LECTURE NOTES

ON

WIRELESS COMMUNICATION AND NETWORKS

IV B. Tech II semester (JNTUH-R18)

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SYLLABUS

UNIT-I

The Cellular Concept-System Design Fundamentals: Introduction, Frequency Reuse, Channel Assignment Strategies, Handoff Strategies- Prioritizing Handoffs, Practical Handoff Considerations, Interference and system capacity — Co channel Interference and system capacity, Channel planning for Wireless Systems, Adjacent Channel interference , Power Control for Reducing interference, Trunking and Grade of Service, Improving Coverage & Capacity in Cellular Systems- Cell Splitting, Sectoring.

UNIT—II

Mobile Radio Propagation: Large-Scale Path Loss: Introduction to Radio Wave Propagation, Free Space Propagation Model, Relating Power to Electric Field, The Three Basic Propagation Mechanisms, Reflection-Reflection from Dielectrics, Brewster Angle, Reflection from prefect conductors, Ground Reflection (Two-Ray) Model, Diffraction-Fresnel Zone Geometry, Knife-edge Diffraction Model, Multiple knife-edge Diffraction, Scattering, Outdoor Propagation Models- Longley-Ryce Model, Okumura Model, Hata Model, PCS Extension to Hata Model, Walfisch and Bertoni Model, Wideband PCS Microcell Model, lndoor Propagation Models-Partition losses (Same Floor), Partition losses between Floors, Log-distance path loss model, Ericsson Multiple Breakpoint Model, Attenuation Factor Model, Signal penetration into buildings, Ray Tracing and Site Specific Modeling.

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UNIT —III

Mobile Radio Propagation: Small —Scale Fading and Multipath: Small Scale Multipath propagation-Factors influencing small scale fading, Doppler shift, Impulse Response Model of a multipath channel-Relationship between Bandwidth and Received power, Small-Scale Multipath Measurements-Direct RF Pulse System, Spread Spectrum Sliding Correlator Channel Sounding, Frequency Domain Channels Sounding, Parameters of Mobile Multipath Channels-Time Dispersion Parameters, Coherence Bandwidth, Doppler Spread and Coherence Time, Types of Small-Scale Fading-Fading effects Due to Multipath Time Delay Spread, Flat fading, Frequency selective fading, Fading effects Due to Doppler Spread-Fast fading, slow fading, Statistical Models for multipath Fading Channels-Clarke's model for flat fading, spectral shape due to Doppler spread in Clarke's model, Simulation of Clarke and Gans Fading Model, Level crossing and fading statistics, Two-ray Rayleigh Fading Model.

UNIT -IV

Equalization and Diversity: Introduction, Fundamentals of Equalization, Training A Generic Adaptive Equalizer, Equalizers in a communication Receiver, Linear Equalizers, Non-linear Equalization-Decision Feedback Equalization (DFE), Maximum Likelihood Sequence Estimation (MLSE) Equalizer, Algorithms for adaptive equalization-Zero Forcing Algorithm, Least Mean Square Algorithm, Recursive least squares algorithm. Diversity Techniques-Derivation of selection Diversity improvement, Derivation of Maximal Ratio Combining improvement, Practical Space Diversity Consideration-Selection Diversity, Feedback or Scanning Diversity, Maximal Ratio Combining, Equal Gain Combining, Polarization Diversity, Frequency Diversity, Time Diversity, RAKE Receiver.

UNIT -V

Wireless Networks: Introduction to wireless Networks, Advantages and disadvantages of Wireless Local Area Networks, WLAN Topologies, WLAN Standard IEEE 802.11 ,IEEE 802.11 Medium Access Control, Comparision
of IEEE 802.11 a,b,g and n standards, IEEE 802.16 and its enhancements, Wireless PANs, Hiper Lan, WLL.

TEXT BOOKS

1. Wireless Communications, Principles, Practice — Theodore, S.Rappaport, 2nd Ed., 2002, PHI.
2. Wireless Communications-Andrea Goldsmith, 2005 Cambridge University Press.
3. Mobile Cellular Communication — Gottapu Sasibhushana Rao, Pearson Education, 2012.

REFERENCE BOOKS

1. Principles of Wireless Networks — Kaveh Pah Laven and P. Krishna Murthy, 2002, PE
2. Wireless Digital Communications — Kamilo Feher, 1999, PHI.
3. Wireless Communication and Networking — William Stallings, 2003,PHI.
4. Wireless Communication — Upen Dalal, Oxford Univ. Press
5. Wireless Communications and Networking — Vijay K. Gary, Elsevier.

UNIT 1

Cellular System

Introduction

In the older mobile radio systems, single high power transmitter was used to provide coverage in the entire area. Although this technique provided a good coverage, but it was virtually impossible in this technique to re-use the same radio channels in the system, and any effort to re-use the radio channels would result in interference. Therefore, in order to improve the performance of a wireless system with the rise in the demand for the services, a cellular concept was later proposed. This chapter will examine several parameters related with the cellular concept.

The Cellular Concept

The design aim of early mobile wireless communication systems was to get a huge coverage area with a single, high-power transmitter and an antenna installed on a giant tower, transmitting a data on a single frequency. Although this method accomplished a good coverage, but it also means that it was practically not possible to reuse the same frequency all over the system, because any effort to reuse the same frequency would result in interference.

The cellular concept was a major breakthrough in order to solve the problems of limited user capacity and spectral congestion. Cellular system provides high capacity with a limited frequency spectrum without making any major technological changes [1]. It is a system-level idea in which a single high-power transmitter is replaced with multiple low- power transmitters, and small segment of the service area is being covered by each transmitter, which is referred to as a cell. Each base station (transmitter) is allocated a part of the total number of channels present in the whole system, and different groups of radio channels are allocated to the neighboring base stations so that all the channels present in the system are allocated to a moderately small number of neighboring base stations.

The mobile transceivers (also called mobile phones, handsets, mobile terminals or mobile stations) exchange radio signals with any number of base stations. Mobile phones are not linked to a specific base station, but can utilize any one of the base stations put up by the company. Multiple base stations covers the entire region in such a way that the user can move around and phone call can be carried on without interruption, possibly using more than one base station. The procedure of changing a base station at cell boundaries is called *handover*. Communication from the Mobile Station (MS) or mobile phones to the Base Station (BS) happens on an uplink channel also called reverse link, and downlink channel or forward link is used for communication from BS to MS. To maintain a bidirectional communication between a MS and BS, transmission resources must be offered in both the uplink and downlink directions. This can take place either using Frequency-Division Duplex (FDD), in which separate frequencies are used for both uplink and downlink channels, or through Time-Division Duplex (TDD), where uplink and downlink communications take place on the same frequency, but vary in time.

FDD is the most efficient technique if traffic is symmetric, and FDD has also made the task of radio planning more efficient and easier, because no interference takes place between base stations as they transmit and receive data on different frequencies. In case of an asymmetry in the uplink and downlink data speed, the TDD performs better than FDD. As the uplink data rate increases, extra bandwidth is dynamically allocated to that, and as the data rate decreases, the allotted bandwidth is taken away.

Some of the important cellular concepts are:

- Frequency reuse
- Channel Allocation
- Handoff
- Interference and system capacity
- Trunking and grade of service
- Improving coverage and capacity

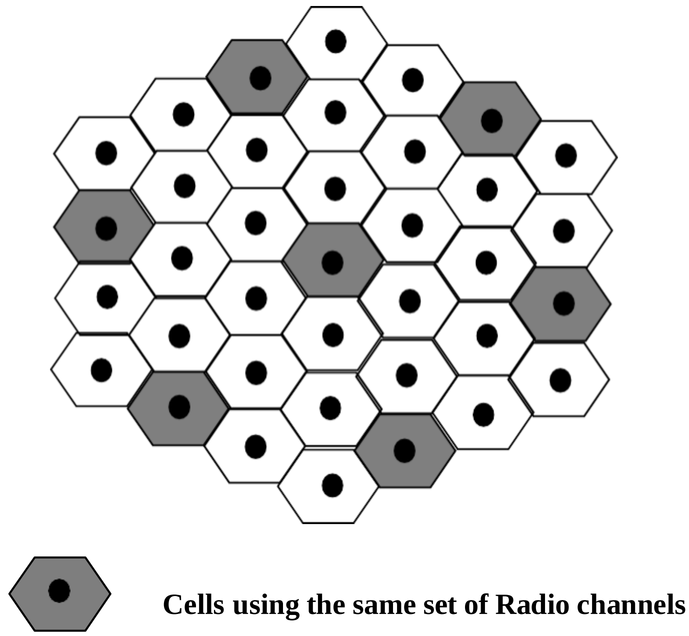


Fig. 1.1: Cellular Network

Frequency Reuse

Conventional communication systems faced the problems of limited service area capability and ineffective radio spectrum utilization. This is because these systems are generally designed to provide service in an autonomous geographic region and by selecting radio channels from a particular frequency band. On the other hand, the present mobile communication systems are designed to offer a wide coverage area and highgrade of service. These systems are also expected to provide a continuous communicationthrough an efficient utilization of available radio spectrum. Therefore, the design of mobile radio network must satisfy the following objectives i.e., providing continuous service, and wide service area, while efficiently using the radio spectrum.

In order to achieve these objectives, the present mobile systems use cellular networks which depend more on an intelligent channel allocation and reuse of channels throughout the region . Each base station is allocated a set of radio channels, which are to be used in a geographic area called a *cell.* Base stations in the neighboring cells are allocated radio channel sets, which are entirely different. The antennas of base station antennas are designed to get the required coverage within the specific cell. By restricting the coverage area of a base station to within the cell boundaries, the same set of radio channels can be used in the different cells that are separated from each other by distances which are large enough in order to maintain interference levels within limits. The procedure of radio sets selection and allocation to all the base stations present within a network is called *frequency reuse* .

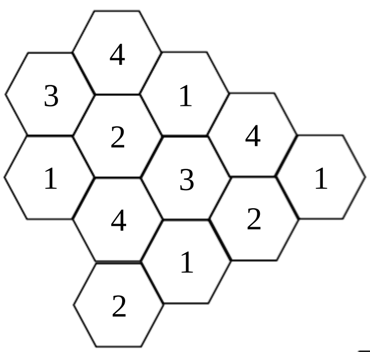
Fig. 1.1 shows the frequency reuse concept in a cell in a cellular network, in which cells utilize the same set of radio channels. The frequency reuse plan indicates where different radio channels are used. The hexagonal shape of cell is purely theoretical and is a simple model of radio coverage for each base station, although it has been globally adopted as the hexagon permits the easy analysis of a cellular system. The radio coverage of a cell can be calculated from field measurements. Although the actual radio coverage is very amorphous, a natural shape of a cell is required for an organized system design. While a circle is generally chosen to represent the coverage area of BS, but the circles present in the neighborhood cannot cover the entire region without leaving gaps or overlapping regions. Therefore, when selecting the cell shapes which can cover the entire geographical region without overlapping, there are three choices possible: a hexagon, square, and triangle. A particular design of the cell chosen in order to serve the weakest mobiles within the coverage area, and these are generally present at the cell boundaries ofthe cell. As hexagon covers the largest area from the center of a polygon to its farthest point, therefore, hexagon geometry can cover the entire geographic region to the fullest with minimum number of cells. When hexagon geometry is used to cover the entire geographic area, the base stations are either put up at the center of the cell, these cells are also called center excited cells or at the three of the six vertices (edge excited cells). Generally, center excited cells use omni-directional antennas and corner excited cells use directional antennas, but practically considerations for placing base stations are not exactly the same as they are shown in the hexagonal layouts.

Channel Reuse Schemes

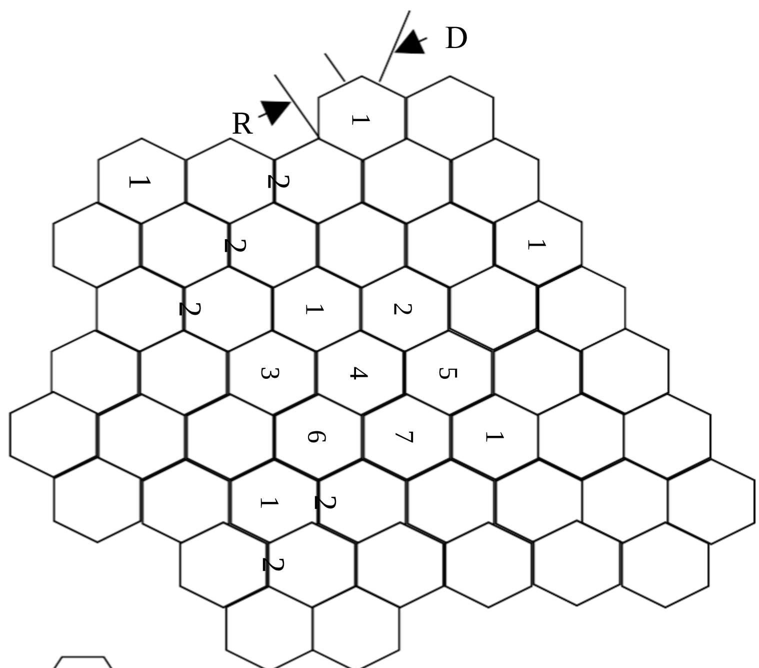
The radio channel reuse model can be used in the time and space domain. Channel reuse in the time domain turns out to be occupation of same frequency in different time slots and is also called Time Division Multiplexing. Channel reuse in the space domain is categorized into:

a) Same channel is allocated in two different areas, e.g. AM and FM radio stations using same channels in two different cities.

b) Same channel is frequently used in same area and in one system the scheme used is cellular systems. The entire spectrum is then divided into K reuse sets.



K=4



q=D/R=4
K=7

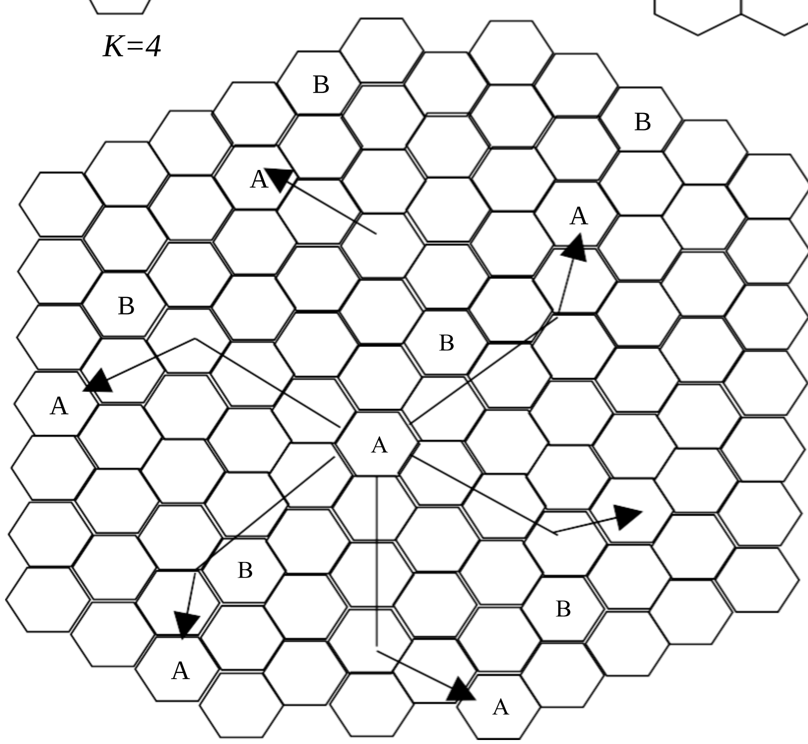


Fig. 1.2: K-Cell Reuse Pattern

Locating Co-channel Cells in a Cellular Network

Cells, which use the same set of channels, are called co-channels cells. For determining the location of co-channel cell present in the neighborhood, two shift parameters *i* and *j* are used where i and j are separated by 60^0, as shown in Fig. 1.3 below. The shift parameters can have any value 0, 1, 2 ...,n.

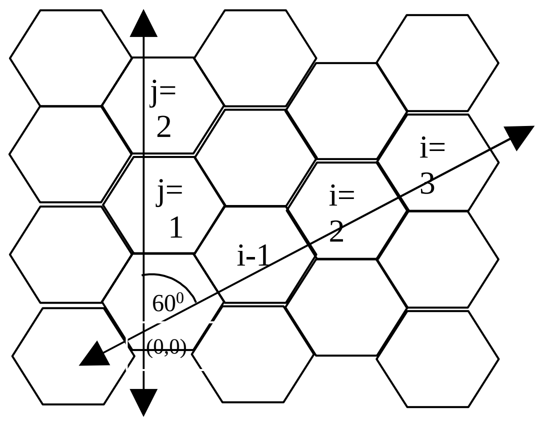


Fig. 1.3: Shift Parameters *i* and *j* in Hexagonal Network

To find the location of nearest co-channel cell, mark the center of the cell as (0, 0) for which co-channel cells are to be located. Define the unit distance as the distance of centre of two adjacent cells, and follow the two steps given below:

Step 1**:** Move *i* number of cells along *i* axis

Step 2**:** Turn 60^0 anti-clockwise and move *j* number of cells

The technique of locating co-channel cells using the preceding procedure is shown in Fig. 2.4 for *i*=3 and *j*=2. The shift parameters *i* and *j* measures the number of neighboring cells between co-channel cells.

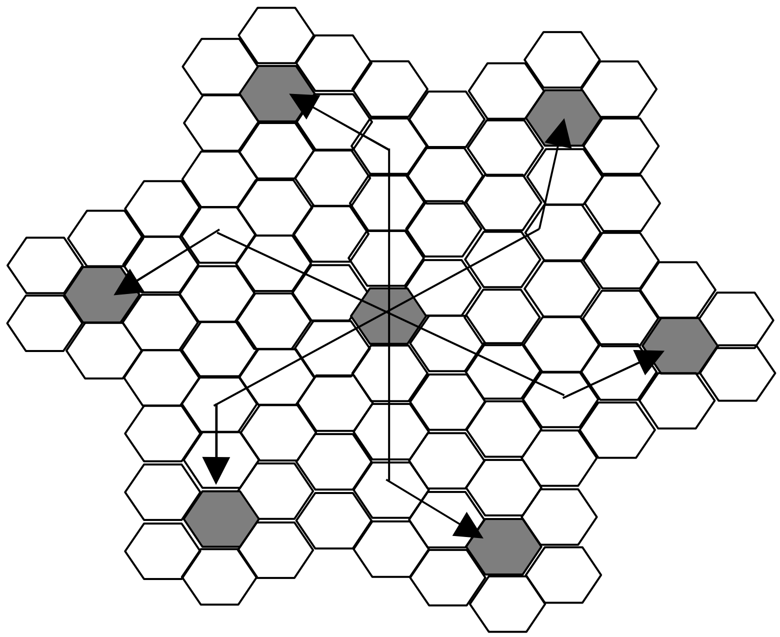


Fig. 1.4: Locating Co-channel Cells when *i=3* & *j=2*

The relationship between cluster size *K* and shift parameters *i* & *j*, is given below:

Let *'R'* be the distance between the center of a regular hexagon to any of its vertex. A regular hexagon is one whose all sides are also of equal length i.e. *'R'*. Let *'d'* be the distance between the centre of two neighboring hexagons, and following steps are followed while calculating the size of a cluster *'K'*.

Step 1: To show that $d = \sqrt{3}R$

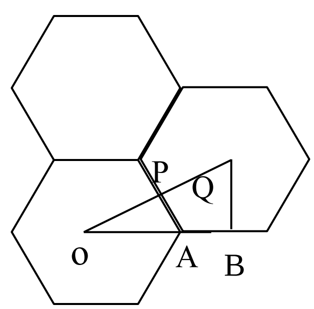


Fig. 1.5: Distance Between two adjacent cells

From the geometry of the Fig. 2.5, OA = R and AB = R/2 (2.1)

$$\text{Then, } OB = OA + AB = R + R/2 = 3R/2 \tag{2.2}$$

Then, in right-angled Δ OAP

$$OP = OA \sin 60^0 = \left(\sqrt{3}/2\right)R \tag{1.3}$$

Let the distance between the centers of two neighboring hexagonal cells, OQ, be denoted by ‘d’, then,

$$OQ = OP + PQ \text{ (where } OP = PQ)$$

Therefore, $$d = \left[\left(\sqrt{3}/2\right)R\right] + \left[\left(3\sqrt{2}\right)R\right]$$

$$\text{Hence, } d = \sqrt{3}R \tag{1.4}$$

Step 2: Area of a small hexagon, $A_{small\ hexagon}$ W

The area of a hexagonal cell with radius R is given as

$$A_{small\ hexagon} = \left(3\sqrt{3}/2\right) \times R^2 \tag{1.5}$$

Step 3: To find the relation between D, d and shift parameters

Let ‘D’ be the distance between the center of a particular cell under consideration to the centre of the nearest co-channel cell.

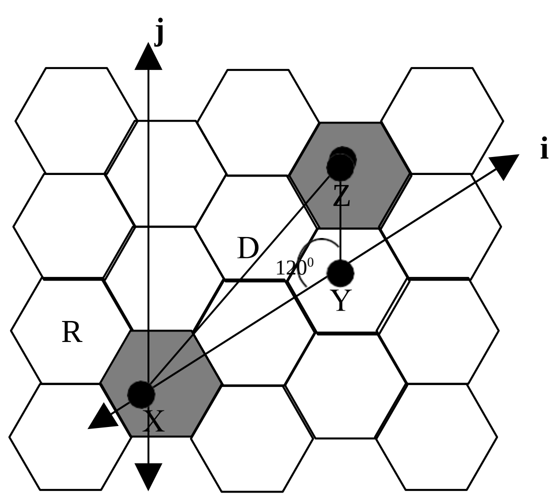


Fig. 1.6: Relationship Between K and Shift Parameters *(i & j)*

Using cosine formula Δ XYZ in Fig. 2.6, we have

$$XZ^2 = XY^2 + YZ^2 - 2 \times XY \times YZ \cos 120^0$$

or, $D^2 = (i \times d)^2 + (j \times d)^2 - 2 \times (i \times d) \times (j \times d) \cos 120^0$ D^2

$= (i \times d)^2 + (j \times d)^2 - 2 \times (i \times d) \times (j \times d) \times (-1/2) D^2 =$

$(i \times d)^2 + (j \times d)^2 + (i \times d) \times (j \times d)$

$$D^2 = d^2 (i^2 + j^2 + i \times j) \qquad (1.6)$$

$$D^2 = 3 \times R^2 \times (i^2 + j^2 + i \times j) \qquad (1.7)$$

Step 4: To find the area of a large hexagon, $A_{large\ hexagon}$

By joining the centers of the six nearest neighboring co-channel cells, a large hexagon is formed with radius equal to D, which is also the co-channel cell separation. Refer Fig. 1.7.

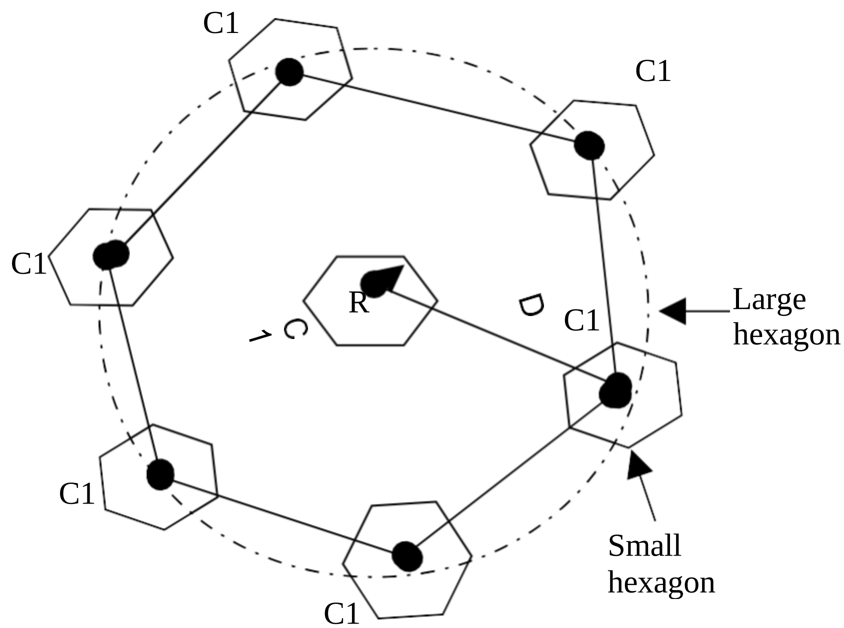


Fig. 1.7: Larger Hexagon in the First Tier

The area of the large hexagon having a radius D can be given as

$$A_{large\ hexagon} = \left(3\sqrt{3}/2\right) \times D^2 \qquad (1.8)$$

Using equation 1.7

$$A_{large\ hexagon} = \left(3\sqrt{3}/2\right) \times 3 \times R^2 \times \left(i^2 + j^2 + i \times j\right) \qquad (1.9)$$

Step 7: To find the number of cells in the large hexagon (L)

Number of cells in large hexagon

$$L = A_{large\ hexagon} / A_{small\ hexagon} \qquad (1.10)$$

Using equations 2.9, 2.5 & 2.10, we get

$$L = 3 \times (i^2 + j^2 + i \times j) \qquad (1.11)$$

Step 8: Find the correlation between L and cluster size K

It can be seen from Fig. 2.8, that the larger hexagon is created by joining the centers of co-channel cells present in the first tier contains 7 cells of the central cluster plus 1/3rd of the number of 7 cells of all the neighboring six clusters. Therefore, it can be calculated that the larger hexagon consisting of the central cluster of *K* cells plus 1/3rd the number of the cells connected with six neighboring clusters present in the first tier.

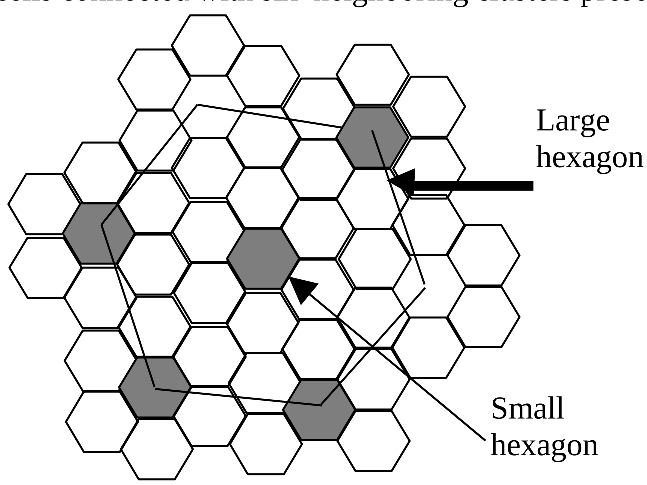


Fig. 2.8: Number of Clusters in the First Tier for N=7

Hence, the total number of cells enclosed by the larger hexagon is

$$L = K + 6 \times [(1 / 3) \times K)$$

$$L = 3 \times K \qquad (1.12)$$

Step 9: To establish relation between K and shift parameters

From equation 1.11 and 1.12, we get

$$3 \times K = 3 \times (i^2 + j^2 + i \times j)$$

$$K = (i^2 + j^2 + i \times j) \qquad (1.13)$$

The Table 1.1 shows the frequency reuse patterns along with the cluster sizes

Table 1.1: Frequency Reuse Pattern and Cluster Size

Frequency Reuse Pattern ***(I, j)***	**Cluster Size** $\boldsymbol{K = (i^2 + j^2 + i \times j)}$
(1, 1)	3
(2, 0)	4
(2, 1)	7
(3, 0)	9
(2, 2)	12
(3, 1)	13
(4, 0)	16
(2, 3)	19
(4, 1)	21
(5, 0)	25

Frequency Reuse Distance

To reuse the same set of radio channels in another cell, it must be separated by a distance called frequency reuse distance, which is generally represented by D.

Reusing the same frequency channel in different cells is restricted by co-channel interference between cells. So, it is necessary to find the minimum frequency reuse distance D in order to minimize the co-channel interference. Fig. 2.9 illustrates the separation of cells by frequency reuse distance in a cluster of 7 cells. In order to derive a formula to compute D, necessary properties of regular hexagon cell geometry are first discussed.

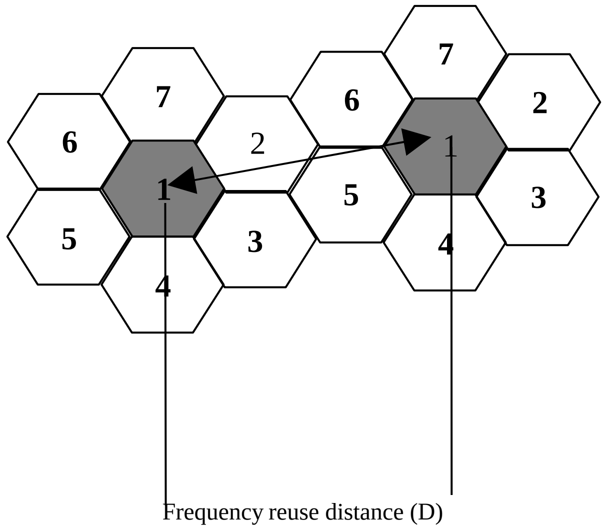


Fig. 1.9: Frequency Reuse Distance

The frequency reuse distance (D), which allows the same radio channel to be reused in co-channel cells, depends on many factors:

- the number of co-channel cells in the neighborhood of the central cell
- the type of geographical terrain
- the antenna height
- the transmitted signal strength by each cell-site

Suppose the size of all the cells in a cellular is approximately same, and it is usually calculated by the coverage area of the proper signal strength in every cell. The co-channel interference does not depend on transmitted power of each, if the cell size is fixed, i.e., the threshold level of received signal at the mobile unit is tuned to the size of the cell.
The co-channel interference depends upon the frequency reuse ratio, q, and is defined as

$$q = D / R$$

Where D is the distance between the two neighboring co-channel cells, and R is the radius of the cells. The parameter q is also referred to as the frequency reuse ratio or co-channel reuse ratio. The following steps are used to find the relationship between frequency reuse ratio q and cluster size K

Fig. 2.10 shows an array of regular hexagonal cells, where R is the cell radius. Due to the hexagonal geometry each hexagon has exactly six equidistant neighbors.

