MOBILE COMMUNICATION AND NETWORKS

UNIT I CELLULAR CONCEPTS

Prepared by

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AIM & OBJECTIVES

- ❖ To understand the issues involved in mobile communication system design and analysis.
- ❖ To understand the concept of frequency reuse.
- ❖ To understand the characteristics of wireless channels.
- ❖ To know the fundamental limits on the capacity of wireless channels.

PRE TEST-MCQ TYPE

1. What is a cell in cellular system?
a) A group of cells
b) A group of subscribers
c) A small geographical area
d) A large group of mobile systems

2. Why the shape of cell is not circle?
a) Omni directionality
b) Small area
c) Overlapping regions or gaps are left
d) Complex design

3. What is the main reason to adopt hexagon shape in comparison to square and triangle?
a) Largest area
b) Simple design
c) Small area
d) Single directional

4. What is a cluster in a cellular system?
a) Group of frequencies
b) Group of cells
c) Group of subscribers
d) Group of mobile systems

5. Capacity of a cellular system is directly proportional to __________
a) Number of cells
b) Number of times a cluster is replicated
c) Number of Base stations
d) Number of users

UNIT I CELLULAR CONCEPTS - CONTENTS

Cellular concepts- Cell structure, frequency reuse, cell splitting, channel assignment, handoff, interference, capacity, power control; Wireless Standards: Overview of 2G and 3G cellular standards.

THEORY

Introduction

Communication is one of the integral parts of science that has always been a focus point for exchanging information among parties at locations physically apart. After its discovery, telephones have replaced the telegrams and letters. Similarly, the term `mobile' has completely revolutionized the communication by opening up innovative applications that are limited to one's imagination. Today, mobile communication has become the backbone of the society. All the mobile system technologies have improved the way of living. Its main plus point is that it has privileged a common mass of society. In this chapter, the evolution as well as the fundamental techniques of the mobile communication is discussed. The first wireline telephone system was introduced in the year 1877.

Mobile communication systems as early as 1934 were based on Amplitude Modulation (AM) schemes and only certain public organizations maintained such systems. With the demand for newer and better mobile radio communication systems during the World War II and the development of Frequency Modulation (FM) technique by Edwin Armstrong, the mobile radio communication systems began to witness many new changes. Mobile telephone was introduced in the year 1946. However, during its initial three and a half decades it found very less market penetration owing to high costs and numerous technological drawbacks. But with the development of the cellular concept in the 1960s at the Bell Laboratories, mobile communications began to be a promising field of expanse which could serve wider populations. Initially, mobile communication was restricted to certain official users and the cellular concept was never even dreamt of being made commercially available.

Moreover, even the growth in the cellular networks was very slow. However, with the development of newer and better technologies starting from the 1970s and with the mobile users now connected to the Public Switched Telephone Network (PSTN), there has been an astronomical growth in the cellular radio and the personal communication systems. Advanced Mobile Phone System (AMPS) was the first U.S. cellular telephone system and it was deployed in 1983. Wireless services have since then been experiencing a 50% per year growth rate. The number of cellular telephone users grew from 25000 in 1984 to around 3 billion in the year 2007 and the demand rate is increasing day by Day.

Mobile Telephony Development

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Cellular Concept:

The power of the radio signals transmitted by the BS decay as the signals travel away from it. A minimum amount of signal strength (say, x dB) is needed in order to be detected by the MS or mobile sets which may the hand-held personal units or those installed in the vehicles. The region over which the signal strength lies above this threshold value x dB is known as the coverage area of a BS and it must be a circular region, considering the BS to be isotropic radiator. Such a circle, which gives this actual radio coverage, is called the foot print of a cell (in reality, it is amorphous). It might so happen that either there may be an overlap between any two such side by side circles or there might be a gap between the coverage areas of two adjacent circles. This is shown in Figure. Such a circular geometry, therefore, cannot serve as a regular shape to describe cells. A regular shape for cellular design over a territory which can be served by 3 regular polygons, namely, equilateral triangle, square and regular hexagon, which can cover the entire area without any overlap and gaps is needed. Along with its regularity, a cell must be designed such that it is most reliable too, i.e., it supports even the weakest mobile with occurs at the edges of the cell. For any distance between the center and the farthest point in the cell from it, a regular hexagon covers the maximum area. Hence regular hexagonal geometry is used as the cells in mobile communication.

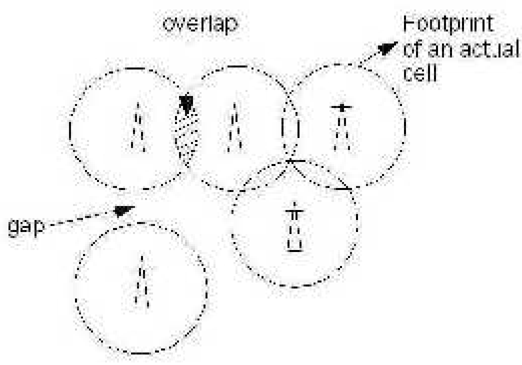

Figure Footprint of cells showing the overlaps and gaps.

Frequency Reuse:

Frequency reuse, or, frequency planning, is a technique of reusing frequencies and channels within a communication system to improve capacity and spectral efficiency. Frequency reuse is one of the fundamental concepts on which commercial wireless systems are based that involve the partitioning of an RF radiating area into cells. The increased capacity in a commercial wireless network, compared with a network with a single transmitter, comes from the fact that the same radio frequency can be reused in a different area for a completely different transmission. Frequency reuse in mobile cellular systems means that frequencies allocated to

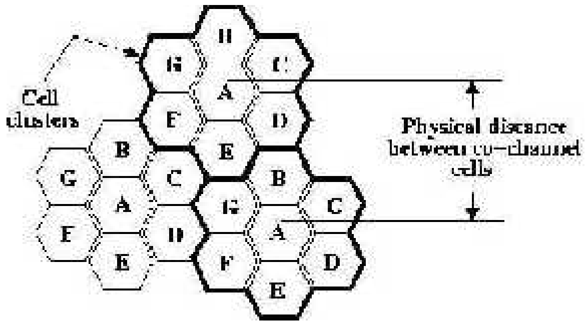

Figure Frequency reuse technique of a cellular system.

the services are reused in a regular pattern of cells, each covered by one base station. The repeating regular pattern of cells is called cluster. Since each cell is designed to use radio frequencies only within its boundaries, the same frequencies can be reused in other cells not far away without interference, in another cluster. Such cells are called `co-channel' cells. The reuse of frequencies enables a cellular system to handle a huge number of calls with a limited number of channels. Figure shows a frequency planning with cluster size of 7, showing the co-channels cells in different clusters by the same letter. The closest distance between the co-channel cells (in different clusters) is determined by the choice of the cluster size and the layout of the cell cluster. Consider a cellular system with S duplex channels available for use and let N be the number of cells in a cluster. If each cell is allotted K duplex channels with all being allotted unique and disjoint channel groups then S = KN under normal circumstances. Now, if the cluster are repeated M times within the total area, the total number of duplex channels, or, the total number of users in the system would be T = MS = KMN. Clearly, if K and N remain constant, then $T \propto M$ and, if T and K remain constant, then

$$N \propto \frac{1}{M}.$$

Hence the capacity gain achieved is directly proportional to the number of times a cluster is repeated, as shown, as well as, for a fixed cell size, small N =25 decreases the size of the cluster with in turn results in the increase of the number of clusters and hence the capacity. However for small N, co-channel cells are located much closer and hence more interference. The value of N is determined by calculating the amount of interference that can be tolerated for a sufficient quality communication.

Hence the smallest N having interference below the tolerated limit is used. However, the cluster size N cannot take on any value and is given only by the following equation

$$N - i^2 + ij + j^2, \qquad\qquad i \geq 0, j \geq 0,$$

Where i and j are integer numbers.

Channel Assignment Strategies

With the rapid increase in number of mobile users, the mobile service providers had to follow strategies which ensure the effective utilization of the limited radio spectrum. With increased capacity and low interference being the prime objectives, a frequency reuse scheme was helpful in achieving these objectives. A variety of channel assignment strategies have been followed to aid these objectives. Channel assignment strategies are classified into two types: fixed and dynamic, as discussed below.

Fixed Channel Assignment (FCA)

In fixed channel assignment strategy each cell is allocated a fixed number of voice channels. Any communication within the cell can only be made with the designated unused channels of that particular cell. Suppose if all the channels are occupied, then the call is blocked and subscriber has to wait. This is simplest of the channel assignment strategies as it requires very simple circuitry but provides worst channel utilization. Later there was another approach in which the channels were borrowed from adjacent cell if all of its own designated channels were occupied. This was named as borrowing strategy. In such cases the MSC supervises the borrowing process and ensures that none of the calls in progress are interrupted.

Dynamic Channel Assignment (DCA)

In dynamic channel assignment strategy channels are temporarily assigned for use in cells for the duration of the call. Each time a call attempt is made from a cell the corresponding BS requests a channel from MSC. The MSC then allocates a channel to the requesting the BS. After the call is over the channel is returned and kept in a central pool. To avoid co-channel interference any channel that in use in one cell can only be reassigned simultaneously to another cell in the system if the distance between the two cells is larger than minimum reuse distance. When compared to the FCA, DCA has reduced the likelihood of blocking and even increased the trunking capacity of the network as all of the channels are available to all cells, i.e., good quality of service. But this type of assignment strategy results in heavy load on switching center at heavy traffic condition.

Handoff Process

When a user moves from one cell to the other, to keep the communication between the user pair, the user channel has to be shifted from one BS to the other without interrupting the call, i.e., when a MS moves into another cell, while the conversation is still in progress, the MSC automatically transfers the call to a new FDD channel without disturbing the conversation. This process is called as handoff. A schematic diagram of handoff is given in Figure Processing of handoff is an important task in any cellular system.

Handoffs must be performed successfully and be imperceptible to the users. Once a signal level is set as the minimum acceptable for good voice quality (Prmin), then a slightly stronger level is chosen as the threshold (PrH)at which handoff has to be made, as shown in Figure.

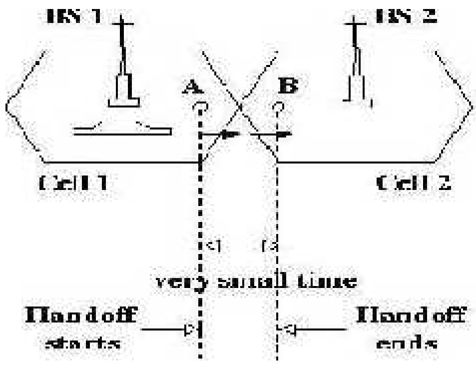

Figure Handoff scenario at two adjacent cell boundary.

A parameter, called power margin, defined as

$$\Delta = P_{r_H} \quad P_{r_{min}}$$

is quite an important parameter during the handoff process since this margin can neither be too large nor too small. If Δis too small, then there may not be enough time to complete the handoff and the call might be lost even if the user crosses the cell boundary. If Δ is too high o the other hand, then MSC has to be burdened with unnecessary handoffs. This is because MS may not intend to enter the other cell. Therefore Δ should be judiciously chosen to ensure imperceptible handoffs and to meet other objectives

Interference & System Capacity

Susceptibility and interference problems associated with mobile communications equipment are because of the problem of time congestion within the electromagnetic spectrum. It is the limiting factor in the performance of cellular systems. This interference can occur from clash with another mobile in the same cell or because of a call in the adjacent cell.

There can be interference between the base stations operating at same frequency band or any other non-cellular system's energy leaking inadvertently into the frequency band of the cellular system. If there is an interference in the voice channels, cross talk is heard will appear as noise between the users. The interference in the control channels leads to missed and error calls because of digital signaling. Interference is more severe in urban areas because of the greater RF noise and greater density of mobiles and base stations. The interference can be divided into 2 parts: co-channel interference and adjacent channel interference.

Co-channel interference (CCI)

For the efficient use of available spectrum, it is necessary to reuse frequency bandwidth over relatively small geographical areas. However, increasing frequency reuse also increases interference, which decreases system capacity and service quality. The cells where the same set of frequencies is used are call co-channel cells. Co-channel interference is the cross talk between two different radio transmitters using the same radio frequency as is the case with the co-channel cells. The reasons of CCI can be because of either adverse weather conditions or poor frequency planning or overly crowded radio spectrum. If the cell size and the power transmitted at the base stations are same then CCI will become independent of the transmitted power and will depend on radius of the cell (R) and the distance between the interfering co-channel cells (D). If D/R ratio is increased, then the effective distance between the co-channel cells will increase 34 and interference will decrease. The parameter Q is called the frequency reuse ratio and is related to the cluster size. For hexagonal geometry

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

From the above equation, small of `Q' means small value of cluster size `N' and increase in cellular capacity. But large `Q' leads to decrease in system capacity but increase in transmission quality. Choosing the options is very careful for the selection of `N', the proof of which is given in the first section. The Signal to Interference Ratio (SIR) for a mobile receiver which monitors the forward channel can be calculated as

$$\frac{S}{I} = \frac{S}{\sum_{i=1}^{i_0} I_i}$$

where i0 is the number of co-channel interfering cells, S is the desired signal power from the baseband station and Ii is the interference power caused by the i-th interfering co-channel base station. In order to solve this equation from power calculations, look into the signal power characteristics. The average power in the mobile radio channel decays as a power law of the distance of separation between transmitter and receiver. The expression for the received power Pr at a distance d can be approximately calculated as

$$P_r = P_0\left(\frac{d}{d_0}\right)^{-n}$$

and in the dB expression as

$$P_r(dB) = P_0(dB) \quad 10n\log\left(\frac{d}{d_0}\right)$$

where P0 is the power received at a close-in reference point in the far field region at a small distance do from the transmitting antenna, and `n' is the path loss exponent. Now calculate the SIR for this system. If Di is the distance of the i-th interferer from the mobile, the received power at a given mobile due to i-th interfering cell is proportional to (Di) n (the value of 'n' varies between 2 and 4 in urban cellular systems). Take that the path loss exponent is same throughout the coverage area and the transmitted power be same, then SIR can be approximated as

$$\frac{S}{I} = \frac{R^{-n}}{\sum_{i=1}^{i_0} D_i^{-n}}$$

where the mobile is assumed to be located at R distance from the cell center. Consider only the first layer of interfering cells and assume that the interfering base stations are equidistant from the reference base station and the distance between the cell centers is 'D' then the above equation can be converted as

$$\frac{S}{I} = \frac{(D/R)^n}{i_0} = \frac{(\sqrt{3N})^n}{i_0}$$

which is an approximate measure of the SIR. Subjective tests performed on AMPS cellular system which uses FM and 30 kHz channels show that sufficient voice quality can be obtained by SIR being greater than or equal to 18 dB. Take n= 4, the value of 'N' can be calculated. Therefore minimum N is 7. The above equations are based on hexagonal geometry and the distances from the closest interfering cells can vary if different frequency reuse plans are used for a more approximate calculation for co-channel SIR. This is the example of a 7 cell reuse case. The mobile is at a distance of D-R from 2 closest interfering cells and approximately D+R/2, D, D-R/2 and D+R distance from other interfering cells in the first tier. Taking n = 4 in the above equation, SIR can be approximately calculated as

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

which can be rewritten in terms frequency reuse ratio Q as

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

Using the value of N equal to 7 (this means Q = 4.6), the above expression yields that worst case SIR is 53.70 (17.3 dB). This shows that for a 7 cell reuse case the worst case SIR is slightly less than 18 dB. The worst case is when the mobile is at the corner of the cell i.e., on a vertex. Therefore N = 12 cluster size should be used. But this reduces the capacity by 7/12 times. Therefore, co-channel interference controls link performance, which in a way controls frequency reuse plan and the overall capacity of the cellular system. The effect of co-channel interference can be minimized by optimizing the frequency assignments of the base stations and their transmit powers. Tilting the base- station antenna to limit the spread of the signals in the system can also be done.

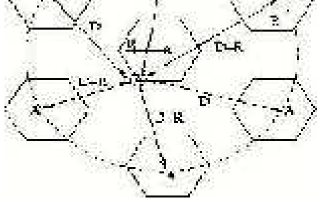

Figure First tier of co-channel interfering cells

Adjacent Channel Interference (ACI)

This is a different type of interference which is caused by adjacent channels i.e. channels in adjacent cells. It is the signal impairment which occurs to one frequency due to presence of another signal on a nearby frequency. This occurs when imperfect receiver filters allow nearby frequencies to leak into the pass band. This problem is enhanced if the adjacent channel user is transmitting in a close range compared to the subscriber's receiver while the receiver attempts to receive a base station on the channel. This is called near-far effect.

The more adjacent channels are packed into the channel block, the higher the spectral efficiency, provided that the performance degradation can be tolerated in the system link budget. This effect can also occur if a mobile close to a base station transmits on a channel close to one being used by a weak mobile. This problem might occur if the base station has problem in discriminating the mobile user from the "bleed over" caused by the close adjacent channel mobile. Adjacent channel interference occurs more frequently in small cell clusters and heavily used cells.

If the frequency separation between the channels is kept large this interference can be reduced to some extent. Thus assignment of channels is given such that they do not form a contiguous band of frequencies within a particular cell and frequency separation is maximized. Efficient assignment strategies are very much important in making the interference as less as possible. If the frequency factor is small then distance between the adjacent channels cannot put the interference level within tolerance limits. If a mobile is 10 times close to the base station than other mobile and has energy spill out of its pass band, then SIR for weak mobile is approximately

$$\frac{S}{I} = 10^{-n}$$

which can be easily found from the earlier SIR expressions. If n = 4, then SIR is 52 dB. Perfect base station filters are needed when close-in and distant users share the same cell. Practically, each base station receiver is preceded by a high Q cavity filter in order to remove adjacent channel interference. Power control is also very much important for the prolonging of the battery life for the subscriber unit but also reduces reverse channel SIR in the system.

Power control is done such that each mobile transmits the lowest power required to maintain a good quality link on the reverse channel.

