INSTITUTE OF AERONAUTICAL ENGINEERING

(Autonomous)

Dundigal, Hyderabad -500 043

ELECTRONICS AND COMMUNICATION ENGINEERING

COURSE LECTURE NOTES

Course Name	**WIRELESS COMMUNICATION AND NETWORK**
Course Code	AEC524
Programme	B.Tech
Semester	VI
Course Coordinator	Mr. A Karthik, Assistant Professor
Course Faculty	Mr. A Karthik, Assistant Professor
Lecture Numbers	1-74
Topic Covered	All

COURSE OBJECTIVES (COs):

The course should enable the students to:	
I	Understand fundamental treatment of wireless communications and the Cellular Concept-System Design, Fundamental concepts like frequency reuse, Radio Wave Propagation Basic Propagation Mechanisms and Diffraction Models.
II	Understand the concept of frequency reuse and be able to apply it in the design of mobile cellular system.
III	Understand the various modulation schemes and multiple access techniques that are used in wireless communications.
IV	Remember the analytical perspective on the design and analysis of the traditional and emerging wireless networks and discuss the nature of and solution methods to the fundamental problems in wireless networking.

COURSE LEARNING OUTCOMES (CLOs):

Students, who complete the course, will have demonstrated the ability to do the following:

CLO Code	CLO’s	At the end of the course, the student will have the ability to:
AEC524.01	CLO 1	Understand the principles and fundamentals of wireless communications.
AEC524.02	CLO 2	Demonstrate cellular system design concepts in wireless mobile communication networks.
AEC524.03	CLO 3	Understand the fundamental Radio Wave Propagation Mechanisms.
AEC524.04	CLO 4	Analyze perspective on Fundamentals of Equalization and Mobile Radio Propagation Multipath Measurements.

CLO Code	CLO's	At the end of the course, the student will have the ability to:
AEC524.05	CLO 5	Analyze various multiple access schemes and techniques used in wireless communication.
AEC524.06	CLO 6	Discuss the Parameters of Mobile Multipath Channels and Types of Small-Scale Fading- Fading effects.
AEC524.07	CLO 7	Examine the perspective on Fundamentals of Equalization, Linear Equalizers, Non-linear Equalization.
AEC524.08	CLO 8	Study and understand the Diversity Techniques and RAKE Receiver in Radio Propagation.
AEC524.09	CLO 9	Demonstrate wireless local area networks and their specifications in communication system.
AEC524.10	CLO 10	Understand the analytical perspective on the design and analysis of the traditional and emerging wireless networks
AEC524.11	CLO 11	Discuss the nature of and solution methods to the fundamental problems in wireless networking.
AEC524.12	CLO 12	Understand the architecture of the various wireless wide area networks such as GSM, IS-95, GPRS and SMS.
AEC524.13	CLO 13	Understand the operation of the various wireless wide area networks such as GSM, IS-95, GPRS and SMS.
AEC524.14	CLO 14	Understand the existing and emerging wireless standards in wireless wide area networks
AEC524.15	CLO 15	Examine the emerging techniques OFDM and its importance in the wireless communications.

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) mode; 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, indoor 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	
UNIT-III	CELLULAR SYSTEM DESIGN FUNDAMENTALS
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, comparison of IEEE 802.11 a,b,g and n standards, IEEE 802.16 and its enhancements, wireless PANs, Hipper LAN, WLL.

Text Books:

1. Theodore .S. Rapport, ―Wireless Communications‖, Pearson Education, 2nd Edition, 2010.
2. Upen Dalal, “Wireless communication”, oxford University press.
3. Kaveh Pahlvan, Prashant Krishnamurthy, “Principle of wireless networks”, A United Approach‖, Pearson Education, 2004.
4. Andrea Goldsmith, “Wireless Communications”, Cambridge University Press, 2005.

Reference Books:

1. P.Nicopolitidis, M.S. Obaidat, G.I.Papadimitria, A.S. Pomportsis,”Wireless Networks” John Wiley & sons, 1st Edition, 2003.
2. Vijay K Garg,”Wireless Communications and Networks”, Morgan Kaufmann Publishers an Imprint of Elsevier, USA 2009 (Indian Reprint).
3. Mark Ciampa Jorge Olenewa, “wireless communication and Networking”, IE, 2009.
4. X.Wang, H.V.Poor ,Wireless communication system, Pearson Education, 2004.
5. Jochen Schiller,”Mobile Communication”, Pearson Education, 2nd Edition, 2003.

UNIT-I

The Cellular Concept System Design Fundamentals

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Introduction

The cellular concept was a major breakthrough in solving the problem of spectral congestion and user capacity. It offered very high capacity in a limited spectrum allocation without any major technological changes. The cellular concept is a system-level idea which calls for replace- in a single, high power transmitter (large cell) with many low power transmitters (small cells), each providing coverage to only a small portion of the service area. Each base station is allocated a portion of the total number of channels available to the entire system, and nearby base stations is assigned different groups of channels so that all the available channels are assigned to a relatively small number of neigh boring base stations. Neigh boring base stations are assigned different groups of channels so that the interference

Frequency Reuse

Cellular radio systems rely on an intelligent allocation and reuse of channels throughout a coverage region. Each cellular base station is allocated a group of radio channels to be used within a small geographic area called a *cell*. Base stations in adjacent cells are assigned channel groups which contain completely different channels than neighboring cells. The base station antennas are designed to achieve the desired coverage within the particular cell. By limiting the coverage area to within the boundaries of a cell, the same group of channels may be used to cover different cells that are separated from one another by distances large enough to keep interference levels within tolerable limits. The design process of selecting and allocating channel groups for all of the cellular base stations within a system is called frequency reuse or frequency planning **frequency Reuse**, the actual cellular footprint is determined by the contour in which a given transmitter serves the mobiles successfully.

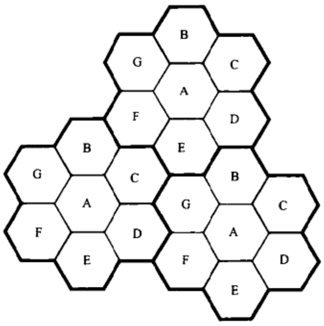
$$S = Kn \tag{3.1}$$

The N cells which collectively use the complete set of available frequencies is called a cluster. If a cluster is replicated M times within the system, the total number of duplex channels, C, can be used as a measure of capacity and is given by

$$C = MkN = MS \tag{3.2}$$

Due to the fact that the hexagonal geometry of Figure 3.1 has exactly six equidistant neighbours and that the lines joining the centres of any cell and each of its neighbours are separated by multiples of 60 degrees, there are only certain cluster sizes and cell layouts which are possible

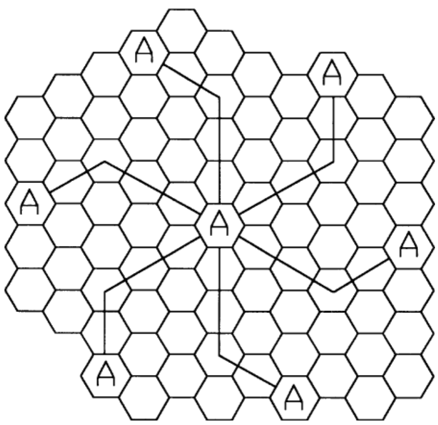
[Mac79].



In order to tessellate to connect without gaps between adjacent cells—the geometry of hexagons is such that the number of cells per cluster, N, can only have values which satisfy Equation (3.3)

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

Method of locating co-channel cells in a cellular system. In this example, $N = 19$ (i.e., $i = 3$, $j = 2$).



(Adapted from [Oet83] © IEEE.)

Handoff Strategies

When a mobile moves into a different cell while a conversation is in progress, the MSC automatically transfers the call to a new channel belonging to the new base station. This handoff operation not only involves identifying a new base station, but also requires that the voice and control signals be allocated to channels associated with the new base station. Processing handoffs is an important task in any cellular radio system.

Many handoff strategies prioritize handoff requests over call initiation requests when allocating unused channels in a cell site. Handoffs must be performed successfully and as infrequently as possible, and be imperceptible to the users. In order to meet these requirements, system designers must specify an optimum signal level at which to initiate a handoff. Once a particular signal level is specified as the minimum usable signal for acceptable voice quality at the base station receiver (normally taken as between –90 dBm and –100 dBm), a slightly stronger signal level is used as a threshold at which a handoff is made. This margin, given by

A = $P_{r\ handoff} - P_{r\ minimum\ usable}$, cannot be too large or too small.

The time over which a call may be maintained within a cell, without handoff, is called the *dwell time* . The dwell time of a particular user is governed by a number of factors, including propagation, interference, distance between the subscriber and the base station, and other time varying effects depending on the speed of the user and the type of radio coverage.

Prioritizing Handoffs

One method for giving priority to handoffs is called the *guard channel concept*, whereby a fraction of the total available channels in a cell is reserved exclusively for handoff requests from ongoing calls which may be handed off into the cell. This method has the disadvantage of reducing the total carried traffic, as fewer channels are allocated to originating calls. Guard channels, however, offer efficient spectrum utilization when dynamic channel assignment strategies, which minimize the number of required guard channels by efficient demand-based allocation, are used.

Practical Handoff Considerations

The practical cellular systems, several problems arise when attempting to design for a wide range of mobile velocities.

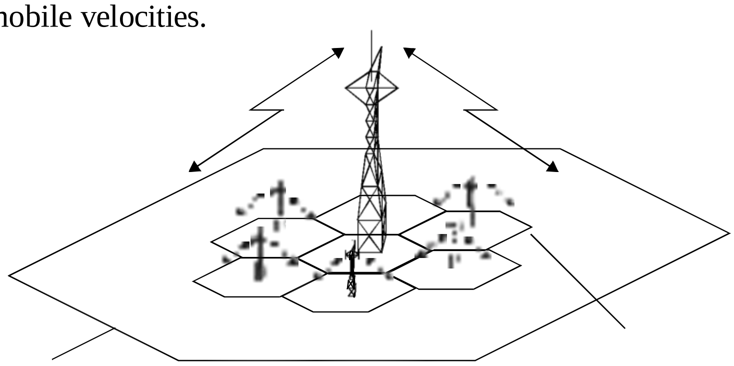


Figure 3.4 The umbrella cell approach.

High speed vehicles pass through the coverage region of a cell within a matter of seconds, whereas pedestrian users may never need a handoff during a call. Particularly with the addition of microcells to provide capacity, the MSC can quickly become burdened if high speed users are constantly being passed between very small cells. Several schemes have been devised to handle the simultaneous traffic of high speed and low speed users while minimizing the handoff intervention from the MSC. Another practical limitation is the ability to obtain new cell sites. Large “umbrella” cell for high speed traffic Small microcells for low speed traffic

Another feature of newer cellular systems is the ability to make handoff decisions based on a wide range of metrics other than signal strength. The co-channel and adjacent channel interference levels may be measured at the base station or the mobile, and this information may be used with conventional signal strength data to provide a multi-dimensional algorithm for determining when a handoff is needed.

Interference and System Capacity

Interference is the major limiting factor in the performance of cellular radio systems. Sources of interference include another mobile in the same cell, a call in progress in a neigh boring cell, other base stations operating in the same frequency band, or any non-cellular system which inadvertently leaks energy into the cellular frequency band. Interference on voice channels causes cross talk, where the subscriber hears interference in the background due to an undesired transmission. On control channels, interference leads to missed and blocked calls due to errors in the digital signalling. Interference is more severe in urban areas, due to the greater RF noise floor and the large number of base stations and mobiles. Interference has been recognized as a major bottleneck in increasing capacity and is often responsible for dropped calls. The two major types of system-generated cellular interference are co-channel interference and adjacent channel interference. Even though interfering signals are often generated within the cellular system,

Trunking and Grade of Service

Cellular radio systems rely on trunking to accommodate a large number of users in a limited radio spectrum. The concept of trunking allows a large number of users to share the relatively small number of channels in a cell by providing access to each user, on demand, from a pool of available channels. In a trunked radio system, each user is allocated a channel on a per call basis, and upon termination of the call, the previously occupied channel is immediately returned to the pool of available channels

The grade of service (GOS) is a measure of the ability of a user to access a trunked system during the busiest hour. The busy hour is based upon customer demand at the busiest hour during a week, month, or year. The busy hours for cellular radio systems typically occur during rush hours, between 4 p.m. and 6 p.m. on a Thursday or Friday evening. The grade of service is a benchmark used to define the desired performance of a particular trunked system by specifying a desired likelihood of a user obtaining channel access given a specific number of channels avail- able in the system. It is the wireless designer's job to estimate the maximum required capacity and to allocate the proper number of channels in order to meet the GOS. GOS is typically given as the likelihood that a call is blocked, or the likelihood of a call experiencing a delay greater than a certain queuing time.

Set-up Time: The time required to allocate a trunked radio channel to a requesting user.

Blocked Call: Call which cannot be completed at time of request, due to congestion. Also referred to as a lost call.

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

Traffic Intensity: Measure of channel time utilization, which is the average channel occupancy measured in Erlangs. This is a dimensionless quantity and may be used to measure the time utilization of single or multiple channels. Denoted by A.

Load: Traffic intensity across the entire trunked radio system,
measured in Erlangs.

Grade of Service (GOS): A measure of congestion which is specified as the probability of a call being blocked (for

Erlang B), or the probability of a call being delayed beyond a certain amount of time (for Erlang C).

Request Rate: The average number of call requests per unit time. Denoted by λ seconds–1.

A number of definitions listed in Table 3.3 are used in trunking theory to make capacity estimates in trunked systems. The traffic intensity offered by each user is equal to the call request rate multiplied by the holding time. That is, each user generates a traffic intensity of Au Erlangs given by

$$Au = H \qquad (3.13)$$

where H is the average duration of a call and λ is the average number of call requests per unit time for each user. For a system containing U users and an unspecified number of channels, the total offered traffic intensity A, is given as

$$A = UA_u \quad (3.14)$$

Furthermore, in a C channel trunked system, if the traffic is equally distributed among the channels, then the traffic intensity per channel, Ac, is given as

$$A_c = UA_u C \quad (3.15)$$

GOS for a trunked system which provides no queuing for blocked calls. The Erlang B formula is derived in Appendix A and is given by

$$Pr[blocking] = \frac{\frac{A^C}{C!}}{\sum_{k=0}^{C} \frac{A^k}{k!}} = GOS \quad (3.16)$$

Where C is the number of trunked channels offered by a trunked radio system and A is the total offered traffic. While it is possible to model trunked systems with finite users, the resulting expressions are much more complicated than the Erlang B result, and the added complexity is not warranted for typical trunked systems which have users that outnumber available channels by orders of magnitude. Furthermore, the Erlang B formula provides a conservative estimate of the GOS,

To find the GOS, it is first necessary to find the likelihood that a call is initially denied access to the system. The likelihood of a call not having immediate access to a channel is determined by the Erlang C formula derived in Appendix A

$$Pr[delay > 0] = \frac{A^C}{A^C + C!\left(1 - \frac{A}{C}\right)\sum_{k=0}^{C-1} \frac{A^k}{k!}} \quad (3.17)$$

If no channels are immediately available the call is delayed, and the probability that the delayed call is forced to wait more than t seconds is given by the probability that a call is delayed, multiplied by the conditional probability that the delay is greater than t seconds. The GOS of a trunked system where blocked calls are delayed is hence given by

$$\begin{aligned} Pr[delay > t] &= Pr[delay > 0]Pr[delay > t \mid delay > 0] \\ &= Pr[delay > 0]\exp(-(C - A)t/H) \end{aligned} \quad (3.18)$$

The average delay D for all calls in a queued system is given by

$$D = Pr[delay > 0]\frac{H}{C - A} \quad (3.19)$$

where the average delay for those calls which are queued is given by *H/(C – A)*.The Erlang B and Erlang C formulas are plotted in graphical form in Figures 3.6 and 3.7.These graphs are useful for determining GOS in rapid fashion, although computer simulations are often used to determine transient behaviours experienced by particular users in a mobile system. To use Figures 3.6 and 3.7, locate the number of channels on the top portion of the graph. Locate the traffic intensity of the system on the bottom portion of the graph. The blocking probability *Pr*[*blocking*] is shown on the abscissa of Figure 3.6, and *Pr*[*delay*] > 0 is shown on the abscissa of Figure 3.7. With two of the parameters specified, it is easy to find the third parameter.