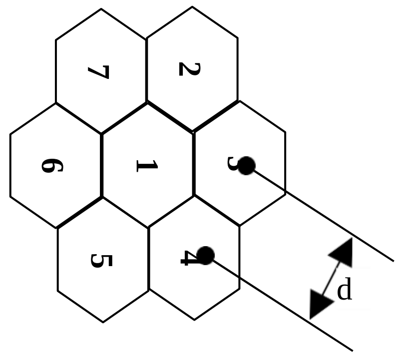


Fig. 1.10: Distance Between Two Adjacent Cells (d)

Let d be the distance between two cell centers of neighboring cells. Therefore,

$$d = \sqrt{3}R$$

The relationship between D, d, and shift parameters is

$$D^2 = 3 \times R^2 \times (i^2 + j^2 + i \times j)$$

As $K = i^2 + j^2 + i \times j$

$$D^2 = 3 \times R^2 \times K$$

$$\frac{D^2}{R^2} = 3 \times K$$

$$\frac{D}{R} = \sqrt{3K}$$

As $q = D/R$

$$q = \sqrt{3K}$$

Thus, the frequency reuse ratio q can be computed from the cluster size K. Table 2.2 shows the frequency reuse ratios for different cluster sizes, K

Table 1.2: Frequency Reuse Ratio and Cluster Size

Cluster Size K	**Frequency Reuse Ratio** $q = \sqrt{3K}$
3	3.00
4	3.46
7	4.58
9	5.20
12	6.00
13	6.24
19	7.55
21	7.94
27	9.00

As the D/R measurement is a ratio, if the cell radius is decreased, then the distance between co-channel cells must also be decreased by the same amount, for keeping co-channel interference reduction factor same. On the other hand, if a cell has a large radius, then the distance between frequency reusing cells must be increased proportionally in order to have the same D/R ratio.

As frequency reuse ratio (q) increases with the increase in cluster size (K), the smaller value of K largely increase the capacity of the cellular system. But it will also increase the co-channel interference. Therefore, the particular value of q (or K) is selected in order to keep the signal-to-cochannel interference ratio at an acceptable level. If all the antennas transmit the same power, then with the increase in K, the frequency reuse distance (D) increases, and reduce the likelihood that co-channel interference may occur. Therefore, the challenge is to get the optimal value of K so that the desired system performance can be achieved in terms of increased system capacity, efficient radio spectrum utilization and signal quality.

Channel Allocation Schemes

For effective utilization of the radio spectrum, a channel reuse scheme is required which must be able to increase the capacity and reduce interference. Several channel allocation schemes have been proposed to address these objectives. Channel allocation schemes are classified into *fixed, dynamic, and hybrid.* The selection of a particular channel allocation scheme influences the performance of the system, mainly how to manage the calls when a call is handed-over from one cell to another [190], [117], [186], [163].

In a fixed channel allocation scheme, a set of nominal channels are permanently allocated to each cell. Any call generated from within the cell can only be served by the idle radio channels present in that cell. If all the radio channels present in that cell are occupied, then the call is *blocked*. However, there exist a several variations of the fixed allocation. In one of the variation, a cell can borrow channels from neighboring cells if its own channels are already busy, and this scheme is called channel borrowing strategy. Such a borrowing procedure is being managed by mobile switching center (MSC) and it try to make sure that the borrowing of a radio channel form neighboring cells does not interferewith any of the existing calls present in the donor cell.

In a dynamic channel allocation scheme, cells are not allocated radio channels permanently. Instead, every time when a call is received, the serving base station (BS) enquires a channel from the MSC. The MSC allocates a channel to the cell after taking into consideration the possibility of future blocking rate of the candidate cell, the re-use distance of the channel, and several other parameters.

Therefore, the MSC then allocates a particular channel if that radio channel is currently not in use in the candidate cell as well in any other neighboring cell which falls inside the minimum channel reuse distance in order to avoid co-channel interference. The Dynamic channel allocation minimizes the possibility of blocking, thereby increasing the trunking capacity of the system, as all the available channels are accessible to all the cells. In Dynamic channel allocation schemes MSC gather information on traffic distribution, channel occupancy of all channels on a regular basis. This results in increased channel utilization with decreased probability of dropped and blocked calls, but at the same time the computational load on the system also increases.

Handoff Strategies

When a mobile moves from one cell to another cell when a call is in progress, the MSC automatically shifts the call to a new channel present in the new cell. This handoff operation requires the identification of a new base station, and channels that are associated with the new base station.

In any cellular network, managing handoff is very important job. Many handoff schemes give high priority to handover requests over new call requests while allocating free channels, and it must be performed successfully and as infrequently as possible. Therefore, in order to satisfy these requirements, optimum signal at which to begin a handoff level must be specified by system designers. When an optimal signal level for acceptable voice quality is specified, a somewhat stronger signal level is used as a threshold at which a handoff is made. This margin is given by A = $P_{r\ handoff}$-$P_{r\ minimam\_usable}$, and it should not be too large or too small. If A is very large, needless handoffs which canburden the MSC may take place, and if A is very small, there may not be a sufficient timeto complete a handoff process, before a call is vanished due to weak signal. Therefore, A should be carefully selected to meet these contradictory requirements. Fig. 1.11 shows a handoff situation. Fig. 1.11(a) presents a case in which a handoff does not take place and the signal strength falls below the minimum acceptable level in order to keep the channel active. This call dropping occurs when there is tremendous delay by the MSC in allocating a handoff, or when the threshold A is too small. During high traffic loads unnecessary delays may take place and this happens either due to computational overloading at the MSC or no free channels are available in any of the neighboring cells and thereby MSC has to wait until a free channel is found in a neighboring cell.

While deciding about handoff initiation time, it is important to make sure that the drop in the signal level is not due to temporary fading but the mobile is in fact moving away from its base station. Therefore, base station observes the signal strength for a definite period of time before a handoff begins. This signal strength measurement must be optimized in order to avoid unwanted handoffs, while ensuring that unwanted handoffs are completed before a call gets dropped. The time required to come to a decision if a handoff is needed, depends on the speed of the vehicle at which it is moving. Information about the speed of vehicle can also be calculated from the fading signal received at the base station.

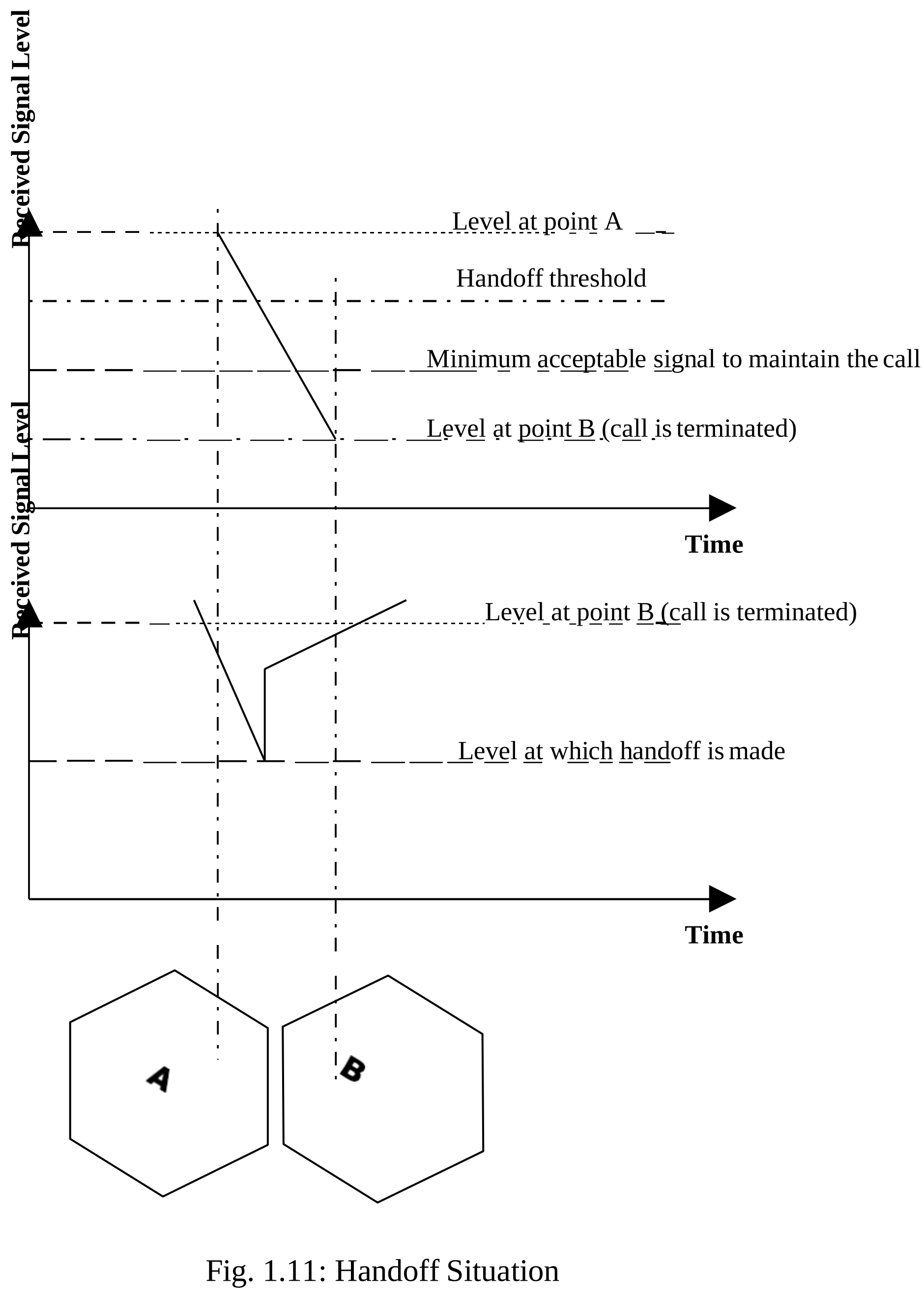


Fig. 1.11: Handoff Situation

The time during which a caller remains within a cell, without any handoff to the neighboring cells, is called the *dwell time* . The dwell time of a call depends upon a number of factors i.e. propagation, interference, distance between the caller and the base station, and several other time varying factors. It has been analyzed that variation of dwell time depends on the speed of the caller and the radio coverage type. e.g., a cell in

which radio coverage is provided to highway callers (using vehicles), a large number of callers have a moderately steady speed and they follow fixed paths with good radio coverage. For such instances, the dwell time for random caller is a random variable having distribution that is highly concentrated on the mean dwell time. Whereas, for callers present in dense, micro-cellular environments, there is normally a huge deviation of dwell time about the mean, and the dwell times in general are shorter than the cell geometry. It is clear that the information of dwell time is very important while designing handoff algorithms.

In first generation cellular systems, signal strength computations are done by the base stations and monitored by the MSC. All the base stations regularly observe the signal strengths of its reverse channels to find out the relative location of each mobile user with respect to the base station. In addition to calculating the radio signal strength indication (RSSI) of ongoing calls in the cell, an extra receiver in each base station, is used to find out signal strengths of mobile users present in the neighboring cells. The extra receiver is controlled by the MSC and is used to examine the signal strength of callers in the neighboring cells, and informs RSSI to the MSC. Based on the RSSI values received from each extra receiver, the MSC determines whether handoff is required or not.

In second generation cellular systems using digital TDMA technology, handoff decisions are *mobile assisted.* In *mobile assisted handoff (MAHO),* each mobile station measures the received power from the neighboring base stations and informs these results to the serving base station. A handoff starts when the power received from the base station of a neighboring cell go above the power received from the present base station. In MAHO scheme, the call to be handed off between different base stations at a lot faster speed than in first generation systems because the handoff computations are done by each mobile and by keeping the MSC out of these computations. MAHO is suitable for micro-cellular network architectures where handoffs are more frequent.

When a call is in progress, if a mobile shifts from one cellular system to an another cellular system managed by a different MSC, an *intersystem handoff* is required. AnMSC performs an intersystem handoff when a signal goes weak in a particular cell and

the MSC fails to find another cell inside its system to which it can move the ongoing call, and several issues should be addressed while intersystem handoff is implemented. e.g. a local call might automatically turn into a long-distance call when the caller shifts out of its home network and enters into a neighboring system.

Various systems have different methods for dealing with hand-off requests. Several systems manage handoff requests in the same way as they manage new call requests. In such systems, the possibility that a handoff call will not be served by a new base station is equivalent to the blocking probability of new calls. However, if a call is terminated unexpectedly while in progress is more frustrating than being blocked occasionally on a new call. Therefore, to improve the quality of service, various methods have been created to give priority to handoff call requests over new call requests while allocating channels.

Prioritizing Handoffs

One scheme for prioritizing handoffs call requests is called the *guard channel concept,* in which a part of the existing channels in a cell is reserved entirely for handoff call requests. The major drawback of this scheme is that it reduces the total carried traffic, as smaller number of channels is allocated to new calls. However, guard channels scheme present efficient spectrum utilization when dynamic channel allocation strategies are used.

Queuing of handoff calls is another way to minimize the forced call terminations due to unavailability of channels in the cell. There is actually a tradeoff between the minimization in the possibility of forced call termination of handoff calls and total carriedtraffic. Handoff call queuing is possible as there is a fixed time interval between the time the received signal strength falls below the handoff threshold and the time the call is terminated due to unavailability of signal strength. The queue size and delay time is calculated from the traffic pattern of the service area. It should be noted that queuing of handoff calls does not promise a zero forced call terminations, because large delays will force the received signal strength to fall below the minimum level required to maintain communication and therefore, lead to forced handoff call termination.

Interference and System Capacity

Interference is one of the major factors affecting the performance of cellular radio systems. Sources of interference consist of another mobile inside the same cell, an ongoing call in a neighboring cell, other base stations transmitting signal in the same fre- quency band, or any non-cellular system which accidentally transmits energy into the cellular frequency band. Interference on voice signals could give rise to cross talk, where the caller hears interference in the background due to the presence of an unwanted transmission. The presence of interference in control channels, gives rise to missed and blocked calls. Interference is very dangerous in urban areas, due to the presence of larger base stations and mobile with greater RF noise. Interference has been accepted as a majorobstruction in increasing the capacity of a system and is largely responsible for dropped calls in a network. The two major types of interferences that are taken consideration while allocating channels to the calls are *co-channel* and *adjacent channel interference.* While interfering signals are generated inside the cellular system by cellular transmitters, but they are difficult to control. The interference due to out-of-band users is very difficult to control, which happens without any word of warning, because of front end overload of subscriber equipment or intermittent inter-modulation products.

Co-channel Interference and System Capacity

The channel reuse approach is very useful for increasing the efficiency of radio spectrum utilization but it results in co-channel interference because the same radio channel is repeatedly used in different co-channel cells in a network. In this case, the quality of a received signal is very much affected both by the amount of radio coverage area and the co-channel interference.

Co-channel interference takes place when two or more transmitters located within a wireless system, or even a neighboring wireless system, which are transmitting on the same radio channel. Co-channel interference happens when the same carrier frequency (base station) reaches the same receiver (mobile phone) from two different transmitters.

This type of interference is generally generated because channel sets have been allocated to two different cells that are nor far enough geographically, and their signals are strong enough to cause interference to each other. Thus, co-channel interference can either modify the receiver or mask the particular signal. It may also merge with the particular signal to cause severe distortions in the output signal.

The co-channel interference can be evaluated by picking any particular channel and transmitting data on that channel at all co-channel sites. In a cellular system with hexagonal shaped cells, there are six co-channel interfering cells in the first tier. Fig. 2.12shows a Test 1 which is set-up to calculate the co-channel interference at the mobile unit, in this test mobile unit is not stationary but is continuously moving in its serving cell.

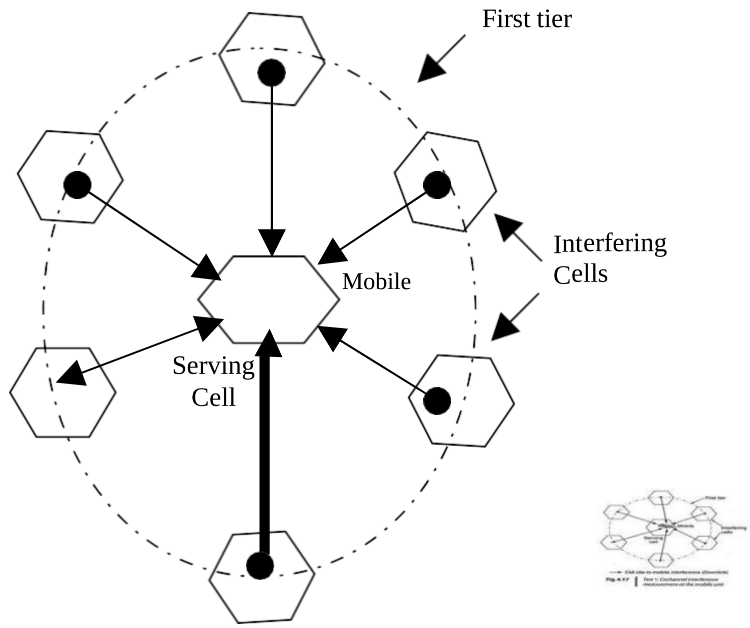


Fig. 1.12: Co-channel Interference Measurement at the Mobile Unit

In a small cell system, interference will be the major dominating factor and thermal noise can be neglected. Thus the S/I can also be written as:

$$\frac{S}{I} = \frac{1}{\sum_{k=1}^{6}\left(\frac{D_k}{R}\right)^{-\gamma}} \quad (1.14)$$

where

S/I = Signal to interference ratio at the desired mobile receiver,

S = desired signal power,

I = Interference power,

$2 \le \gamma \le 5$ is the propagation path-loss slope and γ depends on the terrain environment.

If we assume, for simplification, that D_k is the same for the six interfering cells, i.e., $D = D_k$, then the formula above becomes:

$$\frac{S}{I} = \frac{1}{6(q)^{-\gamma}} = \frac{q^{\gamma}}{6} \qquad (1.15)$$

1

For analog systems using frequency modulation, normal cellular practice is to specify an S/I ratio to be 18 dB or higher based on subjective tests. An S/I of 18 dB is the measured value for the accepted voice quality from the present-day cellular mobile receivers.

Using an S/I ratio equal to 18dB ($10^{18/10} = 63.1$) and γ =4 in the Eq. (1.15), then

$$q = [6 \times 63.1]^{0.25} = 4.41 \qquad (1.17)$$

Substituting q from Eq. (2.17) into Eq. (2.12) yields

$$N = \frac{(4.41)^2}{3} = 6.49 \approx 7\,. \qquad (1.18)$$

Eq. (1.18) indicates that a 7-cell reuse pattern is needed for an S/I ratio of 18 dB.

Therefore, the performance of interference-limited cellular mobile system can be calculated from the following results.

a) If the signal-to-interference ratio (S/I) is greater than 18 dB, then the system is said to be correctly designed.

b) If S/I is less than 18 dB and signal-to-noise ratio (S/N) is greater than 18 dB, then the system is said to be experiencing with a co-channel interference problem.

c) If both S/I and S/N are less than 18 dB and S/I is approximately same as S/N in a cell, then the system has a radio coverage problem.

d) If both S/I and S/N are less than 18 dB and S/I is less than S/N, the system has both co-channel interference and radio coverage problem.

Therefore, the reciprocity theorem can be used to study the radio coverage problem, but it does not give accurate results when used for the study of co-channel interference problem. Therefore, it is suggested to perform Test 2 in order to measure co-channel interference at the cell-site. In Test 2 shown in Fig. 2.13, both the mobile unit present in the serving cell and six other mobile units present in the neighboring cells are transmitting simultaneously at the same channel.

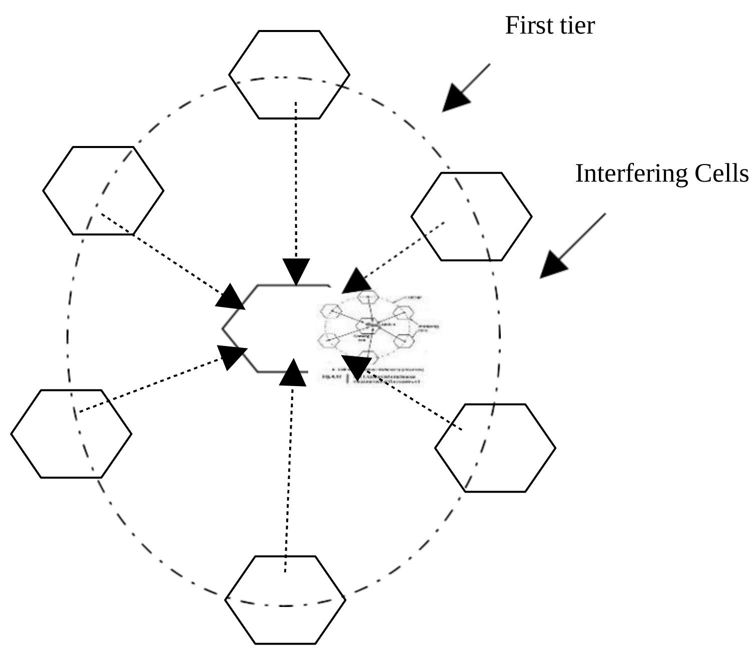


Fig. 1.13: Co-channel Interference Measurement at the Cell-site

The received signal strength measurements are done at the serving cell, under the following conditions:

- When only the mobile unit present in the serving cell transmits (signal measured as S)
- Up to six interference levels are measured at the serving cell-site due to presence of six mobile units in the neighboring cells (the average signal measured as I)
- Noise from sources other than mobile unit (signal measured as N)

Then the received S/I and S/N is computed at the serving cell site. The test results are compared with the Test 1, and from the it can be easily found whether the cellular system has a radio coverage or a co-channel interference problem or both.

Co-channel Interference Reduction Methods

Interference is major factor affecting the performance of cellular communication systems. Sources of interference may consist of a different mobile working in the same or in the neighboring cells, which are operating in the same frequency band that may leak energy into the cellular band.

Cells that use same set of radio channels are called co-channel cells, and the interference caused by the received signals coming from these cells is called co-channel interference. If the different cells in the cellular network use different radio channels then the inter-cell interference should be kept at a minimum level. When the number of mobile users increase and the radio channels available in the system are limited, then, in order to satisfy this high demand, the radio channels have to be reused in various cells. That is why for increasing the capacity, there exist many co-channel cells which can simultaneously serve the large number of users.

In fact, Deployment of radio channel reuse is required to improve the capacity of a system. But, the reuse mechanism brings in co-channel interference from neighboring cells using the same set of radio channels. Therefore, the quality of received signal gets affected by the amount of co-channel interference and the extent of radio coverage. Therefore, frequency reuse should be planned very carefully in order to keep the co- channel interference at an acceptable level.

The co-channel interference can be reduced by the following methods:

a. *Increasing the distance(D) between two co-channel cells, D*

   As D increases, the strength of interfering signal from co-channel interfering cells decreases significantly. But it is not wise to increase D because as D is increased, K must also be increased. High value of K means fewer number of radio channels are available per cell for a given spectrum. This results into decrease of the system capacity in terms of channels that are available per cell.

b. *Reducing the antenna heights*

   Reducing antenna height is a good method to minimize the co-channel interference in some environment, e.g., on a high hill. In the cellular system design effective antenna height is considered rather than the actual antenna height.Therefore, the effective antenna height changes according to the present location of the mobile unit in such a difficult terrain.

   When the antenna is put up on top of the hill, the effective antenna height gets more than the actual antenna height. So, in order to minimize the co-channel interference, antenna with lower height should be used without decreasing the received signal strength either at the cell-site or at the mobile device. Similarly, lower antenna height in a valley is very useful in minimizing the radiated power in a far-off high-elevation area where the mobile user is believed to be present.

   However, reducing the antenna height does not always minimize the co-channel interference, e.g., in forests, the larger antenna height clears the tops of the longest trees in the surrounding area, particularly when they are located very close to the antenna. But reducing the antenna height would not be appropriate for minimizing co-channel interference because unnecessary attenuation of the signal would occur in the vicinity of the antenna as well as in the cell boundary if the height of the antenna is below the treetop level.

c. *Using directional antennas.*

   The use of directional antennas in every cell can minimize the co-channel interference if the co-channel interference cannot be avoided by a fixed division of co-channel cells. This will also improve the system capacity even if the traffic increases. The co-channel interference can be further minimized by smartly setting up the directional antenna.

d. *Use of diversity schemes at the receiver*.

The diversity scheme used at the receiving end of the antenna is an efficient technique for minimizing the co-channel interference because any unwanted action performed at the receiving end to increase the signal interference would notcause further interference. For example, the division of two receiving antennas installed at the cell-site meeting the condition of h/s=11, (where h is the antenna height and s is the division between two antennas), would produce the correlation coefficient of 0.7 for a two-branch diversity system. The two correlated signals can be combined with the use of selective combiner. The mobile transmitter couldsuffer up to 7 dB minimization in power and the same performance at the cell-site can be achieved as a non-diversity receiver. Therefore, interference from the mobile transmitters to the receivers can be significantly reduced.

Adjacent Channel Interference

Signals from neighboring radio channels, also called adjacent channel, leak into the particular channel, thus causing adjacent channel interference. Adjacent channel interference takes place due to the inability of a mobile phone to separate out the signals of adjacent channels allocated to neighboring cell sites (e.g., channel 101 in cell A, and channel 102 in cell B), where both A and D cells are present in the same reuse cluster. The problem of adjacent channel interference can become more serious if a user transmitting on a channel, which is extremely close to a subscriber's receiver channel, while the receiver tries to receive a signal from base station on the desired channel. Thisis called the *near and far* effect, where a neighboring transmitter catches the receiver of the user. Otherwise, the near-far effect occurs when a mobile near to a base station transmits on a channel which is close to the one being used by a weak mobile. The base station may find some trouble in separating out a particular user from the one using adjacent channel

Adjacent channel interference can be reduced through careful and thorough filtering and efficient channel allocations. As each cell is allocated only a portion of the total channels, a cell must not be allocated channels which are located adjacent in frequency. By maintaining the channel separation as large as possible in a given cell, the adjacent channel interference may well be minimized significantly. Hence, instead of allocating contiguous band of channels to each cell, channels are allocated such a way that the fre- quency separation between channels in a given cell should be maximized. With sequentially allocating consecutive channels to various cells, several channel allocation schemes are capable enough to keep apart adjacent channels present in a cell with bandwidth of N channels, where N is the size of a cluster. However, some channel allocation schemes also avoid a secondary source of adjacent channel interference by not using the adjacent channels in neighboring cells.

Trunking and Grade of Service

In cellular mobile communication, the two major aspects that have to be considered with extra care are: trunking, and grade of service. These aspects are to be planned very well in order to get a better system performance. The grade of service is a standard which is used to define the performance of a cellular mobile communication system by specifyinga desired probability of a mobile user acquiring channel access, when a definite number of radio channels are present in the system. The cellular communication network dependson a trunking system to fit large number of mobile users in a limited radio band. The statistical behavior of mobile users is being exploited by trunking so that a fixed number of channels can be allocated to large mobile users. In trunking, large number of mobile users is being accommodated to share the limited radio channels available in a cell.

In trunked cellular communication systems, each mobile user present in network is allocated a channel on the basis of a request. After the call is terminated, the occupied channels immediately go back to the pool of available channels. When a mobile user made a request for channels and if all of the radio channels are occupied, then the incoming call is blocked. In few communication systems, a queue is generally used to keep the requesting mobile users until a channel becomes free. The grade of service (GOS) is used to determine the capability of a user to get access to trunked radio systems during busy hours. The busy hour is generally based on customer’s request for channels during peak load.

It is, therefore, necessary to approximate the maximum required capacity in terms of number of available channels and to allocate the appropriate number of channels in order to meet the GOS. GOS is generally defined as the probability that a call is blocked. A call which cannot get completed after the call request is made by a user is called a blocked or lost call, and it may happen either due to channel congestion or due to the non- availability of a free channel. Therefore, GOS can be computed from channel congestion which is defined as the call blocking probability, or being delayed beyond a certain time.

The traffic intensity (A_u Erlangs) generated by each user is

$$A_u = \lambda H$$

where λ is the average number of calls generated per unit time and H is the average duration of each call. If A is having U users and number of channels are not mentioned, then the total offered traffic intensity A is

$$A = UA_u$$

Additionally, if a trunked system is having C channels, and the traffic is equally divided between the channels, then the intensity of traffic (A_c) for each radio channel is

$$A_c = UA_u/C \qquad (1.19)$$

Note that when the offered traffic goes past the maximum capacity of the system, the total carried traffic gets very limited due to the limited number of channels. The maximum possible carried traffic is the total number of channels, C, in Erlangs. The AMPS system is generally developed for a GOS of 2% blocking and it shows that 2 outof 100 calls will be blocked because channels are occupied during the busiest hour.

Different types of trunked radio systems commonly used in the networks are:

1. In the first type, no queuing is offered for call requests i.e., for each user who requests service, there exists no setup time and if free radio channel is available, it is immediately allocated to the user. If all the channels are busy, then the requesting user is blocked. In this trunking system, it is assumed that call arrival follows a Poisson distribution and the trunking is also called *blocked calls cleared*. Moreover, it

is also assumed that there are unlimited users in the network and having the following additional features:

(a) The channel request can be made at any time by all the mobile users (both new and blocked users); (b) the probability of a user being allocated a channel is exponentially distributed, therefore, occurrence of longer call duration is very unlikely as explained by an exponential distribution; and (c) there are a fixed number of channels present in the trunking pool, and it is known as an M/M/m queue, which leads us to the derivation of the Erlang B formula. The Erlang B formula helps in finding the probability that a call is blocked and also measures the GOS for a trunked radio system which does not provide queuing for blocked calls. The Erlang B formula is

$$Pr\ [blocking] = \frac{\frac{A^C}{C!}}{\sum_{k=0}^{C} \frac{A^k}{k!}} = \text{GOS} \tag{1.20}$$

where C is the number of trunked channels present in the trunked radio system and A is the offered traffic. It is possible to design a trunked systems with fixed number of users, but the final expressions are found to be very complex than the Erlang B, and the added complexity is not acceptable for typical trunked radio systems in which number of users are more than the channels present in the system.. The capacity of a trunked radio system in which blocked calls are lost is shown in Table 2.3.

Table 2.3: 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	0.9OO	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

2. In a second form of trunked networks, a queue is used to keep the blocked calls. If all the channels are presently busy, then the call can be postponed until a free channel is found, and this whole process is *Blocked Calls Delayed,* and the GOS for this type of trunking is the probability that a new call is not allocated to a channel even after waiting a certain time in the queue. The probability that a new call is not allocated a channel immediately is calculated by the Erlang C formula

$$Pr\,[delay > 0] = \frac{A^C}{A^C + C!\left(1 - \frac{A}{C}\right)\sum_{k=0}^{C-1}\frac{A^k}{k!}} \qquad (1.21)$$

If no channel is currently found free then the call is delayed, and the GOS of a trunked system in which the blocked calls are delayed is given by

$$\begin{aligned} Pr[delay>t] &= Pr\,[delay > 0]\,Pr\,[delay > t \mid delay > 0] \qquad (1.22) \\ &= Pr\,[delay > 0]\exp(-(C-A)t/H) \end{aligned}$$

For all the calls n a queued system the average delay D is given by

$$D = Pr\,[delay > 0]\,\frac{H}{C - A} \qquad (1.23)$$

Improving Capacity In Cellular Systems

With the rise in the demand for wireless services, the number of radio channels allocated to each cell could become inadequate in order to satisfy this increase in the demand. Therefore, to increase the capacity (i.e. a cellular system can take up more calls) of a cellular system, it is very important to allocate more number of radio channels to each cell in order to meet the requirements of mobile traffic. Various techniques that are proposed for increasing the capacity of a cellular system is as follows:

i. Cell splitting

ii. Cell sectoring

iii. Repeaters for extending range

iv. Micro zone method

Cell Splitting

Cell splitting is a method in which congested (heavy traffic) cell is subdivided into smaller cells, and each smaller cell is having its own base station with reduction in antenna height and transmitter power. The original congested bigger cell is called macrocell and the smaller cells are called microcells. Capacity of cellular network can be increased by creating micro-cells within the original cells which are having smaller radiusthan macro-cells, therefore, the capacity of a system increases because more channels perunit area are now available in a network.