Cell-Splitting

Cell Splitting is based on the cell radius reduction and minimizes the need to modify the existing cell parameters. Cell splitting involves the process of sub-dividing a congested cell into smaller cells, each with its own base station and a corresponding reduction in antenna size and transmitting power. This increases the capacity of a cellular system since it increases the number of times that channels are reused. Since the new cells have smaller radii than the existing cells, inserting these smaller cells, known as microcells, between the already existing cells results in an increase of capacity due to the additional number of channels per unit area.

There are few challenges in increasing the capacity by reducing the cell radius. Clearly, if cells are small, there would have to be more of them and so additional base stations will be needed in the system. The challenge in this case is to introduce the new base stations without the need to move the already existing base station towers. The other challenge is to meet the generally increasing demand that may vary quite rapidly between geographical areas of the system. For instance, a city may have highly populated areas and so the demand must be supported by cells with the smallest radius. The radius of cells will generally increase when moving from urban to sub urban areas, because the user density decreases on moving towards sub-urban areas. The key factor is to add as minimum number of smaller cells as possible wherever an increase in demand occurs.

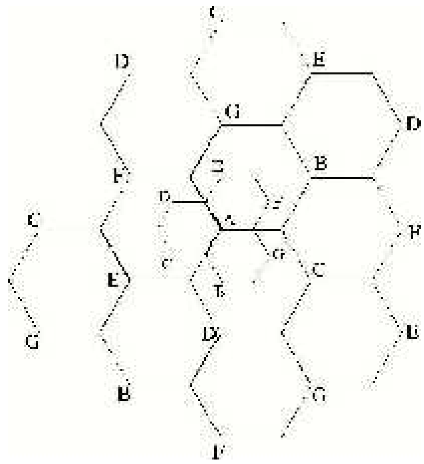

Figure Splitting of congested seven-cell clusters.

The gradual addition of the smaller cells implies that, at least for a time, the cellular system operates with cells of more than one size. Figure shows a cellular layout with seven-cell clusters. Consider that the cells in the center of the diagram are becoming congested, and cell A in the center has reached its maximum capacity.

Figure also shows how the smaller cells are being superimposed on the original layout. The new smaller cells have half the cell radius of the original cells. At half the radius, the new cells will have one-fourth of the area and will consequently need to support one- fourth the number of subscribers. Notice that one of the new smaller cells lies in the center of each of the larger cells. Assume that base stations are located in the cell centers, this allows the original base stations to be maintained even in the new system layout. However, new base stations will have to be added for new cells that do not lie in the center of the larger cells. The organization of cells into clusters is independent of the cell radius, so that the cluster size can be the same in the small-cell layout as it was in the large-cell layout.

Also the signal-to-interference ratio is determined by cluster size and not by cell radius. Consequently, if the cluster size is maintained, the signal-to-interference ratio will be the same after cell splitting as it was before. If the entire system is 41 replaced with new half- radius cells, and the cluster size is maintained, the number of channels per cell will be exactly as it was before, and the number of subscribers per cell will have been reduced. When the cell radius is reduced by a factor, it is also desirable to reduce the transmitted power. The transmit power of the new cells with radius half that of the old cells can be found by examining the received power PR at the new and old cell boundaries and setting them equal.

This is necessary to maintain the same frequency re-use plan in the new cell layout as well. Assume that PT1 and PT2 are the transmit powers of the larger and smaller base stations respectively. Then, assuming a path loss index n=4, the power received at old cell boundary = PT1=R4 and the power received at new cell boundary = PT2=(R=2)4. On equating the two received powers, PT2 = PT1 / 16.

In other words, the transmit power must be reduced by 12 dB in order to maintain the same S/I with the new system lay out. At the beginning of this channel splitting process, there would be fewer channels in the smaller power groups. As the demand increases, more and more channels need to be accommodated and hence the splitting process continues until all the larger cells have been replaced by the smaller cells, at which point splitting is complete within the region and the entire system is rescaled to have a smaller radius per cell. If a cellular layout is replaced entirety by a new layout with a smaller cell radius, the signal-to-interference ratio will not change, provided the cluster size does not change.

Some special care must be taken, however, to avoid co-channel interference when both large and small cell radii coexist. It turns out that the only way to avoid interference between the large-cell and small-cell systems is to assign entirely different sets of channels to the two systems. So, when two sizes of cells co-exist in a system, channels in the old cell must be broken down into two groups, one that corresponds to larger cell reuse requirements and the other which corresponds to the smaller cell reuse requirements. The larger cell is usually dedicated to high speed users as in the umbrella cell approach so as to minimize the number of hand- offs.

Sectoring

Sectoring is basically a technique which can increase the SIR without necessitating an increase in the cluster size. Till now, it has been assumed that the base station is located in the center of a cell and radiates uniformly in all the directions behaving as an omni- directional antenna. However it has been found that 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. In the Figure, a cell is shown which has been split into three 120^o sectors. The base station feeds three 120^o directional antennas, each of which radiates into one of the three sectors. The channel set serving this cell has also been divided, so that each sector is assigned one- third of the available number cell of channels. This technique for reducing co-channel interference wherein by using suit able directional antennas, a given cell would receive interference and transmit with a fraction of available co-channel cells is called 'sectoring'.

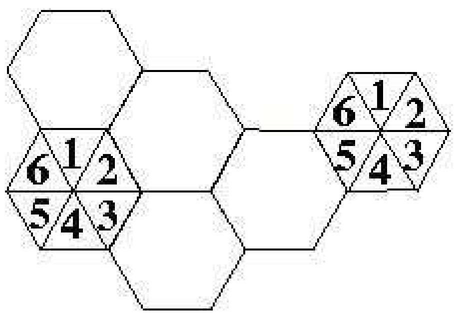

Figure A seven-cell cluster with 60^o sectors

In a seven-cell-cluster layout with 120^o sectored cells, it can be easily understood that the mobile units in a particular sector of the center cell will receive co-channel interference from only two of the first-tier co- channel base stations, rather than from all six. Likewise, the base station in the center cell will receive co-channel interference from mobile units in only two of the co-channel cells. Hence the signal to interference ratio is now modified to

$$\frac{S}{I} = \frac{(\sqrt{3N})^n}{2}$$

where the denominator has been reduced from 6 to 2 to account for the reduced number of interfering sources. Now, the signal to interference ratio for a seven-cell cluster layout using 120^o sectored antennas can be found from equation above to be 23.4 dB which is a significant improvement over the Omni-directional case where the worst-case S/I is found to be 17 dB (assuming a path-loss exponent, n=4). Some cellular systems divide the cells into 60^o sectors. Similar analysis can be performed on them as well.

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. To overcome this problem, a new microcell zone concept has been proposed. As shown in Figure this scheme has a cell divided into three microcell zones, with each of the three zone sites connected to the base station and sharing the same radio equipment. It is necessary to note that all the microcell zones, within a cell, use the same frequency used by that cell; that is no handovers occur between microcells. Thus when a mobile user moves between two microcell zones of the cell, the BS simply switches the channel to a different zone site and no physical re-allotment of channel takes place.

Locating the mobile unit within the cell: An active mobile unit sends a signal to all zone sites, which in turn send a signal to the BS. A zone selector at the BS uses that signal to select a suitable zone to serve the mobile unit - choosing the zone with the strongest signal. Base Station Signals: When a call is made to a cellular phone, the system already knows the cell location of that phone. The base station of that cell knows in which zone, within that cell, the cellular phone is located. Therefore when it receives the signal, the base station transmits it to the suitable zone site. The zone site receives the cellular signal from the base station and transmits that signal to the mobile phone after amplification. By confining the power transmitted to the mobile phone, co-channel interference is reduced between the zones and the capacity of system is increased.

Benefits of the micro-cell zone concept:

1) Interference is reduced in this case as compared to the scheme in which the cell size is reduced.
2) Handoffs arereduced (also compared to decreasing the cell size) since the microcells within the cell operate at the same frequency; no handover occurs when the mobile unit moves between the microcells.
3) Size of the zone apparatus is small. The zone site equipment being small can be mounted on the side of a building or on poles.
4) System capacity is increased.

The new microcell knows where to locate the mobile unit in a particular zone of the cell and deliver the power to that zone. Since the signal power is reduced, the microcells can be closer and result in an increased system capacity. However, in a microcellular system, the transmitted power to a mobile phone within a microcell has to be precise; too much power results in interference between microcells, while with too little power the signal might not reach the mobile phone. This is a drawback of microcellular systems, since a change in the surrounding (a new building, say, within a microcell) will require a change of the transmission power

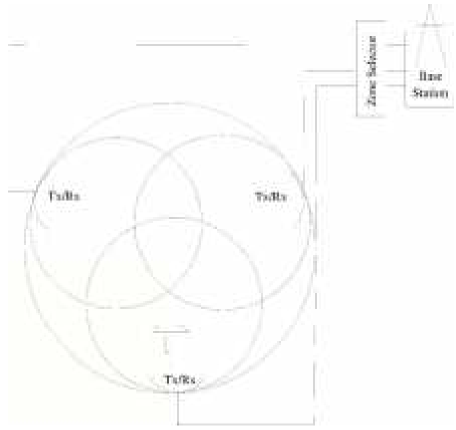

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Figure The micro-cell zone concept

Trunked Radio System

In the previous sections, the frequency reuse plan, the design trade-offs and also explored certain capacity expansion techniques like cell-splitting and sectoring is discussed. Now, look at the relation between the number of radio channels a cell contains and the number of users a cell can support. Cellular systems use the concept of trunking to accommodate a large number of users in a limited radio spectrum. It was found that a central office associated with say, 10,000 telephones 47 requires about 50 million connections to connect every possible pair of users. However, a worst case maximum of 5000 connections need to be made among these telephones at any given instant of time, as against the possible 50 million connections. In fact, only a few hundreds of lines are needed owing to the relatively short duration of a call. This indicates that the resources are shared so that the number of lines is much smaller than the number of possible connections.

A line that connects switching offices and that is shared among users on an as-needed basis is called a trunk. The fact that the number of trunks needed to make connections between offices is much smaller than the maximum number that could be used suggests that at times there might not be sufficient facilities to allow a call to be completed. A call that cannot be completed owing to a lack of resources is said to be blocked. In a trunked radio system, a channel is allotted on per call basis. The performance of a radio system can be estimated in a way by looking at how efficiently the calls are getting connected and also how they are being maintained at handoffs.

Some of the important factors to take into consideration are (i) Arrival statistics, (ii)Service statistics, (iii)Number of servers/channels.

Various Generations of Wireless Networks:

At the initial phase, mobile communication was restricted to certain official users and the cellular concept was never even dreamt of being made commercially available. Moreover, even the growth in the cellular networks was very slow. However, with the development of newer and better technologies starting from the 1970s and with the mobile users now connected to the PSTN, there has been a remarkable growth in the cellular radio. However, the spread of mobile communication was very fast in the 1990s when the government throughout the world provided radio spectrum licenses for Personal Communication Service (PCS) in 1.8 - 2 GHz frequency band.

First Generation Networks

The first mobile phone system in the market was AMPS. It was the first U.S. cellular telephone system, deployed in Chicago in 1983. The main technology of this first generation mobile system was FDMA/FDD and analog FM.

Second Generation Networks

Digital modulation formats were introduced in this generation with the main technology as TDMA/FDD and CDMA/FDD. The 2G systems introduced three popular TDMA standards and one popular CDMA standard in the market. These are as follows:

TDMA/FDD Standards

(a) Global System for Mobile (GSM): The GSM standard, introduced by Groupe Special Mobile, was aimed at designing a uniform pan-European mobile system. It was the first fully digital system utilizing the 900 MHz frequency band. The initial GSM had 200 KHz radio channels, 8 full-rate or 16 half-rate TDMA channels per carrier, encryption of speech, low speed data services and support for SMS for which it gained quick popularity.

(b) Interim Standard 136 (IS-136): It was popularly known as North American Digital Cellular (NADC) system. In this system, there were 3 full-rate TDMA users over each 30 KHz channel. The need of this system was mainly to increase the capacity over the earlier analog (AMPS) system.

(c) Pacific Digital Cellular (PDC): This standard was developed as the counterpart of NADC in Japan. The main advantage of this standard was its low transmission bit rate which led to its better spectrum utilization.

CDMA/FDD Standard

Interim Standard 95 (IS-95): The IS-95 standard, also popularly known as CDMAOne, uses 64 orthogonally coded users and codewords are transmitted simultaneously on each of 1.25 MHz channels. Certain services that have been standardized as a part of IS-95 standard are: short messaging service, slotted paging, over-the-air activation (meaning the mobile can be activated by the service provider without any third party intervention), enhanced mobile station identities etc.

2.5G Mobile Networks

In an effort to retrofft the 2G standards for compatibility with increased throughput rates to support modern Internet application, the new data centric standards were developed to be overlaid on 2G standards and this is known as 2.5G standard. Here, the main upgradation techniques are:

- supporting higher data rate transmission for web browsing ff supporting e-mail traffic
- enabling location-based mobile service

2.5G networks also brought into the market some popular application, a few of which are: Wireless Application Protocol (WAP), General Packet Radio Service (GPRS), High Speed Circuit Switched Dada (HSCSD), Enhanced Data rates for GSM Evolution (EDGE) etc.

3G: Third Generation Networks

3G is the third generation of mobile phone standards and technology, superseding 2.5G. It is based on the International Telecommunication Union (ITU) family of standards under the International Mobile Telecommunications-2000 (IMT-2000). ITU launched IMT-2000 program, which, together with the main industry and standardization bodies worldwide, targets to implement a global frequency band that would support a single, ubiquitous wireless communication standard for all countries, to provide the framework for the definition of the 3G mobile systems. Several radio access technologies have been accepted by ITU as part of the IMT-2000 framework.

3G networks enable network operators to offer users a wider range of more advanced services while achieving greater network capacity through improved spectral efficiency. Services include wide-area wireless voice telephony, video calls, and broadband wireless data, all in a mobile environment.

Additional features also include HSPA data transmission capabilities able to deliver speeds up to 14.4Mbit/s on the down link and 5.8Mbit/s on the uplink. 3G networks are wide area cellular telephone networks which evolved to incorporate high-speed internet access and video telephony. IMT-2000 defines a set of technical requirements for the realization of such targets, which can be summarized as follows:

- high data rates: 144 kbps in all environments and 2 Mbps in low-mobility and indoor environments
- symmetrical and asymmetrical data transmission
- circuit-switched and packet-switched-based services
- speech quality comparable to wire-line quality
- improved spectral efficiency
- several simultaneous services to end users for multimedia services
- seamless incorporation of second-generation cellular systems
- global roaming
- open architecture for the rapid introduction of new services and technology.

3G Standards and Access Technologies

As mentioned before, there are several different radio access technologies defined within ITU, based on either CDMA or TDMA technology. An organization called 3rd Generation Partnership Project (3GPP) has continued that work by defining a mobile system that fulfills the IMT-2000 standard. This system is called Universal Mobile Telecommunications System (UMTS). After trying to establish a single 3G standard, ITU finally approved a family of five 3G standards, which are part of the 3G framework known as IMT-2000:

W-CDMA CDMA2000 TD-SCDMA

Europe, Japan, and Asia have agreed upon a 3G standard called the Universal Mobile Telecommunications System (UMTS), which is WCDMA operating at 2.1 GHz. UMTS and WCDMA are often used as synonyms. In the USA and other parts of America, WCDMA will have to use another part of the radio spectrum.

3G W-CDMA (UMTS)

WCDMA is based on DS-CDMA (direct sequence code division multiple access) technology in which user-information bits are spread over a wide bandwidth (much larger than the information signal bandwidth) by multiplying the user data with the spreading code. The chip (symbol rate) rate of the spreading sequence is 3.84 Mcps, which, in the WCDMA system deployment is used together with the 5-MHz carrier spacing. The processing gain term refers to the relationship between the signal bandwidth and the information bandwidth. Thus, the name wideband is derived to differentiate it from the 2G CDMA (IS-95), which has a chip rate of 1.2288 Mcps. In a CDMA system, all users are active at the same time on the same frequency and are separated from each other with the use of user specific spreading codes. The wide carrier bandwidth of WCDMA allows supporting high user-data rates and also has certain performance benefits, such as increased multipath diversity. The actual carrier spacing to be used by the operator may vary on a 200-kHz grid between approximately 4.4 and 5 MHz, depending on spectrum arrangement and the interference situation.

In WCDMA each user is allocated frames of 10 ms duration, during which the user-data rate is kept constant. However, the data rate among the users can change from frame to frame. This fast radio capacity allocation (or the limits for variation in the uplink) is controlled and coordinated by the radio resource management (RRM) functions in the network to achieve optimum throughput for packet data services and to ensure sufficient quality of service (QoS) for circuit-switched users. WCDMA supports two basic modes of operation: FDD and TDD. In the FDD mode, separate 5-MHz carrier frequencies with duplex spacing are used for the uplink and downlink, respectively, whereas in TDD only one 5-MHz carrier is time shared between the up-link and the downlink. WCDMA uses coherent detection based on the pilot symbols and/or common pilot. WCDMA allows many performance- enhancement methods to be used, such as transmit diversity or advanced CDMA receiver concepts. The support for handovers (HO) between GSM and WCDMA is part of the first standard version. This means that all multi-mode WCDMA/GSM terminals will support measurements from the one system while camped on the other one. This allows networks using both WCDMA and GSM to balance the load between the networks and base the HO on actual measurements from the terminals for different radio conditions in addition to other criteria available.

The world's first commercial W-CDMA service, FoMA, was launched by NTT DoCoMo in Japan in 2001. FoMA is the short name for Freedom of Mobile Multimedia Access, is the brand name for the 3G services being o ered by Japanese mobile phone operator NTT DoCoMo. Elsewhere, W-CDMA deployments have been exclusively UMTS based.

UMTS or W-CDMA, assures backward compatibility with the second generation GSM, IS-136 and PDC TDMA technologies, as well as all 2.5G TDMA technologies. The network structure and bit level packaging of GSM data is retained by W-CDMA, with additional capacity and bandwidth provided by a new CDMA air interface.

