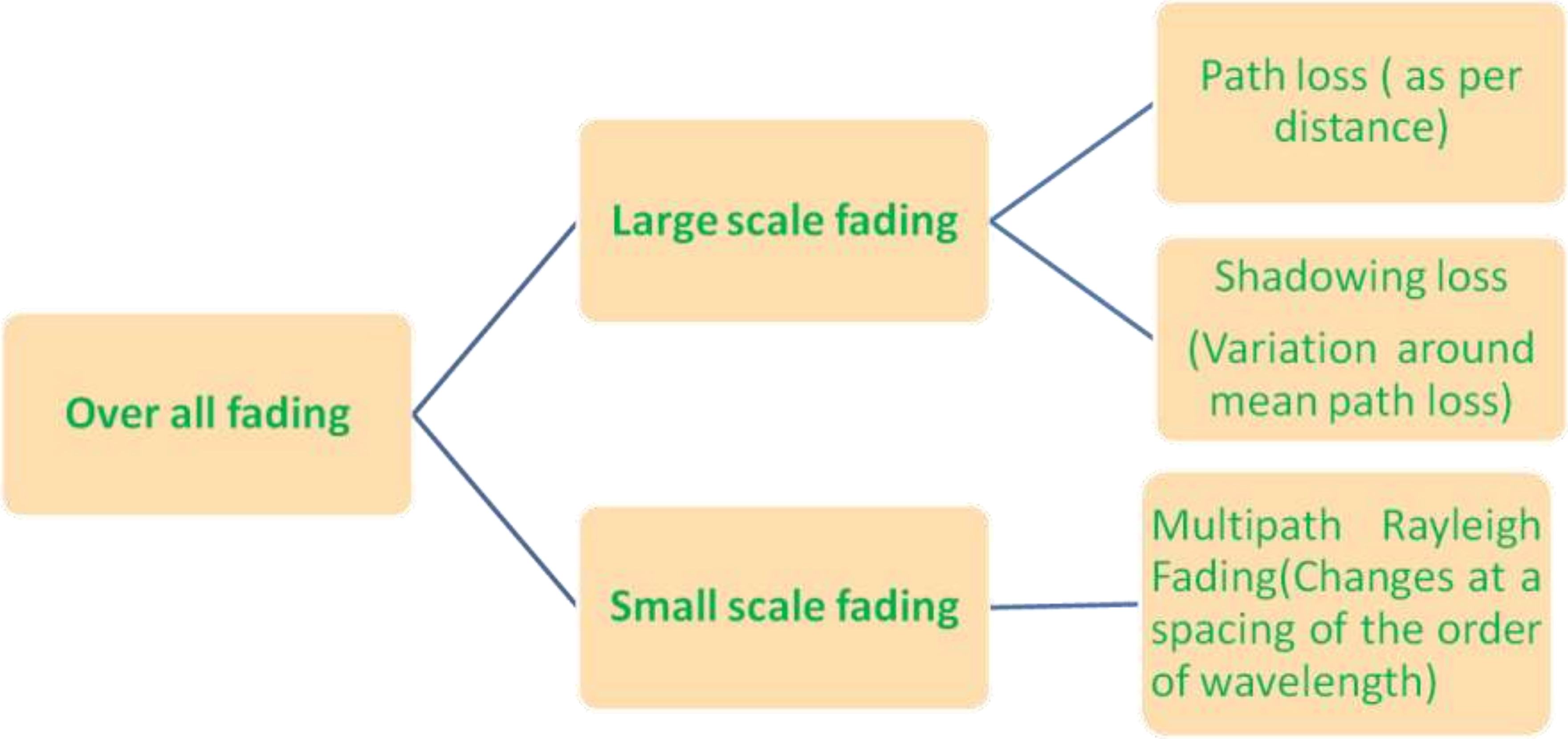
Above relationship means that channel does not change drastically during symbol duration

Most of the modern communication systems are slow fading channels

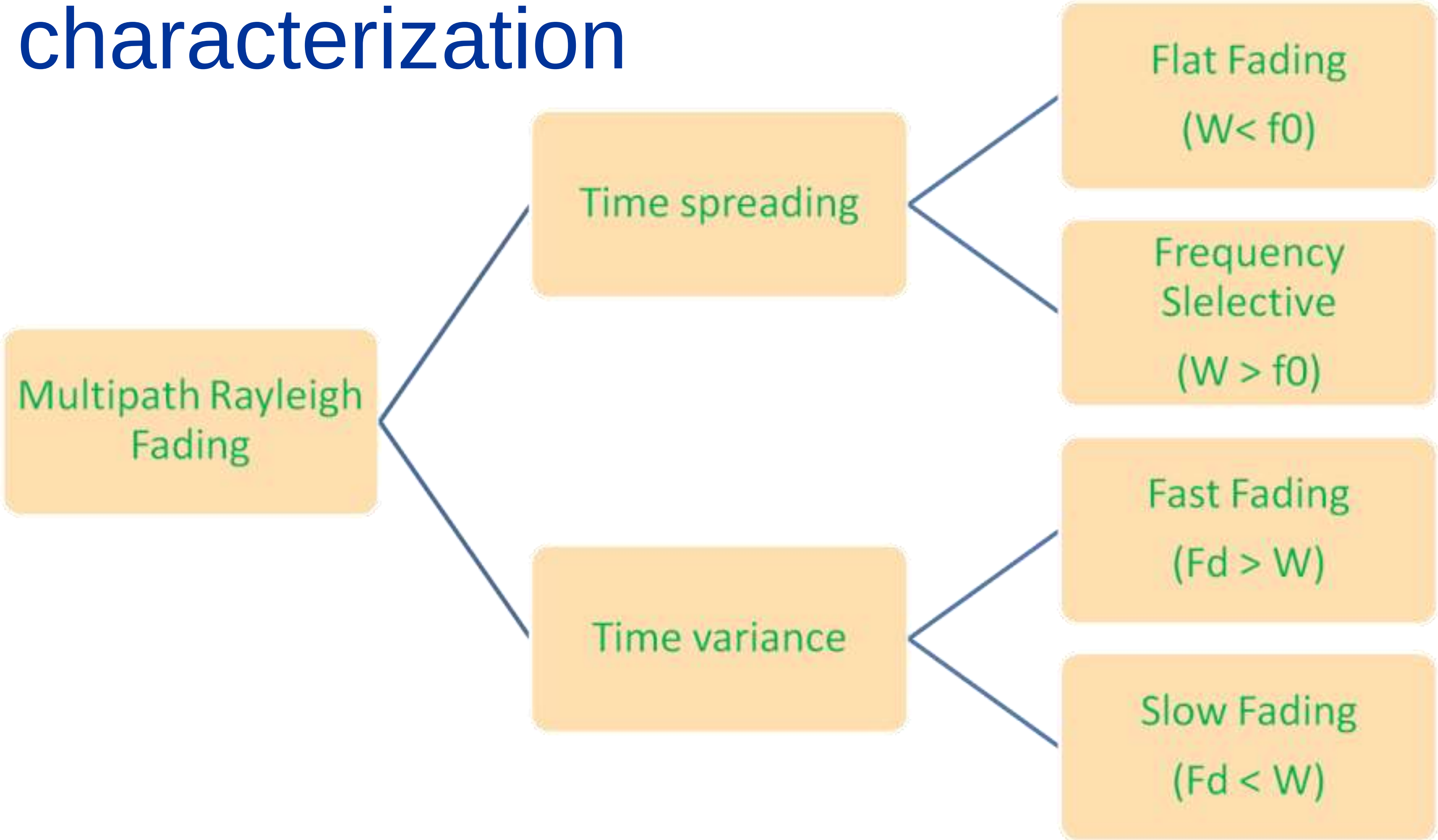
Impact of fast fading:

- Loss in SNR

Summary of Overall Fading



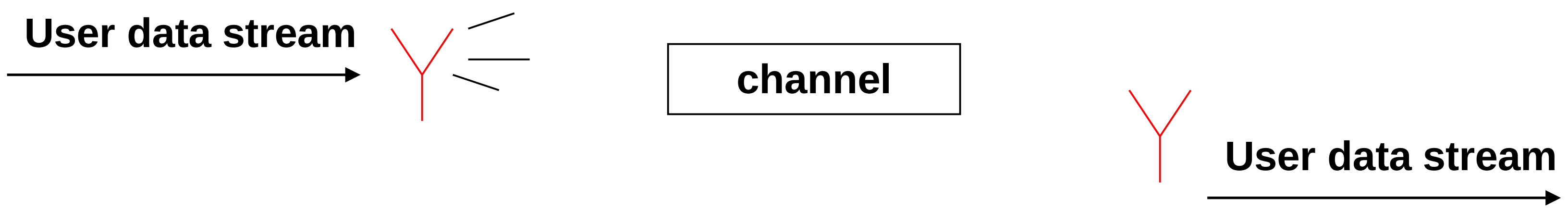
Summary of Multipath Fading characterization



UNIT V
Multiple Antenna Techniques

Antenna Configurations

- Single-Input-Single-Output (SISO) antenna system



- Theoretically, the 1Gbps barrier can be achieved using this configuration if you are allowed to use much power and as much BW as you so please!
- Extensive research has been done on SISO under power and BW constraints. A combination a smart *modulation*, *coding* and *multiplexing* techniques have yielded good results but far from the 1Gbps barrier

MIMO Antenna Configuration

- Use multiple transmit and multiple receive antennas for a single user

- Now this system promises enormous data rates!