Improving Coverage and Capacity in Cellular Systems

As the demand for wireless service increases, the number of channels assigned to a cell eventually becomes insufficient to support the required number of users. At this point,. In order to cover the entire service area with

Figure 3.8 Illustration of cell splitting.

Where Pt1 and Pt2 are the transmit powers of the larger and smaller cell base stations, respectively, and n is the path loss exponent. If we take n = 4 and set the received powers equal to each other, then(3.22)In other words, the transmit power must be reduced by 12 dB in order to fill in the original coverage area with microcells, while maintaining the S/I requirement. In practice, not all cells are split at the same time. This splitting process continues until all the channels in an area are used in the lower power group, at which point cell splitting is complete within the region, and the entire system is rescaled to have a smaller radius per cell. Antenna down tilting, which deliberately focuses radiated energy from the base station toward the ground (rather than toward the horizon), is often used to limit the radio coverage of newly formed microcells.

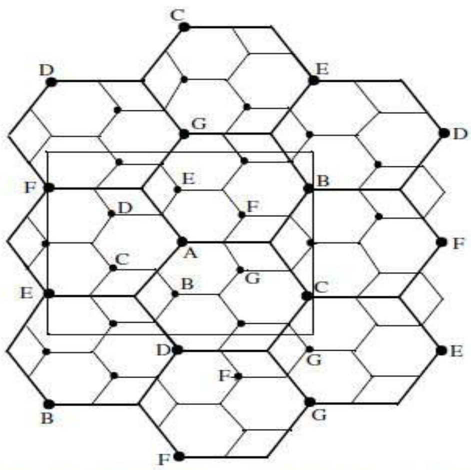


Figure 3.9 Illustration of cell splitting within a 3 km by 3 km square centered around base station A.

Sectoring

As shown in Section cell splitting achieves capacity improvement by essentially rescaling the system. By decreasing the cell radius R and keeping the co-channel reuse ratio D/R unchanged, cell splitting increases the number of channels per unit area. However, another way to increase capacity is to keep the cell radius unchanged and seek methods to decrease the D/R ratio. As we now show, sectoring increases SIR so that the cluster size may be reduced. In this approach, first the SIR is improved using directional antennas, then capacity improvement is achieved by reducing the number of cells in a cluster, thus increasing the frequency reuse.

However, in order to do this successfully, it is necessary to reduce the relative interference without decreasing the transmit power. The co- channel interference in a cellular system may be decreased by replacing a single Omni directional antenna at the base station by several directional antennas, each radiating within a specified sector. By using directional antennas, a given cell will receive interference and transmit with only a fraction of the available co-channel cells. The technique for decreasing co-channel interference and thus increasing system performance by using directional antennas is called sectoring. The factor by which the co-channel interference is reduced depends on the amount of sectoring used. A cell is normally partitioned into three 120° sectors or six 60° sectors as shown in Figure 3.10(a) and (b).

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

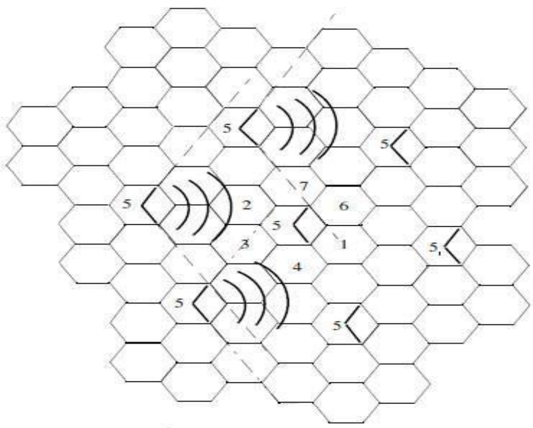


Figure 3.11 Illustration of how 120° sectoring reduces interference from co-channel cells. Out of the 6 co-channel cells in the first tier, only two of them interfere with the center cell. If omnidirectional antennas were used at each base station, all six co-channel cells would interfere with the center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

Repeaters for Range Extension

Repeaters are bidirectional in nature, and simultaneously send signals to and receive signals from a serving base station. Repeaters work using over-the-air signals, so they may be installed anywhere and are capable of repeating an entire cellular or PCS band. Upon receiving signals from a base station forward link, the repeater amplifies and reradiates the base station signals to the specific coverage region.

Unfortunately, the received noise and interference is also reradiated by the repeater on both the forward and reverse link, so care must be taken to properly place the repeaters, and to adjust the various forward and reverse link amplifier levels and antenna patterns. Repeaters can be easily thought of as bidirectional "bent pipes" that retransmit what has been received. In practice, directional antennas or distributed antenna systems (DAS) are connected to the inputs or outputs of repeaters for localized spot coverage, particularly in tunnels or buildings. By modifying the coverage of a serving cell, an operator is able to dedicate a certain amount of the base station's traffic for the areas covered by the repeater. However, the repeater does not add capacity to the system

A Microcell Zone Concept

The increased number of handoffs required when sectoring is employed results in an increased load on the switching and control link elements of the mobile system. A solution to this problem was presented by Lee [Lee91b]. This proposal is based on a microcell concept for seven cell reuse, as illustrated in figure 3.13. In this scheme, each of the three (or possibly more) zone sites The zones are connected by coaxial cable, fibrotic cable, or microwave link to the base station. Multiple zones and a single base station make up a cell. As a mobile travels within the cell, it is served by the zone with the strongest signal. This approach is superior to sectoring since antennas are placed at the outer edges of the cell, and any base station channel may be assigned to any zone by the base station.

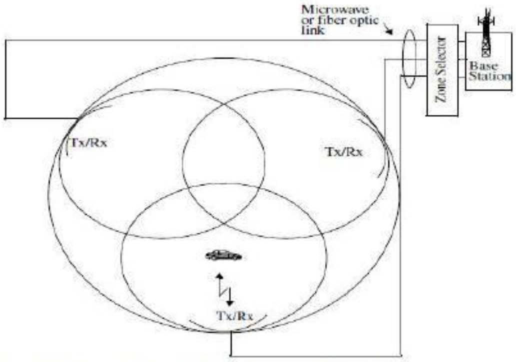


Figure 3.13 The microcell concept [adapted from [Lee91b] © IEEE].

UNIT-II

MOBILE RADIO PROPAGATION

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,

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

Where Pt is the transmitted power, Pr(d) is the received power which is a function of the T-R separation, Gt is the transmitter antenna gain, Gr is the receiver antenna gain, d is the T-R separation distance in meters, L is the system loss factor not related to propagation (L ~ 1), and A is the wavelength in meters. The gain of an antenna is related to its effective aperture, Ae, by

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

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

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

Where *f* is the carrier frequency in Hertz, roc is the carrier frequency in radians per second, and *c* is the speed of light given in meters/s. The values for *P1* and *P r* must be expressed in the same units, and *G1* and *Gr* are dimensionless quantities. The miscellaneous losses *L (L* ~ 1) are usually due to transmission line attenuation, filter losses, and antenna losses in the communication system. A value of *L* = 1indicates no loss in the system hardware.

The Friis free space equation of (4.1) shows that the received power falls off as the square of the T-R separation distance. This implies that the received power decays with distance at a rate of 20 dB/decade.

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

$$EIRP = P_t G_t$$

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

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

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

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

The Friis free space model is only a valid predictor for Pr 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 d1, which is related to the largest linear dimension of the transmitter antenna aperture and the carrier wavelength. The Fraunhofer distance is given by

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

Where D is the largest physical linear dimension of the antenna. Additionally, to be in the far-field region, d *f* must satisfy

$$d_f \gg D$$

$$d_f \gg \lambda$$

The received power in free space at a distance greater than d0 is given by

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

In mobile radio systems, it is not uncommon to find that Pr may change by many orders of magnitude over a typical coverage area of several square kilometers. Because of the large dynamic range of received power levels, often dBm or dBW units are used to express received power levels. Equation (4.8) may be expressed in units of dBm or dBW by simply taking the logarithm of both sides and multiplying by 10. For example, if P r is in units of dBm, the received power is given by

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

Relating Power to Electric Field

The free space path loss model of Section 2.1 is readily derived from first principles. It can be proven that any radiating structure produces electric and magnetic fields. Consider a small linear radiator of length *L that* is placed coincident with the z-axis and has its center at the origin, as shown in Figure 2.1.

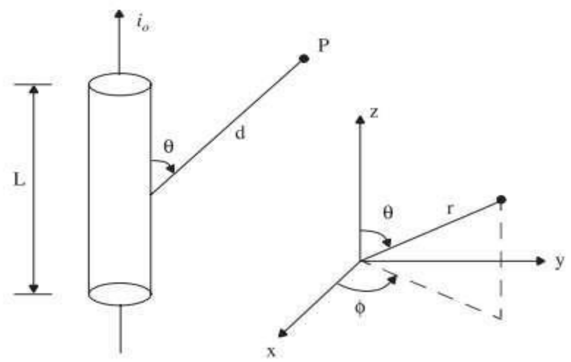


Figure 2.1 Illustration of a linear radiator of length *L(L* « A.), carrying a current of amplitude *i0*and making an angle e with a point, at distance *d*.

$$E_r = \frac{i_0 L \cos\theta}{2\pi\varepsilon_0 c}\left\{\frac{1}{d^2} + \frac{c}{j\omega_c d^3}\right\} e^{j\omega_c(t-d/c)}$$

$$E_\theta = \frac{i_0 L \sin\theta}{4\pi\varepsilon_0 c^2}\left\{\frac{j\omega_c}{d} + \frac{c}{d^2} + \frac{c^2}{j\omega_c d^3}\right\} e^{-j\omega_c(t-d/c)}$$

$$H_\phi = \frac{i_0 L \sin\theta}{4\pi c}\left\{\frac{j\omega_c}{d} + \frac{c}{d^2}\right\} e^{j\omega_c(t-d/c)}$$

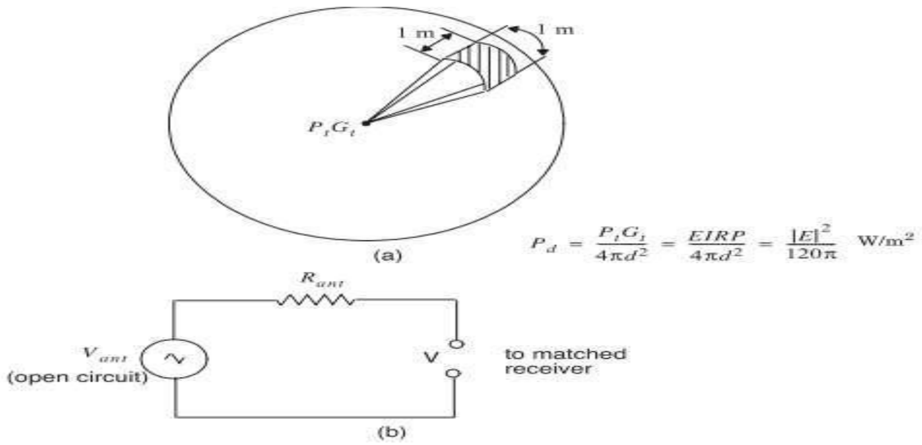
$E_\phi = H_r = H_\theta = 0.$

$$P_d = \frac{EIRP}{4\pi d^2} = \frac{P_t G_t}{4\pi d^2} = \frac{E^2}{R_{fs}} = \frac{E^2}{\eta} \text{ W/m}^2$$

$$P_d = \frac{|E|^2}{377\Omega} \text{ W/m}^2$$

With ,In the above equations, all terms represent the radiation field component,

$$P_r(d) = P_d A_e = \frac{|E|^2}{120\pi} A_e = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2} = \frac{|E|^2 G_r \lambda^2}{480\pi^2} \text{ W}$$



$$P_r(d) = \frac{V^2}{R_{ant}} = \frac{[V_{ant}/2]^2}{R_{ant}} = \frac{V_{ant}^2}{4R_{ant}}$$

The Three Basic Propagation Mechanisms

Reflection, diffraction, and scattering are the three basic propagation mechanisms which impact propagation in a mobile communication system. These mechanisms are briefly explained in this section, and propagation models which describe these mechanisms are discussed subsequently in this chapter.

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.