3G CDMA2000

Code division multiple access 2000 is the natural evolution of IS-95 (cdma One). It includes additional functionality that increases its spectral efficiency and data rate capability.(code division multiple access) is a mobile digital radio technology where channels are defined with codes (PN sequences). CDMA permits many simultaneous transmitters on the same frequency channel. Since more phones can be served by fewer cell sites, CDMA-based standards have a signi cant economic advantage over TDMA- or FDMA-based standards. This standard is being developed by Telecommunications Industry Association (TIA) of US and is is standardized by 3GPP2.

The main CDMA2000 standards are: CDMA2000 1xRTT,CDMA 2000 1xEV and CDMA2000 EV-DV. These are the approved radio interfaces for the ITU's IMT-2000 standard. In the following, a brief discussion about all these standards is given.

CDMA2000 1xRTT: RTT stands for Radio Transmission Technology and the designation "1x", meaning "1 times Radio Transmission Technology", indicates the same RF bandwidth as IS-95.The main features of CDMA2000 1X are as follows:

Supports an instantaneous data rate up to 307kpbs for a user in packet mode and a typical throughput rates of 144kbps per user, depending on the number of user, the velocity of user and the propagating conditions.

Supports up to twice as many voice users a the 2G CDMA standard provides the subscriber unit with up to two times the standby time for longer lasting battery life. ECE department

CDMA2000 EV: This is an evolutionary advancement of CDMA with the following characteristics:

Provides CDMA carriers with the option of installing radio channels with data only (CDMA2000 EV-DO) and with data and voice (CDMA2000 EV-DV .

The cdma2000 1xEV-DO supports greater than 2.4Mbps of instantaneous high-speed packet throughput per user on a CDMA channel, although the user data rates are much lower and highly dependent on other factors.

CDMA2000 EV-DV can offer data rates up to 144kbps with about twice as many voice channels as IS-95B.

CDMA2000 3x is (also known as EV-DO Rev B) is a multi-carrier evolution.

It has higher rates per carrier (up to 4.9 Mbit /s on the downlink per carrier). Typical deployments are expected to include 3 carriers for a peak rate of 14.7 Mbit /s. Higher rates are possible by bundling multiple channels together. It enhances the user experience and enables new services such as high definition

Video streaming

Uses statistical multiplexing across channels to further reduce latency, enhancing the experience for latency-sensitive services such as gaming, video telephony, remote console sessions and web browsing.

It provides increased talk-time and standby time. The interference from the adjacent sectors is reduced by hybrid frequency re-use and improves the rates that can be offered, especially to users at the edge of the cell. It has efficient support for services that have asymmetric download and upload requirements (i.e. different data rates required in each direction) such as le transfers, web browsing, and broadband multimedia content delivery.

3G TD-SCDMA

Time Division-Synchronous Code Division Multiple Access, or TD-SCDMA, is a 3G mobile telecommunications standard, being pursued in the People's Republic of China by the Chinese Academy of Telecommunications Technology (CATT). This proposal was adopted by ITU as one of the 3G options in late 1999. TD-SCDMA is based on spread spectrum technology.

TD-SCDMA uses TDD, in contrast to the FDD scheme used by W-CDMA. By dynamically adjusting the number of timeslots used for downlink and uplink, the system can more easily accommodate asymmetric traffic with different data rate requirements on downlink and uplink than FDD schemes. Since it does not require paired spectrum for downlink and uplink, spectrum allocation exibility is also increased. Also, using the same carrier frequency for uplink and downlink means that the channel condition is the same on both directions, and the base station can deduce the downlink channel information from uplink channel estimates, which is helpful to the application of beam forming techniques.

TD-SCDMA also uses TDMA in addition to the CDMA used in WCDMA. This reduces the number of users in each timeslot, which reduces the implementation complexity of multiuser detection and beam forming schemes, but the non-continuous transmission also reduces coverage (because of the higher peak power needed), mobility (because of lower power control frequency) and complicates radio resource management algorithms.

The "S" in TD-SCDMA stands for "synchronous", which means that uplink signals are synchronized at the base station receiver, achieved by continuous timing adjustments. This reduces the interference between users of the same timeslot using different codes by improving the orthogonality between the codes, therefore increasing system capacity, at the cost of some hardware complexity in achieving uplink synchronization.

Beyond 3G networks, or 4G (Fourth Generation), represent the next complete evolution in wireless communications. A 4G system will be able to provide a comprehensive IP solution where voice, data and streamed multimedia can be given to users at higher data rates than previous generations. There is no formal definition for 4G ; however, there are certain objectives that are projected for 4G. It will be capable of providing between 100 Mbit/s and 1 Gbit/s speeds both indoors and outdoors, with premium quality and high security. It would also support systems like multicarrier communication, MIMO and UWB.

APPLICATIONS

Examples of 2nd Generation (2G) Systems

Examples of 3rd Generation (3G) Systems

POST TEST-MCQ TYPE

1. Why neighbouring stations are assigned different group of channels in cellular system?

a) To minimize interference
b) To minimize area
c) To maximize throughput
d) To maximize capacity of each cell

2. What is frequency reuse?
a) Process of selecting and allocating channels
b) Process of selection of mobile users
c) Process of selecting frequency of mobile equipment
d) Process of selection of number of cells

3. Which of the following is a universally adopted shape of cell?
a) Square
b) Circle
c) Triangle
d) Hexagon

4. Actual radio coverage of a cell is called __________
a) Fingerprint
b) Footprint
c) Imprint
d) Matrix

5. Which type of antenna is used for center excited cells?
a) Dipole antenna
b) Grid antenna
c) Sectored antenna
d) Omnidirectional antenna

6. For a cellular system, if there are N cells and each cell is allocated k channel. What is the total number of available radio channels, S?
a) S=k*N
b) S=k/N
c) S=N/k
d) $S=k^N$

7. A spectrum of 30 MHz is allocated to a cellular system which uses two 25 KHz simplex channels to provide full duplex voice channels. What is the number of channels available per cell for 4 cell reuse factor?
a) 150 channels
b) 600 channels
c) 50 channels
d) 85 channels

8. What is the drawback of dynamic channel assignment?
a) Decrease channel utilization
b) Increase probability of blocked call
c) Cross talk
d) Increase storage and computational load on system

9. RSSI stands for ________
a) Received Signal Strength Indicator
b) Restricted Signal Strength Indicator
c) Radio Signal Strength Indication
d) Restricted System Software Indicator

10. In dynamic channel assignment strategy, base station requests channel from
a) MSC
b) Neighbouring cell
c) Neighbouring cluster
d) Neighbouring base station

11. What is a borrowing strategy in fixed channel assignments?
a) Borrowing channels from neighbouring cell
b) Borrowing channels from neighbouring cluster
c) Borrowing channels from same cell
d) Borrowing channels from other base station in same cell

12. What happens to a call in fixed channel strategy, if all the channels in a cell are occupied?
a) Queued
b) Cross talk
c) Blocked
d) Delayed

13. Which of the following is not an objective for channel assignment strategies?
a) Efficient utilization of spectrum
b) Increase of capacity
c) Minimize the interference
d) Maximize the interference

14. In fixed channel assignment strategy, each cell is allocated a predetermined set of
a) Voice channels
b) Control channels
c) Frequency
d) base stations

15. Soft handoff is also known as
a) MAHO
b) Hand over
c) Break before make
d) Make before break

16. Cell dragging is a problem that occurs due to
a) Pedestrian users
b) Stationary users
c) High speed mobile systems
d) Base stations having same frequency

17. What is the condition for handoff?
a) A mobile moves into a different cell while in conversation
b) A mobile remains in the same cell while in conversation
c) A mobile moves to different cell when idle
d) A mobile remains in the same cell and is idle

18. The time over which a call can be maintained within a cell without handoff is called
a) Run time
b) Peak time
c) Dwell time
d) Cell time

19. Dwell time does not depend on which of the following factor?
a) Propagation
b) Interference
c) Distance between subscriber and base station
d) Mobile station

20. Which of the following is associated with the handoff in first generation analog cellular systems?
a) Locator receiver
b) MAHO
c) Cell dragging
d) Breathing cell

21. What is the condition for intersystem interference?
a) Mobile moves from one cell to another cell
b) Mobile remains in the same cell
c) Mobile moves from one cellular system to another cellular system
d) Mobile remains in the same cluster

22. Which of the following priority handoff method decrease the probability of forced termination of a call due to lack of available channels?
a) Queuing
b) Guard channel
c) Cell dragging
d) Near far effect

23. What is the disadvantage of guard channel?
a) Efficient utilization of spectrum
b) Cross talk
c) Near far effect
d) Reduce total carried traffic

24. Adjacent channel interference can be minimized through
a) Changing frequency of base stations
b) Careful filtering and channel assignments
c) Increasing number of base stations
d) Increasing number of control channels

25. In near-far effect, a nearby transmitter captures the
a) Receiver of the subscriber
b) Transmitter of the subscriber
c) Nearby MSC
d) Neighbouring base station

26. Adjacent channel interference occurs due to ________
a) Power transmitted by Base station
b) MSCs
c) Same frequency of mobile users
d) Imperfect receiver filters

27. Co-channel interference is a function of __________
a) Radius of cell
b) Transmitted power
c) Received power
d) Frequency of mobile user

28. What is the cluster size for CDMA?
a) N=10
b) N=100
c) N=1
d) N=50

29. Which of the following techniques do not help in expanding the capacity of cellular system?
a) Sectoring
b) Scattering
c) Splitting
d) Microcell zone concept

30. Which of the following technology distributes the coverage of the cell and extends the cell boundary to hard-to-reach places?
a) Cell splitting
b) Scattering
c) Sectoring
d) Micro cell zone concept

31. Which of the following increases the number of base stations in order to increase capacity?
a) Cell splitting
b) Sectoring
c) Repeaters
d) Micro cell zone concept

32. Which of the following experience trunking inefficiencies?
a) Cell splitting
b) Micro cell zone technique
c) Sectoring
d) Repeaters

33. The process of subdividing a congested cell into smaller cells is called
a) Cell splitting
b) Sectoring
c) Micro cell technique
d) Repeaters

34. Which of the following technique is used to limit radio coverage of newly formed microcells?
a) Sectoring
b) Splitting
c) Antenna downtilting
d) Scattering

35. Which of the following has range extension capability?
a) Sectoring
b) Repeaters
c) Scattering
d) Micro cell zone concept

36. Which of the following is not an advantage of micro cell zone technique?
a) Reduced co channel interference
b) Improved signal quality
c) Increase in capacity
d) Increasing number of base stations

37. In a micro cell zone concept, when a mobile travels from one zone to another within the cell, it retains the same
a) Power level
b) Base station
c) Channel
d) Receiver

38. Which of the following multiple access techniques are used by second generation cellular systems?
a) FDMA/FDD and TDMA/FDD
b) TDMA/FDD and CDMA/FDD
c) FDMA/FDD and CDMA/FDD
d) FDMA/FDD only

39. Which of the following is a CDMA standard of second generation network?
a) IS-95
b) IS-136
c) ETACS
d) EDGE

40. How many users or voice channels are supported for each 200 KHz channel in GSM?
a) Eight
b) Three
c) Sixty four
d) Twelve

CONCLUSION

In this unit, the fundamental cellular concepts its structure, the frequency reuse concept, cell splitting, channel assignment strategies have been discussed. The fundamental concepts of handoff, trunking efficiency and frequency planning have been presented. Handoffs are required to pass mobile traffic from cell to cell, and there are various ways handoffs are implemented. The capacity of a cellular system is a function of many variables. The trunking efficiency limits the number of users that can access a trunked radio system. Trunking is affected by the number of available channels and how they are partitioned in a trunked cellular system. Cell splitting, sectoring and zone microcell technique are all shown to improve the capacity by increasing S/I. This unit has provided an overview of the modern wireless communication networks that were used throughout the world. The convergence of the second generation mobile telephone standards into third generation standards was described and the interim 2.5 G data solutions were explained for all major mobile technologies.

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ASSIGNMENT

1. Describe the principle of Operation of cellular mobile system and explain the cellular Concept with neat diagram.
2. Explain Cell splitting and Concept of frequency channels.
3. In your home, how many modern wireless communication networks are available to you? Identify the type of services, the types of technologies, the commercial names of the service providers and the commercial names of the equipment manufacturers that offer these wireless capabilities.
4. Compare and contrast the various 2.5G technology paths that each of the major 2G standards provide.
5. Explain the different generation in cellular wireless networks.
6. What are the Main advantages and disadvantages of various cellular structures?

MOBILE COMMUNICATION AND NETWORKS

UNIT II THE WIRELESS CHANNEL

Prepared by

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AIM & OBJECTIVES

- ❖ To understand the issues involved in mobile communication system design and analysis.
- ❖ To understand the characteristics of wireless channels.
- ❖ To know the fundamental limits on the capacity of wireless channels.

PRE TEST-MCQ TYPE

1. The mechanism behind electromagnetic wave propagation cannot be attributed to
a) Reflection
b) Diffraction
c) Scattering
d) Sectoring

2. Which of the following do not undergo free space propagation?
a) Satellite communication system
b) Wired telephone systems
c) Microwave line of sight radio links
d) Wireless line of sight radio links

3. Relation between wavelength and carrier frequency is
a) $\lambda=c/f$
b) $\lambda=c*f$
c) $\lambda=f/c$
d) $\lambda=1/f$

4. Which of the following antenna radiates power with unit gain uniformly in all directions?
a) Directional antenna
b) Dipole antenna
c) Isotropic antenna
d) Loop antenna

5. Which of the following is called an ideal antenna?
a) Dipole antenna
b) Directional antenna
c) Isotropic antenna
d) Loop antenna

UNIT II THE WIRELESS CHANNEL

Signal Propagation-Propagation mechanism- reflection, refraction, diffraction and scattering, large scale signal propagation and lognormal shadowing. Fading channels -Multipath and small scale fading- Doppler shift, statistical multipath channel models, narrowband and wideband fading models, power delay profile, average and rms delay spread, coherence bandwidth and coherence time, flat and frequency selective fading, slow and fast fading, average fade duration and level crossing rate.

THEORY

Introduction

There are two basic ways of transmitting an electromagnetic (EM) signal, through a guided medium or through an unguided medium. Guided mediums such as coaxial cables and fiber optic cables, are far less hostile toward the information carrying EM signal than the wireless or the unguided medium. It presents challenges and conditions which are unique for this kind of transmissions. A signal, as it travels through the wireless channel, undergoes many kinds of propagation effects such as refection, diffractions and scattering, due to the presence of buildings, mountains and other such obstructions. Refection occurs when the EM waves impinge on objects which are much greater than the wavelength of the traveling wave. Diffraction is a phenomena occurring when the wave interacts with a surface having sharp irregularities. Scattering occurs when the medium through the wave is traveling contains objects which are much smaller than the wavelength of the EM wave. These varied phenomena's lead to large scale and small scale propagation losses. Due to the inherent randomness associated with such channels they are best described with the help of statistical models. Models which predict the mean signal strength for arbitrary transmitter receiver distances are termed as large scale propagation models. These are termed so because they predict the average signal strength for large Transmitter- Receiver separations, typically for hundreds of kilometers.

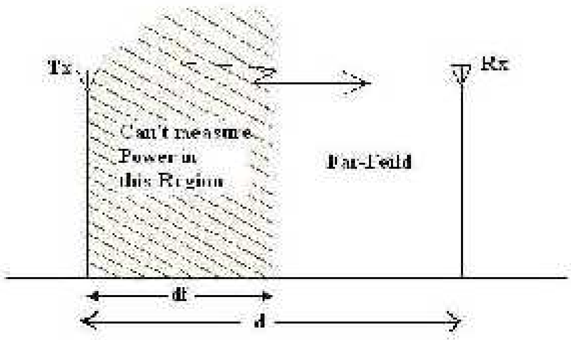


Figure Free space propagation model, showing the near and far fields.

Free Space Propagation Model

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

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

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

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

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

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

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

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

$$EIRP = P_tG_t$$

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

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

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

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

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

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

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

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

$$d_f >> D$$

The Three Basic Propagation Mechanisms

Reflection, diffraction, and scattering are the three basic propagation mechanisms which impact propagation in a mobile communication system. Reflection occurs when a propagating electromagnetic wave impinges upon an object which has very large dimensions when compared to the wavelength of the propagating wave. Reflections occur from the surface of the earth and from buildings and walls.

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

Scattering occurs when the medium through which the wave travels consists of objects with dimensions that are small compared to the wavelength, and where the number of obstacles per unit volume is large. Scattered waves are produced by rough surfaces, small objects, or by other irregularities in the channel. In practice, foliage, street signs, and lamp posts induce scattering in a mobile communications system.

Reflection

When a radio wave propagating in one medium impinges upon another medium having different electrical properties, the wave is partially reflected and partially transmitted. If the plane wave is incident on a perfect dielectric, part of the energy is transmitted into the second medium and part of the energy is reflected back into the first medium, and there is no loss of energy in absorption. lf the second medium is a perfect conductor, then all incident energy is reflected back into the first medium without loss of energy.

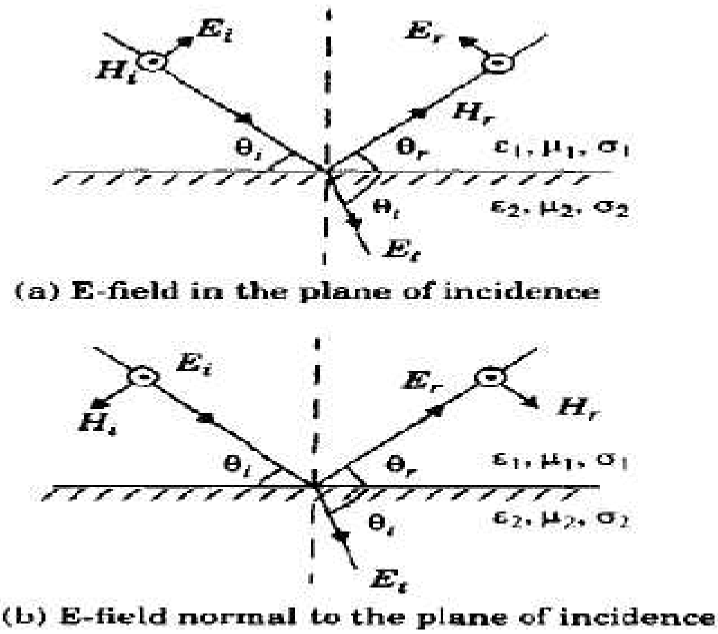


Figure Geometry for calculating the reflection coefficients between two dielectrics

The electric field intensity of the reflected and transmitted waves may be related to the incident wave in the medium of origin through the Fresnel reflection coefficient (Γ). The reflection coefficient is a function of the material properties, and generally depends on the wave polarization, angle of incidence, and the frequency of the propagating wave.