Data Units

Will use the following terms loosely and interchangeably,

- Bits (lowest level): +1 and -1
- Symbols (intermediate): A group of bits
- Packets (highest level): Lots and lots of symbols

Shannon's Capacity (C)

- **Given a unit of BW (Hz), the max error-free transmission rate is**

  $$C = \log_2(1+SNR) \text{ bits/s/Hz}$$

- **Define**

  R: data rate (bits/symbol)

  R_S: symbol rate (symbols/second)

  w: allotted BW (Hz)

- **Spectral Efficiency is defined as the number of bits transmitted per second per Hz**

  $$\frac{R \times R_S}{W} \text{ bits/s/Hz}$$

  As a result of filtering/signal reconstruction requirements, $R_S \leq W$. Hence Spectral Efficiency = R if $R_S = W$

- **If I transmit data at a rate of $R \leq C$, I can achieve an arbitrarily low P_e**

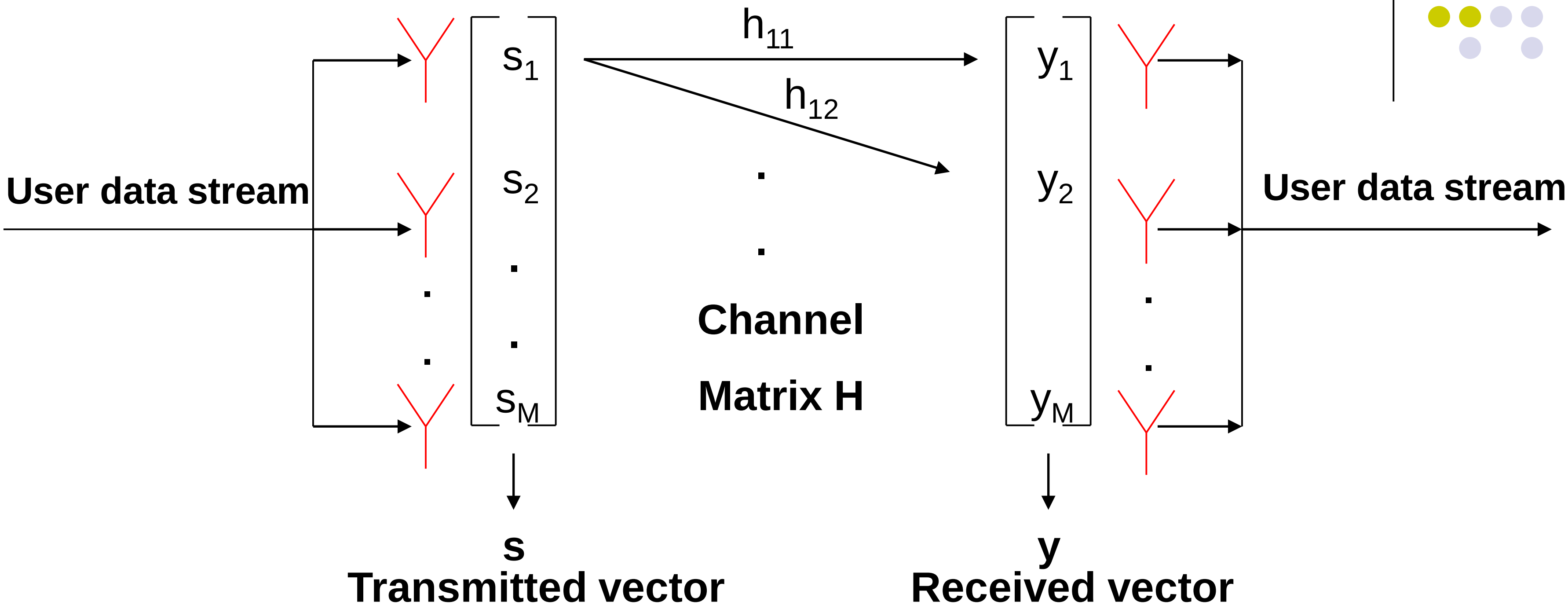
Spectral Efficiency

- Spectral efficiencies of some widely used modulation schemes

Scheme	b/s/Hz
BPSK	1
QPSK	2
16-QAM	4
64-QAM	6

- The Whole point: Given an acceptable P_e , realistic power and BW limits, MIMO Systems using smart modulation schemes provide much higher spectral efficiencies than traditional SISO

MIMO System Model

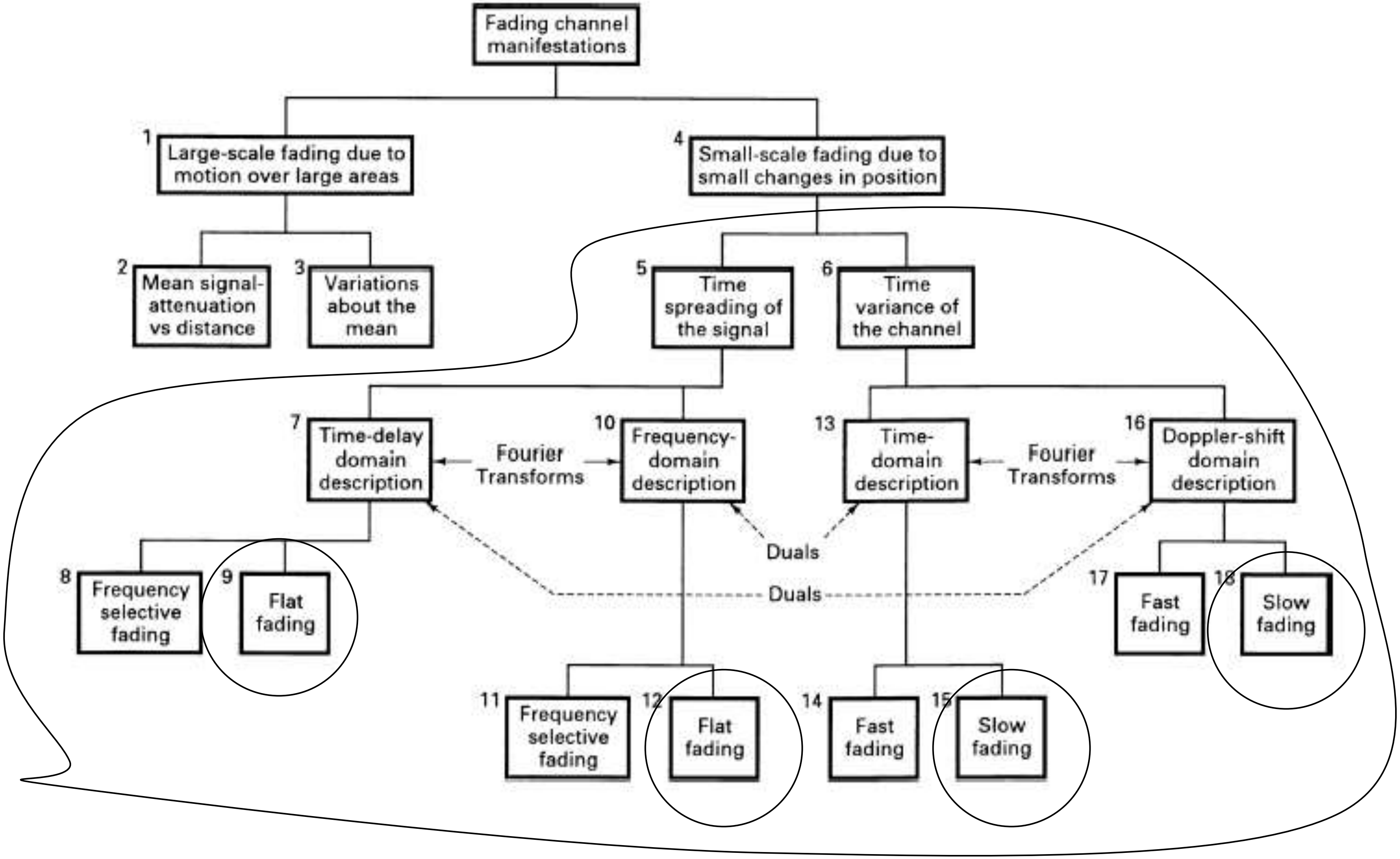


$$\mathbf{y} = \mathbf{H}\mathbf{s} + \mathbf{n}$$

Where $\mathbf{H} = M_R \overbrace{\begin{bmatrix} h_{11} & h_{21} & \cdots\cdots & h_{M1} \\ h_{12} & h_{22} & \cdots\cdots & h_{M2} \\ . & . & \cdots\cdots & . \\ h_{1M} & h_{2M} & \cdots\cdots & h_{MM} \end{bmatrix}}^{M_T}$

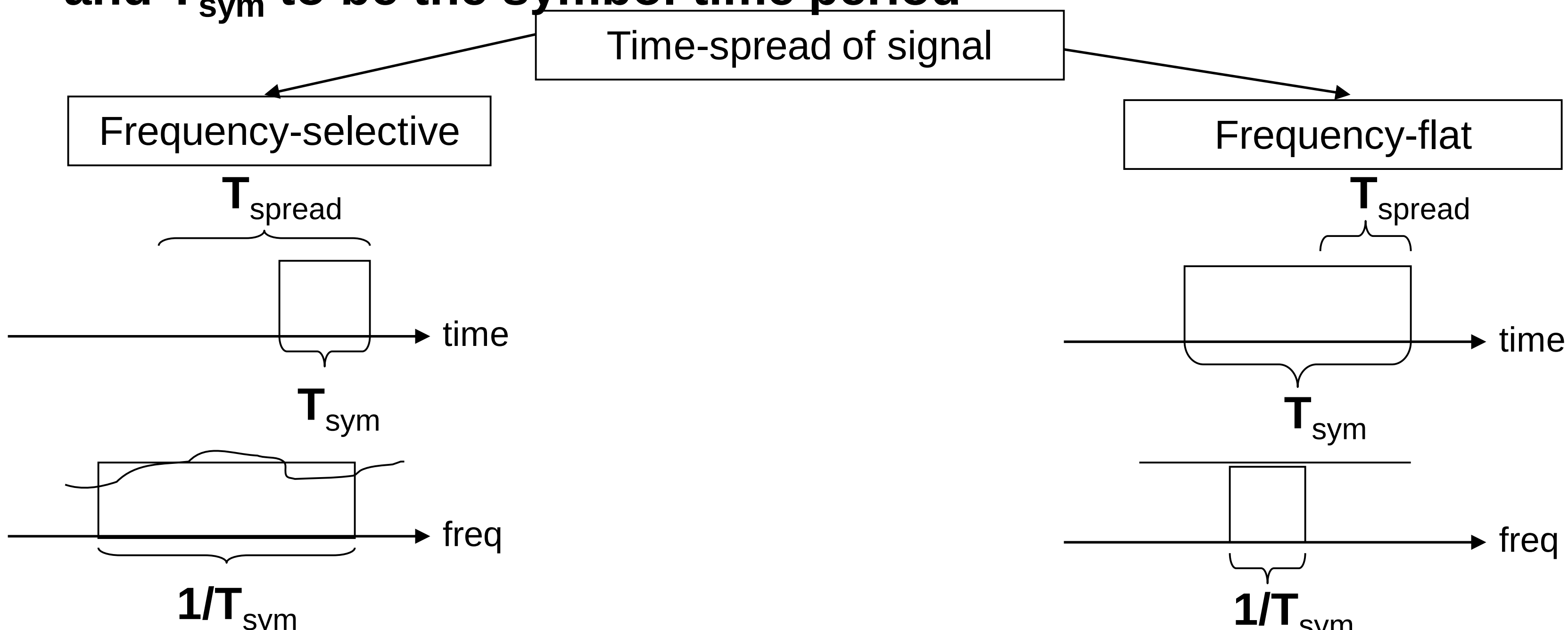
h_{ij} is a Complex Gaussian random variable that models fading gain between the ith transmit and jth receive antenna

Types of Channels



Fading Channels

- *Fading* **refers to changes in signal amplitude and phase caused by the channel as it makes its way to the receiver**
- **Define T_{spread} to be the time at which the last reflection arrives and T_{sym} to be the symbol time period**



Occurs for wideband signals (small T_{sym})

TOUGH TO DEAL IT!