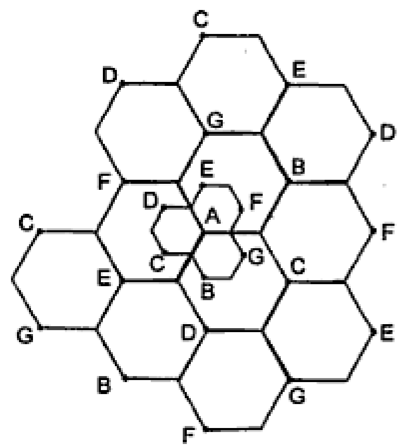


Fig. 1.14: Cell Splitting

Fig. 2.14 shows a cell splitting in which a congested cell, divided into smaller micro- cells, and the base stations are put up at corners of the cells. The micro-cells are to be added in such a way in order to the frequency reuse plan of the system should be preserved. For micro-cells, the transmit power of transmitter should be reduced, and each micro-cell is having half the radius to that of macro-cell. Therefore, transmit power of thenew cells can be calculated by analyzing the received power at the cell boundaries. Thisis required in order to make sure that frequency reuse plan for the micro-cells is also working the same way as it was working for the macro-cells.

$$P_{r\text{-}O} \propto P_{tp} R^{-n}$$

$$P_{r\text{-}N} \propto P_{tN}\left(\frac{R}{2}\right)^{-n}$$

Where P_{tp} is the transmit power of macro-cell

P_{tN} is the transmit power of macro-cell

n is the path loss exponent

$R, \left(\frac{R}{2}\right)$ is the radius of macro and micro-cells

In cell splitting, following factors should be carefully monitored;

1. In cell splitting, allocation of channels to the new cells (micro-cells) must be done very cautiously. So, in order to avoid co-channel interference, cells must follow the minimum reuse distance principle.

2. Power levels of the transmitters for new and old cells must be redesigned. If the transmitter of the old cell has the same power as that of new cells, then the channels in old cell interfere with the channels of new cell. But, if the power level of transmitter is too low then it may result into in sufficient area coverage.

3. In order to overcome the problem of point (2); the channels of macro-cell is divided into two parts. The channels in the first part are for the new cell and other part consists of channel for the old cell. Splitting of cells is done accordingto the number of subscribers present in the areas, and the power levels of the transmitters must be redesigned according to the allocated channels to old and new cells.

4. Antennas of different heights and power levels are used for smooth and easy handoff, and this technique is called Umbrella cell approach. Using this approach large coverage area is provided for high speed users and small coverage area to low speed users. Therefore, the number of call handoffs is maximized for high speed users and provides more channels for slow speed users.

5. The main idea behind cell splitting is the rescaling of entire system. In cell splitting, reuse factor (D/R) is kept constant because by decreasing the radius of cell (R) and, at the same time, the separation between co-channel (D) is also decreased. So, high capacity can be achieved without changing the (D/R) ratio of system.

Sectoring

Another way of improving the channel capacity of a cellular system is to decrease the D/R ratio while keeping the same cell radius. Improvement in the capacity can be accomplished by reducing the number of cells in a cluster, hence increasing the frequency reuse. To achieve this, the relative interference must be minimized without decreasing the transmit power.

For minimizing co-channel interference in a cellular network, a single omni-directional antenna is replaced with multiple directional antennas, with each transmitting within a smaller region. These smaller regions are called sectors and minimizing co-channel interference while improving the capacity of a system by using multiple directional antennas is called sectoring. The amount up to which co-channel interference is minimized depends on the amount of sectoring used. A cell is generally divided either into three 120 degree or six 60 degree sectors. In the three-sector arrangement, three antennas are generally located in each sector with one transmit and two receive antennas. The placement of two receive antennas provide antenna diversity, which is also known as space diversity. Space diversity greatly improves the reception of a signal by efficiently providing a big target for signals transmitted from mobile units. The division between the two receive antenna depends on the height of the antennas above ground.

When sectoring technique is used in cellular systems, the channels used in a particular sector are actually broken down into sectored groups, which are only used inside a particular sector. With 7-cell reuse pattern and 120 degree sectors, the number of interfering cells in the neighboring tier is brought down from six to two. Cell sectoring also improves the signal-to-interference ratio, thereby increasing the capacity of a cellular system. This method of cell sectoring is very efficient, because it utilized the existing system structures. Cell sectoring also minimized the co-channel interference, with the use of directional antennas, a particular cell will get interference and transmit only a fraction of the available co-channel cells.

It is seen that the reuse ratio $q = (N_I \times S/I)^{1/n}$, where N_I depends on the type of antenna used. For an omni-directional antenna with only first-tier of co-channel interferer, the number of co-channel interfering cells $N_I = 6$, but for a 120 degree directional antenna, it is 2

So, the increase in S/I ratio is

$$\frac{(N_I \times S/I)_{120^0}}{(N_I \times S/I)_{omni}} = \frac{q^n_{120^0}}{q^n_{omni}}$$

$$\frac{(S/I)_{120^0}}{(S/I)_{omni}} = 3$$

n = path loss exponent $\quad$ N_I = Number of co-channel interfering cells

q = frequency reuse ratio = D/R

Thus, S/I ratio increases with the increase in number of sectors, but at the cost of additional handoff that might be required for the movement of a user from one sector to another.

Microcell Zone Concept

The micro-cell zone concept is associated with sharing the same radio equipment by different micro-cells. It results in decreasing of cluster size and, therefore, increase in system capacity. The micro-cell zone concept is used in practice to improve the capacity of cellular systems.

To improve both capacity and signal quality of a cellular system, cell sectoring depends upon correct setting up of directional antennas at the cell-site. But it also gives rise to increase in the number of handoffs and trunking inefficiencies. In a 3-sector or 6-sector cellular system, each sector acts like a new cell with a different shape and cell. Channels allocated to the un-sectored cell are divided between the different sectors present in a cell, thereby decreasing number of channels available in each sector. Furthermore, handoff takes place every time a mobile user moves from one sector to another sector of the same cell. This results in significant increase of network load on BSC and MSC of thecellular system. The problem of channel partitioning and increase in network load become very hard if all the 3 or 6-sectored directional antennas are placed at the centre ofthe cell.

As shown in the Fig. 1.15, three directional antennas are put at a point, Z1, also called zone-site, where three adjacent cells C1, C2, and C3 meet with each other. Z1, Z2 and Z3 are three zone-sites of the cell C1, and each zone-site is using three 135 degree directional antennas. All the three zone-sites also behave as receivers, which also receive signals transmitted by a mobile user present anywhere in the cell. All the three zone-sites are linked to one common base station, as shown in Fig. 2.16. This arrangement is knownas Lee's micro-cell zone concept.

In order to avoid delay, these zone-sites are connected through a high-speed fiber link to the base station. The base station first finds out, which of the three zone-sites has the better received signal strength from the mobile user and then that particular zone-site is used to transmit the signal to the mobile user. Therefore, only one zone-site is active at a time for communicating with the user and it also minimizes the co-channel interference experienced by the mobile user.

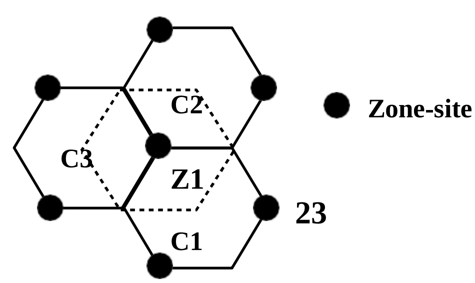


Fig. 1.15: Location of Zone-sites in Sectored Cells

Therefore, micro-cell zone architecture minimizes the co-channel interference, improves system capacity, demands less handoffs, and the system is easy to implement. The system capacity for a system with cluster size k=3 is 2.33 times greater than the present analog cellular system with k=7 for the C/I requirement of 18 dB. This micro-cell system gives improved voice quality than the AMPS cellular system at 850 MHz. The micro-cell zone concept can be used with both digital communication systems and personal communication systems, and is best suited for indoor applications. It is also very useful toprovide services along highways or in crowded urban areas.

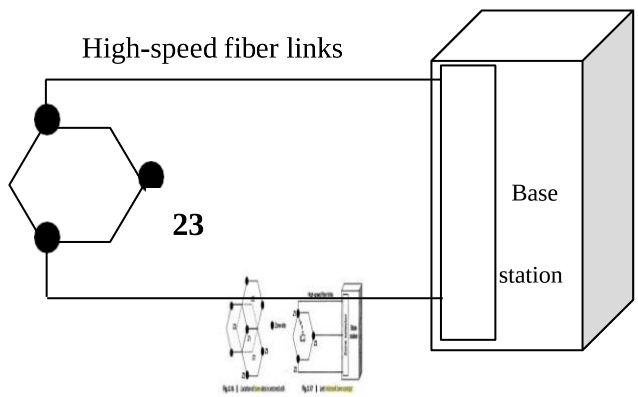


Fig. 1.16: Lee's Microcell Zone Concept

Advantages of micro zone concept:

1. When the mobile user moves from one zone to another within the same cell, the mobile user can keep the same channel for the call progress.
2. The effect of interference is very low due to the installation of low power transmitters.
3. Better signal quality is possible.
4. Fewer handoffs when a call is in progress.

Repeaters for Range Extensions

Wireless operators want to provide dedicated coverage for users located within buildings, or in valleys or tunnels as these areas are sometimes very hard to reach. Radio re-transmitters, also known as repeaters, are frequently used to provide coverage in such areas where range extension capabilities are required. Repeaters are bidirectional devices,as the signals can be concurrently transmitted to and received from a base station.

Repeaters may be installed anywhere as they function using over the air signals, and are able repeat entire frequency band. After receiving signals from base station, the repeater amplifies the signals before it forwards them to the coverage area. As repeaters can also reradiate the received noise, so repeaters must be installed very carefully. Directional antennas or distributed antenna system (DAS) are linked practically to the repeater inputs or outputs for spot coverage, mainly in tunnels or buildings.

A service provider dedicates some amount of cell site's traffic for the areas covered by the repeater by modifying the coverage of cell. As the repeater do not add more channels to the system, it only reradiates the base station signal into specific locations. Repeaters are generally used to provide coverage into those areas, where signal reception has been very weak. Signal penetration inside the building is generally provided by installing micro-cells outside the big building, and installing many repeaters inside the buildings. This technique provides better coverage into targeted areas, but does not increase the capacity that is required with the rise in the indoor and outdoor traffic. Therefore, dedicated base stations inside buildings are required to meet the service demands of large number of cellular users present inside the building. Finding a proper location for repeaters and distributed antennas inside the building needs a very careful planning, mainly due to the interference signals reradiated into the building. Also, repeaters must be able to match the available capacity from the base station. Software SitePlanner helps the engineers to decide the best location for putting up the repeaters and DAS network.

Conclusion

The fundamental concepts of frequency reuse, frequency planning, handoff, and trunking efficiency are presented in this chapter. The performance determining parameters such as grade of service, spectrum efficiency, and radio capacity under diverse situations are also discussed. Handoffs are essential to pass mobile traffic from one cell to another, and there is variety of different ways to implement handoffs. The capacity of a cellular system depends upon several variables. The *S/I* influence the frequency reuse factor of a cellular system, which restricts the number of radio channels within the coverage area. The number of users in a particular area is greatly influe

UNIT II

Mobile Radio Propagation: Link Calculation and Antenna System

Free Space Propagation Model:

The free space propagation model is used to predict received signal strength when the transmitter and receiver have a clear, unobstructed line-of-sight path between them. Satellite communication systems and microwave line-of-sight radio links typically undergo free space propagation. As with most large- scale radio wave propagation models, the free space model predicts that received power decays as a function of the T-R separation distance raised to some power (i.e. a power law function). The free space power received by a receiver antenna which is separated from a radiating transmitter antenna by a distance d, is given by the Friis free space equation,

$$P_r(d)= P_tG_tG_r\lambda^2/(4\pi d)^2$$

where P_t is the transmitted power, P_r (d) is the received power which is a function of the T-R separation, G_t is the transmitter antenna gain, G_r is the receiver antenna gain, d is the T-R separation distance in meters and λ is the wavelength in meters. The gain of an antenna is related to its effective aperture, A_e by,

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

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

$$\lambda = c/f= 2\pi c/\omega_c$$

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

An isotropic radiator is an ideal antenna which radiates power with unit gain uniformly in all directions, and is often used to reference antenna gains in wireless systems. The e/7'ective isotropic radiated power (EIRP) is defined as

$$EIRP = P_tG_t$$

and represents the maximum radiated power available from a transmitter in the direction of maximum antenna gain, as compared to an isotropic radiator. In practice, effective radiated power (ERP) is used instead of EIRP to denote the maximum radiated power as compared to a half-wave dipole antenna (instead of an isotropic antenna).

The path loss, which represents signal attenuation as a positive quantity measured in dB, is defined as the difference (in dB) between the effective transmitted power and the received power, and may or may not include the effect of the antenna gains. The path loss for the free space model when antenna gains are included is given by

$$PL\ (dB) = 10\log(P_t/P_r) = -10\log[P_tG_tG_r\lambda^2/(4\pi d)^2]$$

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

$$PL\ (dB) = 10\log(P_t/P_r) = -10\log[\lambda^2/(4\pi d)^2]$$

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

$$d_f = 2D^2/\lambda$$

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

$$d_f >> D$$

 # 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 Pr(10 km) ? Assume unity gain for the receiver antenna.

The Three Basic Propagation Mechanisms:

Reflection, diffraction, and scattering are the three basic propagation mechanisms which impact propagation in a mobile communication system.

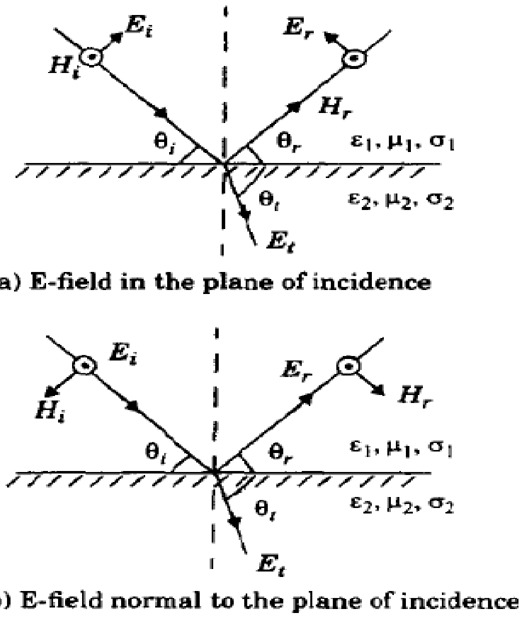
Reflection occurs when a propagating electromagnetic wave impinges upon an object which has very large dimensions when compared to the wavelength of the propagating wave. Reflections occur from the surface of the earth and from buildings and walls.

Diffraction occurs when the radio path between the transmitter and receiver is obstructed by a surface that has sharp irregularities (edges). The secondary waves resulting from the obstructing surface are present throughout the space and even behind the obstacle, giving rise to a bending of waves around the obstacle, even when a line-of-sight path does not exist between transmitter and receiver. At high frequencies, diffraction, like reflection depends on the geometry of the object, as well as the amplitude, phase, and polarization of the incident wave at the point of diffraction.

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. In practice, foliage, street signs, and lamp posts induce scattering in a mobile communications system.

Reflection:

When a radio wave propagating in one medium impinges upon another medium having different electrical properties, the wave is partially reflected and partially transmitted. If the plane wave is incident on a perfect dielectric, part of the energy is transmitted into the second medium and part of the energy is reflected back into the first medium, and there is no loss of energy in absorption. lf the second medium isa perfect conductor, then all incident energy is reflected back into the first medium without loss of energy.The electric field intensity of the reflected and transmitted waves may be related to the incident wave in the medium of origin through the Fresnel reflection coefficient (Γ). The reflection coefficient is a functionof the material properties, and generally depends on the wave polarization, angle of incidence, and the frequency of the propagating wave.



(a) E-field in the plane of incidence

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

Figure 3.4
Geometry for calculating the reflection coefficients between two dielectrics.

Reflection from dielectrics:

Figure 3.4 shows an electromagnetic wave incident at an angle θ_i with the plane of the boundary between two dielectric media. As shown in the figure, part of the energy is reflected back to the first media at an angle θ_r, and part of the energy is transmitted (refracted) into the second media at an angle θ_t. The nature of reflection varies with the direction of polarization of the E-field. The behavior for arbitrary directions of polarization can be studied by considering the two distinct cases shown in Figure

The plane of incidence is defined as the plane containing the incident, reflected, and transmitted rays. In Figure 3.4a, the E—field polarization is parallel with the plane of incidence (that is, the E-field has a vertical polarization, or normal component, with respect to the reflecting surface) and in Figure 3.4b, the E-field polarization is perpendicular to the plane of incidence (that is, the incident E-field is pointing out of the page towards the reader, and is perpendicular to the page and parallel to the reflecting surface).

Because of superposition, only two orthogonal polarizations need be considered to solve general reflection problems. The reflection coefficients for the two cases of parallel and perpendicular E-field polarization at the boundary of two dielectrics are given by

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

$$\Gamma_{\perp} = \frac{E_r}{E_i} = \frac{\eta_2 \sin\theta_i - \eta_1 \sin\theta_t}{\eta_2 \sin\theta_i + \eta_1 \sin\theta_t} \quad \text{(E-field not in plane of incidence)}$$

Where η is the intrinsic impedance of the respective medium.

Or,

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

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

Where ε is the permittivity of the respective medium.

Brewster Angle:

The Brewster angle is the angle at which no reflection occurs in the medium of origin. It occurs when the incident angle BB is such that the reflection coefficient $\Gamma_{\parallel}$ is equal to zero (see Figure 3.6), The Brewster angle is given by the value of θ_B which satisfies

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

For the case when the first medium is free space and the second medium has a relative permittivity ε_r, above equation can be expressed as

$\text{Sin}(\theta_B)= \sqrt{(\varepsilon_r-1)}/\sqrt{(\varepsilon_r^{\,2}-1)}$

Note that the Brewster angle occurs only for vertical (i.e. parallel) polarization.

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

Reflection from Perfect Conductors:

Since electromagnetic energy cannot pass through a perfect conductor a plane wave incident on a conductor has all of its energy reflected. As the electric field at the surface of the conductor must be equal to zero at all times in order to obey Maxwel1's equations, the reflected wave must be equal in magnitude to the incident wave. For the case when E-field polarization is in the plane of incidence, the boundary conditions require that

$\theta_i = \theta_r$

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

Similarly, for the case when the E-field is horizontally polarized, the boundary conditions require that
$\theta_i = \theta_r$
and $E_i = - E_r$ (E-field not in plane of incidence)

Ground Reflection (2-ray) Model:

In a mobile radio channel, a single direct path between the base station and a mobile is seldom the only physical means for propagation, and hence the free space propagation model is in most cases inaccurate when used alone. The 2-ray ground reflection model shown in Figure 3.7 is a useful propagation model that is based on geometric optics, and considers both the direct path and a ground reflected propagation path between transmitter and receiver. This model has been found to be reasonably accurate for predicting the large-scale signal strength over distances of several kilometers for mobile radio systems that use tall towers (heights which exceed 50 m), as well as for line of-sight, microcell channels in urban environments.

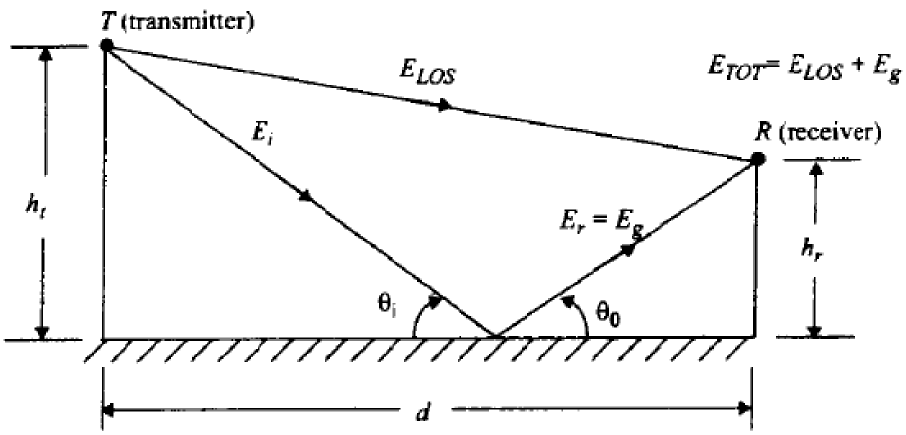


Figure 3.7
Two-ray ground reflection model.

Referring to Figure 3.7, ht is the height of the transmitter and hr is the height of the receiver. lf Eo is the free space E-field (in units of V/m) at a reference distance do from the transmitter, then for d > do , the free space propagating E—field is given by

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

Two propagating waves arrive at the receiver: the direct wave that travels a distance d'; and the reflected wave that travels a distance d”.

The electric field $E_{TOT}(d, t)$ can be expressed as the sum of equations for distances d’ and d” (i.e. direct wave and reflected wave.

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

Diffraction:

Diffraction allows radio signals to propagate around the curved surface of the earth, beyond the horizon, and to propagate behind obstructions. Although the received field strength decreases rapidly as a receiver moves deeper into the obstructed (shadowed) region, the diffraction field still exists and often has sufficient strength to produce a useful signal.

The phenomenon of diffraction can be explained by Huygen`s principle, which states that all points on a wavefront can be considered as point sources for the production of secondary wavelets, and that these wavelets combine to produce a new wavefront in the direction of` propagation. Diffraction is caused by the propagation of secondary wavelets into a shadowed region. The field strength of a diffracted wave in the shadowed region is the vector sum of the electric field components of all the secondary wavelets in the space around the obstacle.

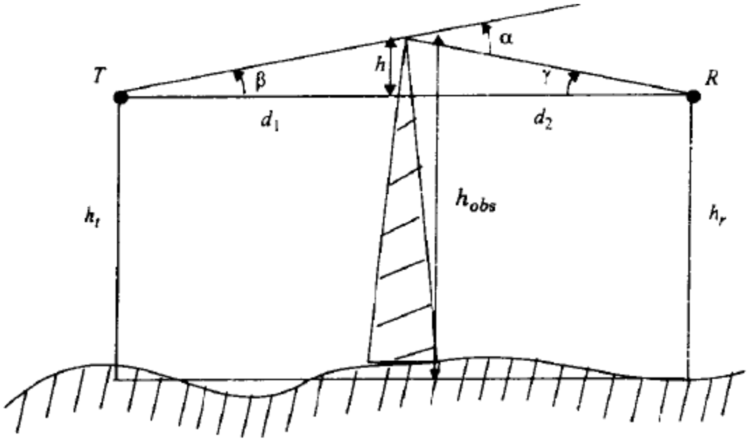
Fresnel Zone Geometry:

Consider a transmitter and receiver separated in free space as shown in Figure 3.10a. Let an obstructing screen of effective height h with infinite width (going into and out of the paper,) be placed between them at a distance d_1 from the transmitter and d_2 from the receiver. It is apparent that the wave propagating from the transmitter to the receiver via the top of the screen travels a longer distance than if a direct line- of-sight path (through the screen) existed. Assuming $h << d_1, d_2$ and $h >> \lambda$, then the difference between the direct path and the diffracted path, called the excess path length (Δ) , can be obtained from the geometry of Figure as

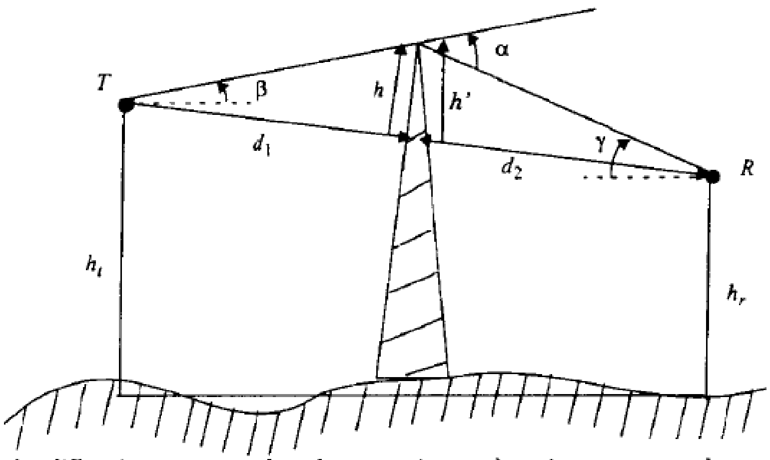
$$\Delta \approx \frac{h^2}{2}\frac{(d_1+d_2)}{d_1 d_2}$$

The corresponding phase difference is given by

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



(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 3.10c.

Knife-edge Diffraction Model:

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

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

Multiple Knife-edge Diffraction:

ln many practical situations, especially in hilly terrain, the propagation path may consist of more than one obstruction, in which case the total diffraction loss due to all of the obstacles must be computed. Bullington suggested that the series of obstacles be replaced by a single equivalent obstacle so that the path loss can be obtained using single knife-edge diffraction models. This method, illustrated in Figure 3.15, oversimplifies the calculations and often provides very optimistic estimates of the received signal strength. In a more rigorous treatment, Millington et. al. gave a wave-theory solution for the field behind two knife edges in series. This solution is very useful and can be applied easily for predicting diffraction losses due to two knife edges. However, extending this to more than two knife edges becomes a formidable mathematical problem. Many models that are mathematically less complicated have been developed to estimate the diffraction losses due to multiple obstructions.

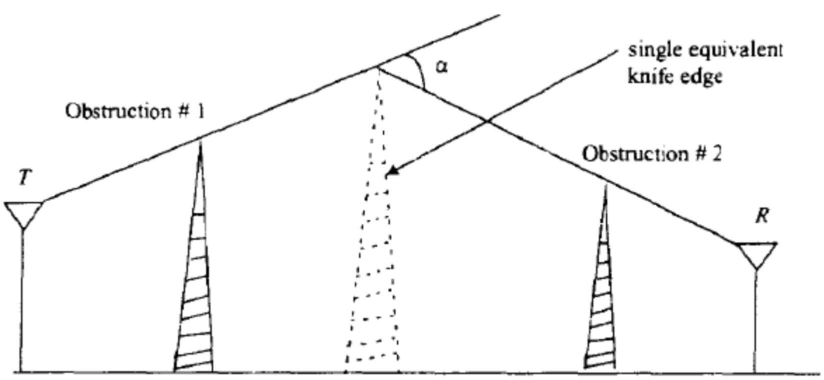


Figure 3.15
Bullington's construction of an equivalent knife edge [From [Bul47] © IEEE].

Scattering:

The actual received signal in a mobile radio environment is often stronger than what is predicted by reflection and diffraction models alone. This is because when a radio wave impinges on a rough surface, the reflected energy is spread out (diffused) in all directions due to scattering. Objects such as lamp posts and trees tend to scatter energy in all directions, thereby providing additional radio energy at a receiver. Flat surfaces that have much larger dimension than a wavelength may be modeled as reflective surfaces. However, the roughness of such surfaces often induces propagation effects different from the specular reflection described earlier in this chapter. Surface roughness is often tested using the Rayleigh criterion which defines a critical height (hc) of surface protuberances for a given angle of incidence i.e. given by

$$hc = \lambda/(8\sin\theta_i)$$

A surface is considered smooth if its minimum to maximum protuberance h is less than hc, and is considered rough if the protuberance is greater than hc. For rough surfaces, the flat surface reflection coefficient needs to be multiplied by a scattering loss factor, ρs, to account for the diminished reflected field.

Outdoor propagation model:

Longley-Rice Model:

The Longley-Rice model is applicable to point-to-point communication systems in the frequency range from 40 MHz to 100 GHz, over different kinds of terrain. The median transmission loss is predicted using the path geometry of the terrain profile and the refractivity of the troposphere. Geometric optics techniques (primarily the 2-ray ground reflection model) are used to predict signal strengths within the radio horizon. Diffraction losses over isolated obstacles are estimated using the Fresnel-Kirchoff knife- edge models. Forward scatter theory is used to make troposcatter predictions over long distances.

The Longley-Rice method operates in two modes. When a detailed terrain path profile is available, the path-specific parameters can be easily determined and the prediction is called a point-to-point mode prediction. On the other hand, if the terrain path profile is not available, the Longley-Rice method provides techniques to estimate the path-specific parameters, and such a prediction is called an area mode prediction.

Okumura Model:

Okumura`s model is one of the most widely used models for signal prediction in urban areas. This model is applicable for frequencies in the range 150 MHz to 1920 MHz (although it is typically extrapolated up to 3000 MHz) and distances of 1 km to 100 km. It can be used for base station antenna heights ranging from 30 m to 1000 m. Okumura developed a set of curves giving the median attenuation relative to free space (Amu), in an urban area over a quasi-smooth terrain with a base station effective antenna height

(h_{te}) of 200 m and a mobile antenna height (h_{re}) of 3 m. These curves were developed from extensive measurements using vertical omni-directional antennas at both the base and mobile, and are plotted as a function of frequency in the range 100 MHz to 1920 MHz and as a function of distance from the base station in the range 1 km to 100 km. To determine path loss using Okumura`s model, the free space path loss between the points of interest is first determined, and then the value of $A_{mu}(f, d)$ (as read from the curves) is added to it along with correction factors to account for the type of terrain. The model can be expressed as

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

where L50 is the 50th percentile (i.e., median) value of propagation path loss, L_F is the free space propagation loss, A_{mu} is the median attenuation relative to free space, $G(h_{te})$ is the base station antenna height gain factor, $G(h_{re})$ is the mobile antenna height gain factor, and G_{AREA} is the gain due to the type of environment. Note that the antenna height gains are strictly a function of height and have nothing to do with antenna patterns.