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

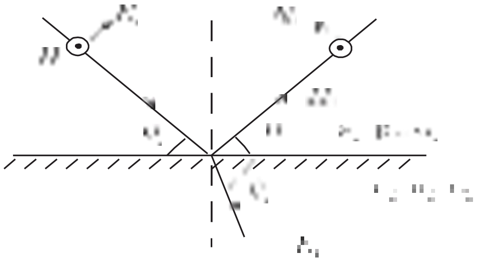
When a radio wave propagating in one medium impinges upon another medium having different electrical Properties, the wave is partially reflected and partially transmitted. If the plane wave is incident on a perfect dielectric, part of the energy is transmitted into the second medium and part of the energy is reflected back into the first medium, and there is no loss of energy in absorption. If the second medium is a 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

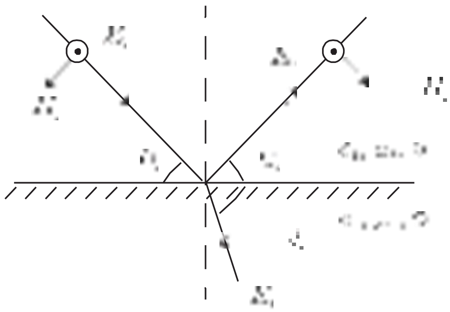
Reflection from Dielectrics

An electromagnetic wave incident at an angle 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 and part of the energy is transmitted (refracted) into the second media at an angle 1 the nature of reflection varies with the direction of polarization

of the E-field. The behaviour for arbitrary directions of polarization can be studied by considering the two distinct cases shown in Figure 4.4. The plane of incidence is defined as the plane containing the incident, reflected, and transmitted rays [Ram65]. In Figure

4.4a, 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 to the incident,





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

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

$$\varepsilon' = \frac{\sigma}{2\pi f}$$

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

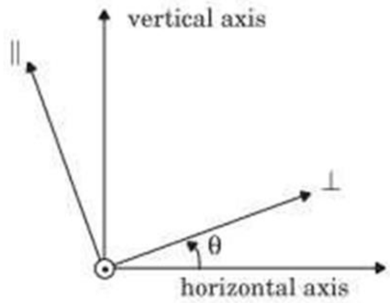
$$\Gamma_{\perp} = \frac{E_r}{E_i} = \frac{\eta_2 \sin\theta_i - \eta_1 \sin\theta_t}{\eta_2 \sin\theta_i + \eta_1 \sin\theta_t} \quad \text{(E-field normal to the plane of incidence)}$$

For the case when the first medium is free space and the reflection coefficients for the two cases of vertical and horizontal polarization can be simplified to

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

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

For the case of elliptical polarized waves, the wave may be broken down (depolarized) into its vertical and horizontal E-field components, and superposition may be applied to determine transmitted and reflected waves. In the general case of reflection or transmission, the horizontal and vertical



Brewster Angle

The *Brewster angle* is the angle at which no reflection occurs in the medium of origin. It occur when the incident angle θ_B is such that the reflection coefficient $\Gamma_{||}$ is equal to zero (se Figure 4.6). The Brewster angle is given by the value of θ_B which satisfies

$$\sin(\theta_B) = \sqrt{\frac{\varepsilon_1}{\varepsilon_1 + \varepsilon_2}} \tag{4.27}$$

For the case when the first medium is free space and the second medium has a relative permittivity ε_r, Equation (4.27) can be expressed as

$$\sin(\theta_B) = \frac{\sqrt{\varepsilon_r - 1}}{\sqrt{\varepsilon_r^2 - 1}} \tag{4.28}$$

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

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 Maxwell'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,

$$\theta_i = \theta_r$$

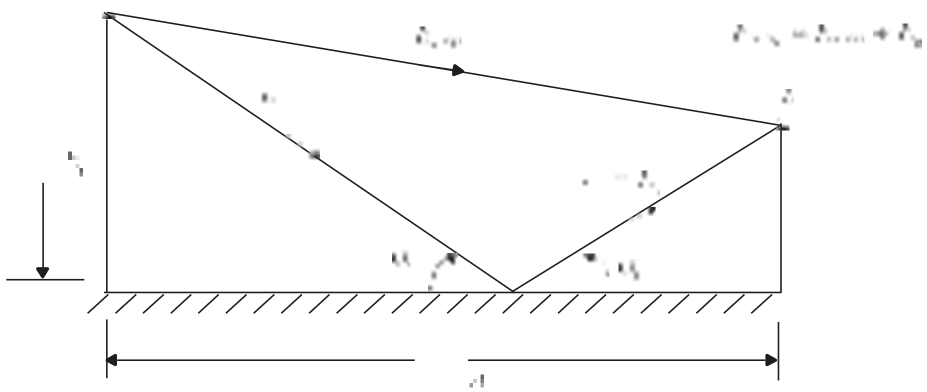
$$E_i = E_r \qquad \text{(E-field in plane of incidence)}$$

Ground Reflection (Two-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, is in most cases inaccurate when used alone. The two-ray ground reflection model shown in Figure 4.7 is a useful propagation model that is based on geometric optics, and considers both the direct path

(receiver)

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 kilometres 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. In most mobile communication systems, the maximum T-R separation distance is at most only a few tens of kilometres, and the earth may be assumed to be flat. The total received E field, ETOT, is then a result of the direct line-of-sight component, E Los, and the ground reflected component, E g •



The free space propagating E-field is given by

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

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

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

and the E-field for the ground reflected wave, which has a propagation distance of *d"*, can be expressed as

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

The electric field *ETOT(d, t)* can be expressed as the sum of Equations

$$E_{TOT}(d,t) = \frac{E_0 d_0}{d}\cos\left(\omega_c\left(t-\frac{d}{c}\right)\right) + (-1)\frac{E_0 d_0}{d}\cos\left(\omega_c\left(t-\frac{d}{c}\right)\right)$$

When the T-R separation distance *d* is very large compared to *hr* + *hr* can be simplified using a Taylor series approximation

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

$$\theta_\Delta = \frac{2\pi\Delta}{\lambda} = \frac{\Delta\omega_c}{c} \tag{4.42}$$

and

$$\tau_d = \frac{\Delta}{c} = \frac{\theta_\Delta}{2\pi f_c} \tag{4.43}$$

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

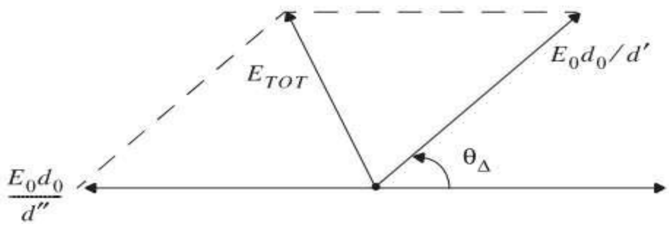
$$\left|\frac{E_0 d_0}{d}\right| \approx \left|\frac{E_0 d_0}{d'}\right| \approx \left|\frac{E_0 d_0}{d''}\right| \tag{4.44}$$

If the received E-field is evaluated at some time, say at $t = d''/c$, Equation (4.39) can be expressed as a phasor sum

$$\begin{aligned} E_{TOT}\left(d, t = \frac{d''}{c}\right) &= \frac{E_0 d_0}{d'}\cos\left(\omega_c\left(\frac{d''-d'}{c}\right)\right) - \frac{E_0 d_0}{d''}\cos 0^\circ \\ &= \frac{E_0 d_0}{d'}\angle\theta_\Delta - \frac{E_0 d_0}{d''} \\ &\approx \frac{E_o d_0}{d}[\angle\theta_\Delta - 1] \end{aligned} \tag{4.45}$$

where d is the distance over a flat earth between the bases of the transmitter and receiver antennas. Referring to the phasor diagram of Figure 4.9 which shows how the direct and ground reflected rays combine, the electric field (at the receiver) at a distance d from the transmitter can be written as

$$|E_{TOT}(d)| = \sqrt{\left(\frac{E_0 d_0}{d}\right)^2(\cos\theta_\Delta - 1)^2 + \left(\frac{E_0 d_0}{d}\right)^2 \sin^2\theta_\Delta} \tag{4.46}$$



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 wave front can be considered as point sources for the production of secondary wavelets, and that these wavelets combine to produce a new wave front 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 4.10a. 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 $h \ll d_1, d_2$ and $h \gg \lambda$, 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 4.10b as

$$\Delta \approx \frac{h^2}{2}\frac{(d_1 + d_2)}{d_1 d_2} \tag{4.54}$$

The corresponding phase difference is given by

$$\phi = \frac{2\pi\Delta}{\lambda} \approx \frac{2\pi}{\lambda}\frac{h^2}{2}\frac{(d_1 + d_2)}{d_1 d_2} \tag{4.55}$$

and when $\tan x \approx x$, then $\alpha = \beta + \gamma$ from Figure 4.10c and

$$\alpha \approx h\left(\frac{d_1 + d_2}{d_1 d_2}\right)$$

(a proof of Equations (4.54) and (4.55) is left as an exercise for the reader).

Equation (4.55) is often normalized using the dimensionless *Fresnel–Kirchoff* diffraction parameter v which is given by

$$v = h\sqrt{\frac{2(d_1 + d_2)}{\lambda d_1 d_2}} = \alpha\sqrt{\frac{2 d_1 d_2}{\lambda(d_1 + d_2)}} \tag{4.56}$$

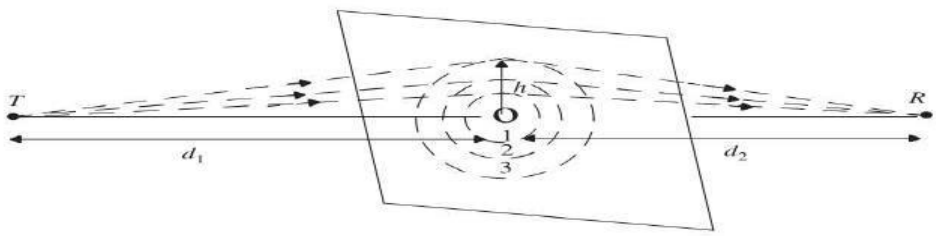
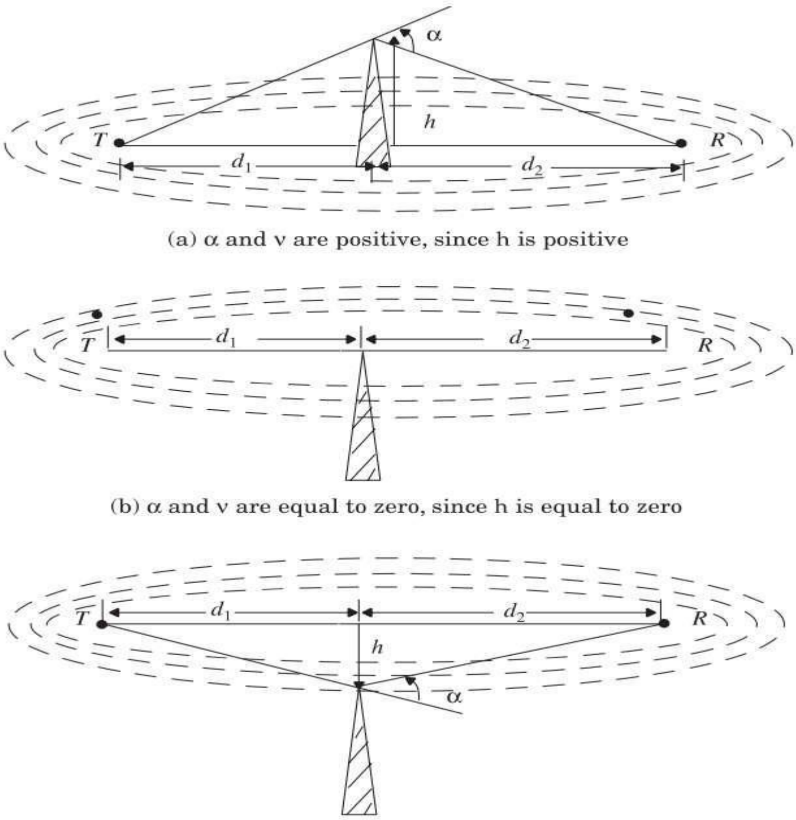


Figure 4.11 Concentric circles which define the boundaries of successive Fresnel zones.

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



(a) α and ν are positive, since h is positive

(b) α and ν are equal to zero, since h is equal to zero

(c) α and ν are negative, since h is negative

Figure 4.12 Illustration of Fresnel zones for different knife-edge diffraction scenarios.

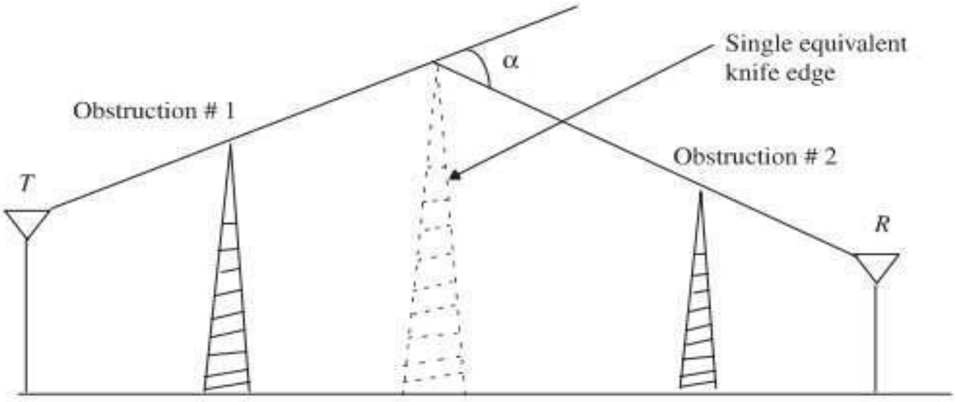
secondary Huygen's sources in the plane above the knife edge. The electric field strength, E_d, of a knife-edge diffracted wave is given by

$$\frac{E_d}{E_o} = F(v) = \frac{(1+j)}{2}\int_{v}^{\infty} \exp((-j\pi t^2)/2)dt \tag{4.59}$$

where E_o is the free space field strength in the absence of both the ground and the knife edge, and $F(v)$ is the complex Fresnel integral. The Fresnel integral, $F(v)$, is a function of the

Multiple Knife-edge Diffraction

In many practical situations, especially in hilly terrain, the propagation path may consist of more than one obstruction, in which case the total diffraction loss due to all of the obstacles must be computed. Bulling ton suggested that the series of obstacles be replaced by a single



Bulling ton’s construction of an equivalent knife edge

Equivalent obstacle so that the path loss can be obtained using single knife-edge diffraction models. This method 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

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 modelled 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 tested using the Rayleigh criterion which defines a critical height (he) of surface protuberances for a given angle of incidence ei ' given by

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

A surface is considered smooth if its minimum to maximum protuberance h is less than he, and is considered rough if the protuberance is greater than he. For rough surfaces, the fiat surface reflection coefficient needs to be multiplied by a scattering loss factor,

Radar Cross Section Model

In radio channels where large, distant objects induce scattering, knowledge of the physical location of such objects can be used to accurately predict scattered signal strengths. The *radar cross section* (RCS) of a scattering object is defined as the ratio of the power density of the signal scattered in the direction of the receiver to the power density of the radio wave incident upon the scattering object, and has units of square meters. Analysis based on the geometric theory of diffraction and physical optics may be used to determine the scattered field strength. For urban mobile radio systems, models based on the bi static radar equation may be used to compute the received power due to scattering in the far field.

$$P_R(\text{dBm}) = P_T(\text{dBm}) + G_T(\text{dBi}) + 20\log(\lambda) + RCS[\text{dB m}^2] - 30\log(4\pi) - 20\log d_T - 20\log d_R$$

Log-normal Shadowing

The surrounding environmental clutter may be vastly different at two different locations having the same T-R separation. This leads to measured signals which are vastly different than the average value predicted by Equation (4.68). Measurements have shown that at any value of d, the path loss PL(d) at a particular location is random and distributed log-normally (normal in dB) about the mean distance-dependent value . That is

$$PL(d)[dB] = \overline{PL}(d) + X_\sigma = \overline{PL}(d_0) + 10n\log\left(\frac{d}{d_0}\right) + X_\sigma \tag{4.69.a}$$

and

$$P_r(d)[dBm] = P_t[dBm] - PL(d)[dB] \quad \text{(antenna gains included in } PL(d)\text{)} \tag{4.69.b}$$

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

The log-normal distribution describes the random *shadowing* effects which occur over a large number of measurement locations which have the same T–R separation, but have different levels of

Clutter on the propagation path. This phenomenon is referred to as log-normal shadowing. Simply put, log-normal shadowing implies that measured signal levels at a specific T-R separation have a Gaussian (normal) distribution about the distance-dependent mean of (4.68), where the measured signal levels have values in dB units.

$$Q(z) = \frac{1}{\sqrt{2\pi}}\int_z^{\infty}\exp\left(-\frac{x^2}{2}\right)dx = \frac{1}{2}\left[1-erf\left(\frac{z}{\sqrt{2}}\right)\right]$$

$$Q(z) = 1 - Q(-z)$$

The probability that the received signal level (in dB power units) will exceed a certain value y can be calculated from the cumulative density function as

$$Pr[P_r(d) > \gamma] = Q\left(\frac{\gamma - \overline{P_r(d)}}{\sigma}\right)$$

Similarly, the probability that the received signal level will be below γ is given by

$$Pr[P_r(d) < \gamma] = Q\left(\frac{\overline{P_r(d)} - \gamma}{\sigma}\right)$$

Appendix F provides tables for evaluating the Q and erf functions.