Occurs for narrowband signals (large T_{sym})

EASIER! Fading gain is complex Gaussian

Multipaths NOT resolvable

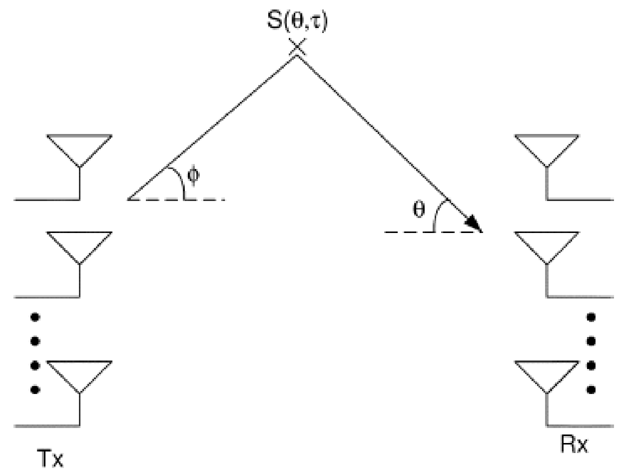
Channel Matrix H

- **In addition, assume *slow fading***
- **MIMO Channel Response**

$$\mathbf{H}(\tau,t)=\begin{bmatrix} h_{1,1}(\tau,t) & h_{1,2}(\tau,t) & \cdots & h_{1,M_T}(\tau,t) \\ h_{2,1}(\tau,t) & h_{2,2}(\tau,t) & \cdots & h_{2,M_T}(\tau,t) \\ \vdots & \vdots & \ddots & \vdots \\ h_{M_R,1}(\tau,t) & h_{M_R,2}(\tau,t) & \cdots & h_{M_R,M_T}(\tau,t) \end{bmatrix}.$$

Channel Time-variance

Time-spread



- **Taking into account slow fading, the MIMO channel impulse response is constructed as,**

$$\mathbf{H}(\tau) = \int_{-\pi}^{\pi}\int_{0}^{\tau_{\max}} S(\theta,\tau')\mathbf{a}(\theta)\mathbf{b}^T(\phi)g(\tau-\tau')d\tau' d\theta$$

- **Because of flat fading, it becomes,**

$$\mathbf{H}(\tau)=\left(\int_{-\pi}^{\pi}\int_{0}^{\tau_{\max}} S(\theta,\tau')\mathbf{a}(\theta)\mathbf{b}^T(\phi)d\tau' d\theta\right)g(\tau)=\mathbf{H}\,g(\tau)$$

a and b are transmit and receive array factor vectors respectively. S is the complex gain that is dependant on direction and delay. g(t) is the transmit and receive pulse shaping impulse response

- **With suitable choices of array geometry and antenna element patterns, $\mathbf{H}(\tau)$ = H which is an M_R x M_T matrix with complex Gaussian i. i. d random variables**
- **Accurate for NLOS rich-scattering environments, with sufficient antenna spacing at transmitter and receiver with all elements identically polarized**

Capacity of MIMO Channels

$$y = Hs + n$$

- Let the transmitted vector s be a random vector to be very general and n is normalized noise. Let the total transmitted power available per symbol period be P. Then,

$$C = \log_2 (I_M + HQH^H) \text{ b/s/Hz}$$

where $Q = E\{ss^H\}$ and trace(Q) < P according to our power constraint

- Consider specific case when we have users transmitting at equal power over the channel and the users are *uncorrelated* (no feedback available). Then,

$$C_{EP} = \log_2 [I_M + (P/M_T)HH^H] \text{ b/s/Hz}$$

Telatar showed that this is the optimal choice for *blind* transmission

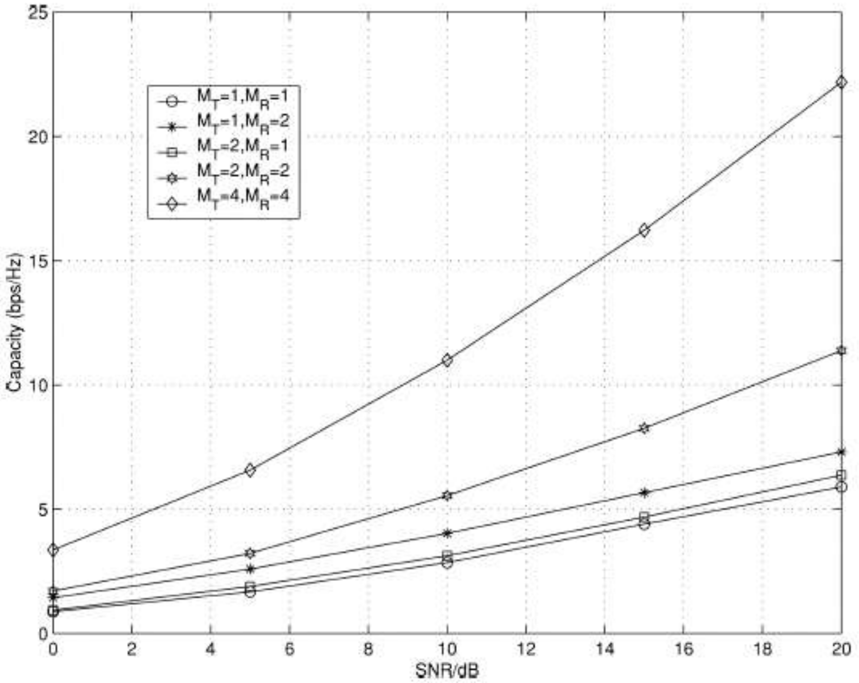
- Foschini and Telatar both demonstrated that as M_T and M_R grow,

$$C_{EP} = \min(M_T, M_R) \log_2 (P/M_T) + \text{constant b/s/Hz}$$

Note: When feedback is available, the *Waterfilling* solution is yields maximum capacity but converges to equal power capacity at high SNRs

Capacity (contd)

- The capacity expression presented was over one realization of the channel. Capacity is a random variable and has to be averaged over infinite realizations to obtain the true *ergodic capacity. Outage capacity* is another metric that is used to capture this



- So MIMO promises enormous rates theoretically! Can we exploit this practically?

MIMO Design Criterion

- **MIMO Systems can provide two types of gain**

Spatial Multiplexing Gain

- **Maximize transmission rate (optimistic approach)**
- **Use rich scattering/fading to your advantage**

Diversity Gain

- **Minimize P_e (conservative approach)**
- **Go for Reliability / QoS etc**
- **Combat fading**

- **If only I could have both! As expected, there is a *tradeoff***
- **System designs are based on trying to achieve either goal or a little of both**

Diversity

- **Each pair of transmit-receive antennas provides a signal path from transmitter to receiver. By sending the SAME information through different paths, multiple independently-faded replicas of the data symbol can be obtained at the receiver end. Hence, more reliable reception is achieved**
- **A diversity gain *d* implies that in the high SNR region, my P_e decays at a rate of $1/SNR^d$ as opposed to 1/SNR for a SISO system**
- **The maximal diversity gain d_{max} is the total number of independent signal paths that exist between the transmitter and receiver**
- **For an (M_R,M_T) system, the total number of signal paths is $M_R M_T$**

$$1 \leq d \leq d_{max} = M_R M_T$$

- **The higher my diversity gain, the lower my P_e**

Spatial Multiplexing

$$y = Hs + n \rightarrow y' = Ds' + n' \text{ (through SVD on H)}$$

where D is a diagonal matrix that contains the eigenvalues of HH^H

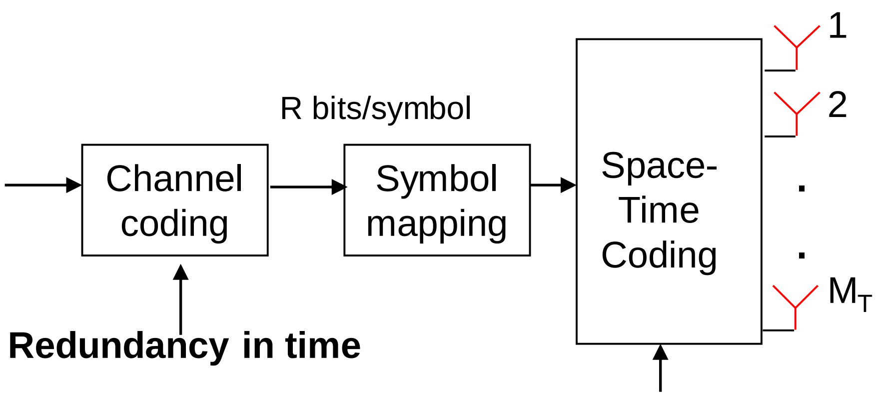
- Viewing the MIMO received vector in a different but equivalent way,

$$C_{EP} = \log_2 [I_M + (P/M_T)DD^H] = \sum_{i=1}^{m} \log_2 [1 + (P/M_T)\lambda_i] \text{ b/s/Hz}$$

- Equivalent form tells us that an (M_T,M_R) MIMO channel opens up $m = \min(M_T,M_R)$ independent SISO channels between the transmitter and the receiver

- So, intuitively, I can send a maximum of m different information symbols over the channel at any given time

Practical System



Redundancy in time

Coding rate = r_c

Space- time redundancy over T symbol periods

Spatial multiplexing gain = r_s

r_s : number of different symbols N transmitted in T symbol periods

r_s = N/T

Non-redundant portion of symbols

Spectral efficiency = $\frac{(R*r_c \text{ info bits/symbol})(r_s)(R_s \text{ symbols/sec})}{w}$