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

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

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

Hata Model:

The Hata model [Hat90] is an empirical formulation of the graphical path loss data provided by Okumura, and is valid from 150 MHz to 1500 MHz. Hata presented the urban area propagation loss as a standard formula and supplied correction equations for application to other situations. The standard formula for median path loss in urban areas is given by

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

where fc is the frequency (in MHz) from 150 MHz to 1500 MHz, h_{te} is the effective transmitter (base station) antenna height (in meters) ranging from 30 m to 200 m,h_{re} is the effective receiver (mobile) antenna height (in meters) ranging from 1 m to 10 m, d is the T-R separation distance (in km), and $a(h_{re})$is the correction factor for effective mobile antenna height which is a function of the size of the coverage area. For a small to medium sized city, the mobile antenna correction factor is given by

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

and for a large city, it is given by

$a(h_{re}) = 8.29(\log 1.54h_{re})^2$ -1.1 dB for $f_c <= 300$ MHz

$a(h_{re}) = 3.2(\log 11.75h_{re})2$ —4.97 dB for $f_c >= 300$ MHz

To obtain the path loss in a suburban area the standard Hata formula in equations are modified as

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

and for path loss in open rural areas, the formula is modified as

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

Indoor Propagation Models:

With the advent of Personal Communication Systems (PCS), there is a great deal of interest in characterizing radio propagation inside buildings. The indoor radio channel differs from the traditional mobile radio channel in two aspects - the distances covered are much smaller, and the variability of the environment is much greater for a much smaller range of T-R separation distances. It has been observed that propagation within buildings is strongly influenced by specific features such as the layout of the building, the construction materials, and the building type. This section outlines models for path loss within buildings.

Indoor radio propagation is dominated by the same mechanisms as outdoor: reflection, diffraction, and scattering. However, conditions are much more variable. For example, signal levels vary greatly depending on whether interior doors are open or closed inside a building. Where antennas are mounted also impacts large-scale propagation. Antennas mounted at desk level in a partitioned office receive vastly different signals than those mounted on the ceiling. Also, the smaller propagation distances make it more difficult to insure far-field radiation for all receiver locations and types of antennas.

Partition Losses (same floor):

Buildings have a wide variety of partitions and obstacles which form the internal and external structure. Houses typically use a wood frame partition with plaster board to form internal walls and have wood or non-reinforced concrete between floors. Office buildings, on the other hand, often have large open areas (open plan) which are constructed by using moveable office partitions so that the space may be reconfigured easily, and use metal reinforced concrete between floors. Partitions that are formed as part of the building structure are called hard partitions, and partitions that may be moved and which do not spanto the ceiling are called soft partitions. Partitions vary widely in their physical and electrical characteristics, making it difficult to apply general models to specific indoor installations.

Partition Losses between Floors:

The losses between floors of a building are determined by the external dimensions and materials of the building, as well as the type of construction used to create the floors and the external surroundings. Even the number of windows in a building and the presence of tinting (which attenuates radio energy) can impact the loss between floors. It can be seen that for all three buildings, the attenuation between one floors of the building is greater than the incremental attenuation caused by each additional floor. After about five or six floor separations, very little additional path loss is experienced.

Propagation Model (Practical Link Budget design)

Free Space Propagation Model

The reduction of the average signal level as the mobile station moves away from the base station is called propagation path loss.

A good path loss prediction model should be able to distinguish between rural, suburban, and urban areas, since there are important differences in radio propagation in these different area types.

Typical Free Space Propagation model is given by

Received Power (P_r) = $P_t + G_t + G_r$ - (32.44+20log (d) +20 log(f)) Where

P_r = Received power in dBm

P_t = Transmit power in dBm

G_t = Transmitting antenna gain in dBi.

G_r = Receiving antenna gain in dBi.

d = Distance in Km

f = Frequency in MHz

Hata Model

Hata has developed three empirical path loss models based on the measurements by Okumura in Tokyo area of Japan. The Hata model for typical path loss for urban area is given as

$$\mathbf{L_u=69.55+26.16\log(f)-13.82\log(h_B)-a(h_M)+[44.9-6.55\log(h_B)]\log(d)}$$

Where

f=propagation frequency in MHz.

h_B=height of the base station antenna in m

h_M=height of the mobile station antenna in m

d=distance between base station and the mobile station in km.

The term $a(h_M)$ is a correction factor, the value of which depends upon the terrain type. The value of $a(h_M)$ for small and medium sized cities can be found in dB as

$$\mathbf{a(h_M)=[1.1\log(f)-0.7]h_M-1.56\log(f)+0.8}$$

and for large cities depending upon frequency as

$$\mathbf{a(h_M)=8.29[\log(1.54\,h_M)]^2-1.10 \qquad \text{when } f\leq 200\text{ MHz}}$$

$$\mathbf{a(h_M)=3.2[\log(11.75\,h_M)]^2-4.97 \qquad \text{when } f\geq 400\text{ MHz}}$$

For typical suburban area we subtract a correction factor

$$\mathbf{L_{su} = L_u - 2\,[\log(f/28)]^2 - 5.4}$$

For open area we have a different correction factor

$$L_r = L_u - 4.78\,[\log(f)]^2 + 18.33\log(f) - 40.94$$

The open area of the Hata model corresponds to flat deserted area. For path loss of typical rural area a margin of 6-10 dB is often added to the path loss predicted by the open area Hata model.

In Hata model a large city is understood to be heavily built with relatively large buildings averaging more than four floors in height. If the city has lower average buildings height, it is considered to be small or medium.

The range of parameter values, where the Hata model is applicable, is

- f = 150 – 1500 MHz
- h_B = 30 – 200 m
- h_M = 1 – 10 m
- d = 1 - 20 km

For higher carrier frequencies of 1500- 2000 MHz the following modification of Hata model for urban area has been proposed. This is also known as Cost231 Hata model and it is the extension of Hata model for *PCS (1900 MHz)*.

$$L_u = 46.3 + 33.9\log(f) - 13.82\log(h_B) - a(h_M) + [44.9 - 6.55\log(h_B)]\log(d) + C$$

Where the correction factor $a(h_M)$ for small and medium-sized cities

$$a(h_M) = [1.1\log(f) - 0.7)]\,h_M - 1.56\log(f) + 0.8$$

These modified equation have been successfully used for cellular mobile network design at 1800 MHz band. However, it should be noted that (modified) Hata model is only valid for macrocell design. The Hata model is not applicable to microcells with d<1 km.

The Hata model also been extended to distances d= 20…100 km with the following modification

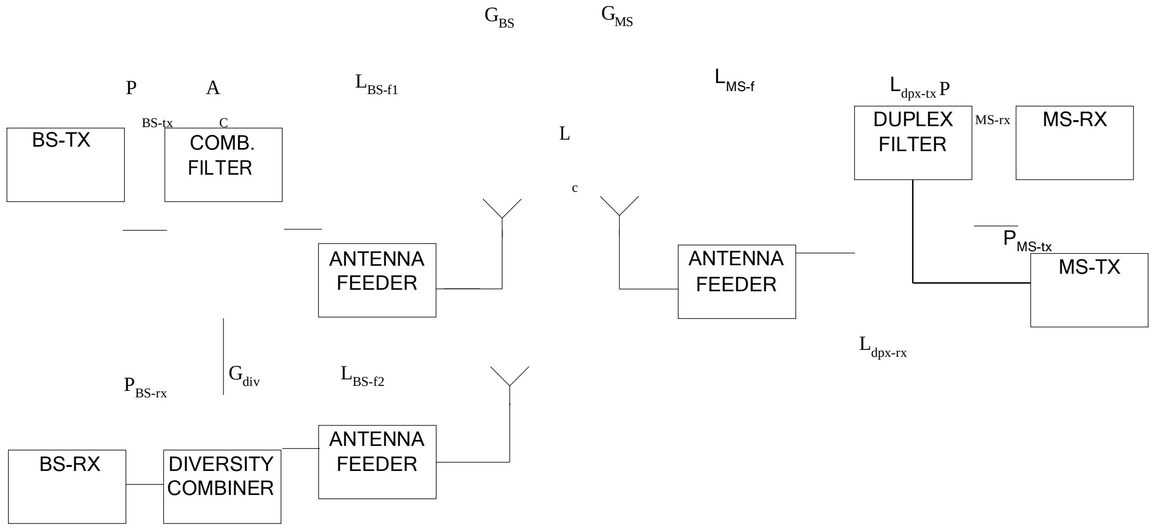
$$\mathbf{L_u = 69.55 + 26.16 \log(f) - 13.82 \log (h_B) - a(h_M) + [\,44.9 - 6.55 \log (h_B)]\,\log(d)]^a}$$

Examples:

1. A transmitter has a power output of 50 W at a carrier frequency of 200MHz. It is connected to an antenna with a gain of 10 dBi. The receiving antenna is 15 Km away and has a gain of 3 dBi. Calculate power delivered to the receiver, assuming free space propagation. Assume also that there are no losses or mismatches in the system.
2. Let us consider a Kathmandu city with different GSM uplink parameters. The MS is transmitting with power 2W .The minimum acceptable received power at BS is -116 dBm. The carrier frequency is 900MHz, the height of base station is 30m and height of mobile station is 1m.Estimate the maximum cell radius and corresponding cell area.

Radio Link Power Budget

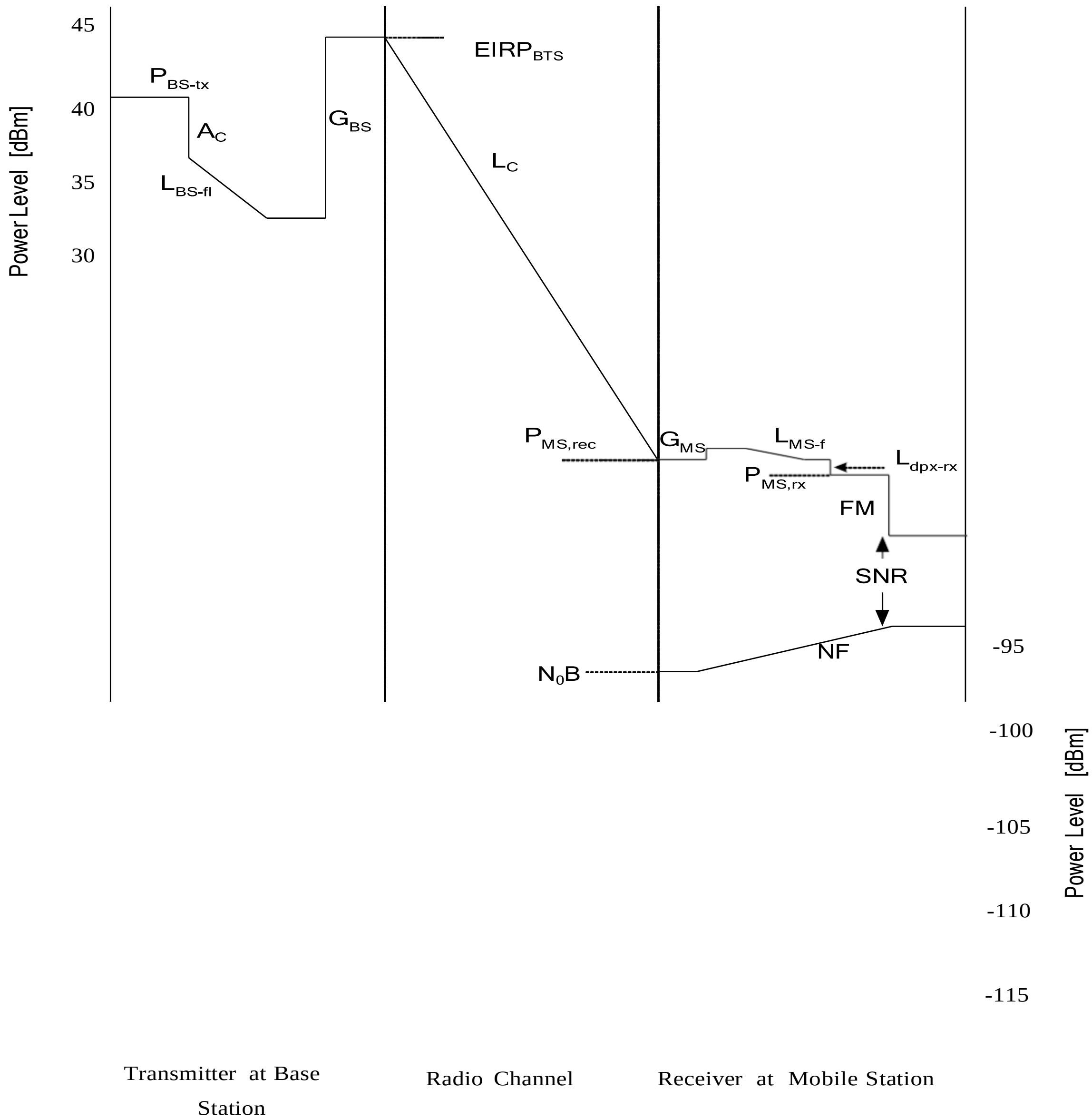
Transmission in uplink and downlink of cellular systems is asymmetric, since the BS transmitter typically uses much higher power than the MS transmitter. However, the transmission quality in uplink and downlink should be equal, especially near the cell edge. Both speech and data services in cellular systems are dimensioned for equal transmission quality in both directions. Transmission quality in uplink and downlink can be determined from the link budget. The terms in the link budget are defined with aid of the block diagram in the adjoining figure.



Where

- $P_{BS\text{-}tx}$ is the output power level of the base station
- A_c is the base station transmitter combining filter loss (dB),
- $L_{BS\text{-}f1}$ is the Base station antenna feeder loss (dB),
- G_{BS} is the base station antenna gain (dBi, relative to an isotropic radiator),
- L_C is the radio path loss between isotropic antennas (dB),
- G_{MS} is the mobile station gain (dBi),
- $L_{MS\text{-}f}$ is the mobile station antenna feeder loss (dB),
- $L_{dpx\text{-}rx}$ is the mobile station duplex filter loss in the downlink direction (dB).
- $L_{dpx\text{-}tx}$ is the duplex filter loss in the uplink direction (dB),
- G_{div} is the base station diversity gain (dB),
- $P_{MS\text{-}rx}$ is the received power level in the mobile station input terminal (dBm),
- $P_{BS\text{-}rx}$ is the received power level in the base station input terminal (dBm).

The following Figure shows a pictorial example of a radio link power budget. In addition to the link budget of the transmitted signal, it also illustrates noise at the receiver. The noise has power NoB at the input of the receiver, and it is amplified by the noise figure (NF) of the receiver.



Power Level [dBm]
45
40
35
30
P_BS-tx
A_C
L_BS-fl
G_BS
EIRP_BTS
L_C
P_MS,rec
G_MS
L_MS-f
L_dpx-rx
P_MS,rx
FM
SNR
NF
N_0B
-95
-100
-105
-110
-115
Power Level [dBm]
Transmitter at Base Station
Radio Channel
Receiver at Mobile Station

The abbreviation FM stands for fading margin for slow fading. Hence, the signal-to-noise ratio(SNR)in the figure corresponds to the SNR that is encountered, when the receiver is shadowed by an obstruction. The average SNR is better that this value. The SNR during a fade should still be larger than the minimum acceptable E_b / N_o defined for the system. The maximum acceptable path loss can be computed as

$$L_c = EIRP_{max} - P_{rec,\ min}$$

Where

EIRP is Effective Isotropic Radiated Power

P_{rec} is Received power with (hypothetical) isotropic antenna

For downlink:

$EIRP_{BTS,\ max}$	$= P_{BS\text{-}tx}$	maximum BTS power (mean power over burst)
	$- A_c$	combiner and filter loss
	$- L_{BS\text{-}f1}$	antenna cable loss
	$+ G_{BS}$	antenna gain
P_{MS}, rec, min	$= P_{MS\text{-}rx}$	MS reference sensitivity (for MS class i)
	$+ L_{MS\text{-}f}$	antenna cable loss
	$- G_{MS}$	antenna gain
	$+ L_{dpx\text{-}rx}$	duplex filter loss

For uplink:

$EIRP_{MS,\,max}$ = $P_{MS\text{-}tx}$ maximum transmission power of MS (mean power over burst for MS class i)

- $L_{dpx\text{-}tx}$ duplex filter loss

- $L_{MS\text{-}f}$ antenna cable loss

+ G_{MS} antenna gain

P_{BTS}, rec, min = $P_{BS\text{-}rx}$ BTS reference sensitivity

+ $L_{BS\text{-}f1}$ antenna cable loss

- G_{BS} antenna gain

- G_{div} diversity gain (if existing)

A balanced power budget is achieved if and only if

$$L_{c,uplink} = L_{c,\,downlink}$$

Which is equivalent to

$$EIRP_{BTS,max} - P_{MS,rec,min} = EIRP_{MS,max} - P_{BTS,rec,min}$$

In reality, balanced power budget is only achieved with an accuracy of ±5 dB, depending primarily on the employed mobile phone class and the initial network design goals. Therefore, uplink and downlink may have different power ranges which may lead to significant performance differences on the cell boundary or at indoor locations.

The acceptable path loss (without interference margin) for any cellular system can be computed separately for uplink and downlink as:

$$L_{c,u} = [P_{MS-tx} - L_{dpx-tx} - L_{MS-f} + G_{MS}] - [P_{BS-rx} + L_{BS-fi} - G_{BS} - G_{div}]$$

$$L_{c,d} = [P_{BS-tx} - A_c - L_{BS-fi} + G_{BS}] - [P_{MS-rx} + L_{MS-f} - G_{MS} + L_{dpx-rx}]$$

Example: power balance / unbalance in a GSM – system with following parameter values:

- P_{BS-tx} = 10 W (40 dBm)
- P_{MS-tx} = 1 W (30 dBm, handheld telephone)
- P_{MS-tx} = 5 W (37 dBm, car mounted telephone)
- S_{BS} = -104 dBm
- S_{MS} = -102 dBm (handheld telephone)
- S_{MS} = -104 dBm (car mounted telephone)
- $L_{dpx-tx} = L_{dpx-rx}$
- G_{div} = 7 dB
- A_c = 3 dB

The power unbalance of the handheld telephone:

$$\Delta L = L_{c,d} - L_{c,u} = [P_{BS-tx} - A_c - L_{BS-fi} + G_{BS}]$$

$$- \quad [P_{MS-rx} + L_{MS-f} - G_{MS} + L_{dpx-rx}]$$

$$- \quad [P_{MS-tx} - L_{dpx-tx} - L_{MS-f} + G_{MS}]$$

$$+ \quad [P_{BS-rx} + L_{BS-fi} - G_{BS} - G_{div}]$$

$$= P_{BS-tx} - A_c - P_{MS-rx} - P_{MS-tx} + P_{BS-rx} - G_{div}$$

$$= 40 - 3 + 102 - 30 - 104 - 7 = -2 \text{ dB}$$

The uplink direction is thus 2 dB better. This difference can be neglected in practical network design.

For the car mounted telephone the power unbalance is:

FORMULA

$$\Delta L = P_{BS-tx} - A_c - P_{MS-rx} - P_{MS-tx} + P_{BS-rx} - G_{div}$$

$$= 40 - 3 + 104 - 37 - 7 - 104 = -7 \text{ dB}$$

The uplink direction is now 7 dB better. From the operator's point of view it would be better if the power unbalance were in favor of the downlink. This would guarantee better network control. However the real time power control used in GSM can easily rectify situation and produce almost perfect power balance.

Factor Influencing Small-Scale fading :→

(i) Multipath Propagation :→

The presence of reflecting objects and scatters in the channel creates a constantly changing enviornment that dissipates the signal energy in amplitude, phase and time.

(ii) Speed of Mobile :→

The relative Motion between the base station and the Mobile result in random frequency modulation due to different Doppler shift on each of the multipath component.

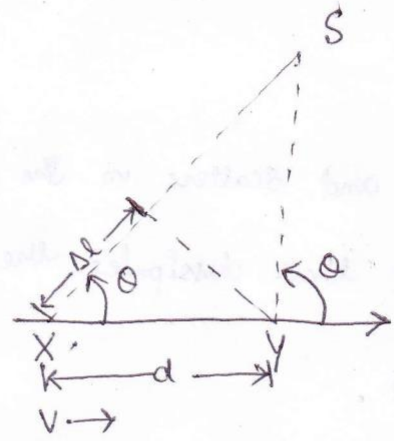
(iii) Speed of Surrounding object :→

If objects in the radio channel are in motion, they induce a time varying Doppler shift on multipath component.

(iv) Transmission Bandwidth of the signal :→

If the transmitted radio signal bandwidth is greater than the bandwidth of the multipath channel, the received signal will be distorted, but the received signal strength will not fade such over a local area.

Doppler shift :-



Consider a mobile moving at a constant velocity v, along a path segement having length d between points X & Y, while it receives signal from a remote Source S. The difference in path lengths travelled by the wave from Source S to the mobile at point X & Y is $\Delta l = d\cos\theta = v\Delta t \cos\theta$, where Δt is the time required for the mobile to travel from X to Y.

The phase change in the received signal due to the difference in path length is 2_0

$$\Delta\phi = \frac{2\pi \Delta l}{d} = \frac{2\pi v \Delta t}{d} \cos\theta$$

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

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

where

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

Coherence Bandwidth :→

While the delay spread is a natural phenomenon caused by reflected and scattered propagation path in the radio channel, the coherence Bandwidth B_C, is defined relation derived from the rms delay spread.

Coherence Bandwidth is a statistical measure of the range of frequencies over which the channel can be considered flat which means the channel that passes all the spectral component approximately equal gain & linear phase.

$$\boxed{B_C = \frac{1}{50\sigma_\tau}}$$

If the defⁿ is relaxed so that the frequency correlation function is above 0.5, then

$$\boxed{B_C = \frac{1}{5\sigma_\tau}}$$

Parameter of Mobile Multipath Channel :→

(i) Time Dispersion parameter (Doppler Spread)

(ii) Coherence Bandwidth

(iii) Coherence Time.

Doppler Spread :→

In order to compare different multipath channels and to develop some general design guidelines for wireless system. The mean excess delay, rms delay spread, excess delay spread, are the multipath channel parameter that can be determined from a power delay profile.

The time dispersive properties of wideband multipath channels are most commonly quantified by their mean excess delay ($\bar{\tau}$) and rms delay spread (σ_τ).

The mean excess delay

$$\bar{\tau} = \frac{\sum_K a_K^2 \tau_K}{\sum_K a_K^2} = \frac{\sum_K P(\tau_K)\tau_K}{\sum_K P(\tau_K)}$$

The rms delay spread

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

Doppler spread and Coherence Time :→

Delay spread and coherence bandwidth are parameters which describe the time dispersive nature of the channel in a local area.

Doppler spread and Coherence time are parameters which describe the time varying nature of the channel in small-scale region.

Doppler spread (B_D) →

It is a measure of the spectral broadening caused by the time rate of change of the mobile radio channel and is defined as the range of frequencies over which the doppler spectrum is "essentially non-zero.

If the baseband signal bandwidth is much greater than B_D, the effect of Doppler spread are negligible at the receiver. This is a slow fading channel.

Coherence Time (T_c) →

It is the time domain dual of Doppler spread and is used to characterize the time varying nature of the frequency dispersiveness of the Channel in the time domain.

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

If Coherence time is defined as the time over which the time Correlation function is above 0.5 then the T_c is

$$T_C = \frac{9}{16\pi f_m}$$

f_m = max^n Doppler shift-

$$f_m = V/\lambda$$

$$T_C = \sqrt{\frac{9}{16\pi f_m^2}} = \frac{0.423}{f_m}$$

Multipath Fading :→

Types of small-scale Fading :→

Depending on the relation between the signal parameters such as bandwidth symbol period etc. and the channel parameter such as rms delay spread and Doppler spread, different transmitted signals will undergo different type of fading.

Multipath delay spread leads to Time dispersion and frequency selective fading, while Doppler spread leads to frequency dispersion and time selective fading.

Fading effect Due to Multipath Time Delay Spread :→

Small-Scale Fading
(Based on multipath time delay spread)

Flat fading

1. BW of Signal < BW of Channel
2. Delay spread < Symbol period

Frequency selective fading

1. BW of signal > BW of channel
2. Delay spread > Symbol period

Flat fading :→

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

A signal undergoes flat fading if

$$B_S << B_C$$

$$T_S >> \sigma_\tau$$

Frequency selective fading :→

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

A signal undergoes frequency selective fading if

$$B_S > B_C$$

$$T_S < \sigma_\tau$$

Fading effect Due to Doppler spread :→

Small scale fading
(Based on Doppler Spread)

Fast fading	Slow fading
1. High Doppler Spread	1. low Doppler Spread
2. coherence time < Symbol period	2. Coherence time > Symbol period
3. Channel Variation faster than baseband signal variation	3. channel variation slower than baseband signal

Fast fading :→

In this, the channel impulse response changes rapidly within the symbol duration i.e the coherence time of the channel is smaller than the symbol period of the transmitted signal. This cause frequency dispersion due to Doppler spreading, which leads to signal distortion.

The signal undergoes fast fading if

$$T_S > T_C$$

$$B_S < B_D$$

Slow fading :→

In this, the channel impulse response changes at rate much slower than the transmitted baseband signal. In this case, the channel may be assumed to be static over one or several reciprocal bandwidth intervals. This implies that the Doppler spread of the channel is much less than the bandwidth of the baseband signal.

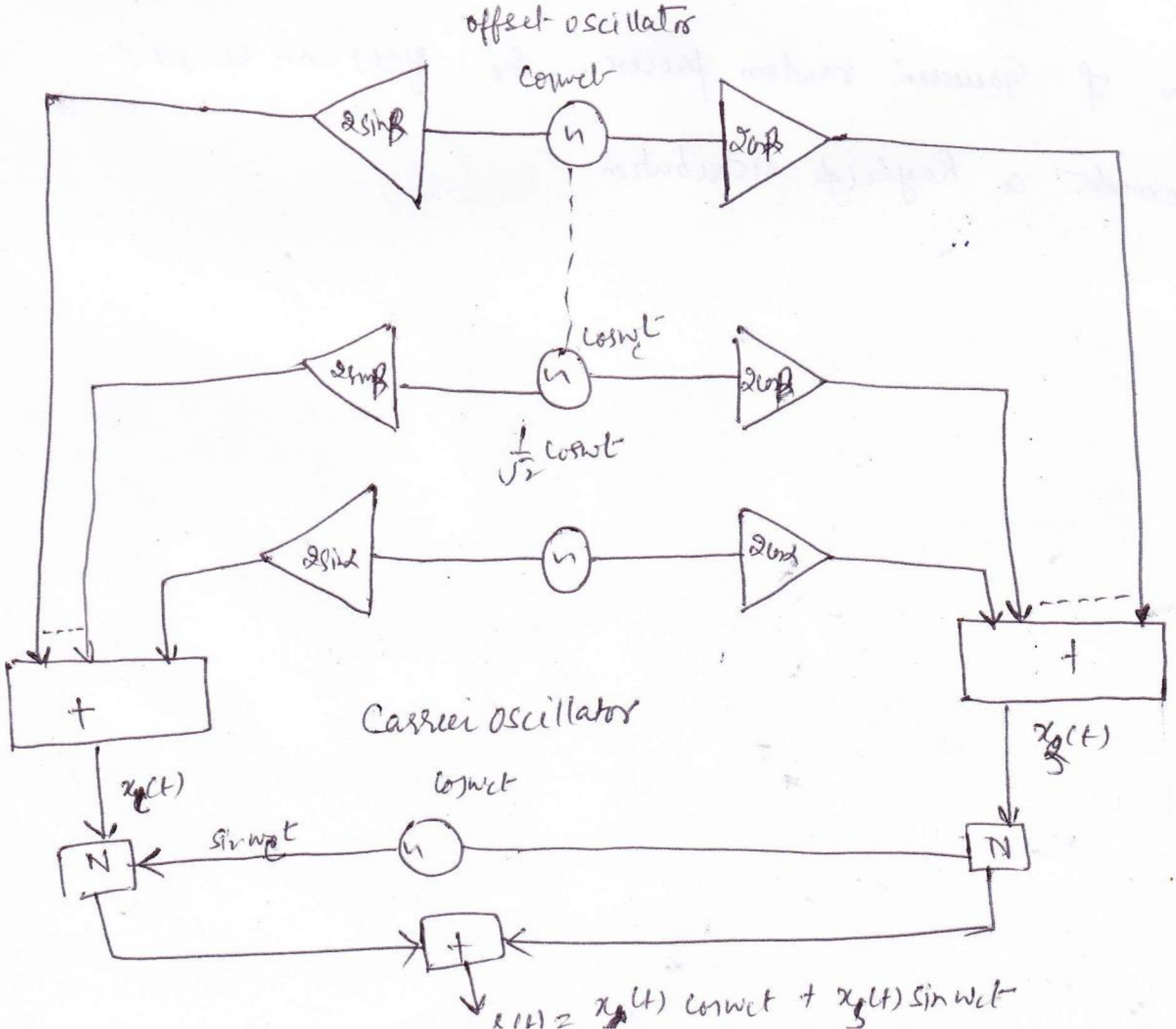
A signal undergoes slow fading if

$$T_S << T_C$$

$$B_S >> B_D$$

Jake's Model :→

Jake's method is a mathematical model suitable to simulated the Rayleigh fading. The computation load is for lower than a model by doing FFT to the Doppler spectrum. It is simplified model of the ring Scattering model, which assome the arrival phases of received multipath Signals are uniformed distributed from 0 to 2π after passing through different Scattering path and angles. In Jake's model, it uses several sinusoids to approximate this effect. The following fig shows the block diagram of Jake's model.



the simmation of the Sinusoid is

$$y(t) = x_c(t)\cos w_c t + x_s(t)\sin(w_c t)$$

$$x_c(t) = 2\sum_{n=1}^{N_0} \cos(\beta_n)\cos(w_n t) + \sqrt{2}\cos(\alpha)\cos(w_m t)$$

$$w_m = \frac{2\pi V}{d}$$

$$w_n = w_m \cos\left(\frac{2\pi n}{N}\right)$$

$$N_0 = \frac{1}{2}\left(\frac{N}{2} - 1\right)$$

It can be observed that Jake's model samples more frequencies near the maximum Doppler shift when $N_0 \leq 8$, $x_c(t)$ and $x_s(t)$ are approximation of Gaussian random process. So $y(t)$ can be used to approximate a Rayleigh distribution.

UNIT III

Mobile Radio Propagation

Multipath Wave Propagation and Fading

Multipath Propagation

In wireless telecommunications, multipath is the propagation phenomenon that re- sults in radio signals reaching the receiving antenna by two or more paths. Causes of multipath include atmospheric ducting, ionospheric reflection and refraction, and reflection from water bodies and terrestrial objects such as mountains and buildings. The effects of multipath include constructive and destructive interference, and phase shifting of the signal. In digital radio communications (such as GSM) multipath can cause errors and affect the quality of communications. We discuss all the related issues in this chapter.