Reflection from dielectrics

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

The plane of incidence is defined as the plane containing the incident, reflected, and transmitted rays. In Figure, 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 , the E-field polarization is perpendicular to the plane of incidence (that is, the incident E-field is pointing out of the page towards the reader, and is perpendicular to the page and parallel to the reflecting surface).

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

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

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

Where η is the intrinsic impedance of the respective medium. Or,

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

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

Where ε is the permittivity of the respective medium.

Brewster Angle

The Brewster angle is the angle at which no reflection occurs in the medium of origin. It occurs when the incident angle BB is such that the reflection coefficient Γ∥ is equal to zero. The Brewster angle is given by the value of θ_B which satisfies

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

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

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

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

Reflection from Perfect Conductors

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

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

Similarly, for the case when the E-field is horizontally polarized, the boundary conditions require that

$\theta_i = \theta_r$ and $E_i = - E_r$ (E-field not in plane of incidence)

Ground Reflection (2-ray) Model

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

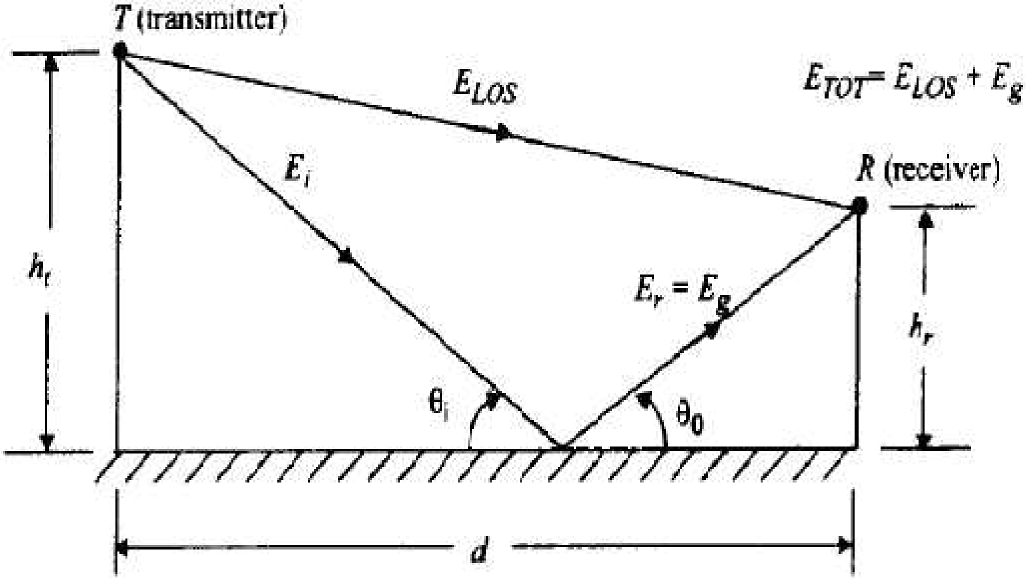


Figure Two ray ground reflection model

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

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

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

The electric field E TOT(d, t) can be expressed as the sum of equations for distances d' and d" (i.e. direct wave and reflected wave.

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

Diffraction

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

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

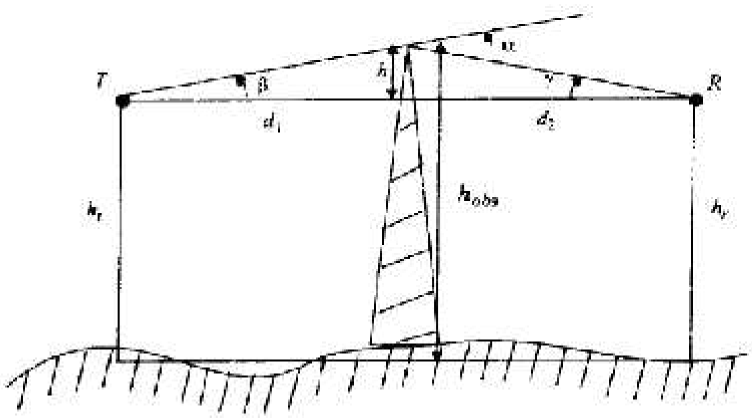
Fresnel Zone Geometry

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

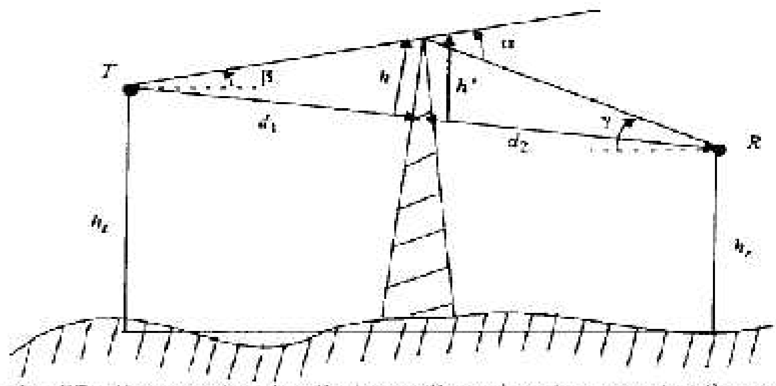
$$\Delta \approx \frac{h^2}{2}\frac{(d_1+d_2)}{d_1 d_2}$$

The corresponding phase difference is given by

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



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



(b) Knife edge diffraction geometry when the transmitter and receiver are not at the same height. Note that if α and β are small and $h \ll d_1$ and d_2, then h and h' are virtually identical and the geometry may be redrawn as shown in Figure 3.10c.

Knife-edge Diffraction Model

Estimating the signal attenuation caused by diffraction of radio waves over hills and buildings is essential in predicting the field strength in a given service area. Generally, it is impossible to make very precise estimates of the diffraction losses, and in practice prediction is a process of theoretical approximation modified by necessary empirical corrections. Though the calculation of diffraction losses over complex and irregular terrain is a mathematically difficult problem, expressions for diffraction losses for many simple cases have been derived.

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

Multiple Knife-edge Diffraction

ln many practical situations, especially in hilly terrain, the propagation path may consist of more than one obstruction, in which case the total diffraction loss due to all of the obstacles must be computed. Bullington suggested that the series of obstacles be replaced by a single equivalent obstacle so that the path loss can be obtained using single knife-edge diffraction models.

This method, illustrated in Figure oversimplifies the calculations and often provides very optimistic estimates of the received signal strength. In a more rigorous treatment, Millington et. al. gave a wave-theory solution for the field behind two knife edges in series. This solution is very useful and can be applied easily for predicting diffraction losses due to two knife edges. However, extending this to more than two knife edges becomes a formidable mathematical problem. Many models that are mathematically less complicated have been developed to estimate the diffraction losses due to multiple obstructions.

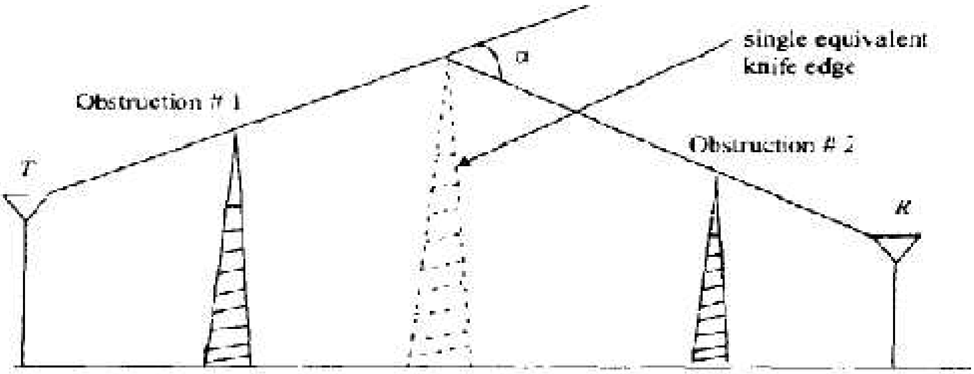


Figure Bullington's construction of an equivalent knife edge

Scattering:

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

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

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

Outdoor propagation model

Longley-Rice Model

The Longley-Rice model is applicable to point-to-point communication systems in the frequency range from 40 MHz to 100 GHz, over different kinds of terrain. The median transmission loss is predicted using the path geometry of the terrain profile and the refractivity of the troposphere. Geometric optics techniques (primarily the 2-ray ground reflection model) are used to predict signal strengths within the radio horizon.

Diffraction losses over isolated obstacles are estimated using the Fresnel-Kirchoff knife- edge models. Forward scatter theory is used to make troposcatter predictions over long distances. The Longley-Rice method operates in two modes. When a detailed terrain path profile is available, the path-specific parameters can be easily determined and the prediction is called a point-to-point mode prediction. On the other hand, if the terrain path profile is not available, the Longley-Rice method provides techniques to estimate the path-specific parameters, and such a prediction is called an area mode prediction.

Okumura Model

Okumura`s model is one of the most widely used models for signal prediction in urban areas. This model is applicable for frequencies in the range 150 MHz to 1920 MHz (although it is typically extrapolated up to 3000 MHz) and distances of 1 km to 100 km. It can be used for base station antenna heights ranging from 30 m to 1000 m. Okumura developed a set of curves giving the median attenuation relative to free space (Amu), in an urban area over a quasi-smooth terrain with a base station effective antenna height (h_{te}) of 200 m and a mobile antenna height (h_{re}) of 3 m. These curves were developed from extensive measurements using vertical omni-directional antennas at both the base and mobile, and are plotted as a function of frequency in the range 100 MHz to 1920 MHz and as a function of distance from the base station in the range 1 km to 100 km.

To determine path loss using Okumura`s model, the free space path loss between the points of interest is first determined, and then the value of Amu(f, d) (as read from the curves) is added to it along with correction factors to account for the type of terrain. The model can be expressed as

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

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

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

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

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

Hata Model

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

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

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

$$a(h_{re}) = (1.11ogf_c - 0.7)h_{re} - (1.56logf_c - 0.8) \text{ dB}$$

and for a large city, it is given by

$$a(h_{re}) = 8.29(log1.54h_{re})^2 - 1.1 \text{ dB for } f_c <= 300 \text{ MHz}$$

$$a(h_{re}) = 3.2(log11.75h_{re})^2 - 4.97 \text{ dB for } f_c >= 300 \text{ MHz}$$

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

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

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

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

Indoor Propagation Models

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

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

Partition Losses (same floor)

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

Partition Losses between Floors

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

Multipath & Small-Scale Fading

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

Fading

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

Multipath Fading Effects

In principle, the following are the main multipath effects

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

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

3. Time dispersion or echoes caused by multipath propagation delays.

Factors Influencing Fading

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

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

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

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

(4) Transmission Bandwidth of the signal - If the transmitted radio signal bandwidth is greater than the bandwidth" of the multipath channel (quantified by coherence bandwidth), the received signal will be distorted.

Types of Small-Scale Fading

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

Fading Effects due to Multipath Time Delay Spread

Flat Fading

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

$$B_S \ll B_C$$

where BS is the signal bandwidth and BC is the coherence bandwidth. Also

$$T_S \gg \sigma_\tau$$

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

Frequency Selective Fading

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

$$B_S \gg B_C$$

$$T_S \ll \sigma_\tau$$

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

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

Fading Effects due to Doppler Spread

Fast Fading

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

$$T_S \gg T_C$$

where T_C is the coherence time and

$$B_S \gg B_D$$

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

Slow Fading

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

$$T_S \ll T_C$$
$$B_S \gg B_D$$

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

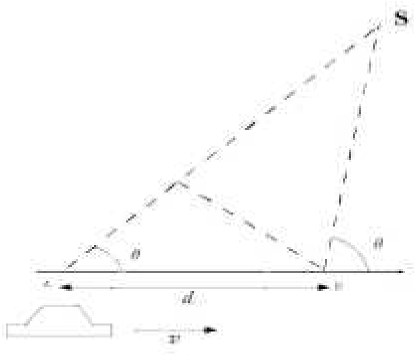


Figure Illustration of Doppler effect

Doppler Shift

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

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

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

In mobile communication, the above equation can be slightly changed according to the convenience since the source (BS) is fixed and located at a remote elevated level from ground. The expected Doppler shift of the EM wave then comes out to be

$$\pm\frac{v_r}{c}f_0 \text{ or, } \pm\frac{v_r}{\lambda}.$$

As the BS is located at an elevated place, a cos factor would also be multiplied with this.. Consider a mobile moving at a constant velocity v, along a path segment length d between points A and B, while it receives signals from a remote BS source S. The difference in path lengths traveled by the wave from source S to the mobile at points A and B is

$$\Delta l = d\cos\theta = v\Delta t\cos\theta, \text{ where } \Delta t$$

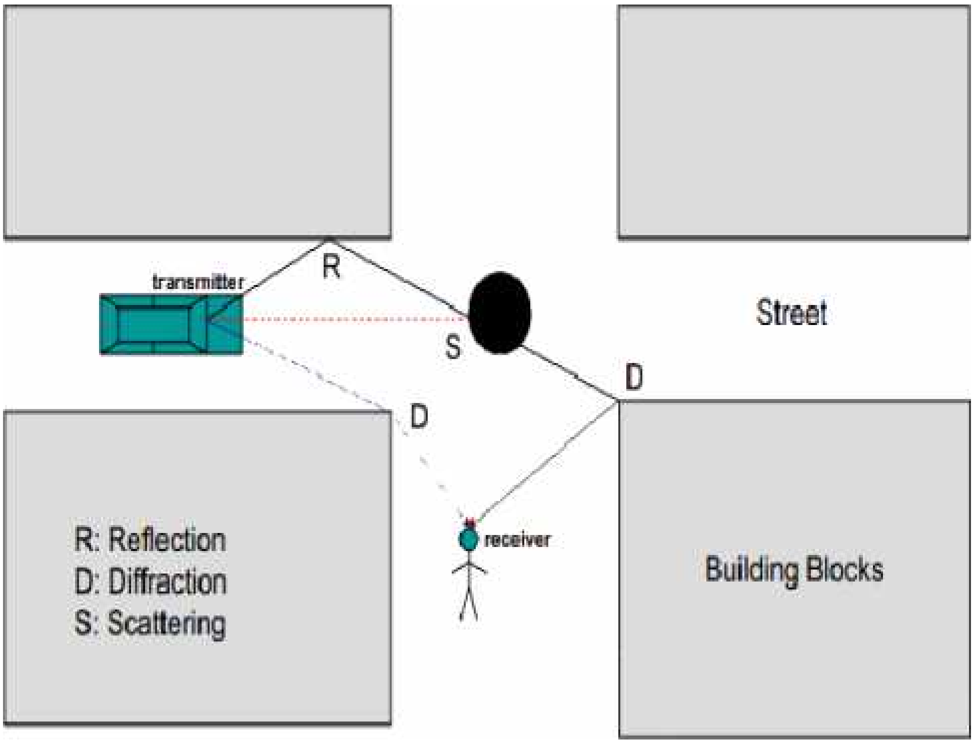
Where Δt is the time required for the mobile to travel from A to B, and f is assumed to be the same at points A and B since the source is assumed to be very far away. The phase change in the received signal due to the difference in path lengths is therefore

$$\Delta\psi = \frac{2\pi\Delta l}{\lambda} = \frac{2\pi v\Delta t}{\lambda}\cos\theta$$

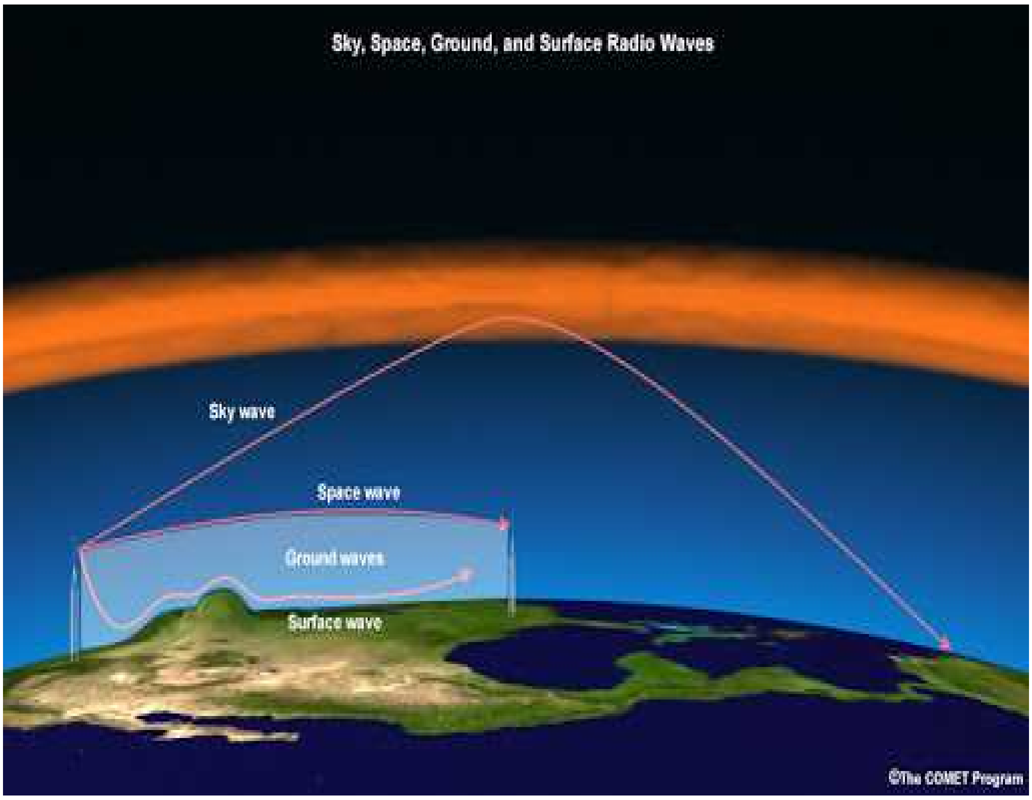
and hence the apparent change in frequency, or Doppler shift (f_d) is