= Rr_cr_s bits/s/Hz assuming $R_s = w$

r_s is the parameter that we are concerned about: $0 \le r_s \le M_T$

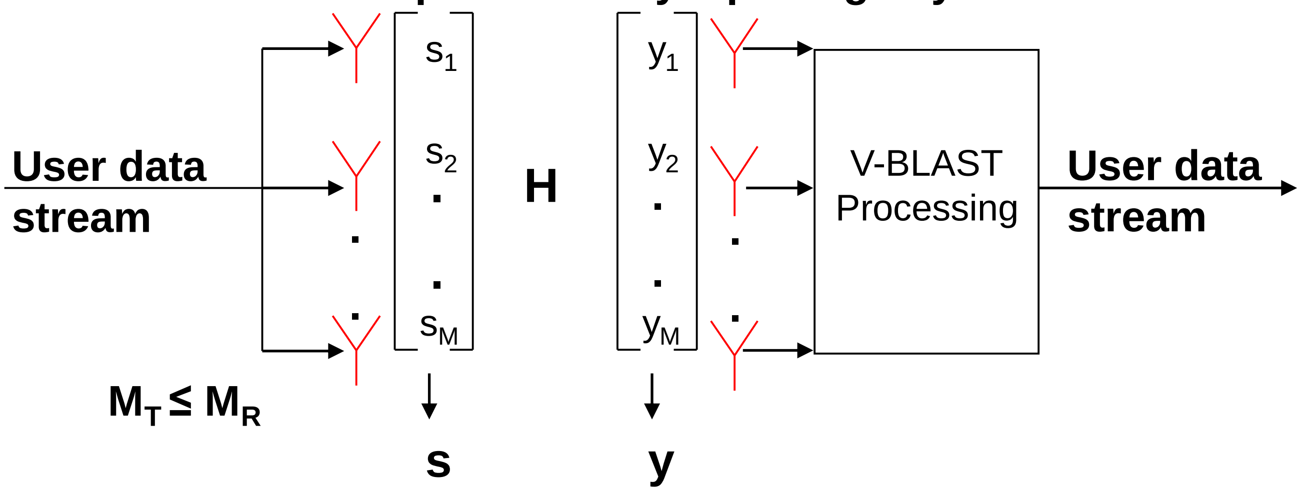
**** If $r_s = M_T$, we are in spatial multiplexing mode (max transmission rate)**

****If $r_s \le 1$, we are in diversity mode**

V-BLAST – Spatial Multiplexing

(Vertical Bell Labs Layered Space-Time Architecture)

- **This is the only architecture that goes all out for maximum rate. Hope the channel helps me out by 'splitting' my info streams!**



initialization:

$$i \leftarrow 1$$
$$\mathbf{G}_1 = \mathbf{H}^{+}$$
$$k_1 = \underset{j}{\operatorname{argmin}} \|(\mathbf{G}_1)_j\|^2$$

recursion:

$$\mathbf{w}_{k_i} = (\mathbf{G}_i)_{k_i}$$
$$y_{k_i} = \mathbf{w}_{k_i}^{T} \mathbf{r}_i$$
$$\hat{a}_{k_i} = Q(y_{k_i})$$
$$\mathbf{r}_{i+1} = \mathbf{r}_i - \hat{a}_{k_i} (\mathbf{H})_{k_i}$$
$$\mathbf{G}_{i+1} = \mathbf{H}^{+}_{\bar{k}_i}$$
$$k_{i+1} = \underset{j \notin \{k_1 \cdots k_i\}}{\operatorname{argmin}} \|(\mathbf{G}_{i+1})_j\|^2$$
$$i \leftarrow i + 1$$

- **Split data into M_T streams → maps to symbols → send**
- **Assume receiver knows H**
- **Uses old technique of *ordered successive cancellation* to recover signals**
- **Sensitive to estimation errors in H**
- **$r_s = M_T$ because in one symbol period, you are sending M_T different symbols**

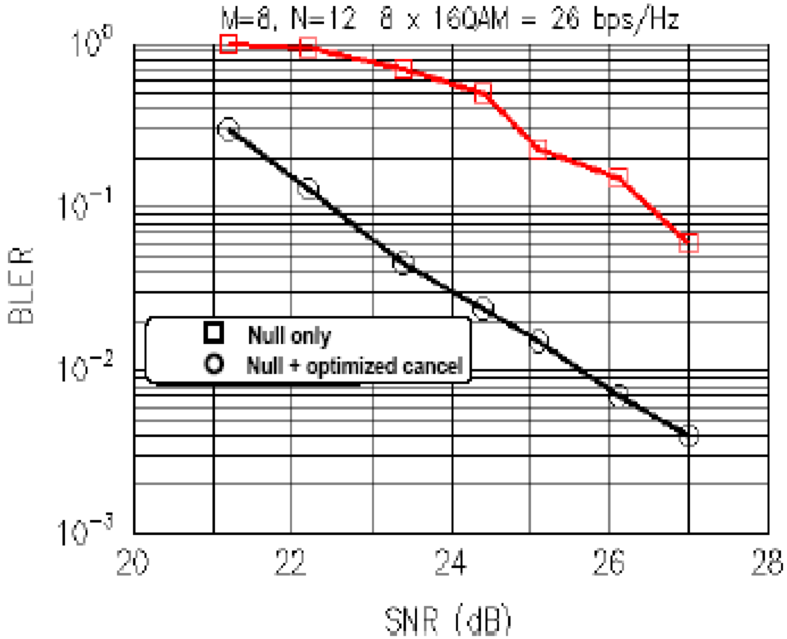
V-BLAST (Experimental Results)

- The prototype in an indoor environment was operated at a carrier frequency of 1.9 GHz, and a symbol rate of 24.3 ksymbols/sec, in a bandwidth of 30 kHz with $M_T = 8$ and $M_R = 12$
- Results shown on Block-Error-Rate Vs average SNR (at one received antenna element); Block = 100 symbols ; 20 symbols for training

• Each of the eight substreams utilized uncoded 16-QAM, i.e. 4 b/symb/trans

• Spec eff = $\dfrac{(8\text{ xmtr})(4\text{ b/sym/xmtr})(24.3\text{ ksym/s})}{30\text{ kHz}}$

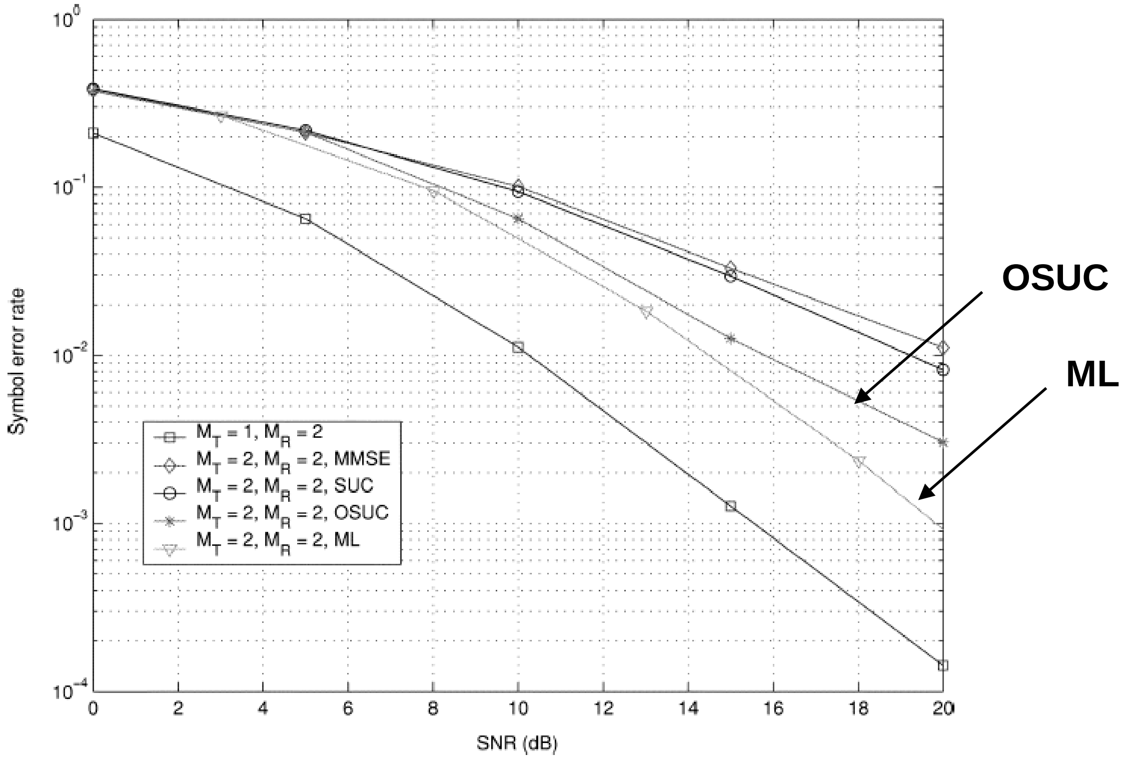
= 25. 9 bps/Hz



- In 30 kHz of bandwidth, I can push across 621Kbps of data!! Wireless spectral efficiencies of this magnitude are unprecedented, and are furthermore unattainable using traditional techniques

Alternate Receivers

- Can replace OSUC by other front-ends; MMSE, SUC, ML for instance



D-BLAST – a little of both

(Diagonal Bell Labs Layered Space-Time Architecture)

- **In D-BLAST, the input data stream is divided into sub streams which are coded, each of which is transmitted on different antennas time slots in a diagonal fashion**
- **For example, in a (2,2) system**

$$\begin{bmatrix} 0 & \boldsymbol{x}_1^{(1)} & \boldsymbol{x}_1^{(2)} & \cdots \\ \boldsymbol{x}_2^{(1)} & \boldsymbol{x}_2^{(2)} & \boldsymbol{x}_2^{(3)} & \cdots \end{bmatrix}$$

$M_T \le M_R$

- **receiver first estimates $x_2^{(1)}$ and then estimates $x_1^{(1)}$ by treating $x_2^{(1)}$ as interference and nulling it out**
- **The estimates of $x_2^{(1)}$ and $x_1^{(1)}$ are fed to a joint decoder to decode the first substream**

- **After decoding the first substream, the receiver cancels the contribution of this substream from the received signals and starts to decode the next substream, etc.**
- **Here, an overhead is required to start the detection process; corresponding to the 0 symbol in the above example**
- **Receiver complexity high**