Outdoor Propagation Models

Radio transmission in a mobile communications system often takes place over irregular terrain. The terrain profile of a particular area needs to be taken into account for estimating the path loss. The terrain profile may vary from a simple curved earth profile to a highly mountainous profile.

The presence of trees, buildings, and other obstacles also must be taken into account. A number of propagation models are available to predict path loss over irregular terrain. While all these models aim to predict signal strength at a particular **receiving** point or in a specific local area (called a sector), the methods vary widely in their approach, complexity, and accuracy. Most of these models are based on a systematic interpretation of measurement data obtained in the service area. Some of the commonly used outdoor propagation models are now discussed.

Longley-Rice Model

The Langley-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 two-ray ground reflection model) are used to predict signal strengths within the radio horizon. Diffraction losses over isolated obstacles are estimated using the Fresnel-Kirchhoff knife-edge models. Forward scatter theory is used to make troposcatter predictions over long distances, and far field diffraction losses in double horizon paths are predicted using a modified Vander Pol-Bremen method. The Langley-Rice propagation prediction model is also referred to as the IT'S irregular terrain model.

Durkin's Model-A Case Study

A classical propagation prediction approach similar to that used by Longley-Rice is discussed by Edwards and Durkin, as well as Dodson. These papers describe a computer simulator, for predicting field strength contours over irregular terrain that was adopted by the Joint Radio Committee (JRC) in the U.K. for the estimation of effective mobile radio coverage areas. Although this simulator only predicts large-scale phenomena (i.e. path loss), it provides an interesting perspective into the nature of propagation over irregular terrain and the losses caused by obstacles in a radio path. An explanation of the Edwards and Durkin method is presented here in order to demonstrate how all of the concepts described in this chapter are used in a single model. Therefore, each point of the reconstructed profile consists of an average of the heights obtained by diagonal, vertical (row), and horizontal (column) interpolation methods.

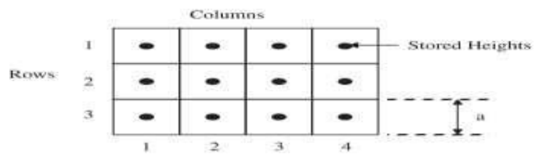
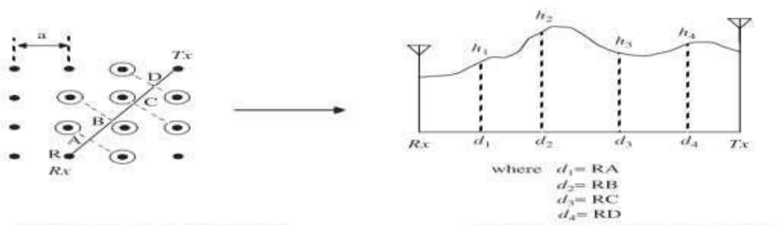


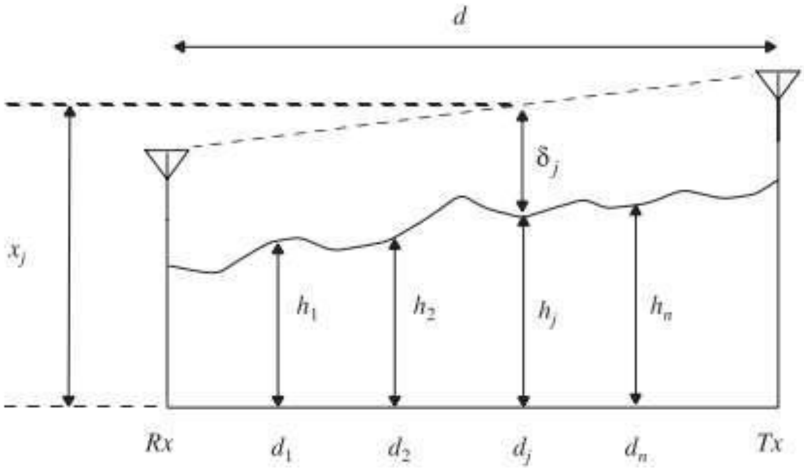
Figure 4.19 Illustration of a two-dimensional array of elevation information.



(a) Top view of interpolated map and line between Tx and Rx

(b) Side view showing reconstructed terrain profile between Tx and Rx

The first step is to decide whether a line-of-sight (LOS) path exists between the transmitter and the receiver.



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 *(Am11)*, in an urban area over a quasi- smooth terrain with a base station effective antenna height *(h1e)* of 200m and a mobile antenna height *(hre)* 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 *Amu(f, d)* (as read from the curves) is added to it along with correction factors to account for the type of terrain.

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

Where L 50 is the 50th percentile (i.e., median) value of propagation path loss, LF is the free space propagation loss, Amu is the median attenuation relative to free space, G(hre) is the base Station antenna height gain factor, G(hre) is the mobile antenna height gain factor, and GAREA 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.

Plots of $A_{mu}(f, d)$ and G_{AREA} for a wide range of frequencies are shown in Figure 4.23 and Figure 4.24. Furthermore, Okumura found that $G(h_{te})$ varies at a rate of 20 dB/decade and $G(h_{re})$ varies at a rate of 10 dB/decade for heights less than 3 m

$$G(h_{te}) = 20\log\left(\frac{h_{te}}{200}\right) \qquad 1000\text{ m} > h_{te} > 30\text{ m} \qquad (4.81.a)$$

$$G(h_{re}) = 10\log\left(\frac{h_{re}}{3}\right) \qquad h_{re} \le 3\text{ m} \qquad (4.81.b)$$

$$G(h_{re}) = 20\log\left(\frac{h_{re}}{3}\right) \qquad 10\text{ m} > h_{re} > 3\text{ m} \qquad (4.81.c)$$

Other corrections may also be applied to Okumura's model. Some of the important terrain related parameters are the terrain undulation height (Δh), isolated ridge height, average slope of the terrain and the mixed land–sea parameter. Once the terrain related parameters are calculated, the necessary correction factors can be added or subtracted as required. All these correction factors are also available as Okumura curves [Oku68].

Hata Model

The Hata model 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

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

where *fc* is the frequency (in MHz) from 150 MHz to 1500 MHz, *h1e* is the effective transmitter (base station) antenna height (in meters) ranging from 30m to 200m, *hre* is the effective receiver (mobile) antenna height (in meters) ranging from 1m to 10m, dis the T-R separation distance (in km),

and for a large city, it is given by

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

$$a(h_{re}) = 3.2(\log 11.75h_{re})^2 - 4.97 \text{ dB} \quad \text{for } f_c \ge 300\text{ MHz}$$

To obtain the path loss in a suburban area, the standard Rata formula in Equation

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

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

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

PCS Extension to Hata Mode

The European Cooperative for Scientific and Technical research (EURO-COST) formed the COST-231 working committee to develop an extended version of the Hata model. COST-231 proposed the following formula to extend Rata's model to 2 GRz. The proposed model for path loss is [EUR91]

$$L_{50}(urban) = 46.3 + 33.9\log f_c - 13.82\log h_{te} - a(h_{re}) + (44.9 - 6.55\log h_{te})\log d + C_M \quad (4.87)$$

where $a(h_{re})$ is defined in Equations (4.83), (4.84.a), and (4.84.b) and

$$C_M = \begin{matrix} 0\ \text{dB} & \text{for medium sized city and suburban areas} \\ 3\ \text{dB} & \text{for metropolitan centers} \end{matrix} \quad (4.88)$$

The COST-231 extension of the Hata model is restricted to the following range of parameters:

f : 1500 MHz to 2000 MHz
h_{te} : 30 m to 200 m
h_{re} : 1 m to 10 m
d : 1 km to 20 km

Walfisch and Bertoni Model

A model developed by Walfisch and Bertoni [Wal88] considers the impact of rooftops and building height by using diffraction to predict average signal strength at street level. The model considers the path loss, S, to be a product of three factors.

$$S = P_0 Q^2 P_1 \quad (4.89)$$

where P_0 represents free space path loss between isotropic antennas given by

$$P_0 = \left(\frac{\lambda}{4\pi R}\right)^2 \quad (4.90)$$

The factor Q^2 gives the reduction in the rooftop signal due to the row of buildings which immediately shadow the receiver at street level. The P_1 term is based upon diffraction and determines the signal loss from the rooftop to the street.

In dB, the path loss is given by

$$S(\text{dB}) = L_0 + L_{rts} + L_{ms} \quad (4.91)$$

where L_0 represents free space loss, L_{rts} represents the "rooftop-to-street diffraction and scatter loss," and L_{ms} denotes multiscreen diffraction loss due to the rows of buildings [Xia92]. Figure 4.25 illustrates the geometry used in the Walfisch Bertoni model [Wal88], [Mac93]. This model is being considered for use by ITU-R in the IMT-2000 standards activities.

Wideband PCS Microcell Model

Work by Feuerstein et al. in 1991 used a 20 MHz pulsed transmitter at 1900 MHz to measure path loss, outage, and delay spread in typical microcellular systems in San Francisco and Oakland. Using base station antenna heights of 3.7 m, 8.5 m, and 13.3 m, and a mobile receiver with an antenna height of 1.7 m above ground, statistics for path loss, multipath, and coverage area

Indoor Propagation Models

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.

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 none in forced 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

Table 4.3 Average Signal Loss Measurements Reported by Various Researchers for Radio Paths Obstructed by Common Building Material

Material Type	Loss (dB)	Frequency	Reference
All metal	26	815 MHz	[Cox83b]
Aluminum siding	20.4	815 MHz	[Cox83b]
Foil insulation	3.9	815 MHz	[Cox83b]
Concrete block wall	13	1300 MHz	[Rap91c]
Loss from one floor	20-30	1300 MHz	[Rap91c]
Loss from one floor and one wall	40-50	1300 MHz	[Rap91c]
Fade observed when transmitter turned a right angle corner in a corridor	10-15	1300 MHz	[Rap91c]
Light textile inventory	3-5	1300 MHz	[Rap91c]
Chain-like fenced in area 20 ft high containing tools, inventory, and people	5-12	1300 MHz	[Rap91c]
Metal blanket — 12 sq ft	4-7	1300 MHz	[Rap91c]
Metallic hoppers which hold scrap metal for recycling — 10 sq ft	3-6	1300 MHz	[Rap91c]
Small metal pole — 6" diameter	3	1300 MHz	[Rap91c]
Metal pulley system used to hoist metal inventory — 4 sq ft	6	1300 MHz	[Rap91c]
Light machinery < 10 sq ft	1-4	1300 MHz	[Rap91c]
General machinery — 10 - 20 sq ft	5-10	1300 MHz	[Rap91c]
Heavy machinery > 20 sq ft	10-12	1300 MHz	[Rap91c]
Metal catwalk/stairs	5	1300 MHz	[Rap91c]
Light textile	3-5	1300 MHz	[Rap91c]
Heavy textile inventory	8-11	1300 MHz	[Rap91c]
Area where workers inspect metal finished products for defects	3-12	1300 MHz	[Rap91c]
Metallic inventory	4-7	1300 MHz	[Rap91c]
Large I-beam — 16 - 20"	8-10	1300 MHz	[Rap91c]
Metallic inventory racks — 8 sq ft	4-9	1300 MHz	[Rap91c]
Empty cardboard inventory boxes	3-6	1300 MHz	[Rap91c]
Concrete block wall	13-20	1300 MHz	[Rap91c]
Ceiling duct	1-8	1300 MHz	[Rap91c]
2.5 m storage rack with small metal parts (loosely packed)	4-6	1300 MHz	[Rap91c]
4 m metal box storage	10-12	1300 MHz	[Rap91c]

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 [Sei92a], [Sei92b]. Even the number of windows in a building and the presence of tinting (which attenuates radio energy) can impact the loss between floors. Table 4.4 illustrates values for floor attenuation factors (FAF) in three buildings in San Francisco [Sei92a]. 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. Table 4.5 illustrates very similar tendencies. Meter about five or six floor separations, very little additional path loss is experienced

Table 4.5 Average Floor Attenuation Factor in dB for One, Two, Three, and Four Floors in Two Office Buildings [Sei92b]

Building	FAF (dB)	τ (dB)	Number of locations
Office Building 1:			
Through One Floor	12.9	7.0	52
Through Two Floors	18.7	2.8	9
Through Three Floors	24.4	1.7	9
Through Four Floors	27.0	1.5	9
Office Building 2:			
Through One Floor	16.2	2.9	21
Through Two Floors	27.5	5.4	21
Through Three Floors	31.6	7.2	21

Log-distance Path Loss Model

Indoor path loss has been shown by many researchers to obey the distance power law in Equation (4.93)

$$PL(\mathrm{dB}) = PL(d_0) + 10n\log\left(\frac{d}{d_0}\right) + X_\sigma \quad (4.93)$$

where the value of n depends on the surroundings and building type, and X_σ represents a normal random variable in dB having a standard deviation of σ dB. Notice that Equation (4.93) is identical in form to the log-normal shadowing model of Equation (4.69.a). Typical values for various buildings are provided in Table 4.6 [And94].

Ericsson Multiple Breakpoint Model

The Ericsson radio system model was obtained by measurements in a multiple floor office building [Ake88]. The model has four breakpoints and considers both an upper and lower bound on the path loss. The model also assumes that there is 30 dB attenuation at d 0 = 1 m, which can be shown to be accurate for *f* = 900 MHz and unity gain antennas. Rather than assuming a log-normal shadowing component, the Ericsson model provides a deterministic limit on the range of path loss at a particular distance. Bernhardt [Ber89] used a uniform distribution to generate path loss values within the maximum and minimum range as a function of distance for in-building simulation. Figure 4.27 shows a plot of in-building path loss based on the Ericsson model as a function of distance.

Signal Penetration into Buildings

The signal strength received inside of a building due to an external transmitter is important for wireless systems that share frequencies with neigh boring buildings or with outdoor systems. As with propagation measurements between floors, it is difficult to determine exact models for penetration as only a limited number of experiments have been published, and they are sometimes difficult to compare. However, some generalizations can be made from the literature. In measurements reported to date, signal strength received inside building increases with height. At the lower floors of a building, the urban clutter induces greater attenuation and reduces the level of penetration. At higher floors, a LOS path may exist, thus causing a stronger incident signal at the exterior wall of the building. RF penetration has been found to be a function of frequency as well as height within the building.

Ray Tracing and Site Specific Modelling

In recent years, the computational and visualization capabilities of computers have accelerated rapidly. New methods for predicting radio signal coverage involve the use of Site Specific (SISP) propagation models and graphical information system (GIS) databases [Rus93]. SISP models support ray tracing as a means of deterministically modelling any indoor or outdoor propagation environment. Through the use of building databases, which may be drawn or digitized using standard graphical software packages, wireless system designers are able to include accurate representations of building and terrain features. For outdoor propagation prediction, ray tracing techniques are used in conjunction with aerial photographs so that three- dimensional (3-D) representations of buildings may be integrated with software that carries out reflection, diffraction, and scattering models

UNIT-III

CELLULAR SYSTEM DESIGN FUNDAMENTALS

In built-up urban areas, fading occurs because the height of the mobile antennas is well below the height of surrounding structures, line-of-sight path to the base station. Even when so there is no single line-of-sight exists, multi- path still occurs due to reflections from the ground and surrounding.