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

APPLICATIONS



Example of Mobile Radio propagation mechanism



Example of Ground Reflection (2-ray) Model used in Remote Sensing

POST TEST-MCQ TYPE

1. The propagation model that estimates radio coverage of a transmitter is called
a) Large scale propagation model
b) Small scale propagation model
c) Fading model
d) Okumura model

2. Small scale propagation model is also known as
a) Hata model
b) Micro scale propagation model
c) Okumura model
d) Fading model

3. Free space propagation model is to predict
a) Received signal strength
b) Transmitted power
c) Gain of transmitter
d) Gain of receiver

4. The free space model predicts that received signal decays as a function of
a) Gain of transmitter antenna
b) T-R separation
c) Power of transmitter antenna
d) Effective aperture of the antenna

5. Relation between gain and effective aperture is given by ______
a) G=(4πAe)/λ2
b) G=(4π λ2)/Ae
c) G=4πAe
d) G=Ae/λ2

6. Path loss in free space model is defined as difference of ________
a) Effective transmitted power and gain
b) Effective received power and distance between T-R
c) Gain and received power
d) Effective transmitter power and receiver power

7. Which of the following mechanism do not impact propagation in mobile communication system?
a) Reflection
b) Diffraction
c) Scattering
d) Refraction

8. What is the dimension of object as compared to wavelength of propagating wave when reflection occurs?
a) Large
b) Small
c) Same
d) Very small

9. Reflection coefficient is not a function of
a) Material property
b) Diffraction loss
c) Wave polarization
d) Angle of incidence

10. Diffraction occurs when radio path between Transmitter and receiver is obstructed by
a) Surface having sharp irregularities
b) Smooth irregularities
c) Rough surface
d) All types of surfaces

11. Diffraction allows radio signals to propagate around ________
a) Continuous surface
b) Smooth surface
c) Curved surface of Earth
d) Does not allow propagation

12. Which principle explains the phenomenon of diffraction?
a) Principle of Simultaneity
b) Pascal's Principle
c) Archimedes' Principle
d) Huygen's principle

13. Difference between the direct path and the diffracted path is called ________
a) Average length
b) Radio path length
c) Excess path length
d) Wavelength

14. The phase difference between a direct line of sight path and diffracted path is function of
a) Height and position of obstruction
b) Only height
c) Operating frequency
d) Polarization

15. In mobile communication system, diffraction loss occurs due to
a) Dielectric medium
b) Obstruction
c) Electric field
d) Operating frequency

16. For predicting the field strength in a given service area, it is essential to estimate ______
a) Polarization
b) Magnetic field
c) Height of transmitter
d) Signal attenuation

17. Scattering occurs when medium consists of objects with dimensions _______ compared to the wavelength.
a) Same
b) Small
c) Large
d) Very large

18. Scattered waves are produced at ________
a**) Rough surface**
b) Shadowed region
c) Smooth surface
d) Horizon

19. A surface is considered rough if protuberance is ________ than critical height.
a) Equal
b) Less
c) Greater
d) No relation

20. Which equation is used to calculate the received power due to scattering for urban mobile radio system?
a) Laplace equation
b) Bistatic radar equation
c) Poisson's equation
d) Maxwell equation

21. In ionosphere propagation, waves arriving at the receiving antenna using the phenomenon of
a) Scattering
b) Refraction
c) Diffraction
d) Radiation

22. Power density is basically termed as ________ power per unit area.
a) Reflected
b) Refracted
c) Radiated
d) Diffracted

23. Small scale fading describes the _________ fluctuations of the amplitude, phases of a signal.
a) Rapid
b) Slow
c) Instantaneous
d) Different

24. Which of the following is not an effect caused by multipath in radio channel?
a) Rapid changes in signal strength
b) Random frequency modulation
c) Power of base station
d) Time dispersion

25. In urban areas, fading occurs due to height of mobile antenna ________ than height of surrounding structure.
a) Same
b) Smaller
c) Greater
d) Very larger

26. Apparent shift in frequency in multipath wave is caused due to relative motion between
a) Base station and MSC
b) Mobile and surrounding objects
c) Mobile and MSC
d) Mobile and base station

27. Which of the following factor does not influence small scale fading?
a) Multipath propagation
b) Power density of base station
c) Speed of mobile
d) Speed of surrounding objects

28. What is a measure of the maximum frequency difference for which signals are strongly correlated in amplitude?
a) Coherence bandwidth
b) Narrow bandwidth
c) Incoherent bandwidth
d) Wide bandwidth

29. The Doppler shift for mobile moving with constant velocity, v is given by
a) $(v*\cos\theta)/\lambda$
b) v/λ
c) $v*\cos\theta$
d) $v*\lambda$

30. Which of the following is not a channel parameter?
a) Bandwidth
b) Coherence time
c) Rms delay spread
d) Doppler spread

31. Which of the following leads to time dispersion and frequency selective fading?
a) Doppler spread
b) Multipath delay spread
c) Time dispersive parameters
d) Frequency delay spread

32. Flat fading channel is also known as _______________
a) Amplitude varying channel
b) Wideband channel
c) Phase varying channel
d) Frequency varying channel

33. Frequency selective fading channels are also known as
a) Narrowband channel
b) Wideband channel
c) Amplitude varying channel
d) Phase varying channel

34. For fast fading channel, the coherence time of the channel is smaller than _______ of transmitted signal.
a) Doppler spread
b) Bandwidth
c) Symbol period
d) Coherence bandwidth

35. In slow fading channel, Doppler spread of the channel is much less than the ________ of baseband signal.
a) Symbol period
b) Phase
c) Coherence time
d) Bandwidth

36. The rapid fluctuations due to small scale fading affect the _________ design.
a) Receiver
b) Transmitter
c) MSC
d) Base station

37. Which of the following is equal to received power?
a) Square of complex voltage
b) Complex voltage
c) Magnitude of complex voltage
d) Magnitude squared of complex voltage

38. Which of the following describes the average fading rate within a local area?
a) Angular spread
b) Angular constriction
c) Azimuthal direction of maximum fading
d) Angle of arrival

39. Which of the following is not a statistical models for multipath fading channels?
a) Clarke’s model for flat fading
b) Saleh and Valenzuela indoor statistical model
c) Two ray Rayleigh fading model
d) Faraday model

40. Which of the following is an important statistics of a Rayleigh fading useful for designing error control codes and diversity schemes?
a) Mobile speed
b) Doppler frequency
c) Level crossing rate (LCR)
d) Power density

CONCLUSION

In this unit, the basic propagation mechanisms that impact propagation in a mobile communication system were discussed. Reflection, refraction, diffraction and scattering are the most important parameters predicted by large scale propagation models. The small scale fading and Multipath propagation models have been discussed. The small scale fading impacts the time delay and the dynamic fading range of signals levels within a small local area at a receiver antenna. This unit demonstrated the important principle that local received power is not a function of bandwidth, but rather the average signal level within a local area is constant regardless of signal bandwidth. Multipath shape factor theory provides an easy, intuitive and accurate method for analyzing small scale fading channels with non- omnidirectional multipath propagation. All of the discussed topics in this unit are critical in the design of a practical air interface , as they impact the selection of modulation data rates and modulation methods for time varying mobile channels.

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

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ASSIGNMENT

1. What do you understand by large scale fading? Explain the 2-ray ground reflection model for path loss prediction.
2. Explain methods of multipath measurements in detail.
3. Explain about types of small scale fading?
4. What are the factors influencing small scale fading?
5. Explain the significance of fading in mobile environment.

MOBILE COMMUNICATION AND NETWORKS

UNIT III ANTENNAS FOR MOBILE TERMINALS

Prepared by

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AIM & OBJECTIVES

- ❖ To understand the issues involved in mobile communication system design and analysis.
- ❖ To understand the characteristics of wireless channels.
- ❖ To know the fundamental limits on the capacity of wireless channels.
- ❖ To acquire knowledge about different types of Diversity receivers.

PRE TEST-MCQ TYPE

1. Coherence bandwidth is a statistical measure of the range of frequencies over which the channel is
a) selective
b) flat
c) slow
d) fast

2. What are the parameters that describe the time varying nature of the channel in a small scale region?
a) Doppler spread only
b) Coherence time only
c) **Doppler spread and Coherence time**
d) Coherence bandwidth

3. Frequency selective fading channels are also known as
a) Amplitude varying channel
b) Phase varying channel
c) Narrow band channels
d) Wide band channels

4. In slow fading channel, Doppler spread of the channel is much less than the ________ of baseband signal.
a) Symbol period
b) Phase
c) Coherence time
d) Bandwidth

5. AWGN stands for
a) Absolute White Gaussian Noise
b) Absolute White Gaussian Number
c) Additive White Gaussian Noise
d) Additive White Gaussian Number

UNIT III ANTENNAS FOR MOBILE TERMINALS

Capacity of flat and frequency selective channels, Antennas - Antennas for mobile terminal monopole antennas, PIFA, base station antennas and arrays.

THEORY

Introduction

The growing demand for wireless communication makes it important to determine the capacity limits of these channels. These capacity limits dictate the maximum data rates that can be transmitted over wireless channels with asymptotically small error probability, assuming no constraints on delay or complexity of the encoder and decoder. Channel capacity was pioneered by Claude Shannon in the late 1940s, using a mathematical theory of communication based on the notion of mutual information between the input and output of a channel. Shannon defined capacity as the mutual information maximized over all possible input distributions. The significance of this mathematical construct was Shannon's coding theorem and converse, which proved that a code did exist that could achieve a data rate close to capacity with negligible probability of error, and that any data rate higher than capacity could not be achieved without an error probability bounded away from zero.

Shannon's ideas were quite revolutionary at the time, given the high data rates he predicted were possible on telephone channels and the notion that coding could reduce error probability without reducing data rate or causing bandwidth expansion.

In time sophisticated modulation and coding technology validated Shannon's theory such that on telephone lines today achieved data rates very close to Shannon capacity with very low probability of error. Unlike in the AWGN case, capacity of a flat-fading channel is not given by a single formula, since capacity depends on what is known about the time-varying channel at the transmitter and/or receiver. Moreover, for different channel information assumptions, there are different definitions of channel capacity, depending on whether capacity characterizes the maximum rate averaged over all fading states or the maximum constant rate that can be maintained in all fading states (with or without some probability of outage).

The flat-fading channel capacity is where only the fading distribution is known at the transmitter and receiver. Capacity under this assumption is typically very difficult to determine, and is only known in a few special cases. Next the channel fade level is known at the receiver only (via receiver estimation) or that the channel fade level is known at both the transmitter and the receiver (via receiver estimation and transmitter feedback) is considered. The fading channel capacity with channel side information at both the transmitter and receiver is achieved when the transmitter adapts its power, data rate, and coding scheme to the channel variation. The optimal power allocation in this case is a "water-filling" in time, where power and data rate are increased when channel conditions are favorable and decreased when channel conditions are not favorable.

The capacity of frequency-selective fading channels for time-invariant frequency-selective channels the capacity is known and is achieved with an optimal power allocation that water-fills over frequency instead of time. The capacity of a time-varying frequency-selective fading channel is unknown in general. However, this channel can be approximated as a set of independent parallel flat-fading channels, whose capacity is the sum of capacities on each channel with power optimally allocated among the channels. The capacity of this channel is known and is obtained with an optimal power allocation that water-fills over both time and frequency.

Capacity in AWGN

Consider a discrete-time additive white Gaussian noise (AWGN) channel with channel input/output relationship $y[i] = x[i] + n[i]$, where $x[i]$ is the channel input at time i, $y[i]$ is the corresponding channel output, and $n[i]$ is a white Gaussian noise random process. Assume a channel bandwidth B and transmit power P. The channel SNR, the power in $x[i]$ divided by the power in $n[i]$, is constant and given by $\gamma = P/(N0B)$, where $N0$ is the power spectral density of the noise. The capacity of this channel is given by Shannon's well-known formula

$$C = B \log2(1 + \gamma)$$

where the capacity units are bits/second (bps). Shannon's coding theorem proves that a code exists that achieves data rates arbitrarily close to capacity with arbitrarily small probability of bit error. The converse theorem shows that any code with rate $R > C$ has a probability of error bounded away from zero. The theorems are proved using the concept of mutual information between the input and output of a channel. For a memoryless time-invariant channel with random input x and random output y, the channel's mutual information is defined as

$$I(X;Y) = \sum_{x \in \mathcal{X}, y \in \mathcal{Y}} p(x,y) \log \left(\frac{p(x,y)}{p(x)p(y)} \right)$$

where the sum is taken over all possible input and output pairs $x \in X$ and $y \in Y$ for X and Y the input and output alphabets. The log function is typically with respect to base 2, in which case the units of mutual information are bits per second. Mutual information can also be written in terms of the entropy in the channel output y and conditional output $y|x$ as $I(X; Y) = H(Y) - H(Y|X)$, where $H(Y) = -\_y{\in}Y\ p(y) \log p(y)$ and $H(Y|X) = -\_x{\in}X{,}y{\in}Y\ p(x, y) \log p(y|x)$. Shannon proved that channel capacity equals the mutual information of the channel maximized over all possible input distributions:

$$C = \max_{p(x)} I(X;Y) = \max_{p(x)} \sum_{x,y} p(x,y) \log \left(\frac{p(x,y)}{p(x)p(y)} \right)$$

For the AWGN channel, the maximizing input distribution is Gaussian, which results in the channel capacity. For channels with memory, mutual information and channel capacity are defined relative to input and output sequences xn and yn. More details on channel capacity, mutual information, and the coding theorem. The proofs of the coding theorem and converse place no constraints on the complexity or delay of the communication system. Therefore, Shannon capacity is generally used as an upper bound on the data rates that can be achieved under real system constraints. At the time that Shannon developed his theory of information, data rates over standard telephone lines were on the order of 100 bps. However, breakthroughs in hardware, modulation, and coding techniques have brought commercial modems of today very close to the speeds predicted by Shannon in the 1950s. In fact, modems can exceed this 30 Kbps Shannon limit on some telephone channels, but that is because transmission lines today are of better quality than in Shannon's day and thus have a higher received power than that used in Shannon's initial calculation. On AWGN radio channels, turbo codes have come within a fraction of a dB of the Shannon capacity limit.

Capacity of Flat-Fading Channels

Channel and System Model

a discrete-time channel with stationary and ergodic time-varying gain _$g[i]$, $0 \leq g[i]$, and AWGN $n[i]$, as shown in Figure. The channel power gain $g[i]$ follows a given distribution $p(g)$, e.g. for Rayleigh fading $p(g)$ is exponential. Assume that $g[i]$ is independent of the channel input. The channel gain $g[i]$ can change at each time i, either as an i.i.d. process or with some correlation over time. In a block fading channel $g[i]$ is constant over some blocklength T after which time $g[i]$ changes to a new independent value based on the distribution $p(g)$. Let P denote the average transmit signal power, $N0/2$ denote the noise power spectral density of $n[i]$, and B denote the received signal bandwidth. The instantaneous received signal-to-noise ratio (SNR) is then $\gamma[i] = Pg[i]/(N0B)$, $0 \leq \gamma[i] < \infty$, and its expected value over all time is $\gamma = Pg/(N0B)$. Since $P/(N0B)$ is a constant, the distribution of $g[i]$ determines the distribution of $\gamma[i]$ and vice versa.

The system model is also shown in Figure, where an input message w is sent from the transmitter to the receiver. The message is encoded into the codeword x, which is transmitted over the time-varying channel as $x[i]$ at time i. The channel gain $g[i]$, also called the channel side information (CSI), changes during the transmission of the codeword.

The capacity of this channel depends on what is known about $g[i]$ at the transmitter and receiver. The three different scenarios regarding this knowledge are as follows

1. Channel Distribution Information (CDI): The distribution of $g[i]$ is known to the transmitter and receiver.

2. Receiver CSI: The value of $g[i]$ is known at the receiver at time i, and both the transmitter and receiver know the distribution of $g[i]$.

3. Transmitter and Receiver CSI: The value of $g[i]$ is known at the transmitter and receiver at time i, and both the transmitter and receiver know the distribution of $g[i]$. Transmitter and receiver CSI allow the transmitter to adapt both its power and rate to the channel gain at time i, and leads to the highest capacity of the three scenarios.

Note that since the instantaneous SNR $\gamma[i]$ is just $g[i]$ multiplied by the constant $P/(N0B)$, known CSI or CDI about $g[i]$ yields the same information about $\gamma[i]$.

Channel Distribution Information (CDI) Known

The case where the channel gain distribution $p(g)$ or, equivalently, the distribution of SNR $p(\gamma)$ is known to the transmitter and receiver. For i.i.d. fading the capacity is given , but solving for the capacity-achieving input distribution, i.e. the distribution achieving the maximum can be quite complicated depending on the fading distribution. Moreover, fading correlation introduces channel memory, in which case the capacity-achieving input distribution is found by optimizing over input blocks, which makes finding the solution even more difficult. For these reasons, finding the capacity-achieving input distribution and corresponding capacity of fading channels under CDI remains an open problem for almost all channel distributions.

The capacity-achieving input distribution and corresponding fading channel capacity under CDI is known for two specific models of interest: i.i.d. Rayleigh fading channels and FSMCs. In i.i.d. Rayleigh fading the channel power gain is exponential and changes independently with each channel use. The optimal input distribution for this channel is to be discrete with a finite number of mass points, one of which is located at zero. This optimal distribution and its corresponding capacity must be found numerically. The lack of closed-form solutions for capacity or the optimal input distribution is somewhat surprising given the fact that the fading follows the most common fading distribution and has no correlation structure. For flat-fading channels that are not necessarily Rayleigh or i.i.d. upper and lower bounds on capacity and these bounds are tight at high SNRs.