Multipath & Small-Scale Fading

Multipath signals are received in a terrestrial environment, i.e., where different forms of propagation are present and the signals arrive at the receiver from transmitter via a variety of paths. Therefore there would be multipath interference, causing multi- path fading. Adding the effect of movement of either Tx or Rx or the surrounding clutter to it, the received overall signal amplitude or phase changes over a small amount of time. Mainly this causes the fading.

Fading

The term **fading**, or, small-scale fading, means rapid fluctuations of the amplitudes, phases, or multipath delays of a radio signal over a short period or short travel distance. This might be so severe that large scale radio propagation loss effects might be ignored.

Multipath Fading Effffects

In principle, the following are the main multipath effects:

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

Factors Influencing Fading

The following physical factors influence small-scale fading in the radio propagation channel:

(1) Multipath propagation – Multipath is the propagation phenomenon that re- sults inradio signals reaching the receiving antenna by two or more paths. The effects of multipath include constructive and destructive interference, and phase shifting of the signal.

(2) Speed of the mobile – The relative motion between the base station and the mobile results in random frequency modulation due to different doppler shifts on each of the multipath components.

(3) Speed of surrounding objects – If objects in the radio channel are in mo- tion, they induce a time varying Doppler shift on multipath components. If the surrounding objects move at a greater rate than the mobile, then this effect dominates fading.

(4) Transmission Bandwidth of the signal – If the transmitted radio signal bandwidth is greater than the “bandwidth” of the multipath channel (quanti- fied by *coherence bandwidth*), the received signal will be distorted.

Types of Small-Scale Fading

The type of fading experienced by the signal through a mobile channel depends on the relation between the signal parameters (bandwidth, symbol period) and the channel parameters (rms delay spread and Doppler spread). Hence we have four different types of fading. There are two types of fading due to the time dispersive nature of the channel.

Fading Effffects due to Multipath Time Delay Spread

Flat Fading

Such types of fading occurs when the bandwidth of the transmitted signal is less than the coherence bandwidth of the channel. Equivalently if the symbol period of the signal is more than the rms delay spread of the channel, then the fading is flat fading.

So we can say that flat fading occurs when

$$B_S \quad B_C$$

where B_S is the signal bandwidth and B_C is the coherence bandwidth. Also

$$T_S \times \sigma_\tau$$

where T_S is the symbol period and σ_τ is the rms delay spread. And in such a case, mobile channel has a constant gain and linear phase response over its bandwidth.

Frequency Selective Fading

Frequency selective fading occurs when the signal bandwidth is more than the co-herence bandwidth of the mobile radio channel or equivalently the symbols duration of the signal is less than the rms delay spread.

$$B_S \times B_C$$

and

$$T_S \quad \sigma_\tau$$

At the receiver, we obtain multiple copies of the transmitted signal, all attenuated and delayed in time. The channel introduces inter symbol interference. A rule of thumb for a channel to have flat fading is if

$$\frac{\sigma_\tau}{T_S} \leq 0.1$$

Fading Effffects due to Doppler Spread

Fast Fading

In a fast fading channel, the channel impulse response changes rapidly within the symbol duration of the signal. Due to Doppler spreading, signal undergoes frequency dispersion leading to distortion. Therefore a signal undergoes fast fading if

$$T_S \times T_C$$

where T_C is the coherence time and

$$B_S \times B_D$$

where B_D is the Doppler spread. Transmission involving very low data rates suffer from fast fading.

Slow Fading

In such a channel, the rate of the change of the channel impulse response is much less than the transmitted signal. We can consider a slow faded channel a channel in which channel is almost constant over atleast one symbol duration. Hence

$$T_S \quad T_C$$

and

$$B_S \times B_D$$

We observe that the velocity of the user plays an important role in deciding whether the signal experiences fast or slow fading.

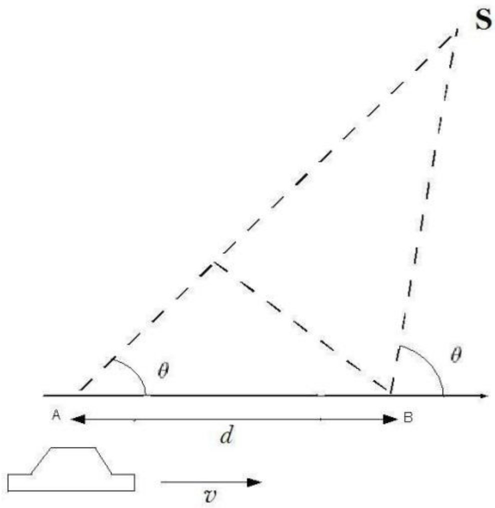


Figure Illustration of Doppler effect.

Doppler Shift

The Doppler effect (or Doppler shift) is the change in frequency of a wave for an observer moving relative to the source of the wave. In classical physics (waves in a medium), the relationship between the observed frequency f and the emitted frequency f_o is given by:

$$f = \left(\frac{v \pm v_r}{v \pm v_s} \right) f_0$$

where v is the velocity of waves in the medium, v_s is the velocity of the source relative to the medium and v_r is the velocity of the receiver relative to the medium.

In mobile communication, the above equation can be slightly changed according to our convenience since the source (BS) is fixed and located at a remote elevated level from ground. The expected Doppler shift of the EM wave then comes out to be $\pm v_r f_o$ or $\lambda \pm v_r$. As the BS is located at an elevated place, a cos φ factor would also be multiplied with this. The exact scenario, as given in Figure 5.1, is illustrated below.

Consider a mobile moving at a constant velocity v, along a path segment length d between points A and B, while it receives signals from a remote BS source S. The difference in path lengths traveled by the wave from source S to the mobile at points A and B is $\Delta l = d \cos\theta = v\Delta t \cos\theta$, where Δt is the time required for the mobile to travel from A to B, and θ is assumed to be the same at points A and B since the source is assumed to be very far away. The phase change in the received signal due to the difference in path lengths is therefore

$$\Delta\phi = \frac{2\pi\Delta l}{\lambda} = \frac{2\pi v\Delta t}{\lambda}\cos\theta \tag{5.11}$$

and hence the apparent change in frequency, or Doppler shift (f_d) is

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

Example 1

An aircraft is heading towards a control tower with 500 kmph, at an elevation of 20°. Communication between aircraft and control tower occurs at 900 MHz. Find out the expected Doppler shift.

Solution As given here,

$$v = 500\ kmph$$

the horizontal component of the velocity is

$$v^j = v\cos\theta = 500 \times \cos 20^\circ = 130\ m/s$$

Hence, it can be written that

$$\lambda = \frac{900 \times 10^6}{3 \times 10^8} = \frac{1}{3}m$$

$$f_d = \frac{130}{1/3} = 390\ Hz$$

If the plane banks suddenly and heads for other direction, the Doppler shift change will be 390 Hz to −390 Hz.

Impulse Response Model of a Multipath Channel

Mobile radio channel may be modeled as a linear filter with time varying impulseresponse in continuous time. To show this, consider time variation due to receiver motionand time varying impulse response $h(d, t)$ and $x(t)$, the transmitted signal.

The received signal $y(d, t)$ at any position d would be

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

For a causal system: $h(d, t) = 0$, for $t < 0$ and for a stable system $\int_{-\infty}^{\infty} |h(d, t)|\, dt < \infty$

Applying causality condition in the above equation, $h(d, t-\tau) = 0$ for $t-\tau < 0$ $\Rightarrow \tau > t$, i.e., the integral limits are changed to

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

Since the receiver moves along the ground at a constant velocity v, the position of the receiver is $d = vt$, i.e.,

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

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

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

It is useful to discretize the multipath delay axis τ of the impulse response into equal time delay segments called *excess delay bins*, each bin having a time delay width equal to $(\tau_{i+1} - \tau_i) = \Delta\tau$ and $\tau_i = i\Delta\tau$ for $i \in \{0, 1, 2, ..N-1\}$, where N represents the total number of possible equally-spaced multipath components, including the first arriving component. The useful frequency span of the model is $2/\Delta\tau$. The model may be used to analyze transmitted RF signals having bandwidth less than $2/\Delta\tau$.
If there are N multipaths, maximum excess delay is given by $N\Delta\tau$.

$$\{y(t) = x(t) * h(t, \tau_i) | i = 0, 1, ...N-1\}$$

Bandpass channel impulse response model is

$$x(t) \rightarrow h(t, \tau) = Re\{h_b(t, \tau)e^{j\omega_c t}\} \rightarrow y(t) = Re\{r(t)e^{j\omega_c t}\}$$

Baseband equivalent channel impulse response model is given by

$$c(t) \rightarrow \frac{1}{2}\bar{h}_b(t, \tau) \rightarrow r(t) = c(t) * \frac{1}{2}\bar{h}_b(t, \tau)$$

Average power is

$$\overline{x^2(t)} = \frac{1}{2}|c(t)|^2$$

The baseband impulse response of a multipath channel can be expressed as

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

where $a_i(t,\tau)$ and $\tau_i(t)$ are the real amplitudes and excess delays, respectively, of the ith multipath component at time t. The phase term $2\pi f_c\tau_i(t) + \phi_i(t,\tau)$ in the above equation represents the phase shift due to free space propagation of the ith multipath component, plus any additional phase shifts which are encountered in the channel.

If the channel impulse response is wide sense stationary over a small-scale time or distance interval, then

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

For measuring $h_b(\tau)$, we use a probing pulse to approximate $\delta(t)$ i.e.,

$$p(t) \approx \delta(t-\tau)$$

Power delay profile is taken by spatial average of $|h_b(t,\tau)|^2$ over a local area. The received power delay profile in a local area is given by

$$p(\tau) \approx k\overline{|h_b(t;\tau)|^2}.$$

Relation Between Bandwidth and Received Power

In actual wireless communications, impulse response of a multipath channel is mea- sured using channel sounding techniques. Let us consider two extreme channel sounding cases.

Consider a pulsed, transmitted RF signal

$$x(t) = Re\{p(t)e^{j2\pi f_c t}\}$$

where $p(t) = \sqrt{\frac{4\tau_{max}}{T_{bb}}}$ for $0 \le t \le T_{bb}$ and 0 elsewhere. The low pass channel output is

$$\begin{aligned} r(t) &= \frac{1}{2}\sum_{i=0}^{N-1} a_i\exp[j\theta_i]p(t-\tau_i) \\ &= \sum_{i=0}^{N-1} a_i\exp[j\theta_i].\sqrt{\frac{\tau_{max}}{T_{bb}}}\,rect\left(t-\frac{T_b}{2}-\tau_i\right). \end{aligned}$$

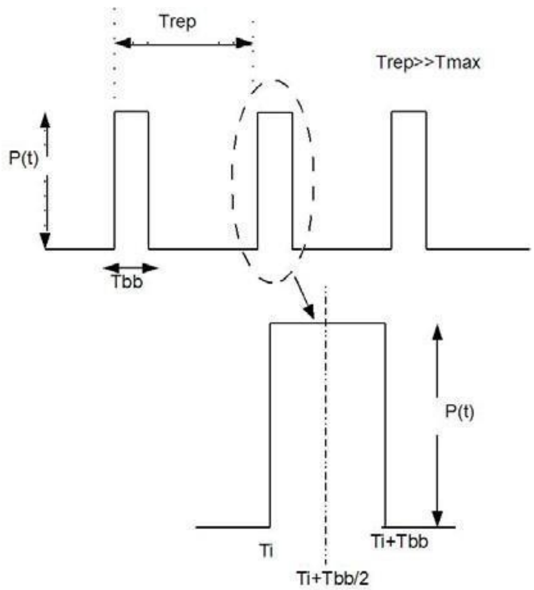


Figure: A generic transmitted pulsed RF signal.

The received power at any time t_0 is

$$|r(t_0)|^2 = \frac{1}{\tau_{max}} \int_0^{\tau_{max}} r(t) r^*(t) dt$$

$$= \frac{1}{\tau_{max}} \int_0^{\tau_{max}} \frac{1}{4} \left(\sum_{k=0}^{N-1} a_k^2(t_0) p^2(t - \tau_k) \right) dt$$

$$= \frac{1}{\tau_{max}} \sum_{k=0}^{N-1} a_k^2(t_0) \int_0^{\tau_{max}} \left\{ \sqrt{\frac{\tau_{max}}{T_{bb}}} \, rect\left(t - \frac{T_b}{2} - \tau_i\right) \right\}^2 dt$$

$$= \sum_{k=0}^{N-1} a_k^2(t_0).$$

Interpretation: If the transmitted signal is able to resolve the multipaths, then average small-scale receiver power is simply sum of average powers received from each multipath components.

$$E_{a,\theta}[P_{WB}] = E_{a,\theta}\left[\sum_{i=0}^{N-1} |a_i \exp(j\theta_i)|^2 \right] \approx \sum_{i=0}^{N-1} \overline{a_i^2}$$

Now instead of a pulse, consider a CW signal, transmitted into the same channel and for simplicity, let the envelope be $c(t) = 2$. Then

$$r(t) = \sum_{i=0}^{N-1} a_i \exp[j\theta_i(t, \tau)]$$

and the instantaneous power is

$$|r(t)|^2 = \left|\sum_{i=0}^{N-1} a_i \exp[j\theta_i(t,\tau)]\right|^2$$

Over local areas, a_i varies little but θ_i varies greatly resulting in large fluctuations.

$$\begin{aligned} E_{a,\theta}[P_{CW}] &= E_{a,\theta}\left[\left|\sum_{i=0}^{N-1} a_i \exp(j\theta_i)\right|^2\right] \\ &\approx \sum_{i=0}^{N-1} \overline{a_i^2} + 2\sum_{i=0}^{N-1}\sum_{i,j\neq i}^{N} r_{ij}\,\overline{\cos(\theta_i - \theta_j)} \end{aligned}$$

where $r_{ij} = E_a[a_i a_j]$.

If, $r_{ij} = \overline{\cos(\theta_i - \theta_j)} = 0$, then $E_{a,\theta}[P_{CW}] = E_{a,\theta}[P_{WB}]$. This occurs if multipath components are uncorrelated or if multipath phases are i.i.d over $[0, 2\pi]$.

Bottomline:

1. If the signal bandwidth is greater than multipath channel bandwidth then fading effects are negligible
2. If the signal bandwidth is less than the multipath channel bandwidth, large fading occurs due to phase shift of unresolved paths.

Linear Time Varying Channels (LTV)

The time variant transfer function(TF) of an LTV channel is FT of $h(t,\tau)$ w.r.t. τ.

$$H(f,t) = FT[h(\tau,t)] = \int_{-\infty}^{\infty} h(\tau,t)e^{-j2\pi f\tau}\,d\tau$$

$$h(\tau,t) = FT^{-1}[H(f,t)] = \int_{-\infty}^{\infty} H(f,t)e^{j2\pi f\tau}\,df$$

The received signal

$$r(t) = \int_{-\infty}^{\infty} R(f,t)e^{j2\pi ft}\,df$$

where $R(f,t) = H(f,t)X(f)$.

For flat fading channel, $h(\tau,t) = Z(t)\delta(\tau - \tau_i)$ where $Z(t) = \sum \alpha_n(t)e^{-j2\pi f_c\tau_n(t)}$. In this case, the received signal is

$$r(t) = \int_{-\infty}^{\infty} h(\tau,t)x(t-\tau)\,d\tau = Z(t)x(t-\tau_i)$$

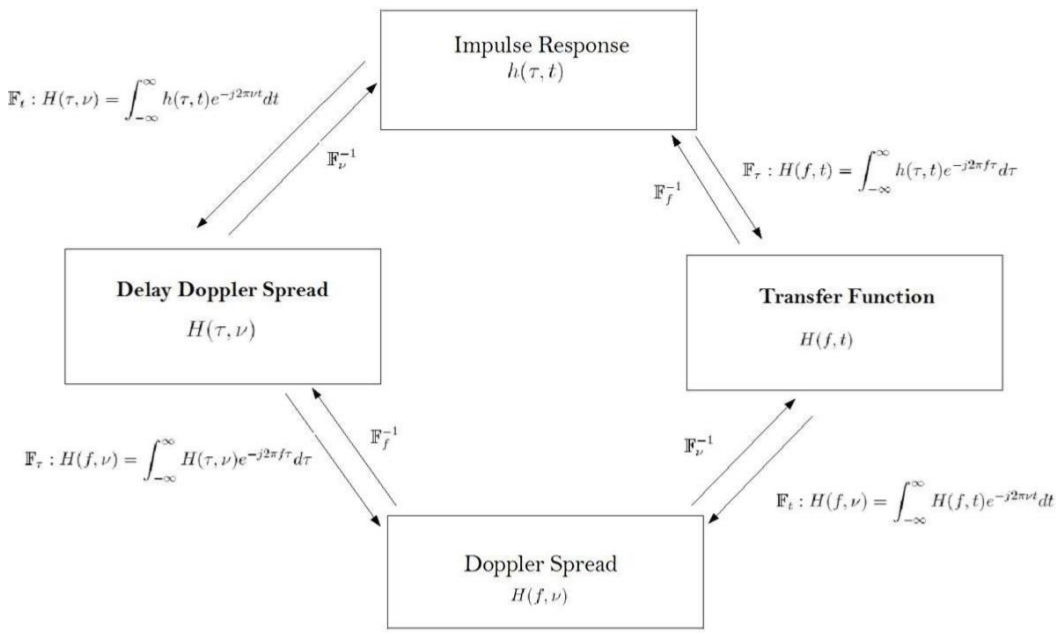


Figure : Relationship among different channel functions.

where the channel becomes multiplicative.

Doppler spread functions:

$$H(f, \nu) = FT\,[H(f, t)] = \int_{-\infty}^{\infty} H(f, t)e^{-j2\pi\nu t}dt$$

and

$$H(f, t) = FT^{-1}[H(f, \nu)] = \int_{-\infty}^{\infty} H(f, \nu)e^{j2\pi\nu t}d\nu$$

Delay Doppler spread:

$$H(\tau, \nu) = FT\,[h(\tau, t)] = \int_{-\infty}^{\infty} h(\tau, t)e^{-j2\pi\nu t}dt$$

Small-Scale Multipath Measurements

Direct RF Pulse System

A wideband pulsed bistatic radar usually transmits a repetitive pulse of width T_{bb} s, and uses a receiver with a wide bandpass filter ($BW = \frac{2}{T_{bb}}$ Hz). The signal is then amplified, envelope detected, and displayed and stored on a high speed oscilloscope. Immediate measurements of the square of the channel impulse response convolved with the probing pulse can be taken. If the oscilloscope is set on averaging mode, then this system provides a local average power delay profile.

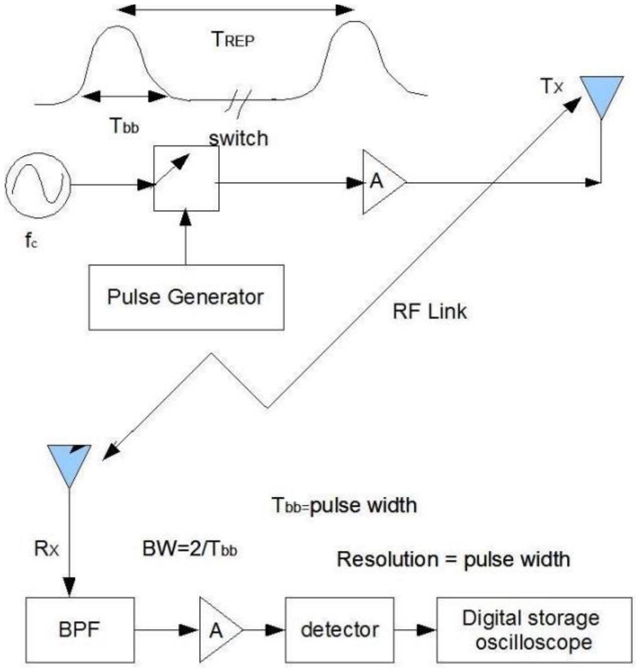


Figure : Direct RF pulsed channel IR measurement.

This system is subject to interference noise. If the first arriving signal is blocked or fades, severe fading occurs, and it is possible the system may not trigger properly.

Frequency Domain Channel Sounding

In this case we measure the channel in the frequency domain and then convert it into time domain impulse response by taking its inverse discrete Fourier transform (IDFT). A vector network analyzer controls a swept frequency synthesizer. An S- parameter test set is used to monitor the frequency response of the channel. The sweeper scans a particular frequency band, centered on the carrier, by stepping through discrete frequencies. The number and spacing of the frequency step impacts the time resolution of the impulse response measurement. For each frequency step, the S-parameter test set transmits a known signal level at port 1 and monitors the received signal at port 2. These signals allow the analyzer to measure the complex response, $S_{21}(\omega)$, of the channel over the measured frequency range. The $S_{21}(\omega)$ measure is the measure of the signal flow from transmitter antenna to receiver

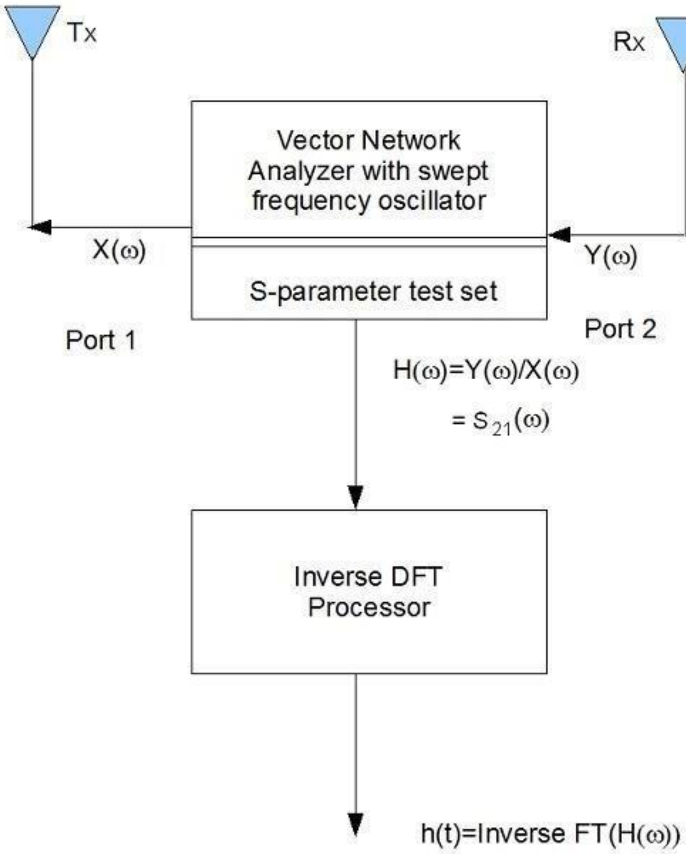


Figure : Frequency domain channel IR measurement.

antenna (i.e., the channel).

This system is suitable only for indoor channel measurements. This system is also non real-time. Hence, it is not suitable for time-varying channels unless the sweep times are fast enough.

Multipath Channel Parameters

To compare the different multipath channels and to quantify them, we define some parameters. They all can be determined from the power delay profile. These pa- rameters can be broadly divided in to two types.

Time Dispersion Parameters

These parameters include the mean excess delay,rms delay spread and excess delay spread. The mean excess delay is the first moment of the power delay profile and is defined as

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

where a_k is the amplitude, τ_k is the excess delay and $P(\tau_k)$ is the power of the individual multipath signals.

The mean square excess delay spread is defined as

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

Since the rms delay spread is the square root of the second central moment of the power delay profile, it can be written as

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

As a rule of thumb, for a channel to be flat fading the following condition must be satisfied

$$\frac{\sigma_\tau}{T_S} \leq 0.1$$

where T_S is the symbol duration. For this case, no equalizer is required at the receiver.

Example 2

1. Sketch the power delay profile and compute RMS delay spread for the follow- ing:
   $P(\tau) = \sum_{n=0} \delta(\tau - n \times 10^{-6})$ (in watts)

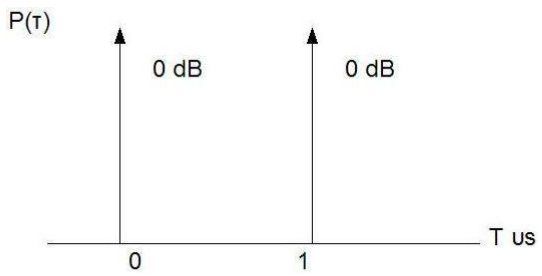
2. If BPSK modulation is used, what is the maximum bit rate that can be sent through the channel without needing an equalizer?

Solution

1. $P(0) = 1$ watt, $P(1) = 1$ watt

$$\bar{\tau} = \frac{(1)(0) + (1)(1)}{1 + 1} = 0.5\mu s$$

$$\overline{\tau^2} = 0.5\mu s^2 \quad \sigma_\tau = 0.5\mu s$$

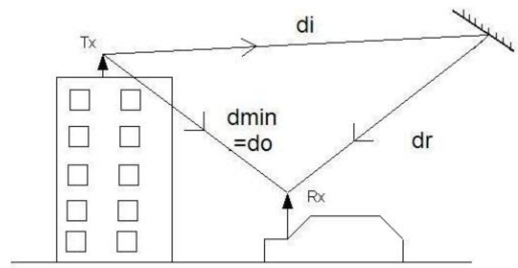


2. For flat fading channel, we need $\frac{\sigma_\tau}{T_s} 0.1 \quad \Rightarrow R_s = \frac{1}{T_s} = 0.2 \times 10^4 = 200\ kbps$

   For BPSK we need $R_b = R_s = 200\ kbps$

Example 3 A simple delay spread bound: **Feher's upper bound**

Consider a simple worst-case delay spread scenario as shown in figure below.



Here $d_{min} = d_0$ and $d_{max} = d_i + d_r$

Transmitted power = P_T, Minimum received power = $P_{R_{\min}} = P_{Threshold}$

$$\frac{P_{R_{\min}}}{P_T} = tt_T\, tt_R \left(\frac{\lambda}{4\pi d_{\max}}\right)^2$$

Put $tt_T = tt_R = 1$ i.e., considering omni-directional unity gain antennas

$$d_{\max} = \left(\frac{\lambda}{4\pi}\right)\left(\frac{P_T}{P_{R_{\min}}}\right)^{\frac{1}{2}}$$

$$\tau_{\max} = \frac{d_{\max}}{c} = \left(\frac{\lambda}{4\pi c}\right)\left(\frac{P_T}{P_{R_{\min}}}\right)^{\frac{1}{2}}$$

$$\tau_{\max} = \left(\frac{1}{4\pi f}\right)\left(\frac{P_T}{P_{R_{\min}}}\right)^{\frac{1}{2}}$$

Frequency Dispersion Parameters

To characterize the channel in the frequency domain, we have the following param- eters.

(1) Coherence bandwidth: it is a statistical measure of the range of frequencies over which the channel can be considered to pass all the frequency components with almost equal gain and linear phase. When this condition is satisfied then we say the channel to be flat.

Practically, coherence bandwidth is the minimum separation over which the two frequency components are affected differently. If the coherence bandwidth is con- sidered to be the bandwidth over which the frequency correlation function is above 0.9, then it is approximated as

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

However, if the coherence bandwidth is considered to be the bandwidth over which the frequency correlation function is above 0.5, then it is defined as

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

The coherence bandwidth describes the time dispersive nature of the channel in the local area. A more convenient parameter to study the time variation of the channel is the coherence time. This variation may be due to the relative motion between the mobile and the base station or the motion of the objects in the channel.

(2) Coherence time: this is a statistical measure of the time duration over which the channel impulse response is almost invariant. When channel behaves like this, it is said to be slow faded. Essentially it is the minimum time duration over which two received signals are affected differently. For an example, if the coherence time is considered to be the bandwidth over which the time correlation is above 0.5, then it can be approximated as

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

where f_m is the maximum doppler spread given be $f_m = \frac{v}{\lambda}$.

Another parameter is the Doppler spread (B_D) which is the range of frequencies over which the received Doppler spectrum is non zero.

Statistical models for multipath propagation

Many multipath models have been proposed to explain the observed statistical na- ture of a practical mobile channel. Both the first order and second order statistics

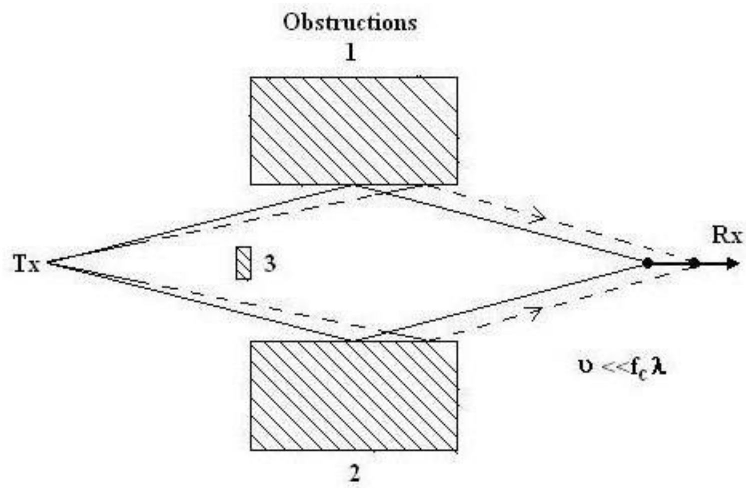


Figure: Two ray NLoS multipath, resulting in Rayleigh fading.

have been examined in order to find out the effective way to model and combat the channel effects. The most popular of these models are Rayleigh model, which de- scribes the NLoS propagation. The Rayleigh model is used to model the statistical time varying nature of the received envelope of a flat fading envelope. Below, we discuss about the main first order and second order statistical models.

NLoS Propagation: Rayleigh Fading Model

Let there be two multipath signals S_1 and S_2 received at two different time instants due to the presence of obstacles as shown in Figure 5.6. Now there can either be constructive or destructive interference between the two signals.