Alamouti's Scheme - Diversity

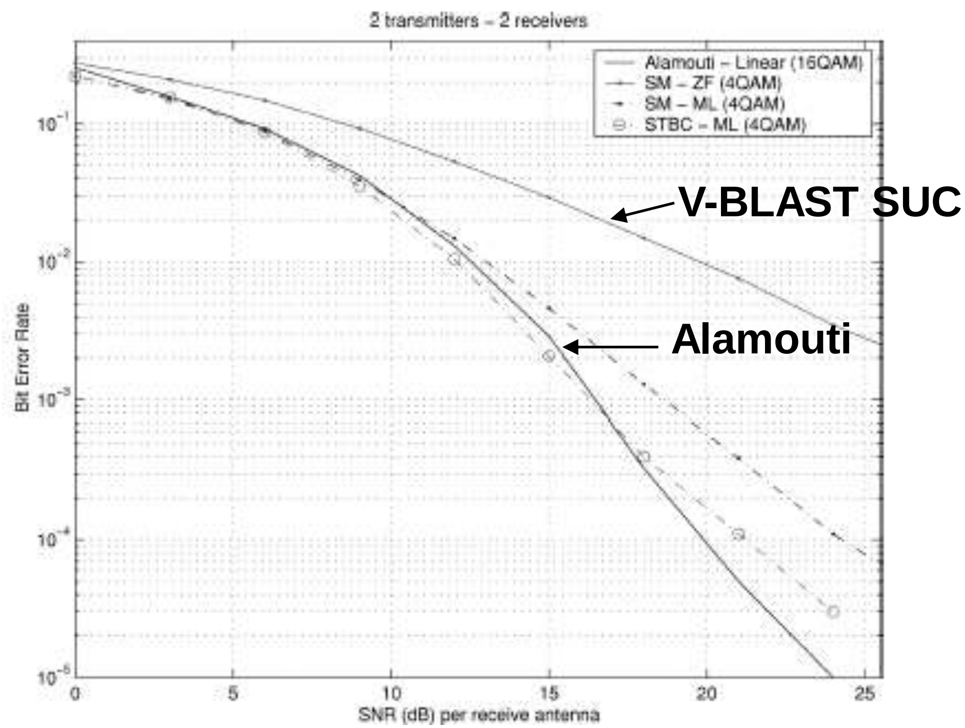
- Transmission/reception scheme easy to implement
- Space diversity because of antenna transmission. Time diversity because of transmission over 2 symbol periods
- Consider (2, M_R) system

$$\begin{bmatrix} x_1 & -x_2^{\dagger} \\ x_2 & x_1^{\dagger} \end{bmatrix}$$

- Receiver uses combining and ML detection
- $r_s = 1$

• If you are working with a (2,2) system, stick with Alamouti!

• Widely used scheme: CDMA 2000, WCDMA and IEEE 802.16-2004 OFDM-256



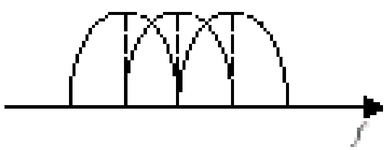
Comparisons

Scheme	Spectral Efficiency	P_e	Implementation Complexity
V-BLAST	HIGH	HIGH	LOW
D-BLAST	MODERATE	MODERATE	HIGH
ALAMOUTI	LOW	LOW	LOW

Orthogonal Frequency Division Multiplexing (OFDM)

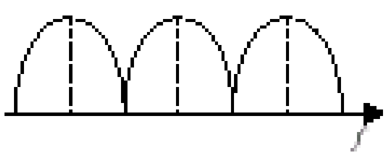
- As the data rate increases in a multipath environment, the interference goes from flat fading to frequency selective (last reflected component arrives after symbol period). This results in heavy degradation
- Most popular solution to compensate for ISI: equalizers
- As we move to higher data rates (i.e.> 1 Mbps), equalizer complexity grows to level of complexity where the channel changes before you can compensate for it!
- Alternate solution: Multi-carrier Modulation (MCM) where channel is broken up into subbands such that the fading over each subchannel becomes flat thus eliminating the problem of ISI

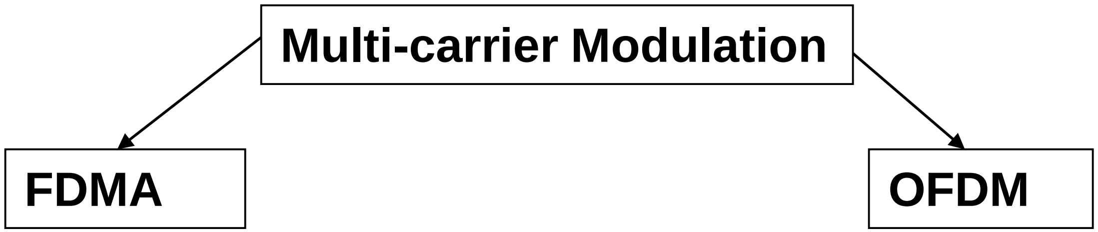
OFDM spectrum



vs.

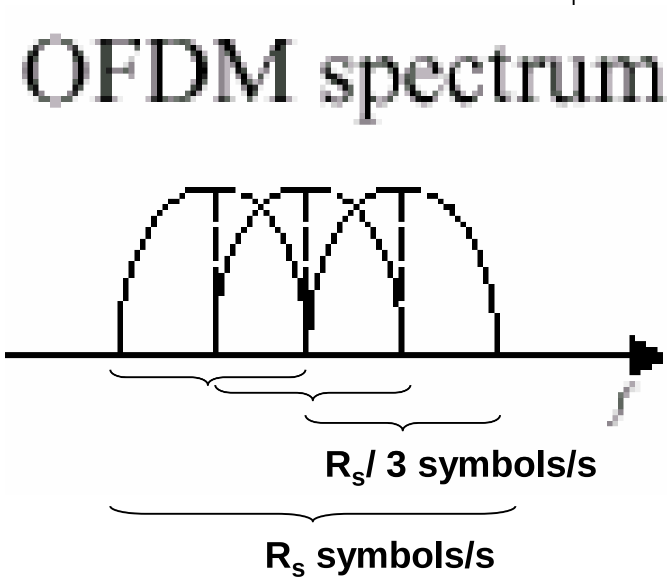
conventional FDM spectrum

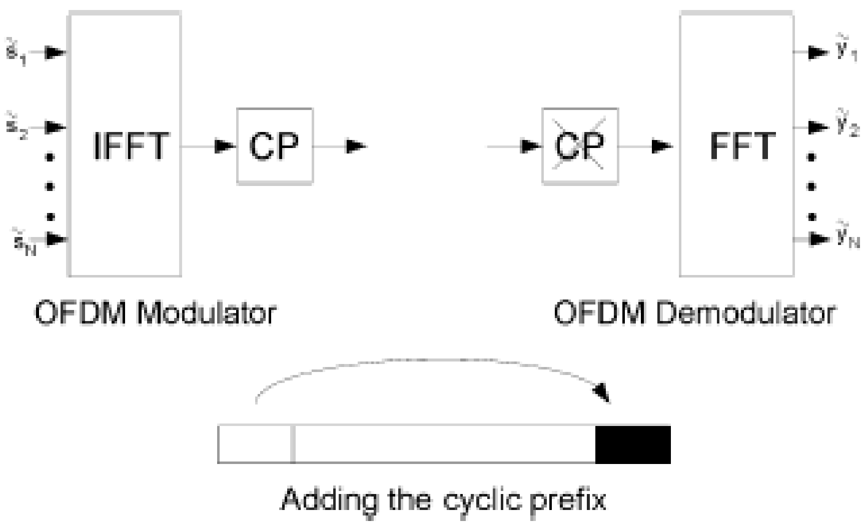




OFDM Spectral Efficiency

- The spectral efficiency of an OFDM-(PSK/ASK) system is same as compared to using the (PSK/ASK) system alone
- Spec eff = $\log_2 M$ bits/s/Hz
- However, you have successfully converted an ugly channel into a channel that you can use