FSMCs to approximate Rayleigh fading channels model approximates the fading correlation as a Markov process. While the Markov nature of the fading dictates that the fading at a given time depends only on fading at the previous time sample, it turns out that the receiver must decode all past channel outputs jointly with the current output for optimal (i.e. capacity-achieving) decoding. This significantly complicates capacity analysis. The capacity of FSMCs has been derived for i.i.d. inputs and for general inputs. Capacity of the FSMC depends on the limiting distribution of the channel conditioned on all past inputs and outputs, which can be computed recursively. As with the i.i.d. Rayleigh fading channel, the complexity of the capacity analysis along with the final result for this relatively simple fading model is very high, indicating the difficulty of obtaining the capacity and related design insights into channels when only CDI is available.

Channel Side Information at Receiver

The case where the CSI $g[i]$ is known at the receiver at time i. Equivalently, $\gamma[i]$ is known at the receiver at time i. Assume that both the transmitter and receiver know the distribution of $g[i]$. In this case there are two channel capacity definitions that are relevant to system design: Shannon capacity, also called ergodic capacity, and capacity with outage. As for the AWGN channel, Shannon capacity defines the maximum data rate that can be sent over the channel with asymptotically small error probability. Note that for Shannon capacity the rate transmitted over the channel is constant: the transmitter cannot adapt its transmission strategy relative to the CSI. Thus, poor channel states typically reduce Shannon capacity since the transmission strategy must incorporate the effect of these poor states. An alternate capacity definition for fading channels with receiver CSI is capacity with outage.

Capacity with outage is defined as the maximum rate that can be transmitted over a channel with some outage probability corresponding to the probability that the transmission cannot be decoded with negligible error probability. The basic premise of capacity with outage is that a high data rate can be sent over the channel and decoded correctly except when the channel is in deep fading. By allowing the system to lose some data in the event of deep fades, a higher data rate can be maintained than if all data must be received correctly regardless of the fading state, as is the case for Shannon capacity. The probability of outage characterizes the probability of data loss or, equivalently, of deep fading.

Shannon (Ergodic) Capacity

Shannon capacity of a fading channel with receiver CSI for an average power constraint *P* can be obtained from

$$C = \int_0^\infty B \log_2(1 + \gamma) p(\gamma) d\gamma$$

Note that this formula is a probabilistic average, i.e. Shannon capacity is equal to Shannon capacity for an AWGN channel with SNR γ, given by $B \log_2(1 + \gamma)$, averaged over the distribution of γ. That is why Shannon capacity is also called ergodic capacity. However, care must be taken in interpreting as an average. In particular, it is incorrect to interpret to mean that this average capacity is achieved by maintaining a capacity $B \log_2(1 + \gamma)$ when the instantaneous SNR is γ, since only the receiver knows the instantaneous SNR $\gamma[i]$, and therefore the data rate transmitted over the channel is constant, regardless of γ. Note, also, the capacity-achieving code must be sufficiently long so that a received codeword is affected by all possible fading states. This can result in significant delay.

By Jensen's inequality

$$\mathbf{E}[B \log_2(1 + \gamma)] = \int B \log_2(1 + \gamma) p(\gamma) d\gamma \leq B \log_2(1 + \mathbf{E}[\gamma]) = B \log_2(1 + \bar{\gamma}),$$

where γ is the average SNR on the channel. Thus see that the Shannon capacity of a fading channel with receiver CSI only is less than the Shannon capacity of an AWGN channel with the same average SNR. In other words, fading reduces Shannon capacity when only the receiver has CSI. Moreover, without transmitter CSI, the code design must incorporate the channel correlation statistics, and the complexity of the maximum likelihood decoder will be proportional to the channel decorrelation time. In addition, if the receiver CSI is not perfect, capacity can be significantly decreased.

Capacity with Outage

Capacity with outage applies to slowly-varying channels, where the instantaneous SNR γ is constant over a large number of transmissions (a transmission burst) and then changes to a new value based on the fading distribution. With this model, if the channel has received SNR γ during a burst then data can be sent over the channel at rate $B \log_2(1 + \gamma)$ with negligible probability of error1. Since the transmitter does not know the SNR value γ, it must fix a transmission rate independent of the instantaneous received SNR.

Capacity with outage allows bits sent over a given transmission burst to be decoded at the end of the burst with some probability that these bits will be decoded incorrectly. Specifically, the transmitter fixes a minimum received SNR γmin and encodes for a data rate $C = B$ log2(1 + γmin). The data is correctly received if the instantaneous received SNR is greater than or equal to γmin. If the received SNR is below γmin then the bits received over that transmission burst cannot be decoded correctly with probability approaching one, and the receiver declares an outage. The probability of outage is thus $pout = p(\gamma < \gamma min)$.

The average rate correctly received over many transmission bursts is $C_o = (1 - p_{out})B \log_2(1 + \gamma_{min})$ since data is only correctly received on $1 - p_{out}$ transmissions. The value of γ_{min} is a design parameter based on the acceptable outage probability.

Channel Side Information at Transmitter and Receiver

When both the transmitter and receiver have CSI, the transmitter can adapt its transmission strategy relative to this CSI, as shown in Figure. In this case there is no notion of capacity versus outage where the transmitter sends bits that cannot be decoded, since the transmitter knows the channel and thus will not send bits unless they can be decoded correctly. This section will derive Shannon capacity assuming optimal power and rate adaptation relative to the CSI, as well as introduce alternate capacity definitions and their power and rate adaptation strategies.

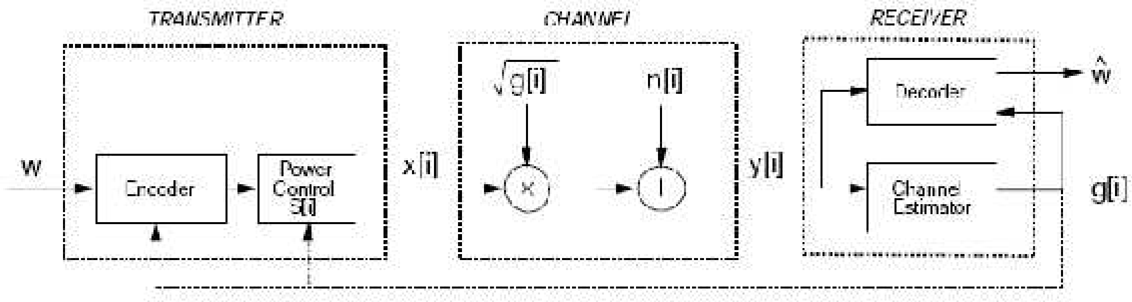

Figure System Model with Transmitter and Receiver CSI

Shannon Capacity

The Shannon capacity is considered when the channel power gain $g[i]$ is known to both the transmitter and receiver at time i. The Shannon capacity of a time-varying channel with side information about the channel state at both the transmitter and receiver was originally considered by Wolfowitz for the following model. Let $s[i]$ be a stationary and ergodic stochastic process representing the channel state, which takes values on a finite set S of discrete memoryless channels. Let C_s denote the capacity of a particular channel $s \in S$, and $p(s)$ denote the probability, or fraction of time, that the channel is in state s. The capacity of this time-varying channel is

$$C = \sum_{s \in S} C_s p(s)$$

the capacity of an AWGN channel with average received SNR γ is $C_\gamma = B \log2(1 + \gamma)$. Let $p(\gamma) = p(\gamma[i] = \gamma)$ denote the probability distribution of the received SNR. The capacity of the fading channel with transmitter and receiver side information is

$$C = \int_0^\infty C_\gamma p(\gamma) d\gamma = \int_0^\infty B \log_2(1 + \gamma) p(\gamma) d\gamma$$

so transmitter side information does not increase capacity unless power is also adapted.

Now allow the transmit power $P(\gamma)$ to vary with γ, subject to an average power constraint $\overline{P}$:

$$\int_0^\infty P(\gamma)p(\gamma)d\gamma \le \overline{P}$$

With this additional constraint, it cannot be applied directly to obtain the capacity. However, expect that the capacity with this average power constraint will be the average capacity given with the power optimally distributed over time. This motivates defining the fading channel capacity with average power constraint as

$$C = \max_{P(\gamma):\int P(\gamma)p(\gamma)d\gamma = \overline{P}} \int_0^\infty B\log_2\left(1 + \frac{P(\gamma)\gamma}{\overline{P}}\right)p(\gamma)d\gamma.$$

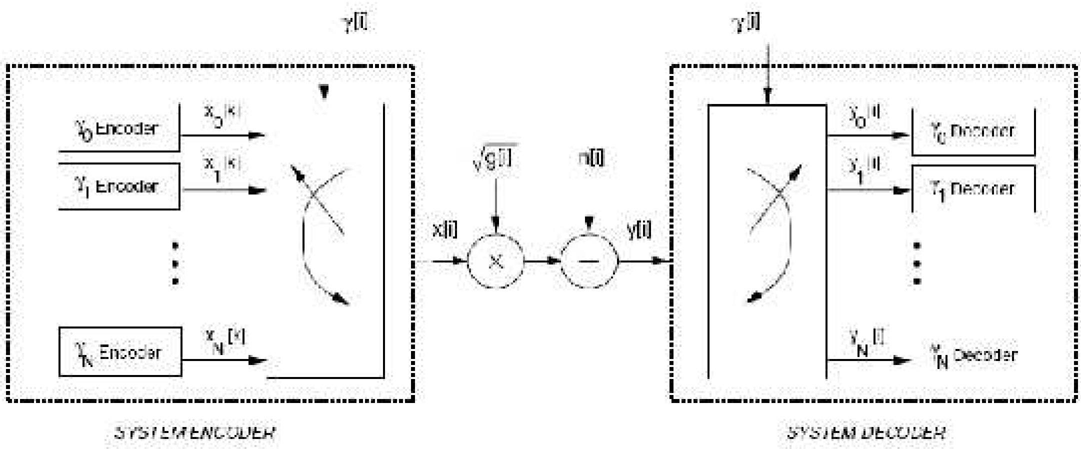

Figure Multiplexed coding and decoding

The main idea behind the proof is a "time diversity" system with multiplexed input and demultiplexed output, as shown in Figure. Specifically, first quantize the range of fading values to a finite set $\{\gamma_j : 1 \le j \le N\}$. For each γ_j, design an encoder/decoder pair for an AWGN channel with SNR γ_j. The input x_j for encoder γ_j has average power $P(\gamma_j)$ and data rate $R_j = C_j$, where C_j is the capacity of a time-invariant AWGN channel with received SNR $P(\gamma_j)\gamma_j/\overline{P}$. These encoder/decoder pairs correspond to a set of input and output ports associated with each γ_j. When $\gamma[i] \approx \gamma_j$, the corresponding pair of ports are connected through the channel. The codewords associated with each γ_j are thus multiplexed together for transmission, and demultiplexed at the channel output.

This effectively reduces the time-varying channel to a set of time-invariant channels in parallel, where the jth channel only operates when $\gamma[i] \approx \gamma_j$. The average rate on the channel is just the sum of rates associated with each of the γ_j channels weighted by $p(\gamma_j)$, the percentage of time that the channel SNR equals γ_j. This yields the average capacity formula.

To find the optimal power allocation $P(\gamma)$, the Lagrangian is formed.

$$J(P(\gamma)) = \int_0^\infty B\log_2\left(1 + \frac{\gamma P(\gamma)}{\overline{P}}\right)p(\gamma)d\gamma - \lambda\int_0^\infty P(\gamma)p(\gamma)d\gamma$$

Next differentiate the Lagrangian and set the derivative equal to zero.

$$\frac{\partial J(P(\gamma))}{\partial P(\gamma)} = \left[\left(\frac{B/\ln(2)}{1+\gamma P(\gamma)/\overline{P}}\right)\frac{\gamma}{\overline{P}} - \lambda\right] p(\gamma) = 0$$

Solving for $P(\gamma)$ with the constraint that $P(\gamma) > 0$ yields the optimal power adaptation that maximizes

$$\frac{P(\gamma)}{\overline{P}} = \begin{cases} \frac{1}{\gamma_0} - \frac{1}{\gamma} & \gamma \geq \gamma_0 \\ 0 & \gamma < \gamma_0 \end{cases}$$

for some "cutoff" value $\gamma 0$. If $\gamma[i]$ is below this cutoff then no data is transmitted over the ith time interval, so the channel is only used at time i if $\gamma 0 \leq \gamma[i] < \infty$.

$$C = \int_{\gamma_0}^{\infty} B \log_2\left(\frac{\gamma}{\gamma_0}\right) p(\gamma) d\gamma.$$

The multiplexing nature of the capacity-achieving coding strategy indicates that it is achieved with a time varying data rate, where the rate corresponding to instantaneous SNR γ is B log2($\gamma/\gamma 0$). Since $\gamma 0$ is constant, this means that as the instantaneous SNR increases, the data rate sent over the channel for that instantaneous SNR also increases. Note that this multiplexing strategy is not the only way to achieve capacity: it can also be achieved by adapting the transmit power and sending at a fixed rate.

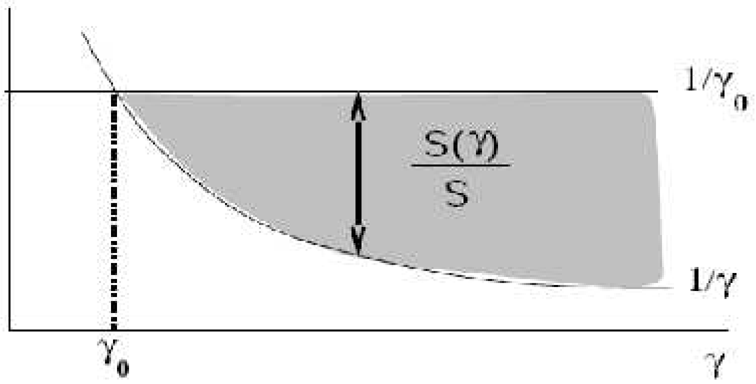

Figure Optimal Power Allocation: Water-Filling

Note that the optimal power allocation policy only depends on the fading distribution $p(\gamma)$ through the cutoff value $\gamma 0$. This cutoff value is found from the power constraint. Specifically, by rearranging the power constraint and replacing the inequality with equality (since using the maximum available power will always be optimal) yields the power constraint

$$\int_0^{\infty} \frac{P(\gamma)}{\overline{P}} p(\gamma) d\gamma = 1.$$

Now substituting the optimal power adaptation into this expression yields that the cutoff value $\gamma 0$ must satisfy

$$\int_{\gamma_0}^{\infty} \left(\frac{1}{\gamma_0} - \frac{1}{\gamma}\right) p(\gamma) d\gamma = 1$$

Note that this expression only depends on the distribution $p(\gamma)$. The value for $\gamma 0$ cannot be solved for in closed form for typical continuous pdfs $p(\gamma)$ and thus must be found numerically. Since γ is time-varying, the maximizing power adaptation policy is a "water-filling" formula in time, as illustrated in Figure. This curve shows how much power is allocated to the channel for instantaneous SNR $\gamma(t) = \gamma$. The water-filling terminology refers to the fact that the line $1/\gamma$ sketches out the bottom of a bowl, and power is poured into the bowl to a constant water level of $1/\gamma 0$. The amount of power allocated for a given γ equals $1/\gamma 0 - 1/\gamma$, the amount of water between the bottom of the bowl ($1/\gamma$) and the constant water line ($1/\gamma 0$). The intuition behind water-filling is to take advantage of good channel conditions: when channel conditions are good (γ large) more power and a higher data rate is sent over the channel. As channel quality degrades (γ small) less power and rate are sent over the channel. If the instantaneous channel SNR falls below the cutoff value, the channel is not used.

Note that the multiplexing argument sketching how capacity is achieved applies to any power adaptation policy, i.e. for any power adaptation policy $P(\gamma)$ with average power P the capacity

$$C = \int_{0}^{\infty} B \log_2 \left(1 + \frac{P(\gamma)\gamma}{P}\right) p(\gamma) d\gamma$$

can be achieved with arbitrarily small error probability. Of course this capacity cannot exceed where power adaptation is optimized to maximize capacity. However, there are scenarios where a suboptimal power adaptation policy might have desirable properties that outweigh capacity maximization.

Zero-Outage Capacity and Channel Inversion

Consider a suboptimal transmitter adaptation scheme where the transmitter uses the CSI to maintain a constant received power, i.e., it inverts the channel fading. The channel then appears to the encoder and decoder as a time-invariant AWGN channel. This power adaptation, called channel inversion, is given by $P(\gamma)/P = \sigma/\gamma$, where σ equals the constant received SNR that can be maintained with the transmit power constraint. The constant σ thus satisfies $\int \sigma/\gamma \, p(\gamma) d\gamma = 1$, so $\sigma = 1/\mathbf{E}[1/\gamma]$. Fading channel capacity with channel inversion is just the capacity of an AWGN channel with SNR σ

$$C = B \log_2 [1 + \sigma] = B \log_2 \left[1 + \frac{1}{\mathbf{E}[1/\gamma]}\right]$$

The capacity-achieving transmission strategy for this capacity uses a fixed-rate encoder and decoder designed for an AWGN channel with SNR σ. This has the advantage of maintaining a fixed data rate over the channel regardless of channel conditions. For this reason the channel capacity given is called zero-outage capacity, since the data rate is fixed under all channel conditions and there is no channel outage. Note that there exist practical coding techniques that achieve near-capacity data rates on AWGN channels, so the zero-outage capacity can be approximately achieved in practice.

Zero-outage capacity can exhibit a large data rate reduction relative to Shannon capacity in extreme fading environments. For example, in Rayleigh fading E[$1/\gamma$] is infinite, and thus the zero-outage capacity given is zero. Channel inversion is common in spread spectrum systems with near-far interference imbalances. It is also the simplest scheme to implement, since the encoder and decoder are designed for an AWGN channel, independent of the fading statistics.