Let E_n be the electric field and Θ_n be the relative phase of the various multipath signals.So we have

$$\tilde{E} = \sum_{n=1}^{N} E_n e^{j\theta_n}$$

Now if N→ ∞(i.e. are sufficiently large number of multipaths) and all the E_n are IID distributed, then by Central Limit Theorem we have,

$$\lim_{N\to\infty} \tilde{E} = \lim_{N\to\infty} \sum_{n=1}^{N} E_n e^{j\theta_n}$$

$$= Z_r + jZ_i = Re^{j\varphi}$$

where Z_r and Z_i are Gaussian Random variables. For the above case

$$R = \sqrt{Z_r^2 + Z_i^2}$$

and

$$\varphi = \tan^{-1} \frac{Z_i}{Z_r}$$

For all practical purposes we assume that the relative phase Θ_n is uniformaly distributed.

$$E[e^{j\theta_n}] = \frac{1}{2\pi} \int^{2\pi} e^{j\theta} d\theta = 0$$

It can be seen that E_n and Θ_n are independent. So,

$$E[\tilde{E}] = E[\sum E_n e^{j\theta_n}] = 0$$

$$E[|\tilde{E}|^2] = E[\sum E_n e^{j\theta_n} \sum E_n^* e^{-j\theta_n}] = E[\sum_m \sum_n E_n E_m e^{j(\theta_n - \theta_m)}] = \sum_{n=1}^{N} E_n^2 = P_0$$

where P_0 is the total power obtained. To find the Cumulative Distribution Func- tion(CDF) of R, we proceed as follows.

$$F_R(r) = P_r(R \leq r) = \int\int_A f_{Z_i,Z_r}(z_i, z_r) dz_i dz_r$$

where A is determined by the values taken by the dummy variable r. Let Z_i and Z_r be zero mean Gaussian RVs. Hence the CDF can be written as

$$F_R(r) = \int\int_A \frac{1}{\sqrt{2\pi\sigma^2}} e^{\frac{-(Z_r^2+Z_i^2)}{2\sigma^2}} dZ_i dZ_r$$

Let $Z_r = p\cos(\Theta)$ and $Z_i = p\sin(\Theta)$ So we have

$$F_R(r) = \int_0^{2\pi}\int_0^{2\pi} \frac{1}{\sqrt{2\pi\sigma^2}} e^{\frac{-p^2}{2\sigma^2}} p dp d\theta$$

$$= 1 - e^{\frac{-r^2}{2\sigma^2}}$$

Above equation is valid for all $r \geq 0$. The pdf can be written as

$$f_R(r) = \frac{r}{\sigma^2} e^{-\frac{r^2}{2\sigma^2}}$$

and is shown in Figure with different σ values. This equation too is valid for all $r \geq 0$. Above distribution is known as Rayleigh distribution and it has been derived

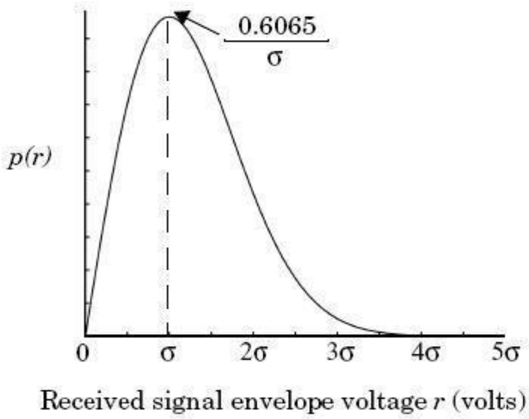


Figure 5.7: Rayleigh probability density function.

for slow fading. However, if f_D 1 Hz, we call it as Quasi-stationary Rayleigh fading. We observe the following:

$$E[R] = \sqrt{\frac{\pi}{2}}\sigma$$

$$E[R^2] = 2\sigma^2$$

$$\mathrm{var}[R] = (2 - \frac{\pi}{2})\sigma^2$$

$$median[R] = 1.77\sigma.$$

LoS Propagation: Rician Fading Model

Rician Fading is the addition to all the normal multipaths a direct LOS path.

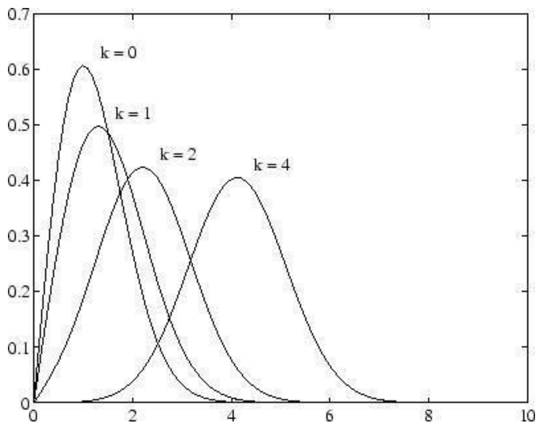


Figure : Ricean probability density function.

$$f_R(r) = \frac{r}{\sigma^2} e^{\frac{-(r^2+A^2)}{2\sigma^2}} I_0\left(\frac{Ar}{\sigma^2}\right)$$

for all $A \geq 0$ and $r \geq 0$. Here A is the peak amplitude of the dominant signal and $I_0(.)$ is the modified Bessel function of the first kind and zeroth order.

A factor K is defined as

$$K_{dB} = 10 \log \frac{A^2}{2\sigma^2}$$

As A $\longrightarrow$ 0 then $K_{dB} \longrightarrow \infty$.

Generalized Model: Nakagami Distribution

A generalization of the Rayleigh and Rician fading is the Nakagami distribution.

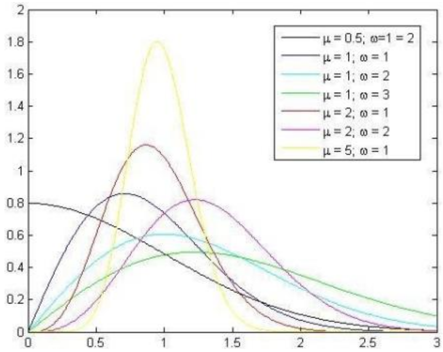


Figure: Nakagami probability density function.

Its pdf is given as,

$$f_R(r) = \frac{2r^{m-1}}{\Gamma(m)} \left(\frac{m^m}{\Omega^m}\right) e^{\frac{-mr^2}{\Omega}}$$

where,

$\Gamma(m)$ is the gamma function

Ω is the average signal power and

m is the fading factor.It is always greater than or equal to 0.5.

When m=1, Nakagami model is the Rayleigh model. When

$$m = \frac{(M+1)^2}{2M+1}$$

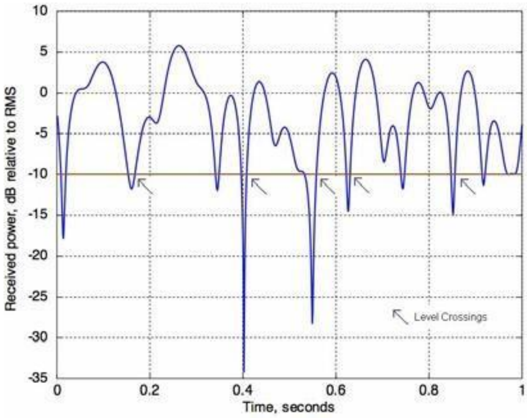


Figure : Schematic representation of level crossing with a Rayleigh fading enve- lope at 10 Hz Doppler spread.

where

$$M = \frac{A}{2\sigma}$$

Nakagami fading is the Rician fading.

As m →∞ Nakagami fading is the impulse channel and no fading occurs.

Second Order Statistics

To design better error control codes, we have two important second order param- eters of fading model, namely the **level crossing rate** (LCR) and **average fade duration** (AFD). These parameters can be utilized to assess the speed of the user by measuring them through the reverse channel. The LCR is the expected rate at which the Rayleigh fading envelope normalized to the local rms amplitude crosses a specific level 'R' in a positive going direction.

$$N_R = \int_0^{\infty} \dot{r}\, p(R, \dot{r})\, d\dot{r} = \sqrt{2\pi} f_D \rho e^{-\rho^2}$$

where $\dot{r}$ is the time derivative of r(t), f_D is the maximum Doppler shift and ρ is the value of the specified level R, normalized to the local rms amplitude of the fading envelope.

The other important parameter, AFD, is the average period time for which the receiver power is below a specified level R.

$$\bar{\tau} = \frac{1}{N_r} P_r(r \leq R)$$

As

$$P_r(r \leq R) = \int_0^R p(r)dr = 1 - e^{-\rho^2},$$

therefore,

$$\bar{\tau} = \frac{1 - e^{-\rho^2}}{\sqrt{2\pi} f_D \rho e^{-\rho^2}}$$

$$= \frac{e^{\rho^2} - 1}{\sqrt{2\pi} f_D \rho}.$$

Apart from LCR, another parameter is fading rate, which is defined as the number of times the signal envelope crosses the middle value (r_m) in a positive going direction per unit time. The average rate is expressed as

$$N(r_m) = \frac{2v}{\lambda}.$$

Another statistical parameter, sometimes used in the mobile communication, is called as depth of fading. It is defined as the ratio between the minimum value and the mean square value of the faded signal. Usually, an average value of 10% as depth of fading gives a marginal fading scenario.

Simulation of Rayleigh Fading Models

Clarke's Model: without Doppler Effffect

In it, two independent Gaussian low pass noise sources are used to produce in-phase and quadrature fading branches. This is the basic model and is useful for slow fading channel. Also the Doppler effect is not accounted for.

Clarke and Gans' Model: with Doppler Effffect

In this model, the output of the Clarke's model is passed through Doppler filter in the RF or through two initial baseband Doppler filters for baseband processing as shown in Figure 5.11. Here, the obtained Rayleigh output is flat faded signal but not frequency selective.

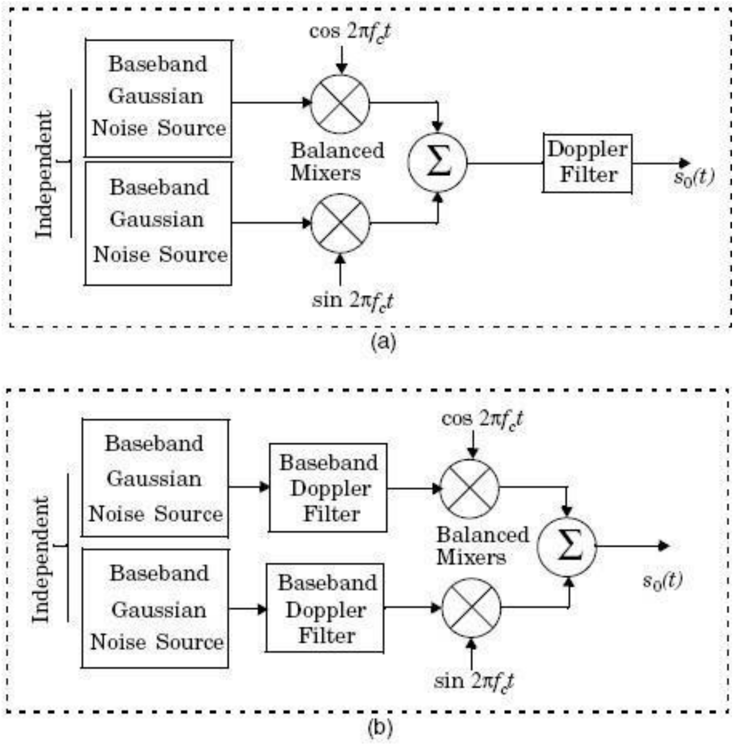


Figure: Clarke and Gan's model for Rayleigh fading generation using quadra- ture amplitude modulation with (a) RF Doppler filter, and, (b) baseband Doppler filter.

Rayleigh Simulator with Wide Range of Channel Conditions

To get a frequency selective output we have the following simulator through which both the frequency selective and flat faded Rayleigh signal may be obtained. This is achieved through varying the parameters a_i and τ_i, as given in Figure

Two-Ray Rayleigh Faded Model

The above model is, however, very complex and difficult to implement. So, we have the two ray Rayleigh fading model which can be easily implemented in software as shown in Figure 5.13.

$$h_b(t) = \alpha_1 e^{j\varphi_1}\delta(t) + \alpha_2 e^{j\varphi_2}\delta(t-\tau)$$

where α_1 and α_2 are independent Rayleigh distributed and φ_1 and φ_2 are indepen- dent and uniformaly distributed over 0 to 2π. By varying τ it is possible to create a wide range of frequency selective fading effects.

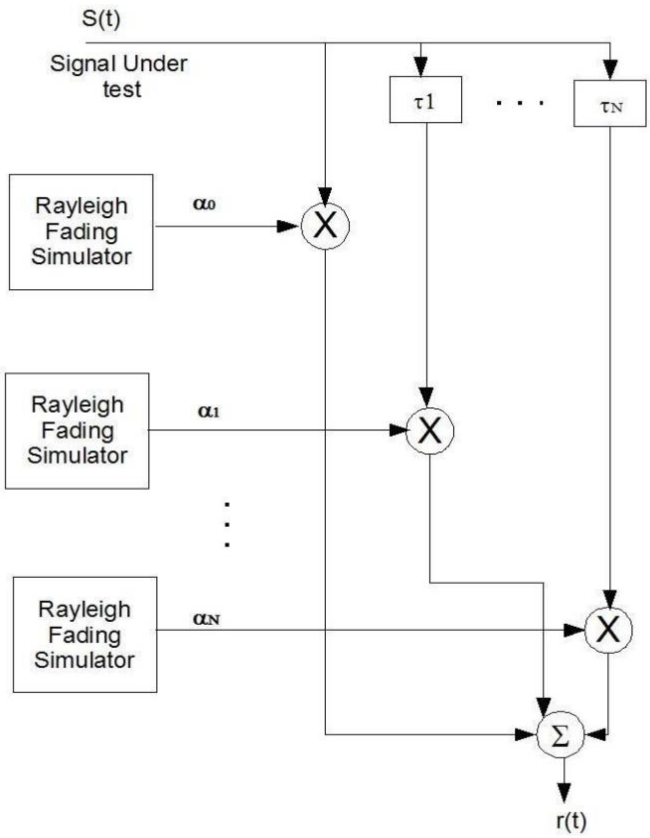


Figure : Rayleigh fading model to get both the flat and frequency selective channel conditions.

Saleh and Valenzuela Indoor Statistical Model

This method involved averaging the square law detected pulse response while sweep- ing the frequency of the transmitted pulse. The model assumes that the multipath components arrive in clusters. The amplitudes of the received components are in- dependent Rayleigh random variables with variances that decay exponentially with cluster delay as well as excess delay within a cluster. The clusters and multipath components within a cluster form Poisson arrival processes with different rates.

SIRCIM/SMRCIM Indoor/Outdoor Statistical Models

SIRCIM (Simulation of Indoor Radio Channel Impulse-response Model) generates realistic samples of small-scale indoor channel impulse response measurements. Sub-

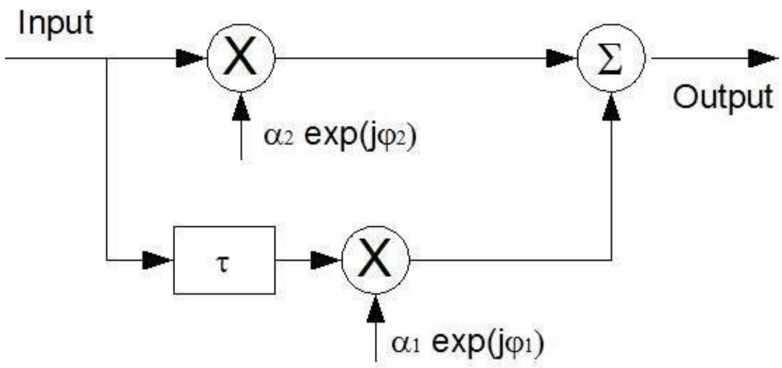


Figure 5.13: Two-ray Rayleigh fading model.

sequent work by Huang produced SMRCIM (Simulation of Mobile Radio Channel Impulse-response Model), a similar program that generates small-scale urban cellu- lar and micro-cellular channel impulse responses.

Conclusion

In this chapter, the main channel impairment, i.e., fading, has been introduced which becomes so severe sometimes that even the large scale path loss becomes insignificant in comparison to it. Some statistical propagation models have been presented based on the fading characteristics. Mainly the frequency selective fading, fast fading and deep fading can be considered the major obstruction from the channel severity view point.

UNIT IV

Equalization and Diversity

Introduction

Apart from the better transmitter and receiver technology, mobile communications require signal processing techniques that improve the link performance. Equaliza- tion, Diversity and channel coding are channel impairment improvement techniques. Equalization compensates for Inter Symbol Interference (ISI) created by multipath within time dispersive channels. An equalizer within a receiver compensates for the average range of expected channel amplitude and delay characteristics. In other words, an equalizer is a filter at the mobile receiver whose impulse response is inverse of the channel impulse response. As such equalizers find their use in frequency selective fading channels. Diversity is another technique used to compensate fast fading and is usually implemented using two or more receiving antennas. It is usually em- ployed to reduce the depths and duration of the fades experienced by a receiver in a flat fading channel. Channel coding improves mobile communication link performance by adding redundant data bits in the transmitted message.At the baseband portion of the transmitter, a channel coder maps a digital message sequence in to another specific code sequence containing greater number of bits than original con- tained in the message. Channel Coding is used to correct deep fading or spectral null. We discuss all three of these techniques in this chapter. A general framework of the fading effects and their mitigation techniques is shown in Figure4.1.

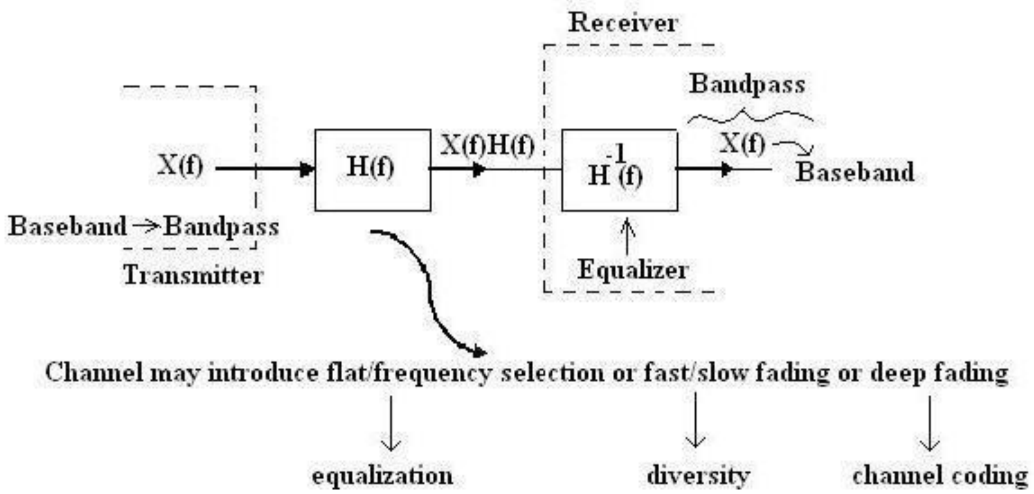


Figure 4.1: A general framework of fading effects and their mitigation techniques.

Equalization

ISI has been identified as one of the major obstacles to high speed data transmission over mobile radio channels. If the modulation bandwidth exceeds the coherence bandwidth of the radio channel (i.e., frequency selective fading), modulation pulses are spread in time, causing ISI. An equalizer at the front end of a receiver compen- sates for the average range of expected channel amplitude and delay characteristics. As the mobile fading channels are random and time varying, equalizers must track the time-varying characteristics of the mobile channel and therefore should be time- varying or adaptive. An adaptive equalizer has two phases of operation: training and tracking. These are as follows.

Training Mode:

- Initially a known, fixed length training sequence is sent by the transmitter so that the receiver equalizer may average to a proper setting.
- Training sequence is typically a pseudo-random binary signal or a fixed, of prescribed bit pattern.
- The training sequence is designed to permit an equalizer at the receiver to acquire the proper filter coefficient in the worst possible channel condition. An adaptive filter at the receiver thus uses a recursive algorithm to evaluate the channel and estimate filter coefficients to compensate for the channel.

Tracking Mode:

- When the training sequence is finished the filter coefficients are near optimal.
- Immediately following the training sequence, user data is sent.
- When the data of the users are received, the adaptive algorithms of the equal- izer tracks the changing channel.
- As a result, the adaptive equalizer continuously changes the filter characteris- tics over time.

A Mathematical Framework

The signal received by the equalizer is given by

$$x(t) = d(t) * h(t) + n_b(t) \tag{4.1}$$

where $d(t)$ is the transmitted signal, $h(t)$ is the combined impulse response of the transmitter,channel and the RF/IF section of the receiver and $n_b(t)$ denotes the baseband noise.

If the impulse response of the equalizer is $h_{eq}(t)$, the output of the equalizer is

$$\hat{y}(t) = d(t) * h(t) * h_{eq}(t) + n_b(t) * h_{eq}(t) = d(t) * g(t) + n_b(t) * h_{eq}(t). \tag{4.2}$$

However, the desired output of the equalizer is $d(t)$ which is the original source data. Assuming $n_b(t)=0$, we can write $y(t) = d(t)$, which in turn stems the following equation:

$$g(t) = h(t) * h_{eq}(t) = \delta(t) \tag{4.3}$$

The main goal of any equalization process is to satisfy this equation optimally. In frequency domain it can be written as

$$H_{eq}(f)\, H(f) = 1 \tag{4.4}$$

which indicates that an equalizer is actually an inverse filter of the channel. If the channel is frequency selective, the equalizer enhances the frequency components with small amplitudes and attenuates the strong frequencies in the received frequency spectrum in order to provide a flat, composite received frequency response and linear phase response. For a time varying channel, the equalizer is designed to track the channel variations so that the above equation is approximately satisfied.

Zero Forcing Equalization

In a zero forcing equalizer, the equalizer coefficients c_n are chosen to force the samples of the combined channel and equalizer impulse response to zero. When each of the delay elements provide a time delay equal to the symbol duration T, the frequency response $H_{eq}(f)$ of the equalizer is periodic with a period equal to the symbol rate 1/T. The combined response of the channel with the equalizer must satisfy Nyquist's criterion

$$H_{ch}(f)\, H_{eq}(f) = 1, |f| < {}^{1}/_{2T} \tag{4.5}$$

where $H_{ch}(f)$ is the folded frequency response of the channel. Thus, an infinite length zero-forcing ISI equalizer is simply an inverse filter which inverts the folded frequency response of the channel.

Disadvantage: Since $H_{eq}(f)$ is inverse of $H_{ch}(f)$ so inverse filter may excessively amplify the noise at frequencies where the folded channel spectrum has high atten- uation, so it is rarely used for wireless link except for static channels with high SNR such as local wired telephone. The usual equalizer model follows a time varying or adaptive structure which is given next.

A Generic Adaptive Equalizer

The basic structure of an adaptive filter is shown in Figure 4.2. This filter is called the transversal filter, and in this case has N delay elements, N+1 taps and N+1 tunable complex multipliers, called weights. These weights are updated continuously by an adaptive algorithm. In the figure the subscript k represents discrete time index. The adaptive algorithm is controlled by the error signal e_k. The error signal is derived by comparing the output of the equalizer, with some signal d_k which is replica of transmitted signal. The adaptive algorithm uses e_k to minimize the cost function and uses the equalizer weights in such a manner thatit minimizes the cost function iteratively. Let us denote the received sequence vector at the receiver and

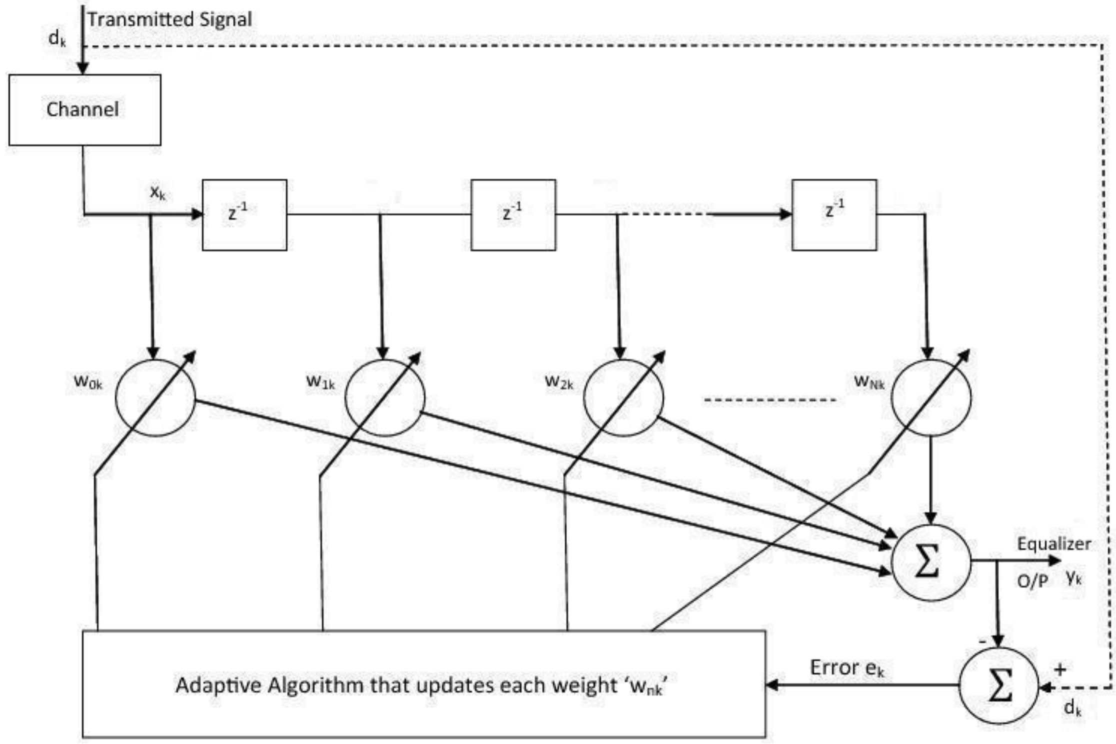


Figure 4.2: A generic adaptive equalizer. the input to the equalizer as

$$\mathbf{x}_k = [x_k, x_{k-1}, \ldots.., x_{k-N}]^T, \tag{4.6}$$

and the tap coefficient vector as

$$\mathbf{w}_k = [w_k^0, w_k^1, \ldots.., w_k^N]^T. \tag{4.7}$$

Now, the output sequence of the equalizer y_k is the inner product of $\mathbf{x}_k$ and $\mathbf{w}_k$, i.e.,

$$y_k = \langle \mathbf{x}_k, \mathbf{w}_k \rangle = \mathbf{x}_k^T \mathbf{w}_k = \mathbf{w}_k^T \mathbf{x}_k. \tag{4.8}$$

The error signal is defined as

$$e_k = d_k - y_k = d_k - \mathbf{x}_k^T \mathbf{w}_k. \tag{4.9}$$

Assuming d_k and $\mathbf{x}_k$ to be jointly stationary, the Mean Square Error (MSE) is given as

$$\begin{aligned} MSE &= E[e_k^2] = E[(d_k - y_k)^2] \\ &= E[(d_k - \mathbf{x}_k^T \mathbf{w}_k)^2] \\ &= E[d_k^2] + \mathbf{w}_k^T E[\mathbf{x}_k \mathbf{x}_k^T]\mathbf{w}_k - 2E[d_k \mathbf{x}_k^T]\mathbf{w}_k \end{aligned} \tag{4.10}$$

where $\mathbf{w}_k$ is assumed to be an array of optimum values and therefore it has been taken out of the $E()$ operator. The MSE then can be expressed as

$$MSE = \xi = \sigma_d^2 + \mathbf{w}_k^T \mathbf{R} \mathbf{w}_k - 2\mathbf{p}^T \mathbf{w}_k \qquad (4.11)$$

where the signal variance $\sigma_d^2 = E[d_k^2]$ and the cross correlation vector $\mathbf{p}$ between the desired response and the input signal is defined as

$$p = E\left[d_k \mathbf{x}_k\right] = E\left[\begin{matrix} d_k x_k & d_k x_{k-1} & d_k x_{k-2} & \cdots & d_k x_{k-N} \end{matrix}\right]^T. \qquad (4.12)$$

The input correlation matrix $\mathbf{R}$ is defined as an $(N + 1) \times (N + 1)$ square matrix, where

$$R = E\left[\mathbf{x}_k \mathbf{x}_k^T\right] = E\begin{bmatrix} x_k^2 & x_k x_{k-1} & x_k x_{k-2} & \cdots & x_k x_{k-N} \\ x_{k-1} x_k & x_{k-1}^2 & x_{k-1} x_{k-2} & \cdots & x_{k-1} x_{k-N} \\ x_{k-2} x_k & x_{k-2} x_{k-1} & x_{k-2}^2 & \cdots & x_{k-2} x_{k-N} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ x_{k-N} x_k & x_{k-N} x_{k-1} & x_{k-N} x_{k-2} & \cdots & x_{k-N}^2 \end{bmatrix}. \qquad (4.13)$$

Clearly, MSE is a function of $\mathbf{w}_k$. On equating $\frac{\partial \xi}{\partial w_k}$

to 0, we get the condition for
minimum MSE (MMSE) which is known as Wiener solution:

$$\mathbf{w}_k = \mathbf{R}^{-1}\mathbf{p}. \tag{4.14}$$

Hence, MMSE is given by the equation

$$MMSE = \xi_{min} = \sigma_d^2 - \mathbf{p}^T\mathbf{w}_k. \tag{4.15}$$

Choice of Algorithms for Adaptive Equalization

Since an adaptive equalizer compensates for an unknown and time varying channel, it requires a specific algorithm to update the equalizer coefficients and track the channel variations. Factors which determine algorithm's performance are:

Rate of convergence: Number of iterations required for an algorithm, in re- sponse to a stationary inputs, to converge close enough to optimal solution. A fast rate of convergence allows the algorithm to adapt rapidly to a stationary environ- ment of unknown statistics.

Misadjustment: Provides a quantitative measure of the amount by which the final value of mean square error, averaged over an ensemble of adaptive filters, deviates from an optimal mean square error.

Computational complexity: Number of operations required to make one com- plete iteration of the algorithm.

Numerical properties: Inaccuracies like round-off noise and representation errors in the computer, which influence the stability of the algorithm.

Three classic equalizer algorithms are primitive for most of today's wireless stan- dards. These include the Zero Forcing Algorithm (ZF), the Least Mean Square Algo- rithm (LMS), and the Recursive Least Square Algorithm (RLS). Below, we discuss a few of the adaptive algorithms.

Least Mean Square (LMS) Algorithm

LMS algorithm is the simplest algorithm based on minimization of the MSE between the desired equalizer output and the actual equalizer output, as discussed earlier. Here the system error, the MSE and the optimal Wiener solution remain the same as given the adaptive equalization framework.

In practice, the minimization of the MSE is carried out recursively, and may be performed by use of the stochastic gradient algorithm. It is the simplest equalization algorithm and requires only 2N+1 operations per iteration. The filter weights are updated by the update equation. Letting the variable n denote the sequence of iteration, LMS is computed iteratively by

$$w_k(n+1) = w_k(n) + \mu e_k(n)\, x(n-k) \tag{4.16}$$

where the subscript k denotes the *k*th delay stage in the equalizer and μ is the step size which controls the convergence rate and stability of the algorithm.

The LMS equalizer maximizes the signal to distortion ratio at its output within the constraints of the equalizer filter length. If an input signal has a time dispersion characteristics that is greater than the propagation delay through the equalizer, then the equalizer will be unable to reduce distortion. The convergence rate of the LMS algorithm is slow due to the fact that there is only one parameter, the step size, that controls the adaptation rate. To prevent the adaptation from becoming unstable, the value of μ is chosen from

$$0 < \mu < \frac{2}{\sum_{i=1}^{N} \lambda_i} \tag{4.17}$$

where λ_i is the i-th eigenvalue of the covariance matrix R.