The incoming radio waves arrive from different directions with different propagation delays. The signal received by the mobile at structures. Any point in space may consist of a large number of plane waves having randomly distributed amplitudes. Phases and angles of arrival. These multipath components combine vector ally at the receiver antenna, and can cause the signal received by the mobile to distort or fade. Even when a mobile receiver is stationary, the received signal may fade due to movement of surrounding objects in the radio channel.

If objects in the radio channel are static, and motion is considered to be only due to that of the mobile, then fading is purely a spatial phenomenon. The spatial variations of the resulting signal are seen as temporal variations by the receiver as it moves through the multipath field. Due to the constructive and destructive effects of multipath waves summing at various points in space, a receiver moving at high speed can pass through several fades in a small period of time. In a more serious case, a receiver may stop at a particular location at which the received signal is in a deep fade. Maintaining good communications can then become very, although passing vehicles or people walking in the vicinity of the mobile can often disturb the field pattern, thereby diminishing the likelihood of the received signal remaining in a deep null for a long period of time.

- Factors Influencing Small-Scale Fading

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

Multipath propagation

The presence of reflecting objects and stutterers in the channel creates a constantly changing environment that dissipates the signal energy in amplitude, phase, and time. These effects result in multiple versions of the transmitted signal that arrive at the receiving antenna, displaced with respect to one another in time and spatial orientation. The

Random phase and amplitudes of the different multipath components cause fluctuations in signal strength, thereby inducing small-scale fading, signal distortion, or both. Multipath propagation often lengthens the time required for the baseband portion of the signal to reach the receiver which can cause signal smearing due to inter symbol interference.

• Speed of the mobile

The relative junction between the base station and the mobile results in random frequency modulation due to different Doppler shifts on each of the multipath components. Doppler shift will be positive negative depending on whether the mobile receiver is moving toward Or away from the base station PDF compression, OCR, web-optimization.

Doppler shift to the mobile velocity and the spatial angle between the direction of motion of the mobile and the direction of arrival of the wave. It can be seen from equation (4.2) that if the mobile is moving toward the direction of arrival of the wave, the Doppler shift is positive (i.e., the apparent received frequency is increased), and if the mobile is moving away from the direction of arrival of the wave, the Doppler shift is negative.

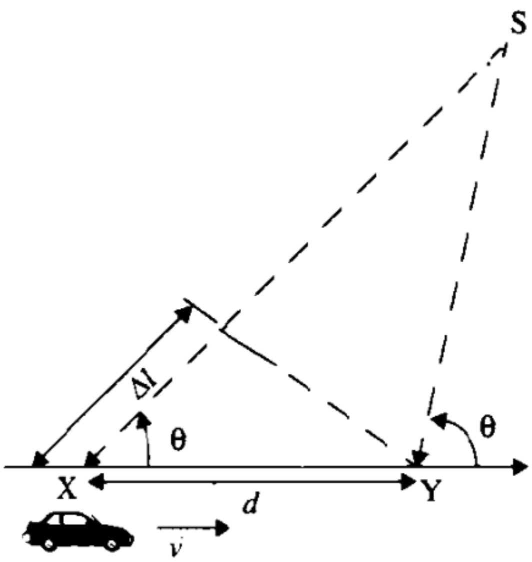


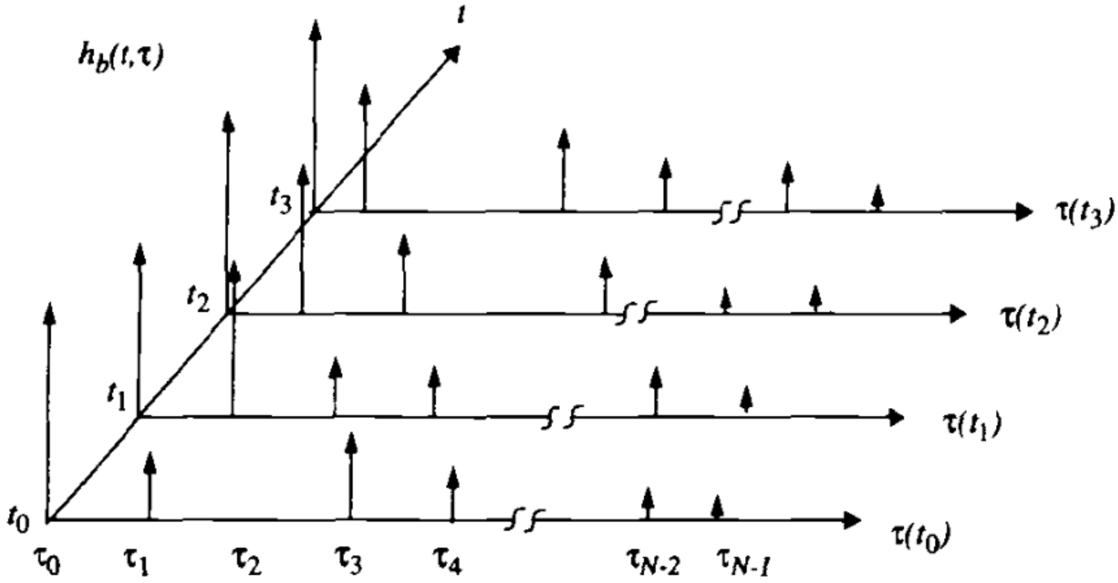
Illustration of Doppler Effect.

Impulse Response Model of a Multipath Channel

The small-scale variations of a mobile radio signal can be directly related to the impulse response of the mobile radio channel. The impulse response is a wideband channel characterization and contains all information necessary to simulate or analyse any type of radio transmission through the channel. This stems from the fact that a mobile radio channel may be modelled as a linear filter with a time varying impulse response, where the time variation is due receiver motion in space. The filtering nature of the channel is caused by the summation of amplitudes and delays of the multiple arriving

Waves at any instant of time The impulse response is a useful characterization of the channel, since it may be used to predict and compare the performance of many different mobile communication systems and transmission bandwidths for mobile channel condition.

Show that a mobile radio channel may be modelled a particular



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

Relationship between Bandwidth and Received Power

In actual wireless communication path channel is measured in the field now consider two extreme channel sounding how the small-scale fading, the impulse response of channel sounding techniques. We multiclass as a means of demonstrating quite differently for two signals with different bandwidths in the identical multipath channel.

Thus it is seen that the received local ensemble average power of wideband and narrowband signals are equivalent. When the transmitted signal has a bandwidth much greater than the bandwidth of the channel, then the multipath structure is completely resolved by the received signal at any time, and the received power varies very little since the individual multipath amplitudes do not change rapidly over a local area. However, if the transmitted signal has a very narrow bandwidth (e.g., the baseband signal has a duration greater than the excess delay of the channel), then multipath is not resolved by the received signal, and large signal fluctuations (fading) occur at the receiver due to the phase shifts of the many unresolved multipath components.

Direct RF Pulse System

This technique allows engineers to determine rapidly the power delay profile of any channel, as demonstrated by Rappaport. Essentially a wide band pulsed bi static radar, this system transmits an repetitive pulse of width tbb s, The signal is then amplified, detected with an envelope detector, and displayed and stored on a high speed oscilloscope. This gives an immediate measurement of the square of the channel impulse response convolved with the probing pulse .If the oscilloscope is set on averaging mode, and then this system can provide a local average power delay profile. Another attractive aspect of this system is the lack of complexity, since off-the-shelf equipment may be used.

Spread Spectrum Sliding Correlate Channel Sounding

When the incoming signal is correlated with the receiver sequence, the signal is collapsed back to the original bandwidth (i.e., "de spread"), envelope detected, and displayed on an oscilloscope. Since different incoming multipath will have different time delays, they will maximally correlate with the receiver PN sequence at different times. The energy of these individual paths will pass through the correlated depending on the time delay. Therefore, after envelope detection, the channel impulse response convolved with the pulse shape of a single chip is displayed on the oscilloscope.

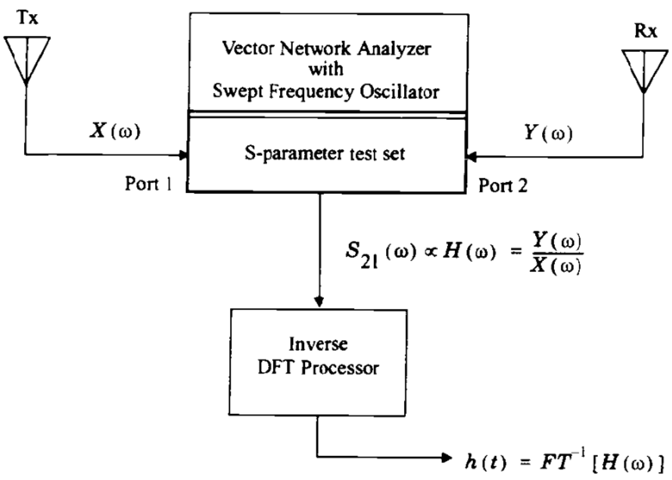
There are several advantages to the spread spectrum channel sounding system. One of the key spread spectrum modulation characteristics is the ability to reject pass band noise, thus improving the coverage range for a given transmitter power. Transmitter and receiver PN sequence synchronization is eliminated by the sliding correlated. Sensitivity is adjustable by changing the sliding factor and the post-correlated filter bandwidth. Also, required transmitter powers can be considerably lower than comparable direct pulse systems due to the inherent "processing gain" of spread spectrum systems.

A disadvantage of the spread spectrum system, as compared to the direct pulse system, is that measurements are not made in real time, but they are compiled as the PN codes slide past one another. Depending on system parameters and measurement objectives, the time required to make power delay profile measurements may be excessive. Another disadvantage of the system described here that a non-coherent detector is used, so that phases of individual multipath components cannot be measured. Even if coherent detection is used, the sweep time of a spread spectrum signal induces delay such that the phases of individual multipath components with different time delays would be measured at substantially different times, during which the channel might change.

Frequency Domain Channel Sounding

Because of the dual relationship between time domain and frequency domain techniques, it is possible to measure the channel impulse response in the frequency domain. Figure shows a frequency domain channel sounder used for measuring channel impulse responses. A vector network analyser controls a synthesized frequency sweeper, and an S-parameter test set is used to monitor the frequency response of the channel. The sweeper scans a particular frequency band (centred on the carrier) by stepping through discrete frequencies. The number and spacing of these frequency steps impact 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 level at port 2.

These signal levels allow the analyser to determine the complex response (i.e., transmissivity S21 (w)) of the channel over the measured frequency range. The transmissivity response is a frequency domain representation of the channel impulse response. This response is then converted to the time domain using inverse discrete Fourier transform (IDFT) processing, giving tied version of the impulse response. In a bandit this technique works well and
Indirectly provides amplitude and phase information in the time domain.



Parameters of Mobile

Multipath channel parameters are derived from the power delay profile, given by equation. Power delay profiles unique discussed in Section 4.4 and are generally represented received power as a function of excess delay with respect are measured using the tech- reference. Power delay profiles are found by averaging instantaneous delay profile measurements over as plots of relative to a fixed time delay a local area in order to determine an average small scale power delay profile. Depending on the time resolution of the probing pulse and the type of multipath channels studied, researchers often sample at spatial separations of power choose to a quarter of a wavelength and over receiver movements no greater than 6 m in outdoor channels and no greater than 2 in in indoor channels in the 450 MHz.

Time Dispersion Parameters

In order to compare different multipath channels and to develop some general design guidelines for wireless systems, parameters which grossly quantify the multipath channel are used. The mean excess delay, rms delay spread, and excess delay spread (X dB) are multipath channel parameters that can be determined from a power delay profile. The time dispersive properties of wide band multipath channels are most commonly quantified by their mean excess delay (0 and rms delay spread (ar). The mean excess delay is the first moment of the power deal) profile and is defined to be

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

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

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

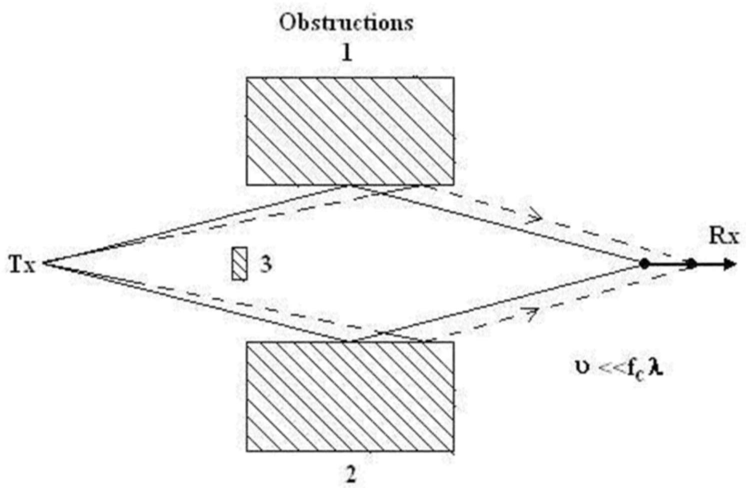
where

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

These delays are measured relative to the first detectable signal arriving at the receiver at to = 0. do not rely on the absolute power level of P (t), but only the relative amplitudes of the multipath components within P (t). Typical values of rms delay spread are on the order of microseconds in outdoor mobile radio channels and on the order of nanoseconds in indoor radio channels. Table 4.1 shows the typical measured values of rms delay spread.It is important to note that the rms delay spread and mean excess delay are defined from a single power delay profile which is the temporal or spatial average of consecutive impulse response measurements collected and averaged over a local area. I5rpically, many measurements are made at many local areas in order to determine a statistical range of multipath channel parameters for mobile communication system over a large-scale area .

In other words, the maximum excess delay is defined as t0 , where t0 is the first arriving signal and is the maximum delay at which a multipath component is within X dB of the strongest arriving multipath signal (which does not necessarily arrive at t0). Figure 4.10 illustrates the computation of the maximum excess delay for multipath components within 10 dB of the maximum. The maximum excess delay (X dB) defines the temporal extent of the multipath that is above a particular threshold. The value of is sometimes called the excess delay spread of a power delay profile, but in all cases must be specified with a threshold that relates the multipath noise floor to the maximum received multipath component.

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



Two ray NLoS multipath, resulting in Rayleigh fading.

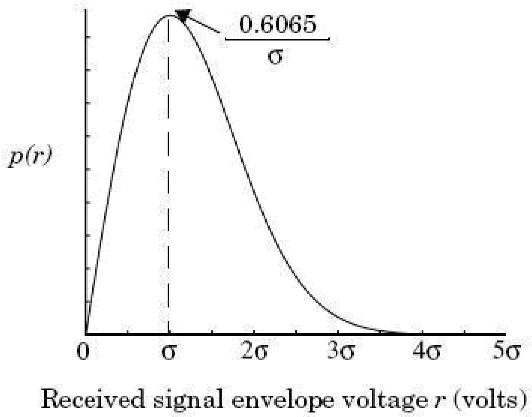
have been examined in order to find out the effective way to model and combat the channel effects. The most popular of these models are Rayleigh model, which describes the NLoS propagation. The Rayleigh model is used to model the statistical time varying nature of the received envelope of a flat fading 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 S1 and S2 received at two different time instants due to the presence of obstacles as shown in Figure. Now there can either be constructive or destructive interference between the two signals. Let En be the electric field and Θn be the relative phase of the various multipath signals. So we have

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

Now if N→ ∞(i.e. are sufficiently large number of multipath) and all the En are IID Distributed, then by Central Limit Theorem we have where Z_r and Z_i are Gaussian Random variables. For the above case with different σ values. This equation too is valid for all r ≥ 0. Above distribution is known as Rayleigh distribution and it has been derived

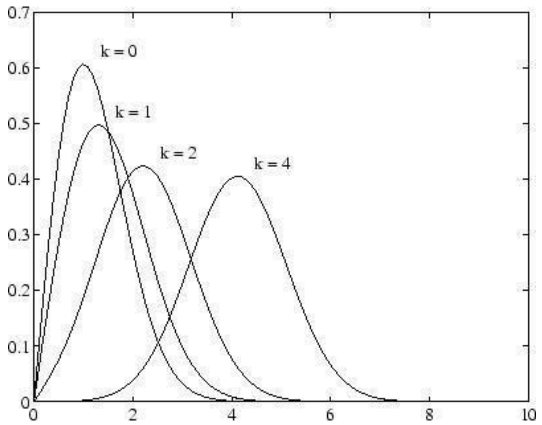


Rayleigh probability density function.

for slow fading. However, if fD 1 Hz, we call it as Quasi-stationary Rayleigh fading.