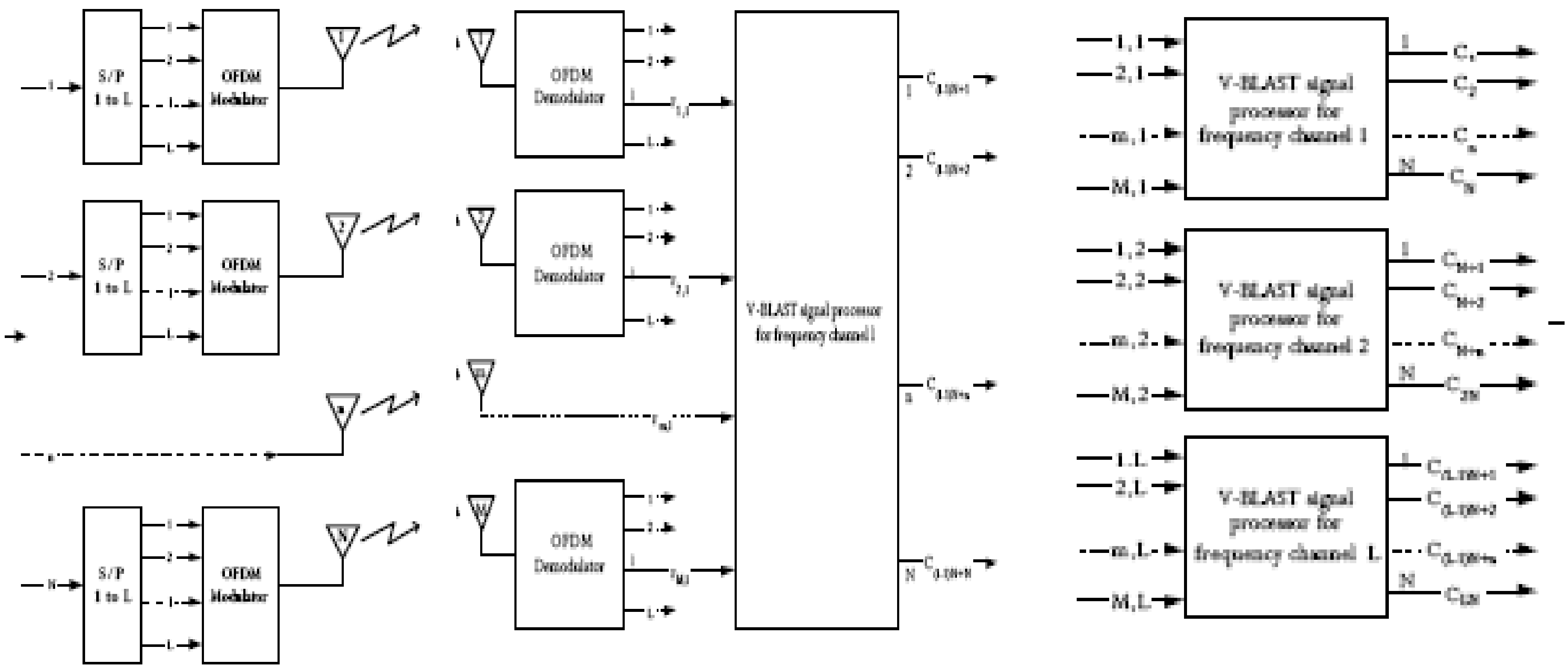


- easy to implement
- Used in IEEE 802.11A, .11G, HiperLAN, IEEE 802.16

MIMO-OFDM

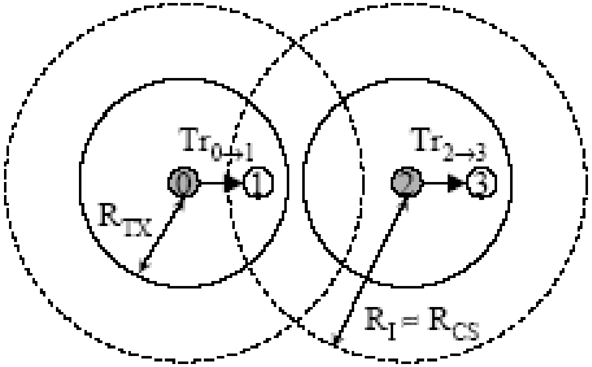
- **OFDM extends directly to MIMO channels with the IFFT/FFT and CP operations being performed at each of the transmit and receive antennas. MIMO-OFDM decouples the frequency-selective MIMO channel into a set of parallel MIMO channels with the input–output relation for the ith (i = 0, 2,…,L-1) tone,**

$$\mathbf{y}_i = \mathbf{H}_i\mathbf{s}_i + \mathbf{n}_i \qquad \mathbf{i = 0, 2,\ldots, L\text{-}1}$$



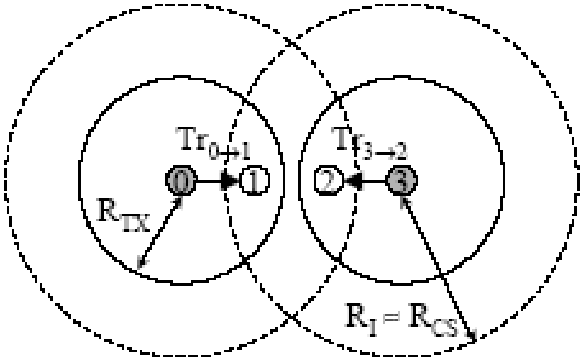
IEEE 802.11 MAC (DCF Mode)

- **As a result of the CSMA/CA with RTS/CTS MAC protocol, two issues arise**
  -the unfairness problem
  -extreme throughput degradation (ETD)



Throughtput $T_{2\to3}$ > Throughtput $T_{0\to1}$

Unfairness



Both throughtput $T_{2\to3}$ and throughtput $T_{0\to1}$ are equally affected

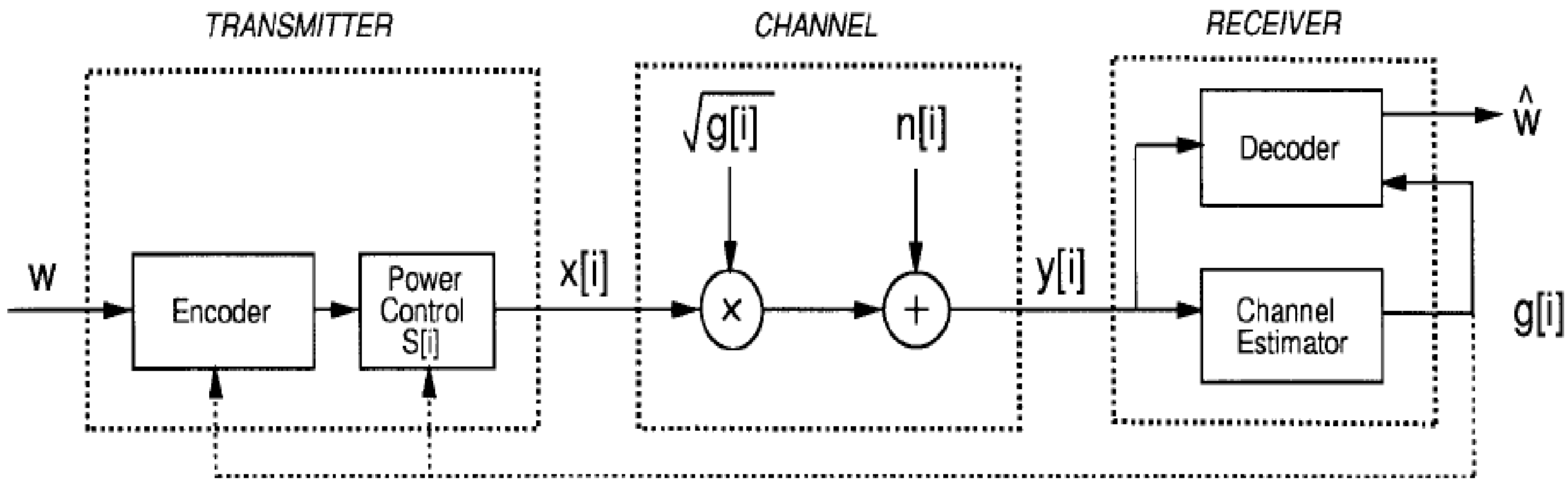
ETD

Capacity of Fading Channels with channel side information

Obtain the Shannon capacity of a single user fading channel with an average power constraint under different channel side information conditions:

- Fading channel with channel side information at the transmitter and the receiver.
- Fading channel with channel side information at the receiver alone.

System Model



- •Fading Channel
- •Block Encoder
- •Power Control
- •Possible Feedback Channel

System Model

- The system model in figure 1 is a discrete-time channel with Stationary ergodic time-varying gain=
- AWGN = n[i]
- Channel power gain = g[i] is independent of the channel input and has an expected value of unity.
- Average transmit signal power =S
- Noise density = №
- Received signal bandwidth =B
- The instantaneous received signal-to-noise ratio (SNR),
- S=(№B).
- The system model, which sends an input message w from the
  transmitter to the receiver.
- The message is encoded into the codeword x, which is transmitted over the time-varying channel as x[i] at time i.
- The channel gain g[i] changes over the transmission of the codeword.
- We assume perfect instantaneous channel estimation so that the channel power gain g[i] is known to the receiver at time i.
- We also consider the case when g[i] is known to both the receiver and transmitter at time i, as might be obtained through an error-free delayless feedback path.
- This allows the transmitter to adapt x[i] to the channel gain at time i, and is a reasonable model for a slowly varying channel with channel estimation and transmitter feedback.

Capacity Analysis
Side Information at the Transmitter and Receiver **Capacity**

$$C = \sum_{s \in \mathcal{S}} C_s p(s). \qquad (1)$$

$$C = \int_\gamma C_\gamma p(\gamma)\, d\gamma = \int_\gamma B \log(1+\gamma) p(\gamma)\, d\gamma. \qquad (2)$$

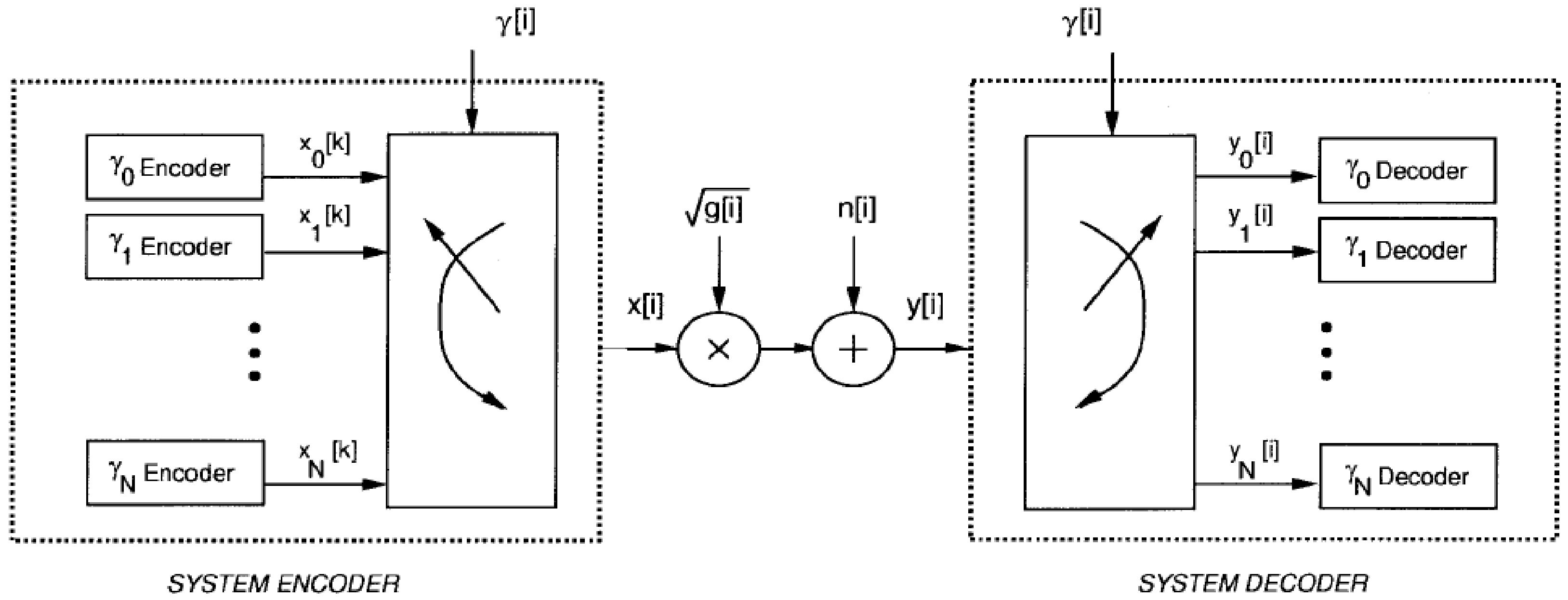
$$\int_\gamma S(\gamma) p(\gamma)\, d\gamma \le S. \qquad (3)$$

$$C(S) = \max_{S(\gamma):\ \int S(\gamma)p(\gamma)\, d\gamma = S} \int_\gamma B \log\left(1 + \frac{S(\gamma)\gamma}{S}\right) p(\gamma)\, d\gamma. \qquad (4)$$

Capacity

- Channel capacity varies with the channel state (SNR).
- The capacity of the fading channel is the sum of the capacities in each fading state weighted by the probability of that state.
- The log function is concave (down), so Jensen's inequality tells us that the expected capacity of the fading channel is less than or equal to the capacity of an AWGN channel with the same average SNR.