Normalized LMS (NLMS) Algorithm

In the LMS algorithm, the correction that is applied to $w_k(n)$ is proportional to the input sample $x(n-k)$. Therefore when $x(n-k)$ is large, the LMS algorithm experiences gradient noise amplification. With the normalization of the LMS step size by $\|\mathbf{x}(n)\|^2$ in the NLMS algorithm, this problem is eliminated. Only when $x(n-k)$ becomes close to zero, the denominator term $\|\mathbf{x}(n)\|^2$ in the NLMS equation becomes very small and the correction factor may diverge. So, a small positive number ε is added to the denominator term of the correction factor. Here, the step size is time varying and is expressed as

$$\mu(n) = \frac{\beta}{\|\mathbf{x}(n)\|^2 + \varepsilon}. \tag{4.18}$$

Therefore, the NLMS algorithm update equation takes the form of

$$w_k(n+1) = w_k(n) + \frac{\beta}{\|\mathbf{x}(n)\|^2 + \varepsilon} e_k(n)\, x(n-k). \tag{4.19}$$

Diversity

Diversity is a method used to develop information from several signals transmitted over independent fading paths. It exploits the random nature of radio propagation by finding independent signal paths for communication. It is a very simple concept where if one path undergoes a deep fade, another independent path may have a strong signal. As there is more than one path to select from, both the instantaneous and average SNRs at the receiver may be improved. Usually diversity decisions are made by receiver. Unlike equalization, diversity requires no training overhead as a training sequence is not required by transmitter. Note that if the distance between two receivers is a multiple of $\lambda/2$, there might occur a destructive interference be- tween the two signals. Hence receivers in diversity technique are used in such a way thatthe signal received by one is independent of the other. Diversity can be of various forms, starting from space diversity to time diversity. We take up the types one by one in the sequel.

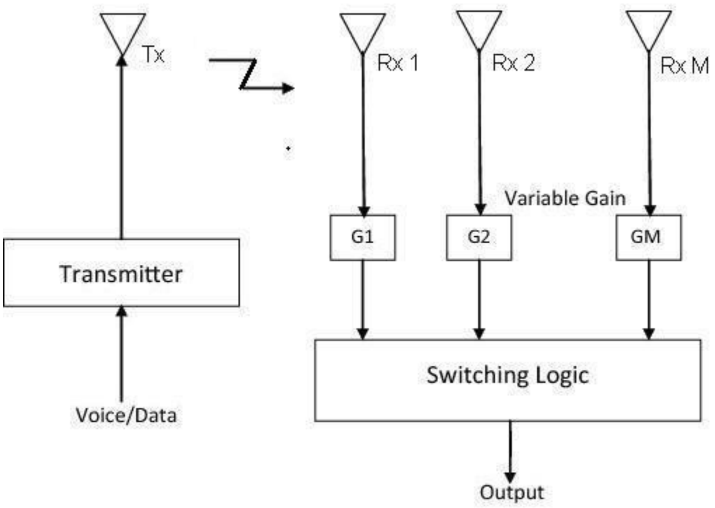


Figure 4.3: Receiver selection diversity, with M receivers.

Different Types of Diversity

Space Diversity

A method of transmission or reception, or both, in which the effects of fading are minimized by the simultaneous use of two or more physically separated antennas, ideally separated by one half or more wavelengths. Signals received from spatially separated antennas have uncorrelated envelopes. Space diversity reception methods can be classified into four categories: selection, feedback or scanning, maximal ratio combining and equal gain combining.

(a) Selection Diversity:

The basic principle of this type of diversity is selecting the best signal among all the signals received from different branches at the receiving end. Selection Diversity is the simplest diversity technique. Figure 7.3 shows a block diagram of this method where 'M' demodulators are used to provide M diversity branches whose gains are adjusted to provide the same average SNR for each branch. The receiver branches having the highest instantaneous SNR is connected to the demodulator.

Let M independent Rayleigh fading channels are available at a receiver. Each channel is called a diversity branch and let each branch has the same average SNR. The signal to noise ratio is defined as

$$SNR = \Gamma = \frac{E_b}{N_0}\alpha^2 \qquad (4.20)$$

where E_b is the average carrier energy, N_0 is the noise PSD, α is a random variable used to represent amplitude values of the fading channel.

The instantaneous SNR(γ_i) is usually defined as γ_i = instantaneous signal power per branch/mean noise power per branch. For Rayleigh fading channels, α has a Rayleigh distribution and so α^2 and consequently γ_i have a chi-square distribution
with two degrees of freedom. The probability density function for such a channel is

$$p(\gamma_i) = \frac{1}{\Gamma} e^{\frac{-\gamma_i}{\Gamma}} \tag{7.21}$$

The probability that any single branch has an instantaneous SNR less than some defined threshold γ is

$$\Pr[\gamma_i \leq \gamma] = \int_0^{\gamma} p(\gamma_i)\, d\gamma_i = \int_0^{\gamma} \frac{1}{\Gamma} e^{\frac{-\gamma_i}{\Gamma}}\, d\gamma_i = 1 - e^{\frac{-\gamma}{\Gamma}} = P(\Gamma). \tag{7.22}$$

Similarly, the probability that all M independent diversity branches receive signals which are simultaneously less than some specific SNR threshold γ is

$$\Pr[\gamma_1, \gamma_2, \ldots, \gamma_M \leq \gamma] = \left(1 - e^{\frac{-\gamma}{\Gamma}}\right)^M = P_M(\gamma) \tag{7.23}$$

where $P_M(\gamma)$ is the probability of all branches failing to achieve an instantaneous SNR $= \gamma$. Quite clearly, $P_M(\Gamma) < P(\Gamma)$. If a single branch achieves SNR $> \gamma$, then the probability that SNR $> \gamma$ for one or more branches is given by

$$\Pr[\gamma_i > \gamma] = 1 - P_M(\gamma) = 1 - \left(1 - e^{\frac{-\gamma}{\Gamma}}\right)^M \tag{7.24}$$

which is more than the required SNR for a single branch receiver. This expression shows the advantage when a selection diversity is used.

To determine of average signal to noise ratio, we first find out the pdf of γ as

$$p_M(\gamma) = \frac{d}{d\gamma} P_M(\gamma) = \frac{M}{\Gamma}\left(1 - e^{-\gamma/\Gamma}\right)^{M-1} e^{-\gamma/\Gamma}. \tag{7.25}$$

The average SNR, $\bar{\gamma}$, can be then expressed as

$$\bar{\gamma} = \int_0^{\infty} \gamma\, p_M(\gamma)\, d\gamma = \Gamma \int_0^{\infty} Mx \left(1 - e^{-x}\right)^{M-1} e^{-x} dx \tag{7.26}$$

where $x = \gamma/\Gamma$ and Γ is the average SNR for a single branch, when no diversity is used.

This equation shows an average improvement in the link margin without requir- ing extra transmitter power or complex circuitry, and it is easy to implement as it needed a monitoring station and an antenna switch at the receiver. It is not an optimal diversity technique as it doesn't use all the possible branches simultaneously.

(b) Feedback or Scanning Diversity:

Scanning all the signals in a fixed sequence until the one with SNR more than a predetermined threshold is identified. Feedback or scanning diversity is very similar to selection diversity except that instead of always using the best of N signals, the N signals are scanned in a fixed sequence until one is found to be above a predetermined threshold. This signal is then received until it falls below threshold and the scanning process is again initiated. The resulting fading statistics are somewhat inferior, but the advantage is that it is very simple to implement(only one receiver is required).

(c) Maximal Ratio Combining:

Signals from all of the m branches are weighted according to their individual signal voltage to noise power ratios and then summed. Individual signals must be cophased before being summed, which generally requires an individual receiver and phasing circuit for each antenna element. Produces an output SNR equal to the sum of all individual SNR. Advantage of producing an output with an acceptable SNR even when none of the individual signals are themselves acceptable. Modern DSP techniques and digital receivers are now making this optimal form, as it gives the best statistical reduction of fading of any known linear diversity combiner. In terms of voltage signal,

$$r_m = \sum_{i=1}^{m} tt_i r_i \qquad (7.27)$$

where tt_i is the gain and r_i is the voltage signal from each branch.

(d) Equal Gain Combining:

In some cases it is not convenient to provide for the variable weighting capability required for true maximal ratio combining. In such cases, the branch weights are all set unity, but the signals from each branch are co-phased to provide equal gain combining diversity. It allows the receiver to exploit signals that are simultaneously received on each branch. Performance of this method is marginally inferior to max- imal ratio combining and superior to Selection diversity. Assuming all the tt_i to be

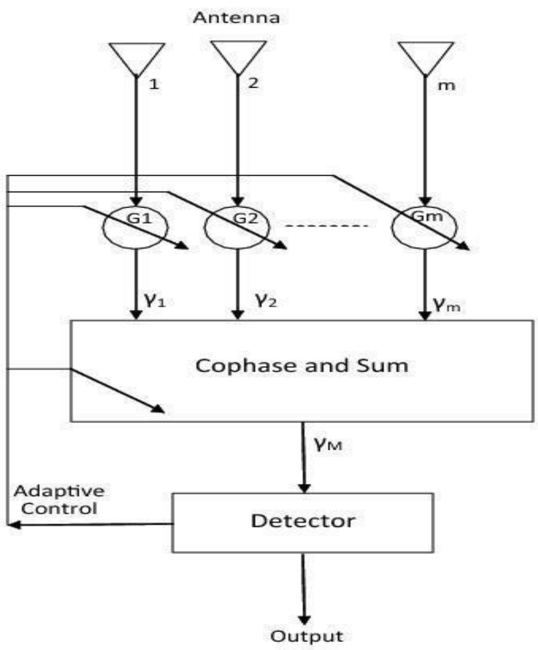


Figure 4.4: Maximal ratio combining technique.

unity, here,

$$r_m = \sum_{i=1}^{m} r_i. \tag{7.28}$$

Polarization Diversity

Polarization Diversity relies on the decorrelation of the two receive ports to achieve diversity gain. The two receiver ports must remain cross-polarized. Polarization Diversity at a base station does not require antenna spacing. Polarization diversity combines pairs of antennas with orthogonal polarizations (i.e. horizontal/vertical, ± slant 45^o, Left-hand/Right-hand CP etc). Reflected signals can undergo polarization changes depending on the channel. Pairing two complementary polarizations, this scheme can immunize a system from polarization mismatches that would otherwise cause signal fade. Polarization diversity has prove valuable at radio and mobile com- munication base stations since it is less susceptible to the near random orientations of transmitting antennas.

Frequency Diversity

In Frequency Diversity, the same information signal is transmitted and received simultaneously on two or more independent fading carrier frequencies. Rationale behind this technique is that frequencies separated by more than the coherence bandwidth of the channel will be uncorrelated and will thus not experience the same fades. The probability of simultaneous fading will be the product of the individual fading probabilities. This method is employed in microwave LoS links which carry several channels in a frequency division multiplex mode (FDM). Main disadvantage is that it requires spare bandwidth also as many receivers as there are channels used for the frequency diversity.

Time Diversity

In time diversity, the signal representing the same information are sent over the same channel at different times. Time diversity repeatedly transmits information at time spacings that exceeds the coherence time of the channel. Multiple repetition of the signal will be received with independent fading conditions, thereby providing for diversity. A modern implementation of time diversity involves the use of RAKE receiver for spread spectrum CDMA, where the multipath channel provides redun- dancy in the transmitted message. Disadvantage is that it requires spare bandwidth also as many receivers as there are channels used for the frequency diversity. Two important types of time diversity application is discussed below.

Application 1: RAKE Receiver

In CDMA spread spectrum systems, CDMA spreading codes are designed to provide very low correlation between successive chips, propagation delay spread in the radio channel provides multiple version of the transmitted signal at the receiver. Delaying multipath components by more than a chip duration, will appear like uncorrelated noise at a CDMA receiver. CDMA receiver may combine the time delayed versions of the original signal to improve the signal to noise ratio at the receiver. RAKE

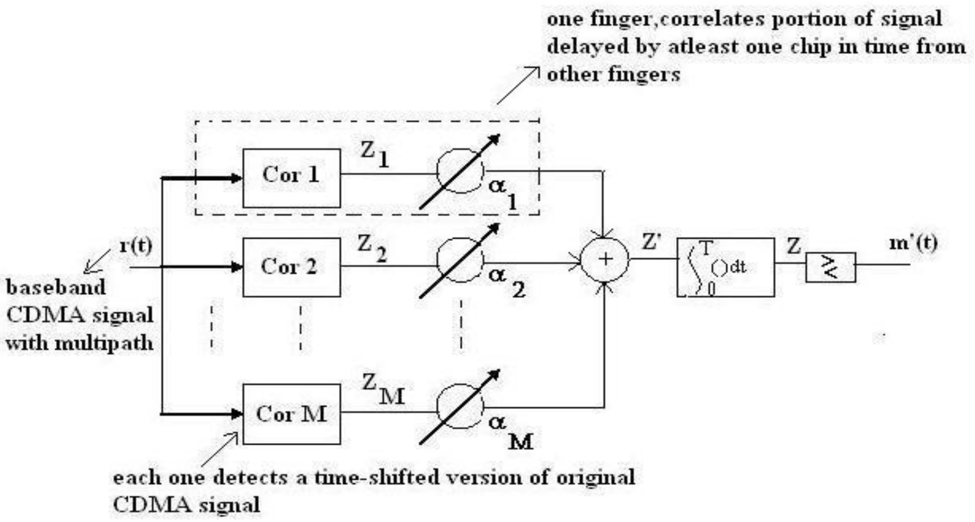


Figure 4.5: RAKE receiver.

receiver collect the time shifted versions of the original signal by providing a sep- arate correlation receiver for M strongest multipath components. Outputs of each correlator are weighted to provide a better estimate of the transmitted signal than provided by a single component. Demodulation and bit decisions are based on the weighted output of the correlators. Schematic of a RAKE receiver is shown in Figure 4.5.

Application 2: Interleaver

In the encoded data bits, some source bits are more important than others, and must be protected from errors. Many speech coder produce several important bits in succession. Interleaver spread these bit out in time so that if there is a deep fade or noise burst, the important bits from a block of source data are not c orruptedat the same time. Spreading source bits over time, it becomes possible to make use of error control coding. Interleaver can be of two forms, a block structure or a convolutional structure.

A block interleaver formats the encoded data into a rectangular array of m rows and n columns, and interleaves nm bits at a time. Each row contains a word of source data having n bits. an interleaver of degree m consists of m rows. source bits are placed into the interleaver by sequentially increasing the row number for each successive bit, and forming the columns. The interleaved source data is then read out row-wise and transmitted over the channel. This has the effect of separating the original source bits by m bit periods. Atthe receiver, de-interleaver stores the received data by sequentially increasing the row number of each successive bit, and then clocks out the data row-wise, one word at a time. Convolutional interleavers are ideally suited for use with convolutional codes.

UNIT -V
WIRELESSNETWORKS

IEEE 802.11 extensions

As the first standard was wrapping up, the creation of a new standards activity begun in the 802.11 standards body. The new activity gave rise to two more standards; IEEE
b and IEEE 802.11a.

- **802.11b:** This standard was developed by IEEE with the support from the consortium Wireless Ethernet Compatibility Alliance (WECA). This standard is backward compatible with the original standard that added two new data rates 5.5 mbps and 11 Mbps using two coding techniques; the mandatory coding mode known as Complementary Coding Keying (CCK) modulation and Packet Binary Convolution Coding (PBCC). Because of backward compatibility with the 802.11, this standard has gained wide popularity with millions of installed base, which is growing rapidly.

802.11a: The successor to 802.11b is 802.11a with greater speed and at a different frequency. It operates at radio frequencies between 5 GHz incorporating a coded multi-carrier scheme known as Orthogonal Frequency Division Multi-carrier (OFDM). The 5 GHz band is currently unlicensed and less congested than the 2.4 GHz ISM band. The 802.11a specifies data speed as high as 54 mbps, also supports 6, 12, 24, and 34 mbps. There is trade off between bandwidth and range - lower bandwidth cases offering increases range. For 54 mbps, the typical range is 20-30 meters. The 802.11a and 802.11b devices can coexist without interference or reduced performance.

- **802.11g:** The success of 802.11b has led to another extension that provides 22 Mbps transmission. It retains backward compatibility with the popular 802.11b standard. This standardwill become 802.11g.

Upper Layers				
802.11 FHSS	802.11 DSSS	802.11a OFDM	802.11b HR-DSSS	802.11g OFDM

WiFi: Any of the above wireless LAN standards are referred to by the brand name **"WiFi"**. It essentially denotes a set of Wireless LAN standards developed by the working group 11 of the IEEE LAN/MAN Standards Committee (IEEE 802).

WiMAX: The story of wireless LAN cannot be complete without the mention of WiMAX, which stands for **W**orldwide **I**nteroperability for **M**icrowave **Acc**ess by the WiMAX Forum. The forum was formed in June 2001 to promote conformance and interoperability of the IEEE 802.16 standard, officially known as Wireless (Metropoliton Area Network) MAN. The Forum describes WiMAX as "a standards-based technology enabling the delivery of last mile wireless broadband access as an alternative to cable and DSL". It supports point to multi-point (PMP) broadband wireless access. WiMAX can deliver a maximum of 70 Mbit/s, over a maximum distance of 70 miles (112.6 kilometers). It has some similarities to DSL in this respect, where one can either have high bandwidth or long range, but not both simultaneously. The other feature to consider with WiMAX is that available bandwidth is shared between users in a given radio sector, so if there are many active users in a single sector, each will get reduced bandwidth.

1. Development of IEEE 802.11

The Physical layer (PHY) and medium access control (MAC) layer were mainly targeted by the IEEE 802 project. When the idea of wireless local area network (WLAN) was first conceived, it was just thought of another PHY of one of the available standards. The first candidate which was considered for this was IEEE's most prominent standard 802.3.

However later findings showed that the radio medium behaved quite different than the conventional well behaved wire. As there was attenuation even over short distances, collisions could not be detected. Hence, 802.3's carrier sense multiple access with collision detection (CSMA/CD) could not be applied.

The next candidate standard considered was

802.4. At that point of time, its coordinated medium access i.e. the token bus concept was believed to be superior to 802.3's contention-based scheme. Hence, WLAN began as 802.4L. Later in 1990 it became obvious that token handling in radio networks was rather difficult. The standardization body realized the need of a wireless communication standard that would have itsown very unique MAC. Finally, on March 21, 1991, the project 802.11 was approved (fig. 1).

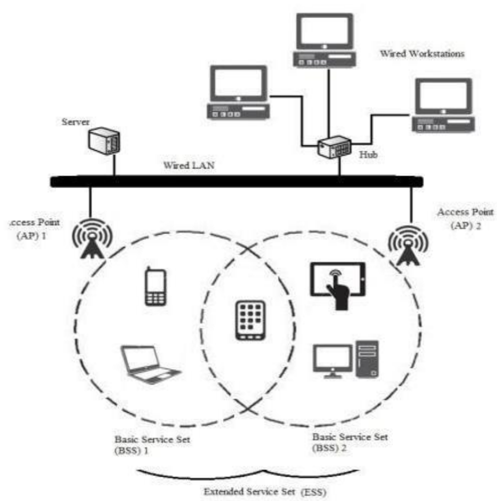


Figure 1 WLAN Network Architecture

2. IEEE 802.11 family

The most widely deployed 802.11 standard has a lot of extension and many more are currently under development. First introduced in 1999,the IEEE 802.11 standards was primarily developed keeping in mind the home and the office environment for wireless local area connectivity. The Initial standards gave a maximum data rate of 2Mbps per AP which increased to 11 Mbps per AP with the deployment of IEEE 802.11b.Newer extensions like IEEE 802.11g and IEEE 802.11a provided maximum data rate of 54Mbps per AP using various methods to boost up the maximum data rates. WLAN devices based on IEEE 802.11g currently offer data rate 100-125Mbps. Similarly, a relatively newer IEEE 802.11n gives a maximum data rate of about 540Mbps .Furthermore, in addition to these, several other standards were deployed which solved many QoS and security issues related with the earlier standards. Additional mechanisms were introduced to remedy QoS support and security problems in IEEE 802.11e [12] and IEEE 802.11i.The IEEE 802.11n standard which we earlier talked about also introduced MAC enhancements to overcome MAC layer limitations in the current standards. The IEEE 802.11s standard added mesh topology support to the IEEE 802.11. The IEEE 802.11u improved internetworking with external non-802.11 networks. The IEEE 802.11w was an added onto 802.11i covering management frame security.

The IEEE 802.11ad standard adds a "fast session transfer" feature, enabling the wireless devices to seamlessly make transition between the legacy 2.4 GHz and 5 GHz bands and the 60 GHz frequency band. The IEEE 802.11ac standard, still under development is expected to provide a multi-station WLAN throughput of at least 1 Gbps and a single link throughput of at least 500 Mbps.

The IEEE 802.11a extension employs a number of channels ranging from 36-161 depending on the frequency band (5.15-5.825 GHz) although it works with a fixed channel centre frequency of 5 GHz. There are 12 non overlapping channels in the frequency band for theIEEE standard in the U.S. and 19 non-overlapping channels in Europe. In contrast, there areonly 3 out 14 non-overlapping in case of 802.11b [2]. IEEE 802.11n uses overlapping channels with channel bandwidth 20 and 40MHz [19]. The 20MHz channel bandwidth is incorporated in every 802.11n device, the 40MHz channel is optional.

Peer to Peer (P2P) WLAN links can be established with the help of directional antennas for a few km ranges. A typical WLAN Access Point (AP) uses omnidirectional antennas with a range of 30-50m (indoors) and 100m (outdoors).This range is greatly affected by the obstacles between the AP and the STA.IEEE 802.11a suffer from increased range and attenuation compared to IEEE 802.11b/g networks, because it operates on the higher frequency range of 5MHz.Use of sectored antennas instead of omnidirectional antennas increases the aggregate WLAN data rate in a given area to 2-3 times .

Medium Access Control (MAC) Layer

IEEE 802.11 uses a contention based scheme known as Distributed Coordinated Function (DCF). In this method the STA linked with the AP scans the air interface for channel availability. If the interface is idle, the STA sends it data to the destination through the AP. If however the air interface is busy or more than one STA tries to access the same AP; a collision occurs. The IEEE 802.11 uses a Carrier Sense Multiple Access/Collision Avoidance (CSMA/CA) to avoid the collisions. IEEE 802.11 uses another MAC technique known as Point Coordination Function (PCF) [18]. This mechanism is divided in to two parts. In the first part, the AP scans all its STA in a round robin fashion and checks to see if any of the STAs has any packets to send. If any of the STAs is not polled during the current period, it will be queued up for polling during the next polling period. The scanned part uses the contention based scheme and it same as DCF.

Moreover, due to polling mechanism in PCF the aggregate throughput of an IEEE 802.11 network decreases. DCF is the default MAC technique used in the IEEE 802.11 standard. While the standard includes both the MAC techniques, PCF is included in the Wi-fi alliance standard and hence not quite as popular as DCF [17]. In both the MAC techniques an automatic response request mechanism is used in this method. Any device in the network receiving data will sendan acknowledgement signal (ACK) back to the sender. In case the receiver receives a corrupt data packet, it issues a NAK (Negative Acknowledgement) and the sender resends the data packet. There is a round trip delay as the sender has to wait for the ACK to transmit the next data packet in the queue.

Request to Send/Clear to Send (RTS/CTS)

In the contention based scheme called DCF if more than two STAs simultaneously try to access the air interface, a collision occurs. To avoid such collision CSMA/CA may result in incorrect medium information. This is called Hidden Node Problem in which collision in the some part of the network cannot be detected [15]. If any two STAs cannot directly communicate, the AP invokes a RTS/CTS mechanism. For each transmission, the source STA issues a RTS message. The destination STA replies to this by sending a CTS message. Upon receiving the CTS message, the source STA starts its data transmission. The medium is assumed to be in use given in the message when they receive RTS and/or CTS message. In PCF using RTS/CTS reduces the network throughput.

Authentication & Encryption

Security is also handled in the MAC layer. To avoid unauthorized access from other STAs, several encryption methods have been used. One of earlier encryption mechanism was Wired Equivalent Piracy (WEP). But the encryption method had security vulnerabilities and the Wi-fi Alliance developed another encryption technique named Wi-Fi protection Access (WPA).

The IEEE 802.11i standard incorporated an enhanced version of WPA (WPA2) [20]. IEEE 802.11i also addressed security issues associated with authentication methods like open standard and shared key authentication and incorporated IEEE 802.1X authentication method which is now used in all the later versions of IEEE 802 family standards. In this method, users can authenticate their identities by a RADIUS or diameter server.

Management Frame

The current 802.11 standards define "frame" types for use in management and control of wireless links. The TGw implemented the IEEE 802.11w standard to implement the Protected Management Frames. The TGw is still working on improving the IEEE 802.11 MAC layer. Security can be enhanced by providing data confidentiality of management frames. These extensions will have interactions with IEEE 802.11r as well as IEEE 802.11u

Advantages of wireless local area network (WLAN) :

- It’s a reliable sort of communication.
- As WLAN reduces physical wires so it’s a versatile way of communication.
- WLAN also reduces the value of ownership.
- It’s easier to feature or remove workstation.
- It provides high rate thanks to small area coverage.
- You’ll also move workstation while maintaining the connectivity.
- For propagation, the sunshine of sight isn’t required.
- The direction of connectivity are often anywhere i.e. you’ll connect devices in any direction unless it’s within the range of access point.
- Easy installation and you would like don’t need extra cables for installation.
- WLAN are often useful in disasters situation e.g. earthquake and fire. WIreless network can connect people in any disaster
- it’s economical due to the tiny area access.
- The amount of power it requires is more as it uses transmitter; therefore, the battery life of laptops can be affected

Disadvantages of wireless local area network (WLAN) :

- WLAN requires license.
- it’s a limited area to hide.
- The Government agenciescan control the flow of signals of WLAN and can also limit it if required. this will affect data transfer from connected devices to the web.
- If the amount of connected devices increases then data transfer rate decreases.
- WLAN uses frequency which may interfere with other devices which use frequency.
- If there’s rain or thunder then communication may interfere.
- Due to Low security as attackers can get access to the transmitted data.
- Signals could also be suffering from the environment as compared to using fiber optics.
- The radiation of WLAN are often harmful to the environment
- Wlan is expensive than wires and hubs as it access points.
- Signals can get from nearest signals by access points.
- it’s required to vary the network card and access point when standard changes.
- LAN cable remains required which acts because the backbone of the WLAN
- Low data transfer rate than wired connection because WLAN uses frequency.

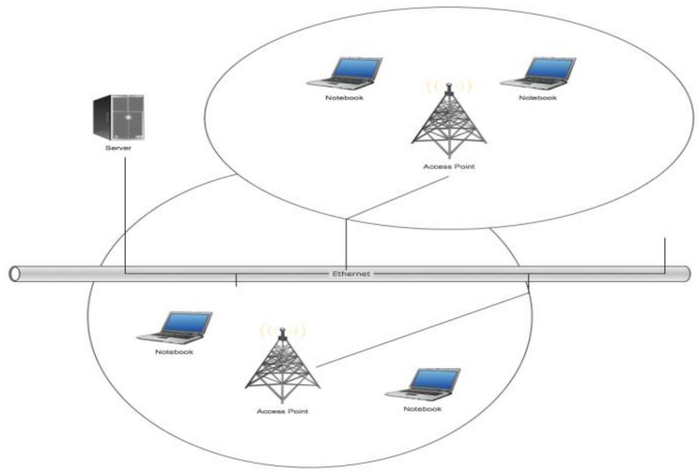
- Chances of errors are high.
- Communication isn't secure and may be accessed by unauthorized users

Wireless Network Topology

Wireless network topology — logical topology.

Wireless network topology shows how the computers connect each other when there is no physical connection. The computers communicate each using the wireless devices.

This sample was created in ConceptDraw DIAGRAM diagramming and vector drawing software using the Wireless Networks solution from Computer and Networks area of ConceptDraw Solution Park.



Example 1. Wireless Network Topology

This sample shows the Wireless network topology.

The infrastructure wireless network topology is a hub and spoke topology. It is also named "one to many" topology. There is a single central wireless access point (WAP) in the infrastructure wireless network topology.

The ad hoc wireless network topology is a "many to many" topology. There is no central access point, every computer of the network communicates directly with other computer in the ad hoc wireless network topology.