LoS Propagation: Rician Fading Model

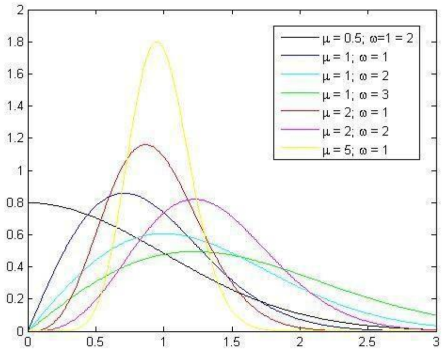
Rican Fading is the addition to all the normal multipath a direct LOS path.



Ricean probability density function.

Generalized Model: Nakagami Distribution

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



Nakagami probability density function.

Its pdf is given as Ω is the average signal$_r$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. Schematic representation of level crossing with a Rayleigh fading envelope at 10 Hz Doppler spread.

Where M = A 2σ 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 parameters 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.

where $\dot{r}$ is the time derivative of r(t), fD 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.= 2πfDρ

Apart from LCR, another parameter is fading rate, which is defined as the number of times the signal envelope crosses the middle value (rm) in a positive going direction per unit time. The average rate is expressed as 2v

N (rm) = λ (5.66)

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 Effect

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 Effect

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.

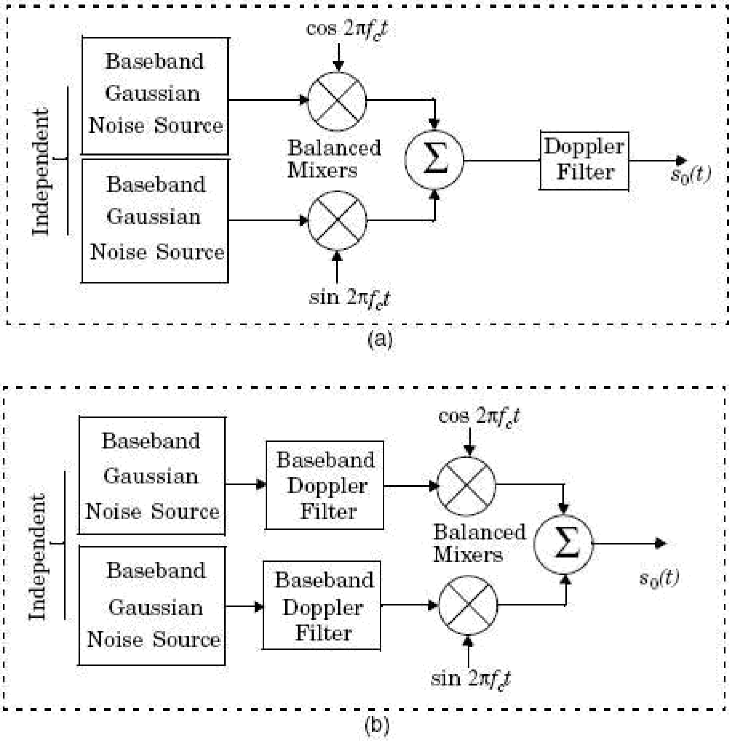


Figure 5.11: Clarke and Gan's model for Rayleigh fading generation using quadrature amplitude modulation with (a) RF Doppler filter, and, (b) baseband Doppler filter. Rayleigh Simulator with Wide Range of Channel Conditions.

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 ai and τi, as given in Figure 5.12.

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.

$$hb(t) = \alpha 1 ej\varphi 1\ \delta(t) + \alpha 2 ej\varphi 2\ \delta(t - \tau) \qquad (5.67)$$

Where α1 and α2 are independent Rayleigh distributed and φ1 and φ2 are independent and uniformly distributed over 0 to 2π. By varying τ it is possible to create a wide range of frequency selective fading effects.

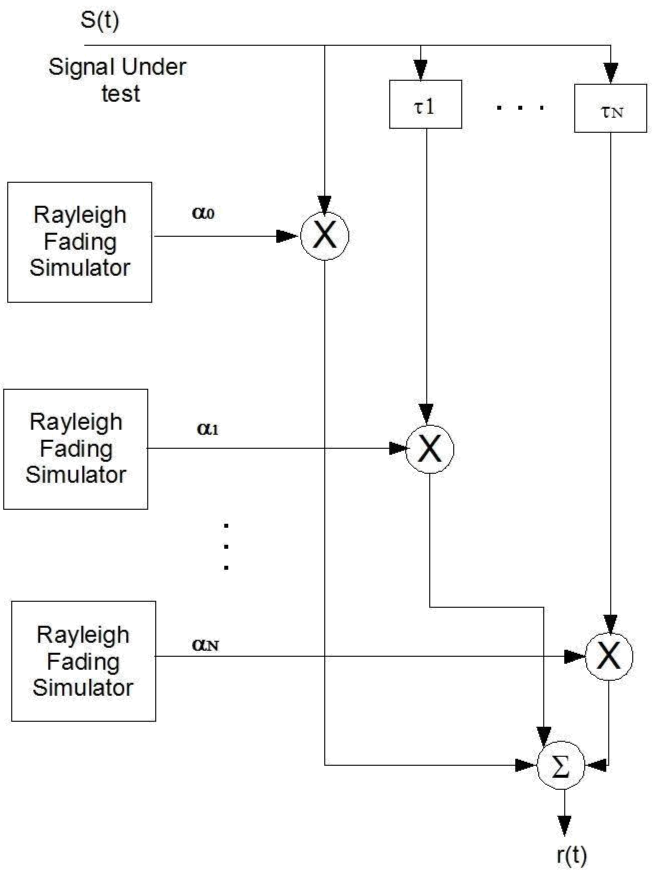


Figure 5.12: Rayleigh fading model to get both the flat and frequency selective channel conditions.

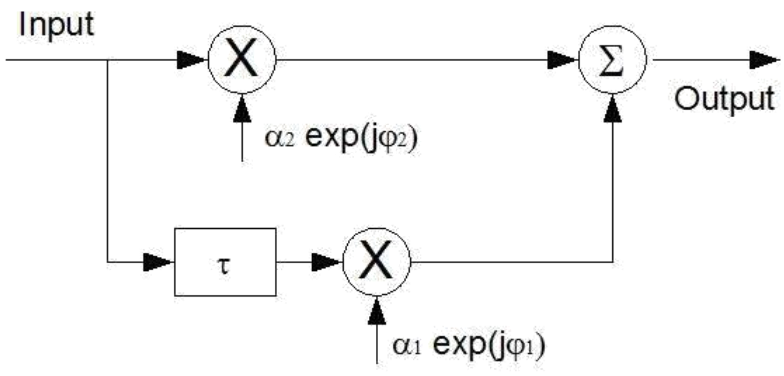
Saleh and Valenzuela Indoor Statistical Model

This method involved averaging the square law detected pulse response while sweeping 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.



Two-ray Rayleigh fading model.

A Few Practical Problems in Hando Scenario

Di erent speed of mobile users: with the increase of mobile users in urban areas, microcells are introduced in the cells to increase the capacity (this will be discussed later in this chapter). The users with high speed frequently crossing the micro-cells become burdened to MSC as it has to take care of hando.

Several schemes thus have been designed to handle the simultaneous tra c of high speed and low speed users while minimizing the hando intervention from the MSC, one of them being the `Umbrella Cell' approach. This technique provides large area coverage to high speed users while providing small area coverage to users traveling at low speed. By using di erent antenna heights and di erent power levels, it is possible to provide larger and smaller cells at a same location. As illustrated in the Figure 3.6, umbrella cell is co-located with few other microcells.

The BS can measure the speed of the user by its short term average signal strength over the RVC and decides which cell to handle that call. If the speed is less, then the corresponding microcell handles the call so that there is good corner coverage. This approach assures that hando is minimized for high speed users and provides additional microcell channels for pedestrian users. (b) Cell dragging problem: this is another practical problem in the urban area with additional microcells. For example, consider there is a LOS path between the MS and

BS1 while the user is in the cell covered by BS2. Since there is a LOS with the BS1, the signal strength received from BS1 would be greater than that received from BS2. However, since the user is in cell covered by BS2, hando cannot take place and as a result, it experiences a lot of interferences. This problem can be solved by judiciously choosing the hando threshold along with adjusting the coverage areas. Several schemes thus have been designed to handle the simultaneous trace of high speed and low speed users while minimizing the hando intervention from the MSC, one of them being the `Umbrella Cell'.

This technique provides large area coverage to high speed users while providing small area coverage to users traveling at low speed. By using decrement antenna heights and di errant power levels, it is possible to provide larger and smaller cells at a same location. As illustrated in the Figure 3.6, umbrella cell is co-located with few other microcells. The BS can measure the speed of the user by its short term average signal strength over the RVC and decides which cell to handle that call. If the speed is less, then the corresponding microcell handles the call so that there is good corner coverage.

This approach assures that hands are minimized for high speed users and provides additional microcell channels for pedestrian users. (c) Cell dragging problem: this is another practical problem in the urban area with additional microcells. For example, consider there is a LOS path between the MS and BS1 while the user is in the cell covered by BS2. Since there is a LOS with the BS1, the signal strength received from BS1 would be greater than that received from BS2. However, since the user is in cell covered by BS2, hand-out cannot take place and as a result, it experiences a lot of interferences. This problem can be solved by judiciously choosing the hando threshold along with adjusting the coverage areas.

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Common channel signalling (CCS) is a digital communications technique that provides simultaneous transmission of user data, signalling data, and other related traffic throughout a network. This is accomplished by using out-of--band signalling channels which logically separate the network data from the user information (voice or data) on the same channel. For second generation wireless communications systems, CCS is used to pass user data and control/supervisory signals between the subscriber and the base station, between the base station and the MSC, and between MSCs.

Even though the concept of CCS implies dedicated, parallel channels, it is implemented in a TDM format for serial data transmissions. Before the introduction of CCS in the 1980s, signaling traffic between the MSC and a subscriber was carried in the same band as the end-user's audio.

The network control data passed between MSCs in the PSTN was also carried in band, requiring that network information be carried within the same channel as the subscriber's voice traffic throughout the PSTN. This technique, called inland signalling, reduced the capacity of the PSTN, since the network signalling data rates were greatly constrained by the limitations of the carried voice channels, and the PSTN was forced to sequentially (not simultaneously) handle signalling and user data for each call. CCS is an out-of-band signalling technique which allows much faster communications between two nodes within the PSTN.

Instead of being constrained to signalling data rates which are on the order of audio frequencies, CCS supports signalling data rates from 56 kbps to many megabits per second. Thus, network signalling data is carried in a seemingly parallel, out-of band, signalling channel while only user data is carried on the PSTN. CCS provides a substantial increase in the number of users which are served by trunked PSTN lines, but requires that a dedicated portion of the trunk time be used to provide a signalling channel used for network traffic. In first generation cellular systems, the SS7 family of protocols, as defined by the Integrated System Digital Network (ISDN) are used to provide CCS.

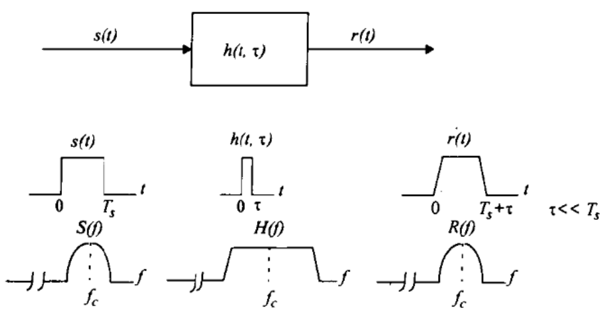
Coherence Bandwidth

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

Doppler Spread and Coherence Time

Delay spread and coherence bandwidth is parameters which describe the time dispersive nature of the channel in a local area. However, they do not offer information about the time varying nature of the channel caused by either relative motion between the mobile and base station, or by movement of objects in the channel. Doppler spread and coherence time are parameters which describe the time varying nature of the channel in a small-scale region. Doppler spread BD is a measure of the spectral broadening caused by the time rate of change of the mobile radio channel and is defined as the range of frequencies over which the received Doppler spectrum is essentially non-zero.

When a pure sinusoidal tone of frequency is transmitted, the received signal spectrum, called the Doppler spectrum, will have components in the range to where fd is the Doppler shift. The amount of spectral broadening depends on which is a function of the relative velocity of the mobile, and the angle 9 between the direction of motion of the mobile and direction of arrival of the scattered waves. If the baseband signal bandwidth is much greater than BD, the effects of Doppler spread are negligible at the receiver. This is a slow fading channel. Coherence time T, is the time domain dual of Doppler spread and is used to characterize the time varying nature of the frequency depressiveness of the channel in the time domain. The Doppler spread and coherence time are inversely proportional to one another. That is,



Flat fading channel characteristics.

Coherence Bandwidth

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

It is important to note that an exact relationship between coherence band- width and rms delay spread does not exist, and equations (4.38) and (4.39) are "ball park estimates". In general, spectral analysis techniques and simulation are required to determine the exact impact that time varying multipath has on a particular transmitted signal. For this reason, accurate multipath channel models must be used in the design of specific modems for wireless applications.