CODING


γ[i]
γ_0 Encoder
γ_1 Encoder
γ_N Encoder
$x_0[k]$
$x_1[k]$
$x_N[k]$
$\sqrt{g[i]}$
n[i]
x[i]
y[i]
$y_0[i]$
$y_1[i]$
$y_N[i]$
γ_0 Decoder
γ_1 Decoder
γ_N Decoder
SYSTEM ENCODER
SYSTEM DECODER

CODING

- The optimal encoder adjusts its rate according to the channel state: better rates for better channels.
- The rate adjustment is made in discrete steps by specifying a codebook of a particular size for quantized SNR intervals.
- Coding works in conjunction with optimal power control.

Optimal Adaptive Technique

- The fading-channel capacity with channel side information at the receiver and transmitter is achieved when the transmitter adapt its power, data rate, and coding scheme to the channel variation.
- The optimal power allocation is a “water-pouring”.
- They required a feedback path between the transmitter and receiver and some complexity in the transmitter.
- It uses variable-rate and power transmission, and the complexity of its decoding technique is comparable to the complexity of decoding a sequence of additive white Gaussian noise(AWGN) channels in parallel.
- The optimal adaptive technique has the highest capacity.

Nonadaptive Technique

- If g[i] is known at the decoder then by scaling, the fading channel with power gain g[i] is equivalent to an AWGN channel with noise power N0B/g[i].
- If the transmit power is fixed at S and g[i] is i.i.d. then the input distribution at time i which achieves capacity is an i.i.d. Gaussian distribution with average power S.
- The channel capacity with i.i.d. fading and receiver side information only is given by:

$$C(S) = \int B \log(1+\gamma) p(\gamma)\, d\gamma \qquad (8)$$

Suboptimal adaptive techniques

1. Channel inversion.
2. Truncated channel inversion.

For the suboptimal techniques :

- They adapt the transmit power and keep the transmission rate constant.
- They have very simple encoder and decoder designs , but they exhibit a capacity penalty in sever fading.
- The suboptimal power control schemes selected for comparison are not really comparable. They're designed for completely different purposes.

Channel Inversion (SubOptimal)

- We now consider a suboptimal transmitter adaptation scheme where the transmitter uses the channel side information to maintain a constant received power, i.e., it inverts the channel fading.
- The channel then appears to the encoder and decoder as a time-invariant AWGN channel.

$$C(S) = B \log[1+\sigma] = B \log\left[1 + \frac{1}{\mathbf{E}[1/\gamma]}\right]. \qquad (9)$$

- Channel inversion is common in spread-spectrum systems with near–far interference imbalances.
- It is also very simple to implement, since the encoder and decoder are designed for an AWGN channel, independent of the fading statistics.
- It can exhibit a large capacity penalty in extreme fading environments.

Truncated channel inversion.

- A truncated inversion policy can only compensates for fading above a certain cutoff fade depth of SNR:

$$\frac{S(\gamma)}{S} = \begin{cases} \frac{\sigma}{\gamma}, & \gamma \geq \gamma_0 \\ 0, & \gamma < \gamma_0. \end{cases} \quad (10)$$

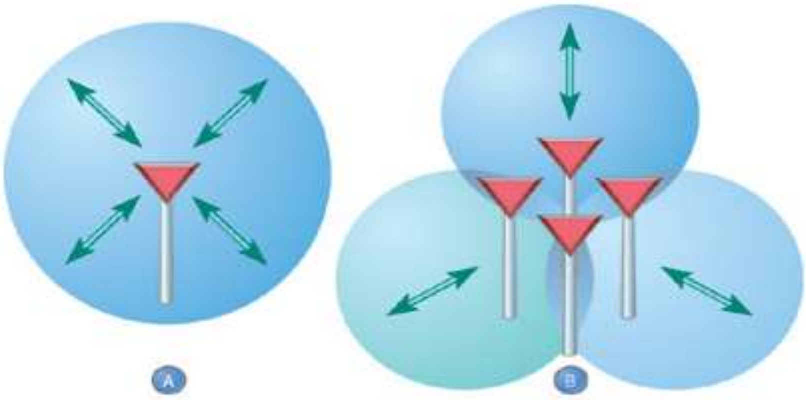
Since the channel is only used when $\gamma \geq \gamma_0$, the power constraint (3) yields $\sigma = 1/\boldsymbol{E}_{\gamma_0}[1/\gamma]$, where

$$\boldsymbol{E}_{\gamma_0}[1/\gamma] \triangleq \int_{\gamma_0}^{\infty} \frac{1}{\gamma} p(\gamma)\, d\gamma. \quad (11)$$

For decoding this truncated policy, the receiver must know when $\gamma < \gamma_0$. The capacity in this case, obtained by maximizing over all possible γ_0, is

$$C(S) = \max_{\gamma_0} B \log \left[1 + \frac{1}{\boldsymbol{E}_{\gamma_0}[1/\gamma]}\right] p(\gamma \geq \gamma_0). \quad (12)$$

Nonsmart-antennas system



➢Traditional omni-directional antennas, as shown in Figure A above act as transducers (that is, they convert electromagnetic energy into electrical energy) and are not an effective way to combat inter-cell and intra-cell interferences.

➢One cost-effective solution to this interference challenge is to split up the wireless cell into multiple sectors using sectorized antennas. As Figure B illustrates, sectorized antennas transmit and receive in a limited portion of the cell, typically one-third of the circular area, thereby reducing the overall interference in the system.

➢Efficiency can increase still further by using either spatial diversity or by focusing a narrow beam on a single user. The second approach is known as ***beam forming***, and it requires an array of antennas that together perform "smart" transmission and reception of signals, via the implementation of advanced signal processing algorithms.

➢Combination of FPGAs, digital signal processing IP, and embedded processors that implement beam-forming applications.

➢The methods used to implement such applications and the benefits of improved processing speed, system flexibility, and reduced risk that this approach can deliver.

Smart antennas

Compared with traditional omni-directional and sectorized antennas, smart-antenna systems can provide:

- Greater coverage area for each cell site
- Better rejection of co-channel interference
- Reduced multipath interference via increased directionality
- Reduced delay spread as fewer scatterers are allowed into the beam
- Increased frequency reuse with fewer base stations
- Higher range in rural areas
- Improved building penetration
- Location information for emergency situations
- Increased data rates and overall system capacity
- Reduction in dropped calls

How is it done?

- A linearly arranged and equally spaced array of antennas forms the basic structure of a beam former.

- In order to form a beam, each user's information signal is multiplied by a set of complex weights (where the number of weights equals the number of antennas) and then transmitted from the array.

- The important point in this transmission is that the signals emitted from different antennas in the array differ in phase (which is determined by the distance between antenna elements) as well as amplitude (determined by the weight associated with that antenna).

- Changing the direction of the beam, therefore, involves changing the weight set as the spacing between the antenna elements is fixed.

- The rest of this Presentation describes two such schemes known as switched and adaptive beam forming.

Switched and adaptive beam

- If the complex weights used are selected from a library of weights that form beams in specific, predetermined directions, the process is called *switched beam forming*.
- In this process, a hand-off between beams is required as users move tangentially to the antenna array.
- If the weights are computed and adaptively updated in real time, the process is known as *adaptive beam forming*.
- The adaptive process permits narrower beams and reduced output in other directions, significantly improving the signal-to-interference-plus-noise ratio (SINR).
- With this technology, each user's signal is transmitted and received by the base station only in the direction of that particular user. This drastically reduces the overall interference in the system.
- A smart-antenna system, as shown in Figure, includes an array of antennas that together direct different transmission/reception beams toward each cellular user in the system.

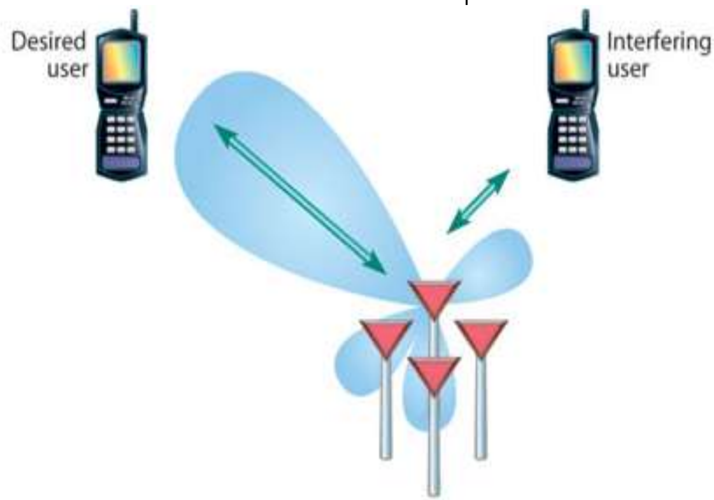


Figure : A beam-forming smart-antennas system

Implementing adaptive beam

➢Adaptive beam forming can be combined with the well known Rake receiver architectures that are widely used in CDMA-based 3G systems, to provide processing gains in both the temporal and spatial domains.