Outage Capacity and Truncated Channel Inversion

The reason zero-outage capacity may be significantly smaller than Shannon capacity on a fading channel is the requirement to maintain a constant data rate in all fading states. By suspending transmission in particularly bad fading states (outage channel states) and can maintain a higher constant data rate in the other states and thereby significantly increase capacity. The outage capacity is defined as the maximum data rate that can be maintained in all nonoutage channel states times the probability of nonoutage. Outage capacity is achieved with a truncated channel inversion policy for power adaptation that only compensates for fading above a certain cutoff fade depth γ_0:

$$\frac{P(\gamma)}{\overline{P}} = \begin{cases} \frac{\sigma}{\gamma} & \gamma > \gamma_0 \\ 0 & \gamma < \gamma_0 \end{cases}$$

where γ_0 is based on the outage probability: $p_{out} = p(\gamma < \gamma_0)$. Since the channel is only used when $\gamma \geq \gamma_0$, the power constraint yields $\sigma = 1/\mathbf{E}_{\gamma_0}[1/\gamma]$, where

$$\mathbf{E}_{\gamma_0}[1/\gamma] \triangleq \int_{\gamma_0}^{\infty} \frac{1}{\gamma} p(\gamma) d\gamma$$

The outage capacity associated with a given outage probability p_{out} and corresponding cutoff γ_0 is given by

$$C(p_{out}) = B \log_2 \left(1 + \frac{1}{\mathbf{E}_{\gamma_0}[1/\gamma]}\right) p(\gamma \geq \gamma_0).$$

The maximum outage capacity by maximizing outage capacity over all possible γ_0

$$C = \max_{\gamma_0} B \log_2 \left(1 + \frac{1}{\mathbf{E}_{\gamma_0}[1/\gamma]}\right) p(\gamma \geq \gamma_0)$$

This maximum outage capacity will still be less than Shannon capacity since truncated channel inversion is a suboptimal transmission strategy. However, the transmit and receive strategies associated with inversion or truncated inversion may be easier to implement or have lower complexity than the water-filling schemes associated with Shannon capacity.

Capacity with Receiver Diversity

Receiver diversity is a well-known technique to improve the performance of wireless communications in fading channels. The main advantage of receiver diversity is that it mitigates the fluctuations due to fading so that the channel appears more like an AWGN channel. More details on receiver diversity and its performance will be given in unit 4. Since receiver diversity mitigates the impact of fading, an interesting question is whether it also increases the capacity of a fading channel. The capacity calculation under diversity combining first requires that the distribution of the received SNR $p(\gamma)$ under the given diversity combining technique be obtained. Once this distribution is known it can be substituted into any of the capacity formulas above to obtain the capacity under diversity combining. The specific capacity formula used depends on the assumptions about channel side information, e.g. for the case of perfect transmitter and receiver CSI the formula would be used.

Capacity under both maximal ratio and selection combining diversity for these different capacity formulas was computed. It was found that, as expected, the capacity with perfect transmitter and receiver CSI is bigger than with receiver CSI only, which in turn is bigger than with channel inversion. The performance gap of these different formulas decreases as the number of antenna branches increases. This trend is expected, since a large number of antenna branches makes the channel look like AWGN, for which all of the different capacity formulas have roughly the same performance.

Recently there has been much research activity on systems with multiple antennas at both the transmitter and the receiver. The excitement in this area stems from the breakthrough results in indicating that the capacity of a fading channel with multiple inputs and outputs (a MIMO channel) is M times larger then the channel capacity without multiple antennas, where $M = \min(Mt,Mr)$ for Mt the number of transmit antennas and Mr the number of receive antennas.

Capacity of Frequency-Selective Fading Channels

In this section, the Shannon capacity of frequency-selective fading channels is considered. First consider the capacity of a time-invariant frequency-selective fading channel. This capacity analysis is similar to that of a flat fading channel with the time axis replaced by the frequency axis..

Time-Invariant Channels

Consider a time-invariant channel with frequency response $H(f)$, as shown in Figure. Assume a total transmit power constraint P. When the channel is time-invariant it is typically assumed that $H(f)$ is known at both the transmitter and receiver: capacity of time-invariant channels under different assumptions of this channel knowledge

First assume that $H(f)$ is block-fading, so that frequency is divided into subchannels of bandwidth B, where $H(f) = Hj$ is constant over each block, as shown in Figure. The frequency-selective fading channel thus consists of a set of AWGN channels in parallel with SNR $|H_j|2P_j/(N0B)$ on the j^{th} channel, where P_j is the power allocated to the j^{th} channel in this parallel set, subject to the power constraint _j $P_j \leq P$.

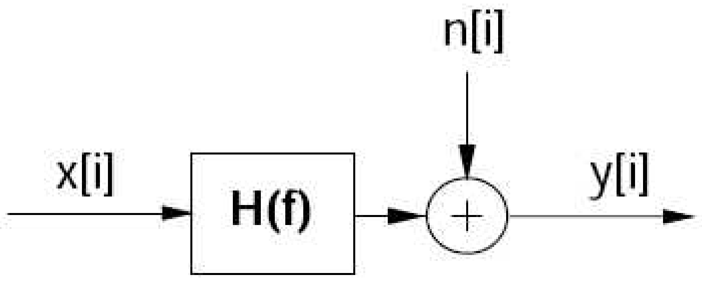

Figure Time-Invariant Frequency-Selective Fading Channel

The capacity of this parallel set of channels is the sum of rates associated with each channel with power optimally allocated over all channels

$$C = \sum_{\max P_j : \sum_j P_j \leq P} B \log_2 \left(1 + \frac{|H_j|^2 P_j}{N_0 B} \right)$$

Note that this is similar to the capacity and optimal power allocation for a flat-fading channel, with power and rate changing over frequency in a deterministic way rather than over time in a probabilistic way. The optimal power allocation is found via the same Lagrangian technique used in the flat-fading case, which leads to the water-filling power allocation.

$$\frac{P_j}{P} = \begin{cases} \frac{1}{\gamma_0} - \frac{1}{\gamma_j} & \gamma_j \geq \gamma_0 \\ 0 & \gamma_j < \gamma_0 \end{cases}$$

for some cutoff value *γ*0, where *γj* = |*Hj* |2*P/*(*N*0*B*) is the SNR associated with the *j*th channel assuming it is allocated the entire power budget.

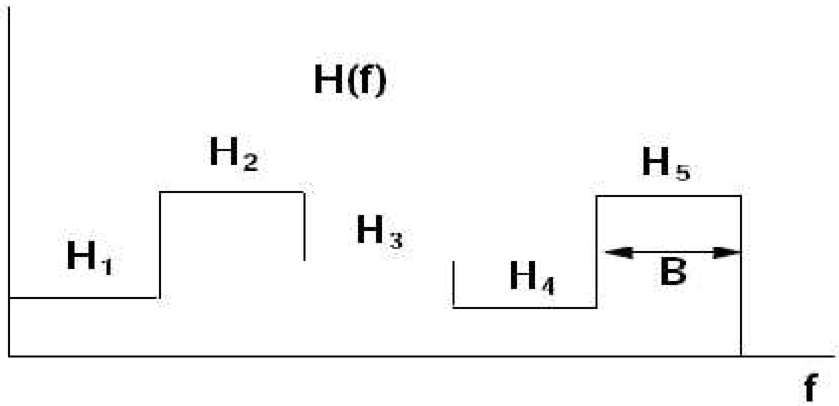

Figure Block Frequency-Selective Fading

This optimal power allocation is illustrated in Figure. The cutoff value is obtained by substituting the power adaptation formula into the power constraint, so *γ*0 must satisfy

$$\sum_j \left(\frac{1}{\gamma_0} - \frac{1}{\gamma_j}\right) = 1.$$

Then the capacity becomes

$$C = \sum_{j:\gamma_j \geq \gamma_0} B \log_2(\gamma_j/\gamma_0).$$

This capacity is achieved by sending at different rates and powers over each subchannel. Multicarrier modulation uses the same technique in adaptive loading

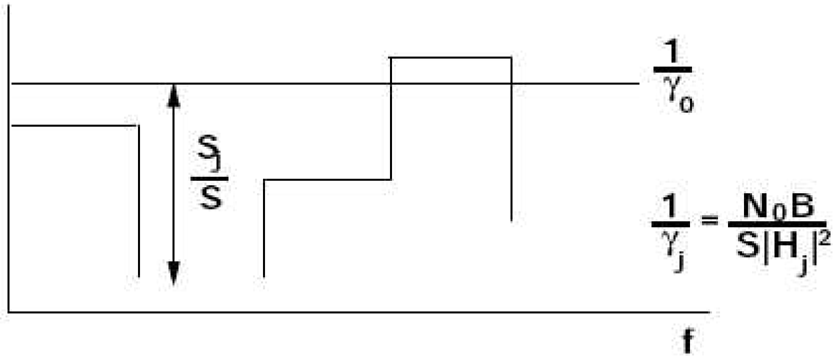

Figure Water-Filling in Block Frequency-Selective Fading

When $H(f)$ is continuous the capacity under power constraint P is similar to the case of the block-fading channel, with some mathematical intricacies needed to show that the channel capacity is given by

$$C = \max_{P(f): \int P(f)df \leq P} \int \log_2\left(1 + \frac{|H(f)|^2 P(f)}{N_0}\right) df.$$

The equation inside the integral can be thought of as the incremental capacity associated with a given frequency f over the bandwidth df with power allocation $P(f)$ and channel gain $|H(f)|^2$. This result is formally proven using a Karhunen-Loeve expansion of the channel $h(t)$ to create an equivalent set of parallel independent channels. The power allocation over frequency, $P(f)$, that maximizes is found via the Lagrangian technique. The resulting optimal power allocation is water-filling over frequency:

$$\frac{P(f)}{P} = \begin{cases} \frac{1}{\gamma_0} - \frac{1}{\gamma(f)} & \gamma(f) \geq \gamma_0 \\ 0 & \gamma(f) < \gamma_0 \end{cases}$$

This results in channel capacity

$$C = \int_{f:\gamma(f) \geq \gamma_0} \log_2(\gamma(f)/\gamma_0) df.$$

Time-Varying Channels

The time-varying frequency-selective fading channel is similar to the model shown in Figure except that $H(f) = H(f, i)$, i.e. the channel varies over both frequency and time. It is difficult to determine the capacity of time-varying frequency-selective fading channels, even when the instantaneous channel $H(f, i)$ is known perfectly at the transmitter and receiver, due to the random effects of self-interference (ISI). In the case of transmitter and receiver side information, the optimal adaptation scheme must consider the effect of the channel on the past sequence of transmitted bits, and how the ISI resulting from these bits will affect future transmissions. The capacity of time-varying frequency-selective fading channels is in general unknown, however upper and lower bounds and limiting formulas exist.

The channel capacity in time-varying frequency-selective fading is approximated by taking the channel bandwidth B of interest and divide it up into subchannels the size of the channel coherence bandwidth Bc. Then assume that each of the resulting subchannels is independent, time-varying, and flat-fading with $H(f, i) = Hj\,[i]$ on the jth subchannel.

Under this assumption, the capacity for each of these flat-fading subchannels based on the average power Pj that will allocate to each subchannel, subject to a total power constraint P is obtained. Since the channels are independent, the total channel capacity is just equal to the sum of capacities on the individual narrowband flat fading channels subject to the total average power constraint, averaged over both time and frequency

$$C = \max_{\{\overline{P}_j\}:\sum_j \overline{P}_j \le \overline{P}} \sum_j C_j(\overline{P}_j),$$

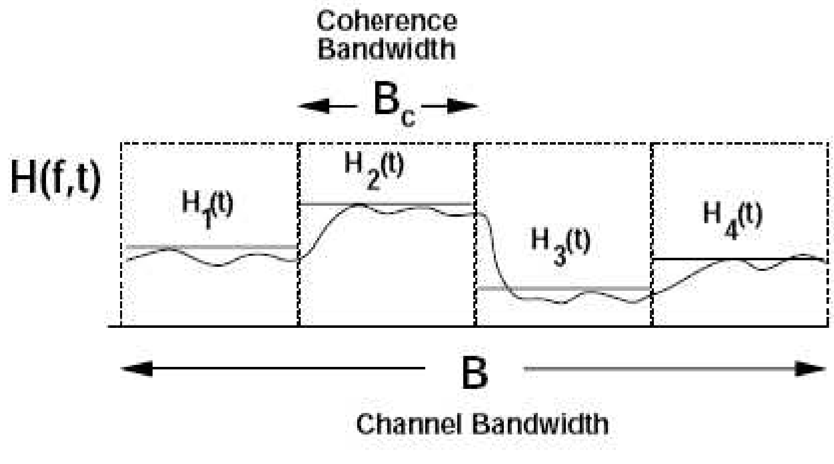

Figure Channel Division in Frequency-Selective Fading

where $Cj(Pj)$ is the capacity of the flat-fading subchannel with average power $P\ j$ and bandwidth Bc given for Shannon capacity under different side information and power allocation policies. $Cj(Sj)$ is as a capacity versus outage if only the receiver has side information.

Shannon capacity assuming perfect transmitter and receiver channel CSI, since this upper bounds capacity under any other side information assumptions or suboptimal power allocation strategies. Fix the average power per subchannel, the optimal power adaptation follows a water-filling formula. The optimal average power to be allocated to each subchannel should also follow a water-filling, where more average power is allocated to better subchannels. Thus the optimal power allocation is a two-dimensional water-filling in both time and frequency. This optimal two-dimensional water-filling and the corresponding Shannon capacity is obtained.

Define $\gamma_j[i] = |H_j[i]|^2 P/(N_0 B)$ to be the instantaneous SNR on the jth subchannel at time i assuming the total power P is allocated to that time and frequency. Now, allow the power $P_j(\gamma_j)$ to vary with $\gamma_j[i]$. The Shannon capacity with perfect transmitter and receiver CSI is given by optimizing power adaptation relative to both time (represented by $\gamma_j[i] = \gamma_j$) and frequency (represented by the subchannel index j):

$$C = \max_{P_j(\gamma_j):\sum_j \int_0^\infty P_j(\gamma_j)p(\gamma_j)d\gamma_j \le \overline{P}} \sum_j \int_0^\infty B_c \log_2\left(1 + \frac{P_j(\gamma_j)\gamma_j}{\overline{P}}\right) p(\gamma_j)d\gamma_j$$

To find the optimal power allocation $P_j(\gamma_j)$, the Lagrangian is formed as

$$J(P_j(\gamma_j)) = \sum_j \int_0^\infty B_c \log_2\left(1 + \frac{P_j(\gamma_j)\gamma_j}{\overline{P}}\right) p(\gamma_j)d\gamma_j - \lambda \sum_j \int_0^\infty P_j(\gamma_j)p(\gamma_j)d\gamma_j$$

Differentiating the Lagrangian and setting this derivative equal to zero eliminates all terms except the given subchannel and associated SNR

$$\frac{\partial J(P_j(\gamma_j))}{\partial P_j(\gamma_j)} = \left[\left(\frac{B/\ln(2)}{1 + \gamma_j P(\gamma_j)/\overline{P}}\right) \frac{\gamma_j}{\overline{P}} - \lambda\right] p(\gamma_j) = 0$$

Solving for $P_j(\gamma_j)$ yields the same water-filling as the flat fading case

$$\frac{P_j(\gamma_j)}{\overline{P}} = \begin{cases} \frac{1}{\gamma_0} - \frac{1}{\gamma_j} & \gamma_j \ge \gamma_0 \\ 0 & \gamma_j < \gamma_0 \end{cases},$$

where the cutoff value γ_0 is obtained from the total power constraint over both time and frequency

$$\sum_j \int_0^\infty P_j(\gamma)p_j(\gamma)d\gamma_j = \overline{P}.$$

Thus, the optimal power allocation is a two-dimensional waterfilling with a common cutoff value γ_0. Dividing the constraint by $\overline{P}$ and substituting in the optimal power allocation to get that γ_0 must satisfy

$$\sum_j \int_{\gamma_0}^\infty \left(\frac{1}{\gamma_0} - \frac{1}{\gamma_j}\right) p(\gamma_j)d\gamma_j = 1.$$

It is interesting to note that in the two-dimensional water-filling the cutoff value for all subchannels is the same. This implies that even if the fading distribution or average fade power on the subchannels is different, all subchannels suspend transmission when the instantaneous SNR falls below the common cutoff value $\gamma 0$. Substituting the optimal power allocation into the capacity expression yields

$$C = \sum_{j} \int_{\gamma_0}^{\infty} B_c \log_2 \left(\frac{\gamma_j}{\gamma_0} \right) p(\gamma_j) d\gamma_j .$$

Antennas for mobile terminal

The typical trend is personalization, which has been accelerated by the personal use of mobile phone systems to access personal information and data easily. This trend is observed in the use of wireless mobile systems. Antennas used for wireless mobile systems, including mobile phones, must be small in size, compact, and light in weight, and yet functional. In areas where 2G and 3G systems coexist, multiband antennas are required and, accordingly, small, compact, and lightweight multiband antennas have been used in both mobile terminals and base stations. Increased use of multimedia services has sped deployment of high speed, high data-rate transmission systems, for which advanced antenna systems like adaptive arrays and multiple-input multiple-output (MIMO) arrays have been developed.

Antennas for Base Stations

The basic design concept of 2G base station antennas does not essentially change for 3G systems. However, 3G systems use different modulation schemes, by which the transmission performance differs from that of 2G systems; thus there should be some differences in the antenna design.

The figure indicates general items that should be taken into consideration when a base station antenna is designed. Figure provides design issues in conjunction with requirements and antenna technologies. A major subject that affects antenna design in 3G systems is the pattern synthesis, which concerns beam shaping, multiband operation, downsizing, and sophistication of antenna systems. Beam shaping includes beam tilting and low sidelobe beams in the vertical plane, along with uniform coverage patterns, sector beams, and multibeams in the horizontal plane.

Multiband antennas are required for systems that provide services on both 2G and 3G systems. Downsizing has become a serious concern in constructing base stations, since with the remarkable increase in the number of subscribers, cell sizes have become smaller to increase channel capacity, and accordingly the number of base stations has increased. Thus base stations often must be installed in limited spaces and places where heavy weight is not allowed. As a result, base-station antennas are often required to be downsized by substituting smaller, more compact, and lighter-weight structures. In order to reduce the number of antennas, because of the installation in confined spaces, multiband and multibeam antennas are being developed.