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

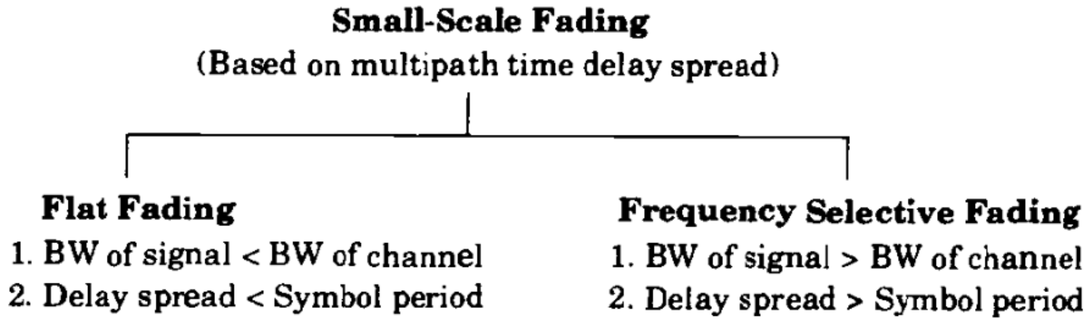
Coherence tine is actually a statistical measure of the time duration over which the channel impulse response is essentially invariant, and quantifies the similarity of the channel response at different times. In other words, coherence time is the time duration over which two received signals have a strong potential for amplitude correlation. If the reciprocal bandwidth of the baseband signal is greater than the coherence time of the channel, then the channel will change during the transmission of the baseband message, thus causing distortion at the receiver. If the coherence time is defined as the time over which the time correlation function is above 0.5, then the coherence time is approximately

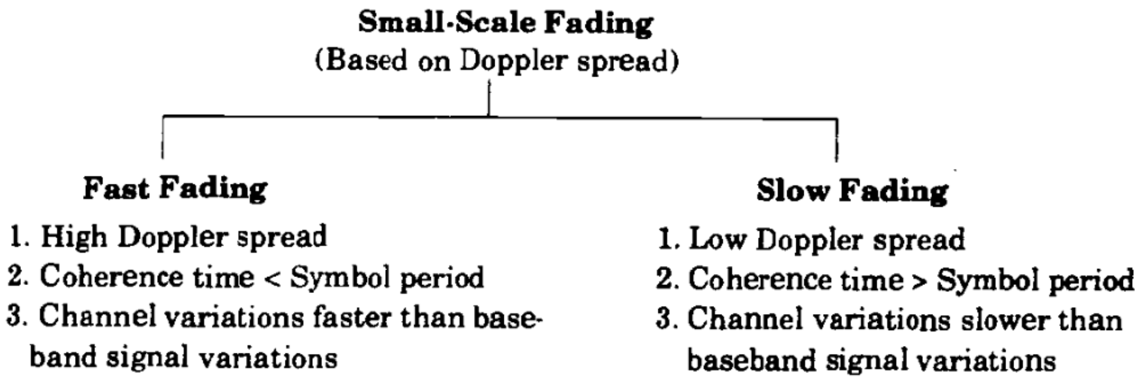
$$T_C \approx \frac{9}{16\pi f_m}$$

The definition of coherence time implies that two signals arriving with a time separation greater than are affected differently by the channel. For example, for a vehicle traveling 60 mph using a 900 MHz carrier, a conservative value of can be shown to be 2.22 ms from (4.40.b). If a digital transmission system is used, then as long as the symbol rate is greater than I= 454 bps, the channel will not cause distortion due to motion (however, distortion could result from multipath time delay spread, depending on the channel impulse response). Using the practical formula of (4.40.c), = 6.77 ms and the symbol rate must exceed 150 bits/s in order to avoid distortion due to frequency dispersion.

Types of Small-Scale Fading

Demonstrated that the type of fading experienced by a signal propagating through a mobile radio channel depends on the nature of the transmitted signal with respect to the characteristics of the channel. Depending on the relation between the signal parameters (such as bandwidth, symbol period, etc.) and the channel parameters (such as rms delay spread and Doppler spread), different transmitted signals will undergo different types of fading. The time dis persion and frequency dispersion mechanisms in a mobile radio channel lead to four possible distinct effects, which are manifested depending on the nature of the transmitted signal, the channel, and the velocity. While multipath delay spread leads to time dispersion and frequency selective fading, Doppler spread leads to frequency dispersion and time selective fading. The two propagation mechanisms are independent of one another.





Flat fading

If the mobile radio channel has a constant gain and linear phase response over a bandwidth which is greater than the bandwidth of the transmitted signal, then the received signal will undergo flat fading. This type of fading is historically the most common type of fading described in the technical literature. In flat fading, the multipath structure of the channel is such that the spectral characteristics of the transmitted signal are preserved at the receiver. However the strength of the received signal changes with time, due to fluctuations in the gain of the channel caused by multipath.

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. Under such conditions the channel impulse response has a multipath delay spread which is greater than the reciprocal bandwidth of the transmitted message waveform. When this occurs, the received signal includes multiple versions of the transmitted waveform which are attenuated (faded) and delayed in time, and hence the received signal is distorted. Frequency selective fading is due to tine dispersion of the transmitted symbols within the channel. Thus the channel induces inter symbol interference (IS!). Viewed in the frequency domain, certain frequency components in the received signal spectrum have greater gains than others.

Fading Effects Due to Doppler Spread

Depending on how rapidly the transmitted baseband signal changes as compared to the rate of change of the channel, a channel may be classified either as a fast fading or slow fading channel. In a fast fading channel, the channel impulse response changes rapidly within the symbol duration. That is, the coherence time of the channel is smaller than the symbol period of the transmitted signal. This causes frequency dispersion (also called time selective fading) due to Doppler spreading, which leads to signal distortion.

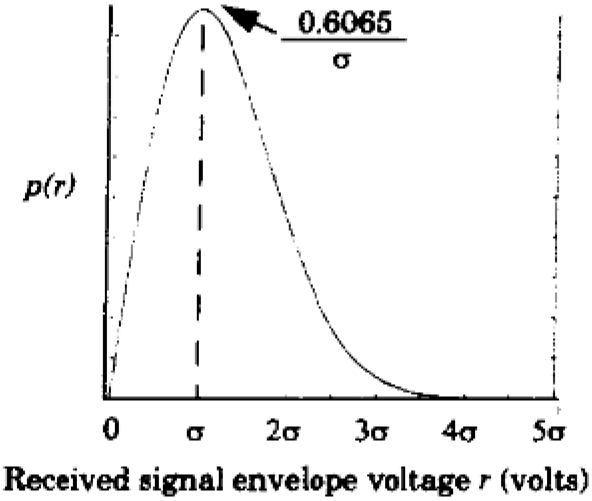
Viewed in the frequency domain, signal distortion due to fast fading increases with increasing Doppler spread relative to the bandwidth of the transmitted signal. Therefore, a signal undergoes fast fading if It should be noted that when a channel is specified as a fast or slow fading channel, it does not specify whether the channel is flat fading or frequency selective in nature. Fast fading only deals with the rate of change of the channel due to motion. En the case of the flat fading channel, we can approximate the impulse response to be simply a delta function (no time delay).

Hence, a flat fading, fast fading channel is a channel in which the amplitude of the delta function varies faster than the rate of change of the transmitted baseband signal. In the case of a frequency selective, fast fading channel, the amplitudes, phases, and time delays of any one of the multipath components vary faster than the rate of change of the transmitted signal. In practice, fast fading only occurs for very low data rates.

Slow Fading

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

Thus the mean and the median differ by only 0.55 dB in a Rayleigh fading signal. Note that the median is often used in practice, since fading data are usually measured in the field and a particular distribution cannot be assumed. By using median values instead of mean values it is easy to compare different fading distributions which may have widely varying means. Figure the Rayleigh pdf. The corresponding Rayleigh cumulative distribution function (CDF) is shown in Figure.



Saleh and Valenzuela Indoor Statistical Model

The results obtained by Saleh and Valenzula show that: (a) the indoor channel is quasi-static or very slowly time varying, and (b) the statistics of the channel impulse response are independent of transmitting and receiving antenna polarization, if there is no line-of-sight path between them. They reported a maximum multipath delay spread of 100ns to 200 ns within the rooms of a building, and 300 ns in hallways. The measured mis delay spread within rooms had a median of 25 ns and a maximum of 50 ns. The large-scale path loss with of-sight path was found to vary over a 60 dB range and obey a log distance power law (see equation (3.68)) with an exponent between 3 and 4.

Saleh and Valenzuela developed a simple multipath model for indoor channels based on measurement results. The model assumes that the multipath components arrive in clusters. The amplitudes of the received components are independent Rayleigh random variables with variances that decay exponentially with cluster delay as well as excess delay within a cluster. The corresponding phase angles are independent uniform random variables no lineover [0,2it}. The clusters and multipath components within a cluster form Poisson arrival processes with different rates. The clusters and multipath components within have exponentially distributed interracial times. The formation of the clusters is related to the building structure, while the components within the cluster formed by multiple reflections from objects in the vicinity of the transmitter and the receiver.

UNIT-IV

EQUALIZATION AND DIVERSITY

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 company 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 equalizer tracks the changing channel.
- As a result, the adaptive equalizer continuously changes the filter characteristics over time.

A Mathematical Framework

The signal received by the equalizer is given by

$$x(t) = d(t) \quad h(t) + nb(t) \qquad (7.1)$$

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 $nb(t)$ denotes the baseband noise.If the impulse response of the equalizer is $heq(t)$, the output of the equalizer is

$$\hat{y}(t) = d(t) \quad h(t) \quad heq(t) + nb(t) \quad heq(t) = d(t) \quad g(t) + nb(t) \quad heq(t) . \qquad (7.2)$$

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

The main goal of any equalization primary. In frequency domain it can be written as(7.3)

$$Heq(f)\, H(f) = 1 \qquad (7.4)$$

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 cn 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 $Heq(f)$ 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

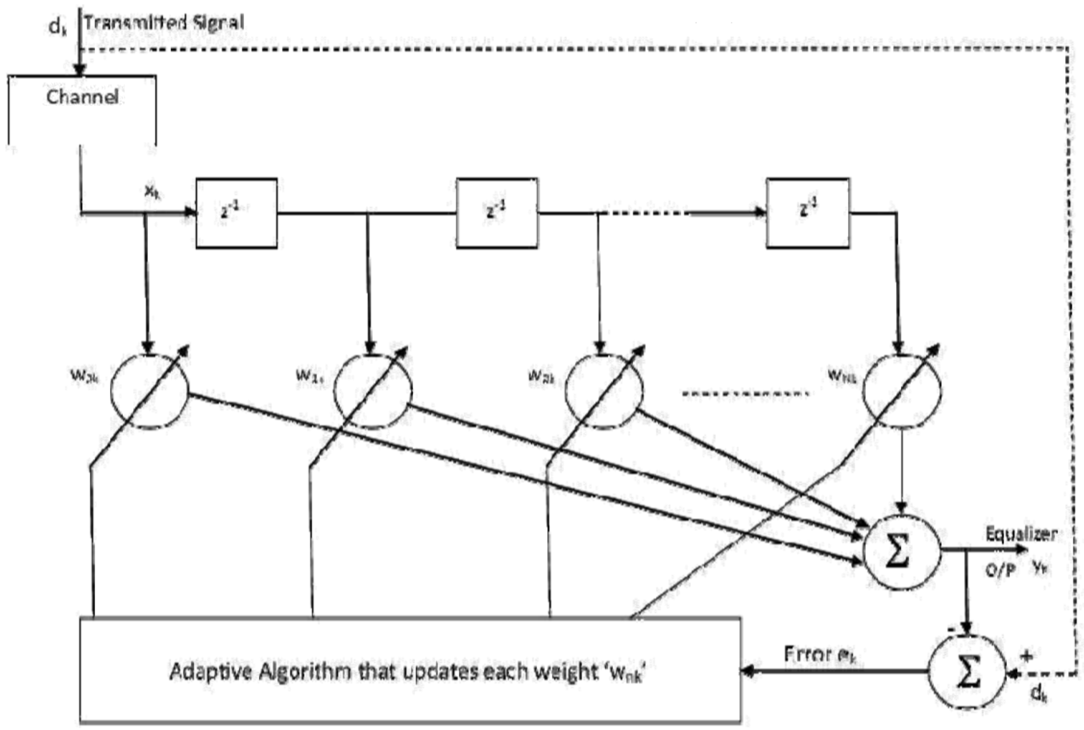
$$Hch(f)\, Heq(f) = 1, \; |f| < {}^{1}/{2T} \qquad (7.5)$$

where $Hch(f)$ 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 $Heq(f)$ is inverse of $Hch(f)$ so inverse filter may excessively amplify the noise at frequencies where the folded channel spectrum has high attenuation, 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 7.2. This filter is called the transversal filter, and in this case has N delay elements, N+1 taps and N+1 tenable 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 *ek*.



Assuming *dk* and x*k* to be jointly stationary, the Mean Square Error (MSE) is given as

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 complete 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 standards. These include the Zero Forcing Algorithm (ZF), the Least Mean Square Algorithm (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 periteration. The filter weights are updated by the update equation. Letting the variable n denote the sequence of iteration, LMS is computed iteratively by

$$w_k(n+1) = w_k(n) + \mu e_k(n)\, x(n-k) \quad (7.16)$$

where the subscript k denotes the kth 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 where λi 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

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 that the 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.

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. Figure7.3 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 are 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 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 have 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 .The probability that any single branch has an instantaneous SNR less than some defined threshold γ is Similarly, the probability that all M independent diversity branches receive signals which are simultaneously less than some specific SNR threshold γ is $= P^M(\gamma)$ (7.23) where $PM(\gamma)$ is the probability of all branches failing to achieve an instantaneous SNR = γ. Quite clearly, $PM(\Gamma) < P(\Gamma)$. If a single branch achieves SNR > γ, then the probability that SNR > γ for one or more branches is given by 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 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 requiring 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.

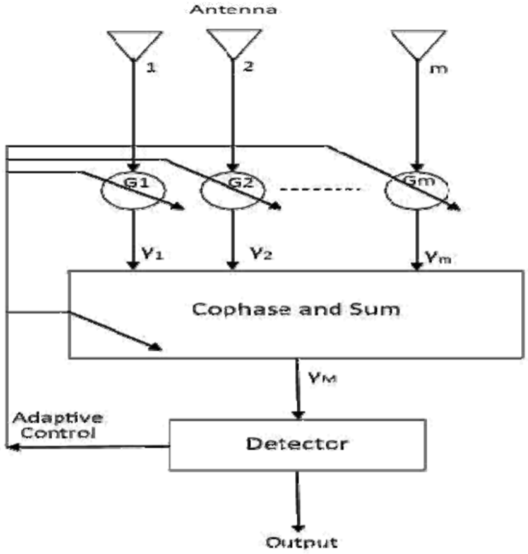
(b) Feedback or Scanning Diversity:

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,

(d) l 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 maximal ratio combining and superior to Selection



Polarization Diversity

Polarization Diversity relies on the de correlation of the two receive ports to achieve diversity gain. The two receiver ports must remain cross-polarized. Polarization Diversityat 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 communication 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 is sent over the same channel at different times. Time diversity repeatedly transmits information at time spacings that exceeds the coherence .

A modern implementation of time diversity involves the use of RAKE receiver for spread spectrum CDMA, where the multipath channel provides redundancy 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 are 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 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

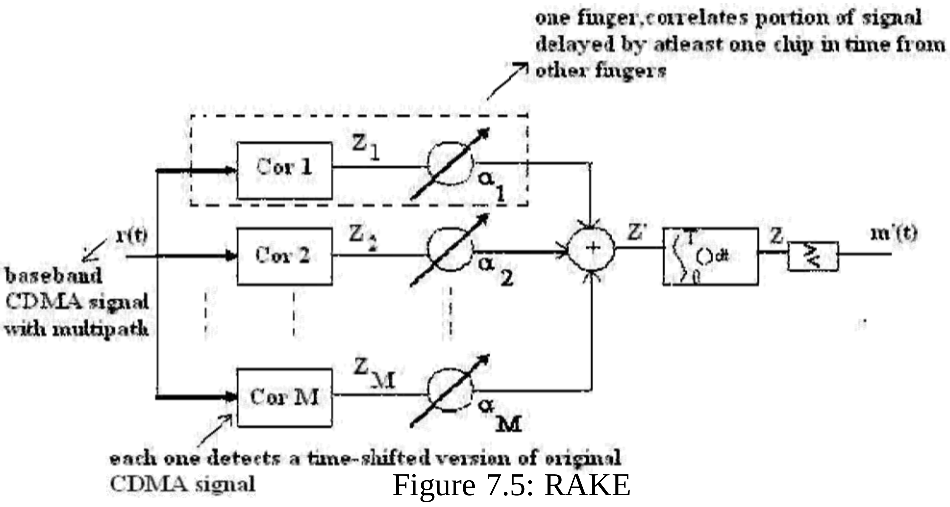


Figure 7.5: RAKE receiver.