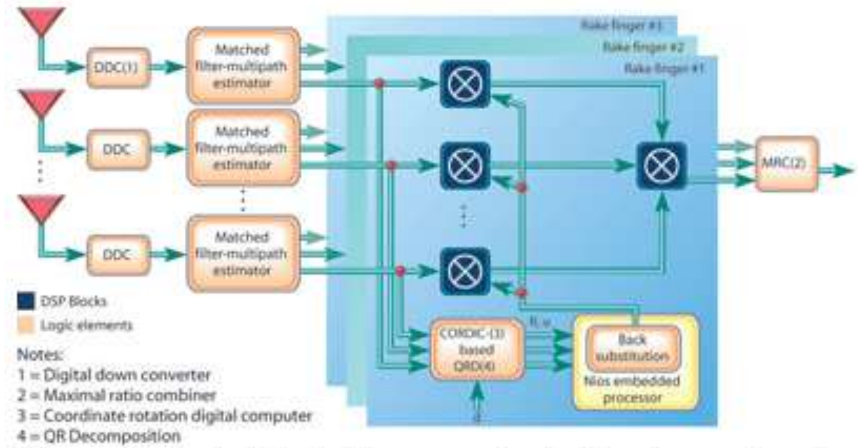


Figure 3: Basic block diagram of adaptive beam forming with FPGA

➢This section describes the implementation of a *Rake beam-former structure*, also known as a *two-dimensional Rake*, which performs joint space-time processing. As illustrated in Figure 3, the signal from each receiving antenna is first down-converted to baseband, processed by the matched filter-multipath estimator, and accordingly assigned to different Rake fingers.

➢The beam-forming unit on each Rake finger then calculates the corresponding beam-former weights and channel estimate using the pilot symbols that have been transmitted through the dedicated physical control channel (DPCCH).

➢The QR-decomposition-(QRD)-based recursive least squares (RLS) algorithm is usually used as the weight-update algorithm for its fast convergence and good numerical properties.

➢The updated beam-former weights are then used for multiplication with the data that has been transmitted through the dedicated physical data channel (DPDCH).

➢ Maximal ratio combining (MRC) of the signals from all fingers is then performed to yield the final soft estimate of the DPDCH data.

Implementing adaptive beam

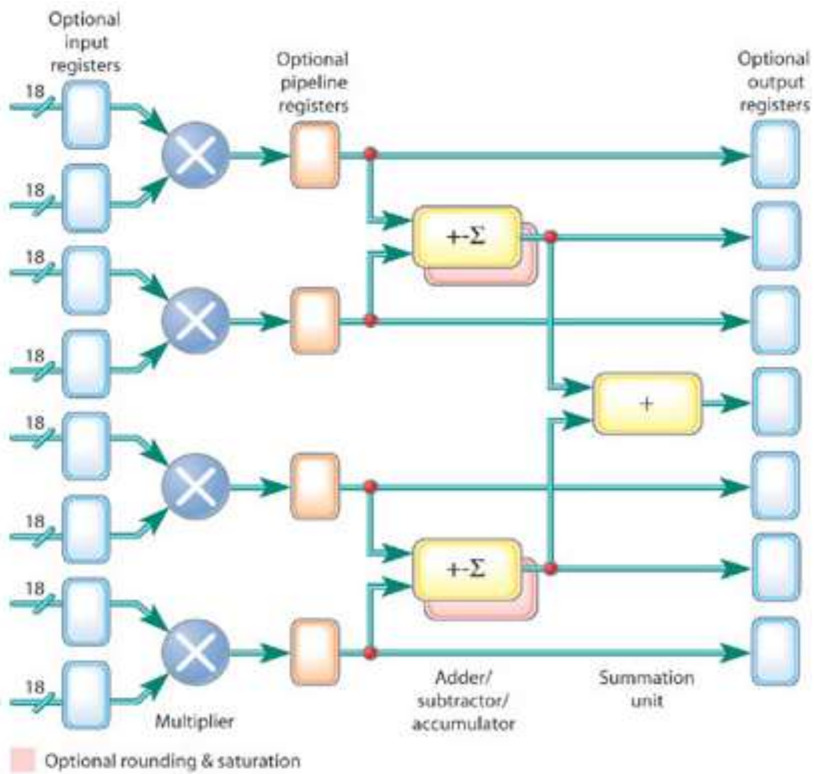


Figure 4: Example DSP block architecture

➢Applying complex weights to the signals from different antennas involves complex multiplications that map well onto the embedded DSP blocks available for many FPGAs. The example in Figure 4 shows DSP blocks with a number of multipliers, followed by adder/subtractor/accumulators, with registers for pipelining. Such a structure lends itself to complex multiplication and routing required in beam-forming designs.

Adaptive algorithms

➢Adaptive signal processing algorithms such as least mean squares (LMS), normalized LMS (NLMS), and recursive least squares (RLS) have historically been used in a number of wireless applications such as equalization, beam forming and adaptive filtering. These all involve solving for an over-specified set of equations, as shown below, where *m* > *N*:

$$x_1(1)c_0 + x_2(1)c_1 + \ldots\ldots x_N(1)c_N = y(1) + e(1)$$
$$x_1(2)c_0 + x_2(2)c_1 + \ldots\ldots x_N(2)c_N = y(2) + e(2)$$
$$\vdots$$
$$x_1(m)c_0 + x_2(m)c_1 + \ldots\ldots x_N(m)c_N = y(m) + e(m)$$

➢Among the different algorithms, the recursive least squares algorithm is generally preferred for its fast convergence. The least squares approach attempts to find the set of coefficients that minimizes the sum of squares of the errors, in other words:

$$\left\{ \min \sum_m e(m)^2 \right\}$$

➢Representing the above set of equations in the matrix form, we have:

$$X\mathbf{c} = \mathbf{y} + \mathbf{e} \quad ..(1)$$

➢where *X* is a matrix (*m*x*N*, with *m*>*N*) of noisy observations, **y** is a known training sequence, and **c** is the coefficient vector to be computed such that the error vector **e** is minimized

Adaptive algorithms

$$X\mathbf{c} = \mathbf{y} + \mathbf{e} \quad \ldots(1)$$

- Direct computation of the coefficient vector **c** involves matrix inversion, which is generally undesirable for hardware implementation due to numerical instability issues.

- Matrix decomposition based on least squares schemes, such as Cholesky, LU, SVD, and QR-decompositions, avoid explicit matrix inversions and are hence more robust and well suited for hardware implementation.

- Such schemes are being increasingly considered for high-sample-rate applications such as digital predistortion, beam forming, and MIMO signal processing. FPGAs are the preferred hardware for such applications because of their ability to deliver enormous signal-processing bandwidth.

- FPGAs provide the right implementation platform for such computationally demanding applications with their inherent parallel-processing benefits (as opposed to serial processing in DSPs) along with the presence of embedded multipliers that provide throughputs that are an order of magnitude greater than the current generation of DSPs.

- The presence of embedded soft processor cores within FPGAs gives designers the flexibility and portability of high-level software design while maintaining the performance benefits of parallel hardware operations in FPGAs.

QRD-RLS algorithm

$$X\mathbf{c} = \mathbf{y} + \mathbf{e} \quad \ldots(1)$$

The least squares algorithm attempts to solve for the coefficient vector **c** from *X* and *y*. To realize this, the QR-decomposition algorithm is first used to transform the matrix *X* into an upper triangular matrix *R* (*N* x *N* matrix) and the vector **y** into another vector **u** such that *R***c**=**u**. The coefficients vector **c** is then computed using a procedure called *back substitution*, which involves solving these equations:

$$c_N = \frac{u_N}{R_{NN}} \quad (2)$$

$$c_i = \frac{1}{R_{ii}}\left(u_i - \sum_{j=i+1}^{N} R_{ij} C_j\right) \text{ for } i = N-1, ..1 \quad (3)$$

The QRD-RLS algorithm flow is depicted in Figure 5.

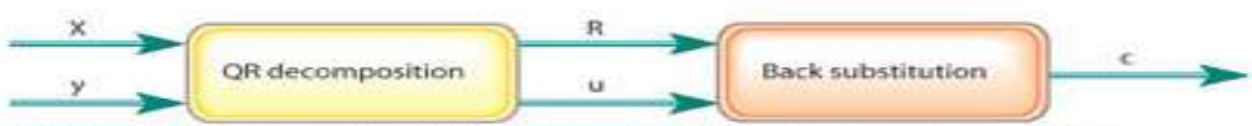


Figure 5: QR-decomposition-based least squares

QRD-RLS algorithm

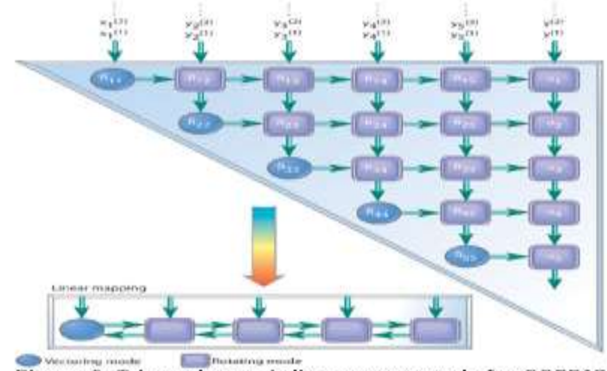


Figure 6: Triangular systolic array example for CORDIC-based QRD-RLS

- The QR-decomposition of the input matrix *X* can be performed, as illustrated in Figure 6, using the well-known systolic array architecture.

- The rows of matrix *X* are fed as inputs to the array from the top along with the corresponding element of the vector **y**. The **R** and **u** values held in each of the cells once all the inputs have been passed through the matrix are the outputs from QR-decomposition. These values are subsequently used to derive the coefficients using back substitution technique.

- Each of the cells in the array can be implemented as a coordinate rotation digital computer (CORDIC) block. CORDIC describes a method of performing a number of functions, including trigonometric, hyperbolic, and logarithmic functions.2 The algorithm is iterative and uses only add, subtract, and shift operations, making it attractive for hardware implementations.

- The number of iterations depends on the input and output precision, with more iterations being needed for more bits.

- For complex inputs, only one CORDIC block is required per cell. Many applications involve complex inputs and outputs to the algorithm, for which three CORDIC blocks are required per cell. In such cases, a single CORDIC block can be efficiently timeshared to perform the complex operations

Weights and measures

➢The beam-former weights vector **c** is related to the **R** and **u** outputs of the triangular array as **Rc=u**. **R** being an upper triangular matrix, **c** can be solved using a procedure called back substitution.

➢As outlined in Haykin and Zhong Mingqian et al., the back-substitution procedure operates on the outputs of the QR-decomposition and involves mostly multiply and divide operations that can be efficiently executed in FPGAs with embedded soft processors.

➢Some FPGA-resident processors can be configured with a 16x16 -> 32-bit integer hardware multipliers.

➢The software can then complete the multiply operation in a single clock cycle. Since hardware dividers generally are not available, the divide operation can be implemented as custom logic block that may or may not become part of the FPGA-resident microprocessor. Between the multiply and divide accelerators, back-substitution becomes easy and efficient.

THANK YOU