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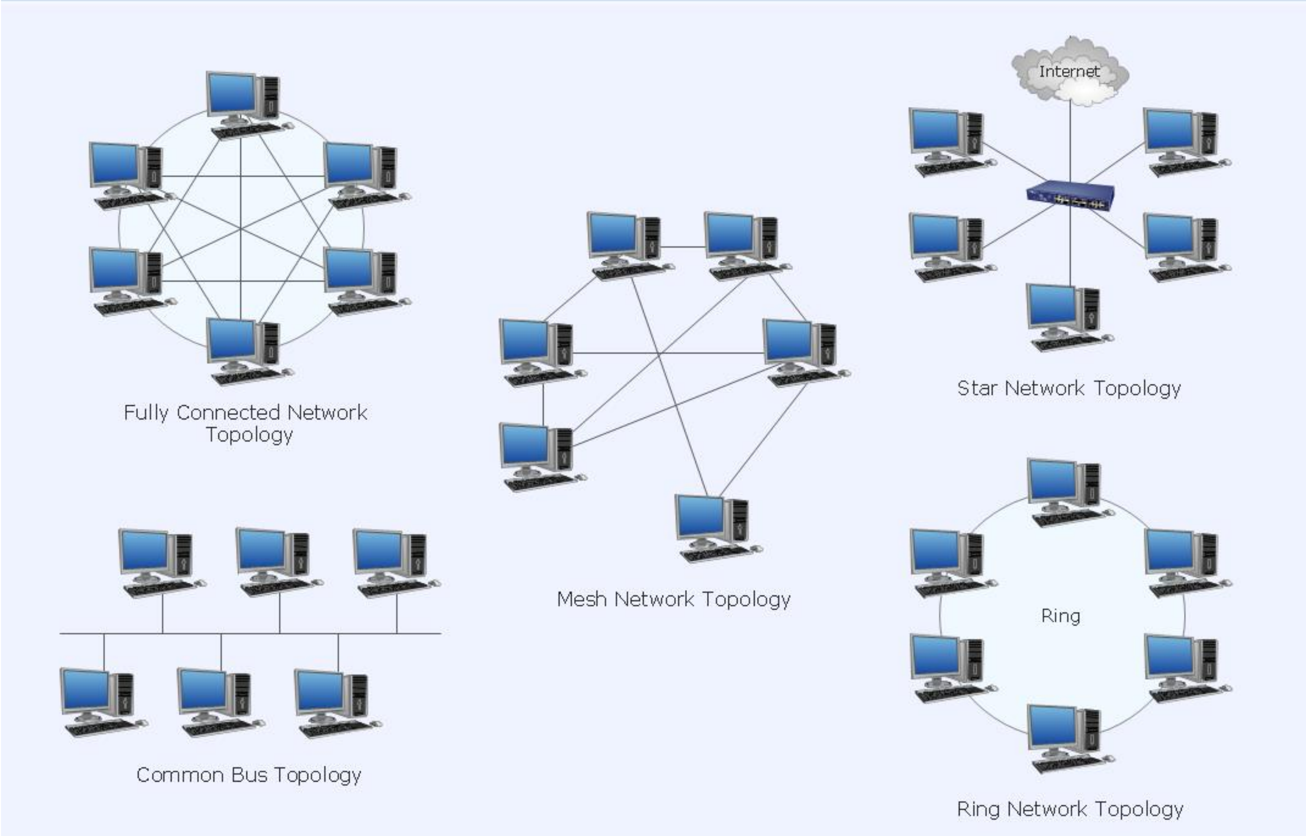
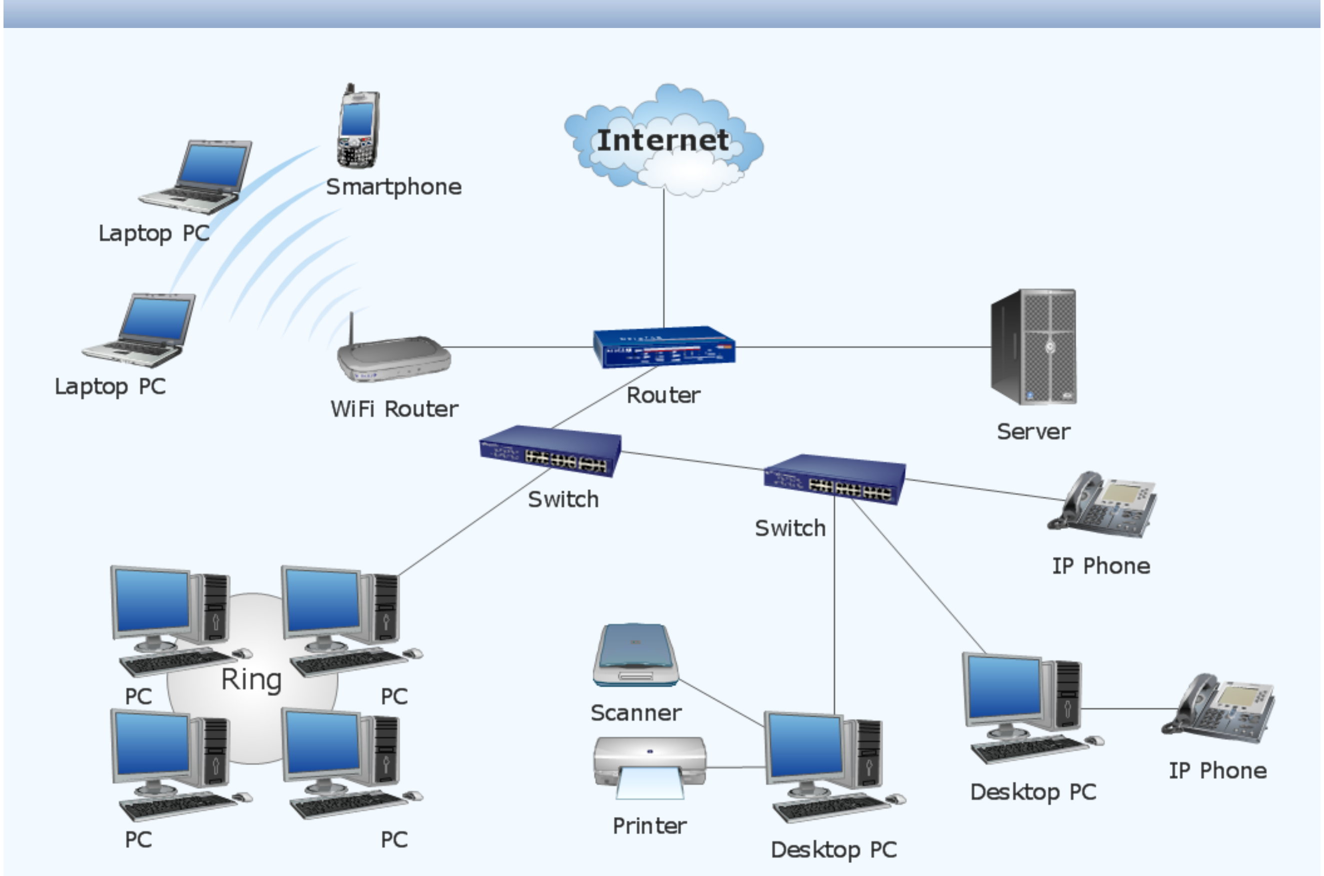
Network Topologies

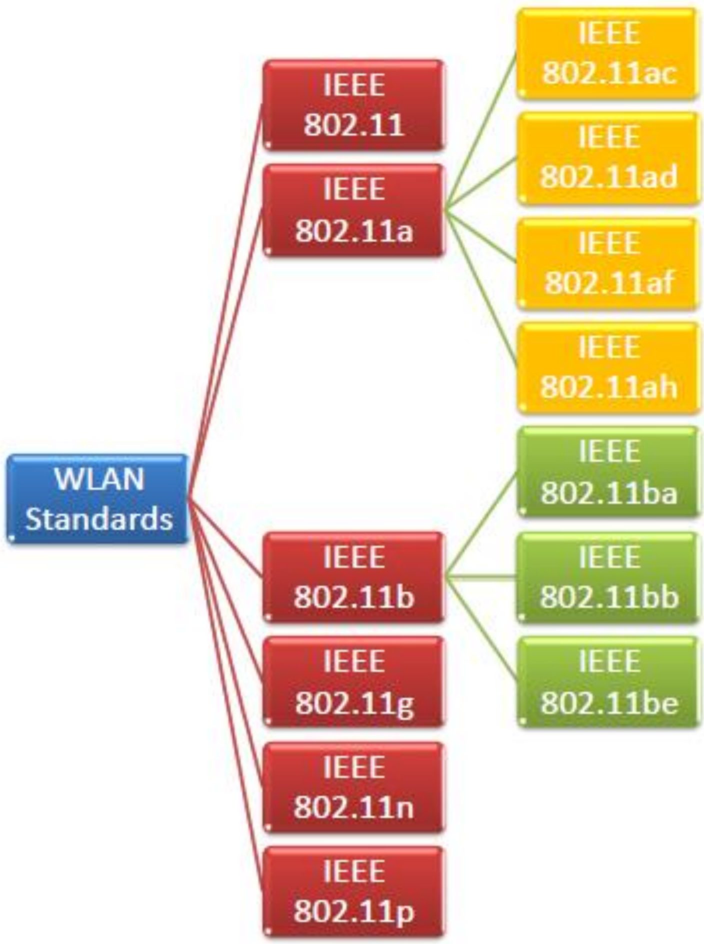




wireless LANs IEEE 802.11 standard

wireless LANs IEEE 802.11 standard, popularly known as WiFi, lays down the architecture and specifications of wireless LANs (WLANs). WiFi or WLAN uses high frequency radio waves for connecting the nodes.

There are several standards of IEEE 802.11 WLANs. The prominent among them are 802.11, 802.11a, 802.11b, 802.11g, 802.11n and 802.11p. All the standards use carrier-sense multiple access with collision avoidance (CSMA/CA). Also, they have support for both centralised base station based as well as ad hoc networks.



IEEE 802.11

IEEE 802.11 was the original version released in 1997. It provided 1 Mbps or 2 Mbps data rate in the 2.4 GHz band and used either frequency-hopping spread spectrum (FHSS) or direct-sequence spread spectrum (DSSS). It is obsolete now.

IEEE 802.11a

802.11a was published in 1999 as a modification to 802.11, with orthogonal frequency division multiplexing (OFDM) based air interface in physical layer instead of FHSS or DSSS of 802.11. It provides a maximum data rate of 54 Mbps operating in the 5 GHz band. Besides it provides error correcting code. As 2.4 GHz band is crowded, relatively sparsely used 5 GHz imparts additional advantage to 802.11a.

Further amendments to 802.11a are 802.11ac, 802.11ad, 802.11af, 802.11ah, 802.11ai, 802.11aj etc.

IEEE 802.11b

802.11b is a direct extension of the original 802.11 standard that appeared in early 2000. It uses the same modulation technique as 802.11, i.e. DSSS and operates in the 2.4 GHz band. It has a higher data rate of 11

Mbps as compared to 2 Mbps of 802.11, due to which it was rapidly adopted in wireless LANs. However, since 2.4 GHz band is pretty crowded, 802.11b devices faces interference from other devices.

Further amendments to 802.11b are 802.11ba, 802.11bb, 802.11bc, 802.11bd and 802.11be.

IEEE 802.11g

802.11g was indorsed in 2003. It operates in the 2.4 GHz band (as in 802.11b) and provides a average throughput of 22 Mbps. It uses OFDM technique (as in 802.11a). It is fully backward compatible with 802.11b. 802.11g devices also faces interference from other devices operating in 2.4 GHz band.

IEEE 802.11n

802.11n was approved and published in 2009 that operates on both the 2.4 GHz and the 5 GHz bands. It has variable data rate ranging from 54 Mbps to 600 Mbps. It provides a marked improvement over previous standards 802.11 by incorporating multiple-input multiple-output antennas (MIMO antennas).

IEEE 802.11p

802.11 is an amendment for including wireless access in vehicular environments (WAVE) to support Intelligent Transportation Systems (ITS). They include network communications between vehicles moving at high speed and the environment. They have a data rate of 27 Mbps and operate in 5.9 GHz band.

IEEE 802.11 standard, popularly known as WiFi, lays down the architecture and specifications of wireless LANs (WLANs). WiFi or WLAN uses high frequency radio waves instead of cables for connecting the devices in LAN. Users connected by WLANs can move around within the area of network coverage.

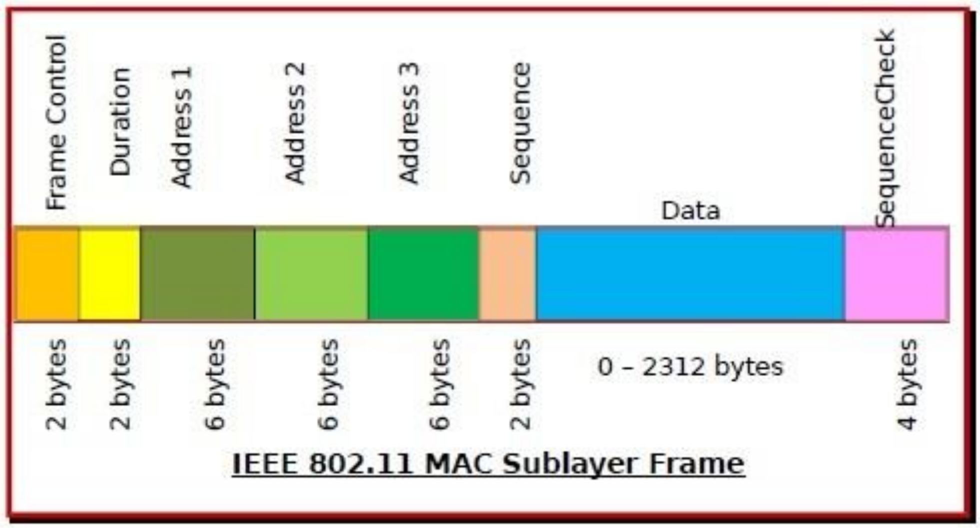
The 802.11 MAC sublayer provides an abstraction of the physical layer to the logical link control sublayer and upper layers of the OSI network. It is responsible for encapsulating frames and describing frame formats.

IEEE 802.11 MEDIUM ACCESS CONTROL

MAC Sublayer frame of IEEE 802.11

The main fields of a frame of wireless LANs as laid down by IEEE 802.11 are −

- **Frame Control −** It is a 2 bytes starting field composed of 11 subfields. It contains control information of the frame.
- **Duration −** It is a 2-byte field that specifies the time period for which the frame and its acknowledgement occupy the channel.
- **Address fields −** There are three 6-byte address fields containing addresses of source, immediate destination and final endpoint respectively.
- **Sequence −** It a 2 bytes field that stores the frame numbers.
- **Data −** This is a variable sized field carries the data from the upper layers. The maximum size of data field is 2312 bytes.
- **Check Sequence −** It is a 4-byte field containing error detection information.



Avoidance of Collisions by 802.11 MAC Sublayer

In wireless systems, the method of collision detection does not work. It uses a protocol called carrier sense multiple access with collision avoidance (CSMA/CA).

The method of CSMA/CA is −

- When a frame is ready, the transmitting station checks whether the channel is idle or busy.
- If the channel is busy, the station waits until the channel becomes idle.
- If the channel is idle, the station waits for an Inter-frame gap (IFG) amount of time and then sends the frame.
- After sending the frame, it sets a timer.
- The station then waits for acknowledgement from the receiver. If it receives the acknowledgement before expiry of timer, it marks a successful transmission.
- Otherwise, it waits for a back-off time period and restarts the algorithm.

Co-ordination Functions in 802.11 MAC Sublayer

IEEE 802.11 MAC Sublayer uses two co-ordination functions for collision avoidance before transmission −

- **Distributed Coordination Function (DCF)** −
  - It is a mandatory function used in CSMA/CA.
  - It is used in distributed contention-based channel access.
  - It is deployed in both Infrastructure BSS (basic service set) as well as Independent BSS.
- **Point Coordination Function (PCF)** −
  - It is an optional function used by 802.11 MAC Sublayer.
  - It is used in centralized contention-free channel access.
  - It is deployed in Infrastructure BSS only.

IEEE 802.16

- **IEEE 802.16** is a series of wireless broadband standards written by the Institute of Electrical and Electronics Engineers (IEEE). The IEEE Standards Board established a working group in 1999 to develop standards for broadband for wireless metropolitan area networks. The Workgroup is a unit of the IEEE 802 local area network and metropolitan area network standards committee.

- Although the 802.16 family of standards is officially called WirelessMAN in IEEE, it has been commercialized under the name "WiMAX" (from "Worldwide Interoperability for Microwave Access") by the WiMAX Forum industry alliance. The Forum promotes and certifies compatibility and interoperability of products based on the IEEE 802.16 standards.
- The 802.16e-2005 amendment version was announced as being deployed around the world in 2009.[1] The version **IEEE 802.16-2009** was amended by IEEE 802.16j-2009.

Standard	Description	Status
802.16	Fixed Broadband Wireless Access (10–66 GHz)	Superseded
802.16.2	Recommended practice for coexistence	Superseded
802.16c	System profiles for 10–66 GHz	Superseded
802.16a	Physical layer and MAC definitions for 2–10 GHz	Superseded
P802.16b	License-exempt frequencies (Project withdrawn)	Withdrawn
P802.16d	Maintenance and System profiles for 2–11 GHz (Project merged into 802.16-2004)	Merged
802.16	Air Interface for Fixed Broadband Wireless Access System (rollup of 802.16–2001, 802.16a, 802.16c and P802.16d)	Superseded
P802.16.2a	Coexistence with 2–11 GHz and 23.5–43.5 GHz (Project merged into 802.16.2-2004)	Merged
802.16.2	IEEE Recommended Practice for **Local and metropolitan area networks Coexistence of Fixed Broadband Wireless Access Systems** (Maintenance and rollup of 802.16.2–2001 and P802.16.2a) Released on 2004-March-17.	Current
802.16f	Management Information Base (MIB) for 802.16-2004	Superseded
802.16-2004/Cor 1–2005	Corrections for fixed operations (co-published with 802.16e-2005)	Superseded
802.16e	Mobile Broadband Wireless Access System	Superseded

802.16k	IEEE Standard for **Local and Metropolitan Area Networks: Media Access Control (MAC) Bridges** **Amendment 2: Bridging of IEEE 802.16** (An amendment to IEEE 802.1D) Released on 2007-August-14.	Current
802.16g	Management Plane Procedures and Services	Superseded
P802.16i	Mobile Management Information Base (Project merged into 802.16-2009)	Merged
802.16-2009	Air Interface for Fixed and Mobile Broadband Wireless Access System (rollup of 802.16–2004, 802.16-2004/Cor 1, 802.16e, 802.16f, 802.16g and P802.16i)	Superseded
802.16j	Multihop relay	Superseded
802.16h	Improved Coexistence Mechanisms for License-Exempt Operation	Superseded
802.16m	Advanced Air Interface with data rates of 100 Mbit/s mobile and 1 Gbit/s fixed. Also known as *Mobile WiMAX Release 2* or *WirelessMAN-Advanced*. Aiming at fulfilling the ITU-R IMT-Advanced requirements on 4G systems.	Superseded[2]
802.16-2012	IEEE Standard for **Air Interface for Broadband Wireless Access Systems** It is a rollup of 802.16h, 802.16j and Std 802.16m (but excluding the WirelessMAN-Advanced radio interface, which was moved to IEEE Std 802.16.1). Released on 2012-August-17.	Superseded
802.16.1	IEEE Standard for **WirelessMAN-Advanced Air Interface for Broadband Wireless Access Systems** Released on 2012-September-07.	Current
802.16p	IEEE Standard for **Air Interface for Broadband Wireless Access Systems** **Amendment 1: Enhancements to Support Machine-to-Machine Applications** Released on 2012-October-08.	Current
802.16.1b	IEEE Standard for **WirelessMAN-Advanced Air Interface for Broadband Wireless Access Systems** **Amendment 1: Enhancements to Support Machine-to-Machine Applications** Released on 2012-October-10.	Current

802.16n	IEEE Standard for **Air Interface for Broadband Wireless Access Systems Amendment 2: Higher Reliability Networks** Approved on 2013-March-06.	Current
802.16.1a	IEEE Standard for **WirelessMAN-Advanced Air Interface for Broadband Wireless Access Systems** **Amendment 2: Higher Reliability Networks** Approved on 2013-March-06.	Current
802.16-2017	IEEE Standard for **Air Interface for Broadband Wireless Access Systems** It is a rollup of 802.16p, 802.16n, 802.16q and Std 802.16s Released on 2017-September.	Current

WIRELESS Personal Area Networks

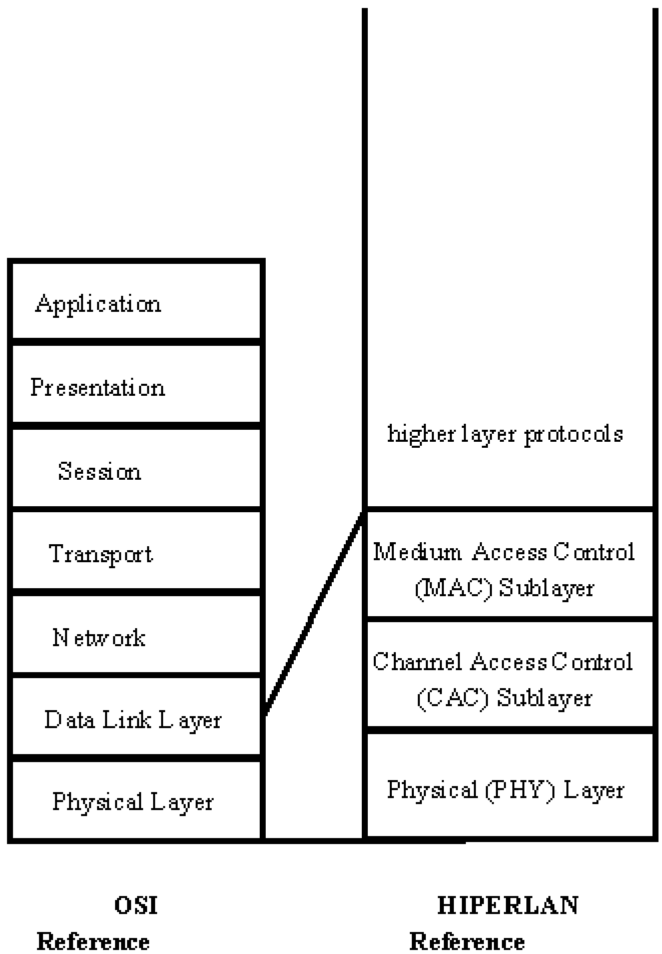
A personal area network is a network concerned with the exchange of information in the vicinity of a person. Typically, these systems are wireless and involve the transmission of data between devices such as smartphones, personal computers, tablet computers, etc. The purpose of such a network is usually to allow either transmission of data or information between such devices or to server as the network that allows further up link to the Internet. Developments in the area of Personal Area Networks (PANs) are largely overseen by the IEEE 802.15 working group
Personal Area Network (PAN): It is an interconnection of personal technology devices to communicate over a short distance, which is less than 33 feet or 10 meters or within the range of an individual person, typically using some form of wireless technologies. Some of these technologies are:

- Bluetooth technology: The idea behind Bluetooth is to embed a low cost transceiver chip in each device, making it possible for wireless devices to be totally synchronized without the user having to initiate any operation. The chips would communicate over a previously unused radio frequency at up to 2 Mbps. The overall goal of Bluetooth might be stated as enabling ubiquitous connectivity between personal technology devices without the use of cabling as written in Mckeown (2003a).
- High rate W-PANs: As per standard IEEE 802.15 TG3, launched in 2003, these technologies use higher power devices (8 dBm) than regular Bluetooth equipment (0 dBm) to transmit data at a rate of up to 55 Mbps and over a range of up to 55 m Ailisto et al (2003).
- Low power W-PANs: As per standard IEEE 802.15 TG4, these technologies are particularly useful for handheld devices since energy consumption for data transmission purposes, and costs, are extremely low. The range of operation of up to 75 m is higher than current Bluetooth applications, but the data transfer rate is low (250 Kbps) Ailisto et al (2003).

HIPER LAN

HIPERLAN is a European family of standards on digital high speed wireless communication in the 5.15-5.3 GHz and the 17.1-17.3 GHz spectrum developed by ETSI. The committee responsible for HIPERLAN is RES-10 which has been working on the standard since November 1991.

The standard serves to ensure the possible interoperability of different manufacturers' wireless communications equipment that operate in this spectrum. The HIPERLAN standard only describes a common air interface including the physical layer for wireless communications equipment, while leaving decisions on higher level configurations and functions open to the equipment manufacturers.



The choice of frequencies allocated to HIPERLAN was part of the 5-5.30 GHz band being allocated globally to aviation purposes. The Aviation industry only used the 5-5.15GHz frequency, thus making the 5.15-5.30 frequency band accessible to HIPERLAN standards.

HIPERLAN is designed to work without any infrastructure. Two stations may exchange data directly, without any interaction from a wired (or radio-based) infrastructure. The simplest HIPERLAN thus consists of two stations. Further, if two HIPERLAN stations are not in radio contact with each other, they may use a third station (i.e. the third station must relay messages between the two communicating stations).

Products compliant to the HIPERLAN 5 GHz standard shall be possible to implement on a PCMCIA Type III card. Thus the standard will enable users to truly take computing power on the road.

The HIPERLAN standard has been developed at the same time as the development of the SUPERnet standard in the United States.

HIPERLAN requirements

- Short range - 50m

- Low mobility - 1.4m/s
- Networks with and without infrastructure
- Support isochronous traffic
- audio 32kbps, 10ns latency
- video 2Mbps, 100ns latency
- Support asynchronous traffic
- data 10Mbps, immediate access

Quality of service

Performance is one of the most important factors when dealing with wireless LANs. In contrast to other radio-based systems, data traffic on a local area network has a randomized bursty nature, which may cause serious problems with respect to throughput.

Many factors have to be taken into consideration, when quality of service is to be measured. Among these are:

- The topography of the landscape in general
- Elevations in the landscape that might cause shadows, where connectivity is unstable or impossible.
- Environments with many signal-reflection surfaces
- Environments with many signal-absorbing surfaces
- Quality of the wireless equipment
- Placement of the wireless equipment
- Number of stations
- Proximity to installations that generate electronic noise
- and many more

The sheer number of factors to take into consideration means, that the physical environment will always be a factor in trying to asses the usefulness of using a wireless technology like HIPERLAN.

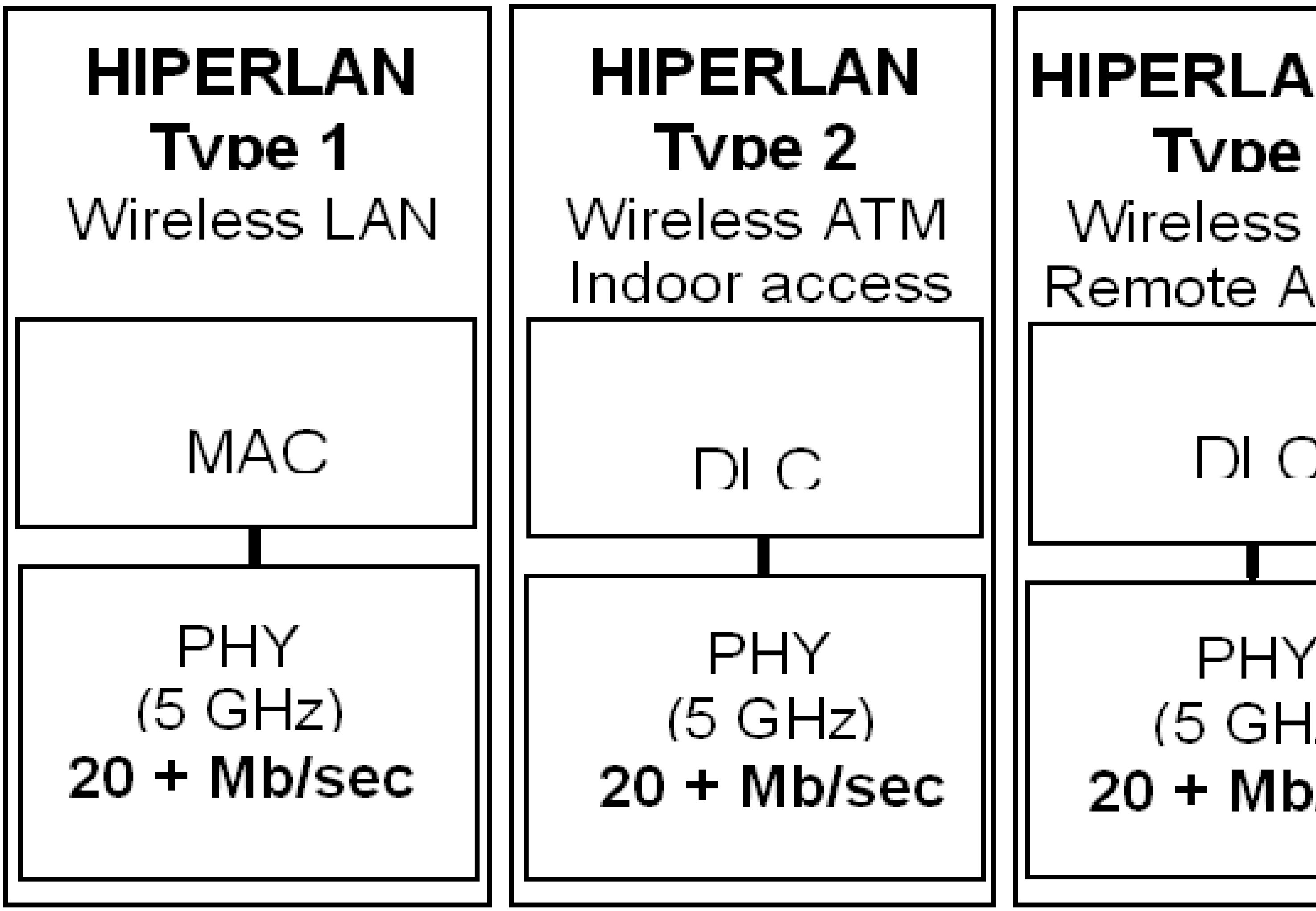
Simulations show that the HIPERLAN MAC can simultaneously support

- 25 audio links at 32kbit/s, 10ms delivery
- 25 audio links at 16kbit/s, 20ms delivery
- 1 video link at 2Mbit/s, 100ms delivery
- Asynch file transfer at 13.4Mbit/s

New HIPERLAN standards ahead

A second set of standards have been constructed for a new version of HIPERLAN - HIPERLAN2. The idea of HIPERLAN2 is to be compatible with ATM.

There is also undergoing work to establish global sharing rules. The WINForum for NII/SUPERNET in the US aim to support HIPERLAN 1 and HIPERLAN 2. This effort involves interaction between ETSI RES10, WINForum, ATM Forum.



Wireless Local Loop (WLL)

Wireless Local Loop (WLL) is a generic word for an access system that connects users to the local telephone company's switch via wireless links rather than traditional copper cables. This system, also known as *fixed wireless access* (FWA) or *fixed radio*, provides telephone, facsimile, and data services to business and residential subscribers using analog or digital radio technology.

- WLL systems enable the rapid deployment of basic phone service in areas where geography or telecommunications development makes traditional wireline service prohibitively expensive.
- WLL systems are easy to integrate with a modified public telephone network (PSTN), and they can usually be installed within a month of acquiring equipment, much faster than traditional wiring, which can take months to set up and years to increase the capacity to meet the growing demand for communication services.
- Analog systems for medium- to low-density and rural applications are among WLL's offerings.
- There are WLL systems based on Code Division Multiple Access for high-density, high-growth urban and suburban settings (CDMA). Telecommunications systems such as TDMA (Time Division Multiple Access) and GSM (Global System for Mobile) are also available.

- Digital WLL systems can offer higher-speed fax and data services in addition to providing better speech quality than analog systems.
- Existing operations support systems (OSS) and transmission and distribution systems are also compatible with WLL technology.

What are the most common wireless access methods?

Frequency division multiple access (FDMA), time division multiple access (TDMA), and code division multiple access (CDMA) are all used to accomplish WLL (CDMA).

CDMA is the one that is utilized in India. This is a full-fledged mobile phone system. In fact, in nations like the United States and Korea, it is the most widely used technology for mobile phone services.

Benefits of Using WLL

WLL systems are scalable, allowing operators to continue to use their existing infrastructure as the system grows. WLL customers use a radio unit connected to the PSTN via a local base station to obtain phone service.

A transceiver, power source, and antenna make up the radio unit. It runs on AC or DC power, can be mounted indoors or out, and usually comes with a battery backup for when the power goes out. The radio unit connects to the premise's cabling on the customer side, allowing customers to utilize their current phones, modems, fax machines, and answering machines.

Following are some of the benefits of using WLL −

- It eliminates the need to build a network connection for the first or final mile.
- Since no copper cables are used, the cost is low.
- Wireless communication is much more secure because of digital encryption technology.
- It is very scalable since it does not require the installation of more wires to scale.