Receiver collects the time shifted versions of the original signal by providing a separate correlation receiver for M strongest multipath components. Outputs of each correlate 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 correlates. Schematic of a RAKE receiver is shown in Figure 7.5.

UNIT-V

WIRELSEES NETWORKS

Introduction to IEEE 802.11x Technologies

802.11X authentication involve three parties: a supplicant, an authenticator, and an authentication server. The supplicant is a client device (such as a laptop) that wishes to attach to the LAN/WLAN - though the term 'supplicant' is also used interchangeably to refer to the software running on the client that provides credentials to the authenticator. The authenticator is a network device, such as an Ethernet switch or wireless access point; and the authentication server is typically a host running software supporting the RADIUS and EAP protocols.

The authenticator acts like a security guard to a protected network. The supplicant (i.e., client device) is not

Allowed access through the authenticator to the protected side of the network until the supplicant’s identity has been validated and authorized. An analogy to this is providing a valid visa at the airport's arrival immigration before being allowed to enter the country. With 802.1X port-based authentication, the supplicant provides credentials, such as user name / password or digital certificate, to the authenticator, and the authenticator forwards the credentials to the authentication server for verification. If the authentication server determines the credentials are valid, the supplicant (client device) is allowed to access resources located on the protected side of the network.

Evolution of Wireless LANs

Wireless LANs have gone through rapid changes with respect to their security architecture in recent years. One view has been to incorporate WLANs under already existing VPN umbrellas and to view them merely as an alternative access method --- thus preserving existing VPN infrastructure. Another view has been to address the security of the airwaves which has been demonstrated to be extremely vulnerable. The evolution of security standardisation based upon the work of the IEEE has evolved from WEP to WPA which introduced new key management and integrity mechanisms through to WAP2 (IEEE 802.11i) which maintains the management and integrity mechanisms of WPA but introduces AES encryption as well as moving much of the security functionality to the hardware. This paper traces the evolution and development of this new WLAN security architecture. Initialization On detection of a new supplicant, the port on the switch (authenticator) is enabled and set to the "unauthorized" state. In this state, only 802.1X traffic is allowed; other traffic, such as the Internet Protocol (and with that TCP and UDP), is dropped.

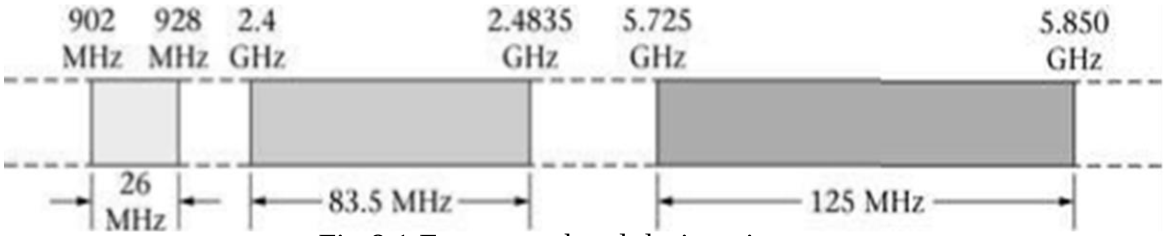


Fig 8.1 Frequency band designation

- Extensions to 802.11
  - 802.11b/a/g
  - 802.11d
  - 802.11e
  - 802.11f
  - 802.11h
- Extensions to 802.11
  -
  - 802.11i
  - 802.11j
  - 802.11k
  - 802.11ma
  - 802.11n
- Extensions to 802.11
  - 802.11p
  - 802.11r
  - 802.11s
  - 802.11u
  - 802.11v
- Layer 1: Overview
  - WLAN radio cards
  - WLAN access points
  - Ad hoc or peer-to-peer connection
  - WLAN radio link

Introduction to 802.15X technologies in PAN applications and architecture.

Bluetooth is a wireless technology standard for exchanging data over short distances (using short-wavelength radio transmissions in the ISM band from 2400–2480 MHz) from fixed and mobile devices, creating personal area networks (PANs) with high levels of security. Created by telecom vendor Ericsson in 1994, it was originally conceived as a wireless alternative to RS-232 data cables. It can connect several devices, overcoming problems of synchronization.

Bluetooth uses a radio technology called frequency-hopping spread spectrum, which chops up the data being sent and transmits chunks of it on up to 79 bands (1 MHz each; cantered from 2402 to 2480 MHz) in the range 2,400– 2,483.5 MHz (allowing for guard bands). This range is in the globally unlicensed Industrial, Scientific and Medical (ISM) 2.4 GHz short- range radio frequency band. It usually performs 800 hops per second, with Adaptive Frequency-Hopping (AFH) enabled.[9]

Communication and connection

A master Bluetooth device can communicate with a maximum of seven devices in a picante (an ad-hoc computer network using Bluetooth technology), though not all devices reach this maximum. The devices can switch roles, by agreement, and the slave can become the master (for example, a headset initiating a connection to a phone will necessarily begin as master, as initiator of the connection; but may subsequently prefer to be slave).

Uses:

Bluetooth is a standard wire-replacement communications protocol primarily designed for low power consumption, with a short range (power-class-dependent, but effective ranges vary in practice; see table below) based on low-cost transceiver microchips in each device. Because the devices use a radio (broadcast) communications system, they do not have to be in visual line of sight of each other, however a quasi-optical wireless path must be viable

Bluetooth profiles

To use Bluetooth wireless technology, a device has to be able to interpret certain Bluetooth profiles, which are definitions of possible applications and specify general behaviours that Bluetooth enabled devices use to communicate with other Bluetooth devices. These profiles include settings to parameterise and to control the communication from start. Adherence to profiles saves the time for transmitting the parameters anew before the bi-directional link becomes effective. There are a wide range of Bluetooth profiles that describe many different types of applications or use cases for devices.

A typical Bluetooth mobile phone headset.

- Wireless control of and communication between a mobile phone and a hands free headset. This was one of the earliest applications to become popular.
- Replacement of previous wired RS-232 serial communications in test equipment, GPS receivers, medical equipment, bar code scanners, and traffic control devices.

Bluetooth vs. Wi-Fi (IEEE 802.11)

Bluetooth and Wi-Fi (the brand name for products using IEEE 802.11 standards) have some similar applications: setting up networks, printing, or transferring files. Wi-Fi is intended as a replacement for cabling for general local area network access in work areas. This category of applications is sometimes called wireless local area networks (WLAN). Bluetooth was intended for portable equipment and its applications. The category of applications is outlined as the wireless personal area network (WPAN). Bluetooth is a replacement for cabling in a variety of personally carried applications in any setting and also works for fixed location applications such as smart energy functionality in the home (thermostats, etc.).

Wi-Fi is a wireless version of a common wired Ethernet network, and requires configuration to set up shared resources, transmit files, and to set up audio links (for example, headsets and hands-free devices). Wi-Fi uses the same radio frequencies as Bluetooth, but with higher power, resulting in higher bit rates and better range from the base station. The nearest equivalents in Bluetooth are the DUN profile, which allows devices to act as modem interfaces, and the PAN profile, which allows for ad-hoc networking

Bluetooth protocols simplify the discovery and setup of services between devices.[20] Bluetooth devices can advertise all of the services they provide. This makes using services easier because more of the security, network address and permission configuration can be automated than with many other network types

Air interface

The protocol operates in the license-free ISM band at 2.402–2.480 GHz. To avoid interfering with other protocols that use the 2.45 GHz band, the Bluetooth protocol divides the band into 79 channels (each 1 MHz wide) and changes channels, generally 800 times per second. Implementations with versions 1.1 and 1.2 reach speeds of 723.1

kbit/s. Version 2.0 implementations feature Bluetooth Enhanced Data Rate (EDR) and reach Mbit/s. Technically, version 2.0 devices have a higher power consumption, but the three times faster rate reduces the transmission times, effectively reducing power consumption to half that of 1.x devices

Because ZigBee nodes can go from sleep to active mode in 30 ms or less, the latency can be low and devices can be responsive, particularly compared to Bluetooth wake-up delays, which are typically around three seconds. Because ZigBee nodes can sleep most of the time, average power consumption can be low, resulting in long battery life.

Application profiles

The current list of application profiles either published, or in the works are:

- Released specifications
- ZigBee Home Automation
- ZigBee Smart Energy 1.0
- ZigBee Telecommunication Services
- ZigBee Health Care
- ZigBee RF4CE – Remote Control
- ZigBee RF4CE – Input Device
- ZigBee Light Link
- Specifications under development
- ZigBee Smart Energy 2.0
- ZigBee Building Automation
- ZigBee Retail Services
- HomeGrid Forum responsible for marketing and certifying ITU-T G.hn technology and products
- HomePlug Powerline Alliance
- International Society of Automotive Engineers SAE International
- IPSO Alliance
- SunSpec Alliance
- Wi-Fi Alliance.

In 2009 the RF4CE (Radio Frequency for Consumer Electronics) Consortium and ZigBee Alliance agreed to jointly deliver a standard for radio frequency remote controls. ZigBee RF4CE is designed for a wide range of consumer electronics products, such as TVs and set-top boxes. It promises many advantages over existing remote control solutions, including richer communication and increased reliability, enhanced features and flexibility, Interoperability and no line-of-sight barrier. The ZigBee RF4CE specification lifts off some networking weight and does not support all the mesh features, which is traded for smaller memory configurations for lower cost devices, such as remote control of consumer electronics. With the introduction of second Zigbee RF4CE application profile in 2012, and increased momentum in MSO market, Zigbee RF4CE team provided an overview on current status of standard, applications, and future of the technology.

Configurable functionality

A number of network properties can be pre-configured. The network is initialised by the Co-ordinator, at which time these configuration values are taken into account. These properties determine the maximum size (in terms of the maximum number of nodes) and shape of the network, and are as follows:

Network Depth: The depth of a device in a network is the number of nodes from the root of the network tree (the Co-ordinator) to the device. The maximum network depth is then the maximum number of hops from the Co- coordinator to the most distant device in the network. This determines the overall diameter for the network. Note that a Star network has a network depth of 1.

Number of Children: Each Router in the network can have a number of child devices attached to it. These may be either Routers or End Devices. The Co-ordinator specifies the maximum number of child devices allowed per Router. Number of Child Routers: In addition to the number of children per Router, a limit is put on how many of these children may be Routers themselves. The Co-ordinator uses the above information during initialisation to allocate blocks of network addresses to the branches of the network tree. In turn, the Routers use it to allocate subsets of these address blocks to their children.

Forming a ZigBee Network: The Co-ordinator is responsible for starting a ZigBee network. Network initialisation involves the following steps:Search for a Radio Channel

- The Co-ordinator first searches for a suitable radio channel (usually the one which has least activity). This search can be limited to those channels that are known to be usable - for example, by avoiding frequencies in which it is known that a wireless LAN is operating. Assign PAN ID
- The Co-ordinator starts the network, assigning a PAN ID (Personal Area Network identifier) to the network. The PAN ID can be pre-determined, or can be obtained

dynamically by detecting other networks operating in the same frequency channel and choosing a PAN ID that does not conflict with theirs. At this stage, the Co-ordinator also assigns a network (short) address to itself. Usually, this is the address 0x0000.

- Start the Network

The Co-ordinator then finishes configuring itself and starts itself in Co-ordinator mode. It is then ready to respond to queries from other devices that wish to join the network.Joining a ZigBee Network: Once the network has been created by the Co-ordinator, other devices (Routers and End Devices) can join the network. Both Routers and the Co- coordinator have the capability to allow other nodes to join the network. The join process is as follows:

- Search for Network

The new node first scans the available channels to find operating networks and identifies which one it should join. Multiple networks may operate in the same channel and are differentiated by their PAN IDs.

- Select Parent

The node may be able to ‘see’ multiple Routers and a Co-ordinator from the same network, in which case it selects which one it should connect to. Usually, this is the one with the best signal.

- Send Join Request

The node then sends a message to the relevant Router or Co-ordinator asking to join the network.

- Accept or Reject Join Request

The Router or Co-ordinator decides whether the node is a permitted device, whether the Router/Co-ordinator is currently allowing devices to join and whether it has address space available. If all these criteria are satisfied, the Router/Co-ordinator will then allow the device to join and allocate it an address. Typically, a Router or Co-ordinator can be configured to have a time-period during which joins are allowed. The join period may be initiated by a user action, such as pressing a button. An infinite join period can be set, so that child nodes can join the parent node at any time.

Message Propagation: The way that a message propagates through a ZigBee network depends on the network topology. However, in all topologies, the message usually needs to pass through one or more intermediate nodes before reaching its final destination. The message therefore contains two destination addresses:

- Address of the final destination
- Address of the node which is the next “hop”

The way these addresses are used in message propagation depends on the network topology, as follows:

- Star Topology: All messages are routed via the Co-ordinator. Both addresses are needed and the “next hop” address is that of the Co-ordinator.
- Tree Topology: A message is routed up the tree until it reaches a node that can route it back down the tree to the destination node. Both addresses are needed and the initial “next hop” address is that of the parent of the sending node. The parent node then resends the message to the next relevant node - if this is the target node itself, the “final destination” address is used. The last step is then repeated and message propagation continues in this way until the target node is reached.

Route Discovery: The ZigBee stack network layer supports a “route discovery” facility in which a mesh network can be requested to find the best available route to the destination, when sending a message. Route discovery is initiated when requested by a data transmission request.

Route Discovery Options There are three options related to route discovery for a mesh network (the required option being indicated in the message):

- SUPPRESS route discovery: The message is routed along the tree.
- ENABLE route discovery: The message is routed along an already discovered mesh route, if one exists, otherwise the Router initiates a route discovery. Once this is complete, the message will be sent along the calculated route. If the Router does not have the capacity to store the new route, it will direct the message along the tree.

FORCE route discovery: If the Router has the route capacity, it will initiate a route discovery, even if a known route already exists. Once this is complete, the message will be sent along the calculated route. If the Router does not have the route capacity, it will route the message along the tree. Use of this option should be restricted, as it generates a lot of network traffic.

Route Discovery Mechanism: The mechanism for route discovery between two End

Devices involves the following steps:

A route discovery broadcast is sent by the parent Router of the source End Device. This broadcast contains the network address of the destination End Device.

All Routers eventually receive the broadcast, one of which is the parent of the destination End Device. The parent Router of the destination node sends back a reply addressed to the parent Router of the source.

- As the reply travels back through the network, the hop count and a signal quality measure for each hop are recorded. Each Router in the path can build a routing table entry containing the best path to the destination End Device.
- Eventually, each Router in the path will have a routing table entry and the route from source to destination End Device is established. Note that the corresponding route from destination to source is not known – the route discovered is unidirectional.

The choice of best path is usually the one with the least number of hops, although if a hop on the most direct route has a poor signal quality (and hence a greater chance that retries will be needed), a route with more hops may be chosen.

There are two types of discovery, Device an Service Discovery:

Device Discovery: Device Discovery involves interrogating a remote node for address information. The retrieved information can be either:

- the MAC (IEEE) address of the node with a given network address
- the network address of the node with a given MAC address.