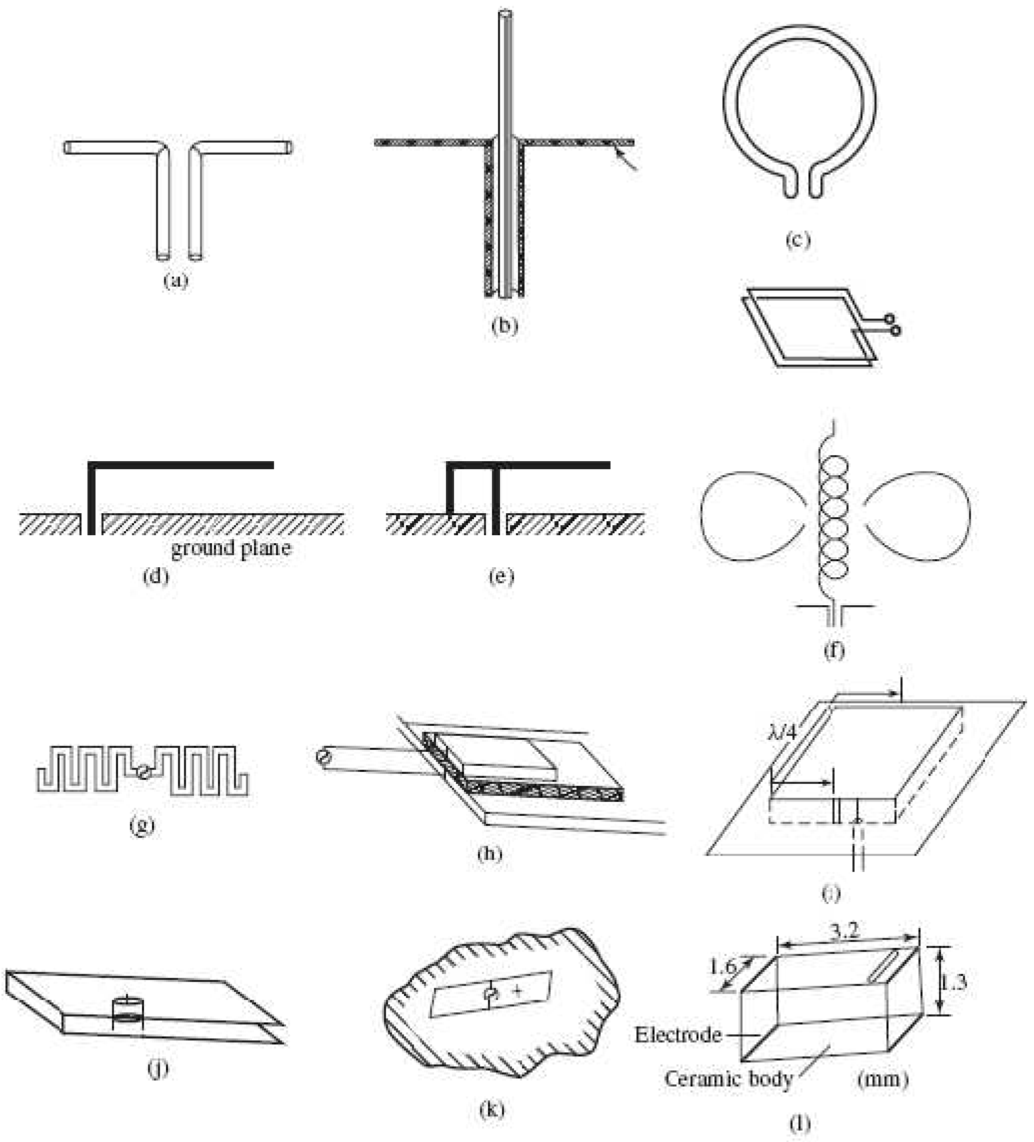

Figure Typical antenna elements: (a) dipole, (b) monopole, (c), loop, (d) inverted-L antenna (ILA), (e) inverted-F antenna (IFA), (f) normal mode helical antenna (NMHA), (g) meander line antenna (MLA), (h) microstrip antenna (MSA),(i) planar inverted-F antenna (PIFA), (j) parallel plate, (k) slot, and (l) chip antenna.

The figure depicts the design steps of base station antennas. Performance and cost sometimes are considered first, while determination of the electrical and mechanical parameters is taken in the second step. There also have been urgent requirements for increasing coverage areas in 3G systems, so that service areas can include closed areas like rooms in buildings, underpasses, inside tunnels, subway stations, and so forth. In order to service such areas, small relay stations have been developed, and again, small, compact, and lightweight antennas have been used for these stations.

Antennas used for these systems have planar structure. In areas where 2G and 3G systems coexist, the number of antenna elements may be increased to serve three bands, for example, 0.8, 1.5, and 2 GHz, depending on national standards. However, since space to install antennas often is limited to narrow areas, the total number of antennas should be reduced. To meet this requirement, multiband antennas that are small in size have been developed. Multibeam antennas have also been developed for covering sector zone areas.

Now that the mobile phone system is evolving toward 4G systems, and other wireless mobile systems are also seeing rapid growth, feasibility studies for smart antenna systems such as MIMO and adaptive arrays are being done and an operational test is being performed in the Tokyo metropolitan area.

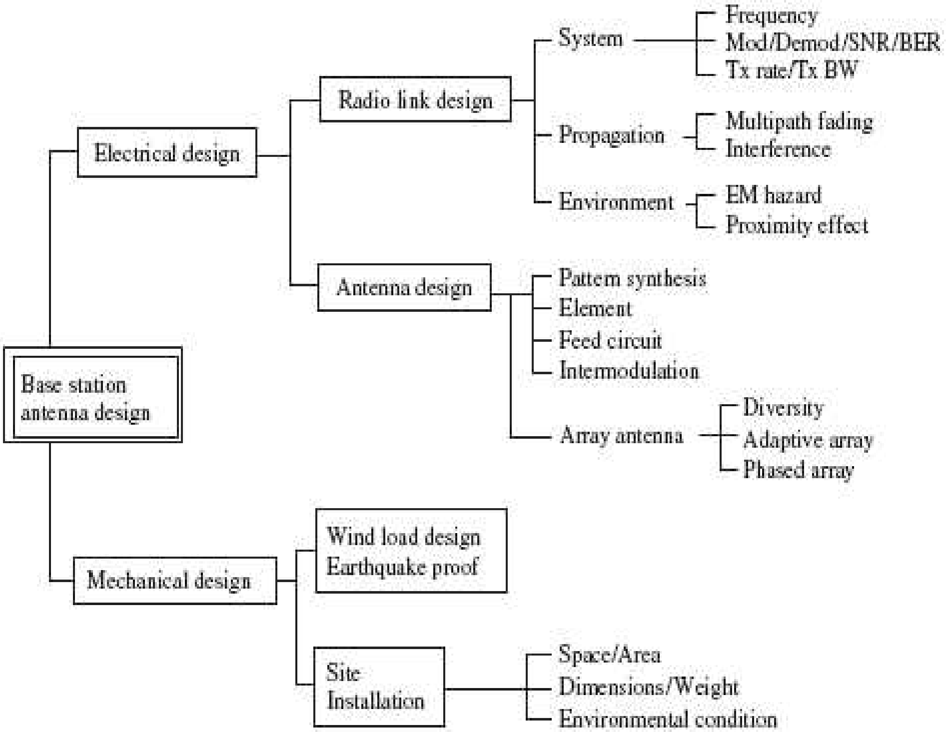

Figure Key items in designing a base station antenna

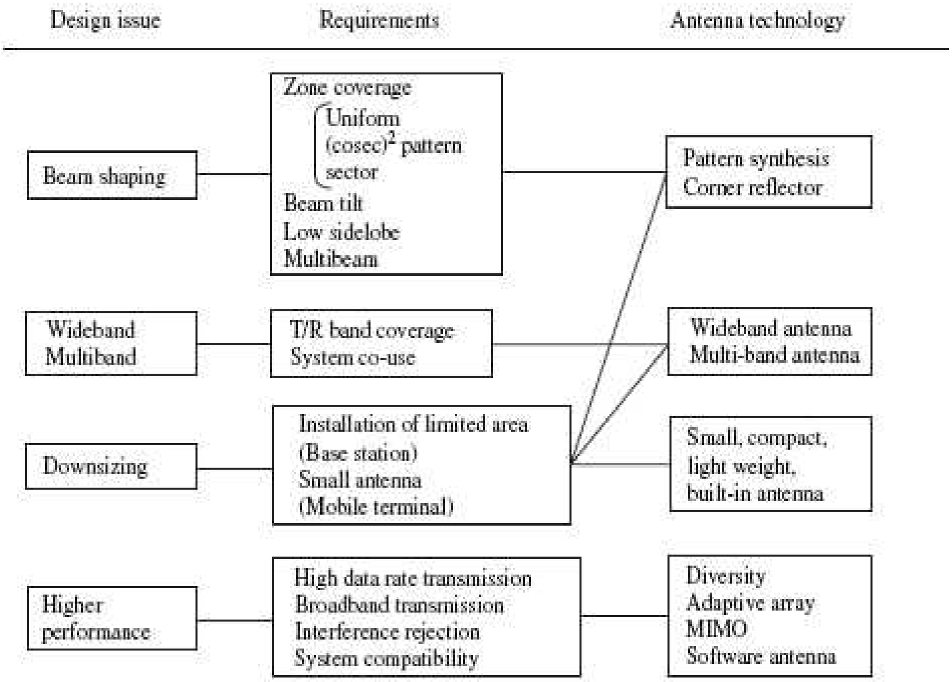

Figure Base station antenna design issues, requirements, and antenna technology

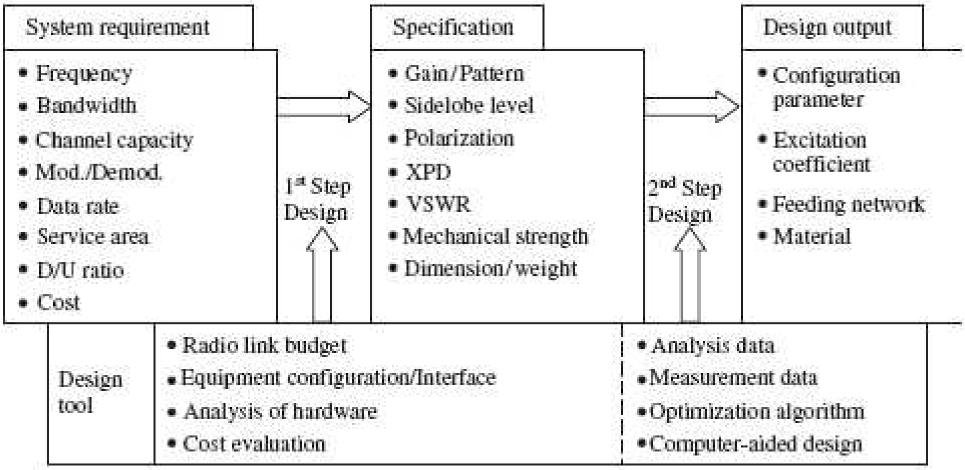

Figure Design steps of base station antennas.

Propagation Problems

Propagation problems differ depending on environmental conditions. Propagation in mobile communications occurs within a diffraction region and differs from that in free space. Path loss in free-space propagation is simply proportional to the square of the distance, while that in mobile communications generally depends on the operating frequency, antenna height, and particular environmental conditions around the antenna as well as the propagation path.

Mobile propagation environments are very complicated but are roughly categorized into four kinds

1. **Open Area:** There are few obstacles such as high trees or buildings in the propagation path. It can be said roughly that free spaces of about 300–400 m in length lie between the base and mobile stations. The propagation path is always on the line of sight (LoS).

2. **Suburban Area**: There are some obstacles around the mobile stations, but they are not dense. Roughly speaking, it is an area of trees and low houses.

3. **Urban Area:** There are many buildings or other high structures, and hence it is an area with high, close buildings, or a densely mixed area of buildings and high trees. No line of sight (NLoS) may exist in a typical propagation path.

4. **Closed Area:** Propagation is confined in limited areas like in buildings, tunnels, subway stations, and underpasses.

Base Station Antenna Techniques

Requirements for Base Station Antenna Systems

The Figure illustrates typical technology necessary for designing base station antennas. In order for the base station to communicate with the mobile stations located in the service area, base station antennas must radiate uniformly inside the area. Moreover, antenna gain should be as high as possible. Since the pattern of the service area is specified, antenna gain cannot be increased by narrowing the beamwidth in the horizontal plane.

Therefore it is customary that the antenna beam is narrowed in the vertical plane to increase gain; a vertical array of linear antennas is used for this purpose. Antennas with a gain in the range of 10–20 dBd are normally used for base station antennas in cellular systems.

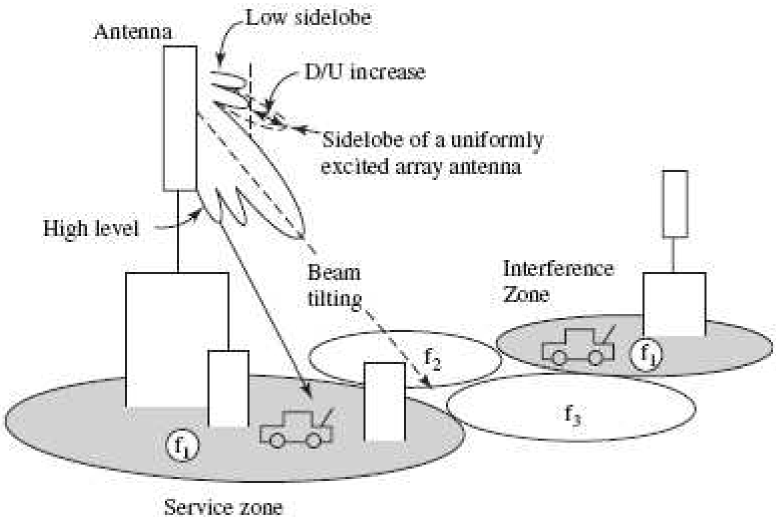

Figure Effect of side reduction for frequency reuse

Since a base station communicates with many mobile stations simultaneously, multiple channels must be handled. This requires wide-frequency characteristics and a function for branching and/or combining the channels. Hence base stations in the cellular system use one antenna for both transmitting and receiving. The required bandwidth of the antenna, for example, in the 800-MHz band Japanese PDC system, is more than 7% for the specified voltage standing wave ratio (VSWR) of less than 1.5. In addition, antennas should be designed for use with several systems (e.g., analog and digital systems). With that requirement, the antenna for the PDC in the 800-MHz band then needs frequency coverage between 810 and 960 MHz, a 17% relative bandwidth.

Furthermore, since 3G systems have started services in areas where 2G systems remain in service, antennas for both 2G and 3G systems should be designed. In 2G systems, the frequency-reuse concept is adopted in order to improve the channel capacity, and beam tilting is employed in base station antennas in order to avoid the co-channel interference between cells where the same frequency is used. In 3G systems, since the modulation scheme is CDMA, every cell uses the same frequency, and the beam-tilting concept is still applied to base station antennas. Interference suppression is still needed so as to increase the channel capacity. Shaped beam technology is still a major design subject.

Another important issue, which needs specific consideration in mobile communications, is to mitigate multipath fading, which deteriorates the signal quality in the narrowband modulation schemes. One way to overcome this problem is to apply diversity antenna systems.

Various antenna technologies for base stations other than diversity systems have been developed to enhance system performance. The adaptive array is one of them. Major subjects in the process of adaptive array development are communication frame format for beam forming, beam forming algorithm, array calibration, and simplification of hardware.

The world's first application of adaptive arrays for mobile systems was to the PHS base station antennas in Japan in 1998. The PHS system is a likely system to apply adaptive arrays, because PHS employs a TDD (time division duplex) system and allows slow movement of mobile terminals. Thus the estimation of the propagation channel property for the downlink can be used for the uplink, which makes feasible the application of adaptive arrays to mobile systems. The iBurst system evolved from the PHS system, and it adopted adaptive arrays for the purpose of interference rejection and SDMA since 2004.

Types of Antenna

Base station antenna configurations depend on the size and shape of the service area, the number of cells, and the number of channels. For limited service areas within a restricted angle in the horizontal plane, a corner reflector antenna is often used. When the service area is wide, as in a macrozone system, a linear array antenna, which has high directivity in the vertical plane, is used.

In the early stage of cellular system development, the base station antenna was mainly designed to achieve higher gain, and uniformly excited array antennas were usually used. However, the design concept of the base station antenna has been shifted from attainment of high gain to a greater ratio of desired-to-undesired signal strength (D/U), as cells have been divided into smaller subcells in order to increase the effectiveness of frequency reuse. Main-beam tilting, either mechanically or electrically, has been adopted throughout the world and co-channel interference can be reduced by about 10 dB. The beam tilting was recognized to be essential for enhancing frequency reuse. Suppression of sidelobes adjacent to the main beam, achieved by synthesizing array antenna patterns appropriately, was also effective in decreasing the distance between cells.

As for diversity antennas, space diversity, in which two antennas are used with separation of 5–10 wavelengths, has commonly been used. Other diversity schemes, such as pattern diversity and polarization diversity, have also been applied to base station antennas in commercial systems. Polarization diversity has gained a new understanding: better performance compared to space diversity has been recognized in urban areas, particularly in dense areas like central Tokyo. This is attributed to an increase in cross-polarization components in mobile phone propagation with the increase in mobile phone users. When a mobile phone is held in a normal talk position, the unit is slanted and produces both vertical and horizontal polarization components. Consequently, polarization diversity performs better than space diversity. Polarization diversity has been employed in CDMA-One systems and presently in IMT-2000 systems.

Base Station Antennas for Cellular Systems

The cell size has been made smaller in cellular mobile phone systems in order to increase the channel capacity in accordance with a rapid increase in the number of mobile phone users. Emphasis in designing antennas in 3G systems has been shifted from 2G systems to obtain lower sidelobes in the vertical plane and narrower sector beams in the horizontal plane in addition to beam tilting and multibeam capability. Downsizing is another significant design problem and thinning the antenna radome, in which an antenna array is housed, has made progress.

In order to keep the number of antenna elements from increasing, multiband antennas, which cover both 2G and 3G frequency bands, and multibeam antennas, which illuminate both 60° and 120° sectors, have been developed.

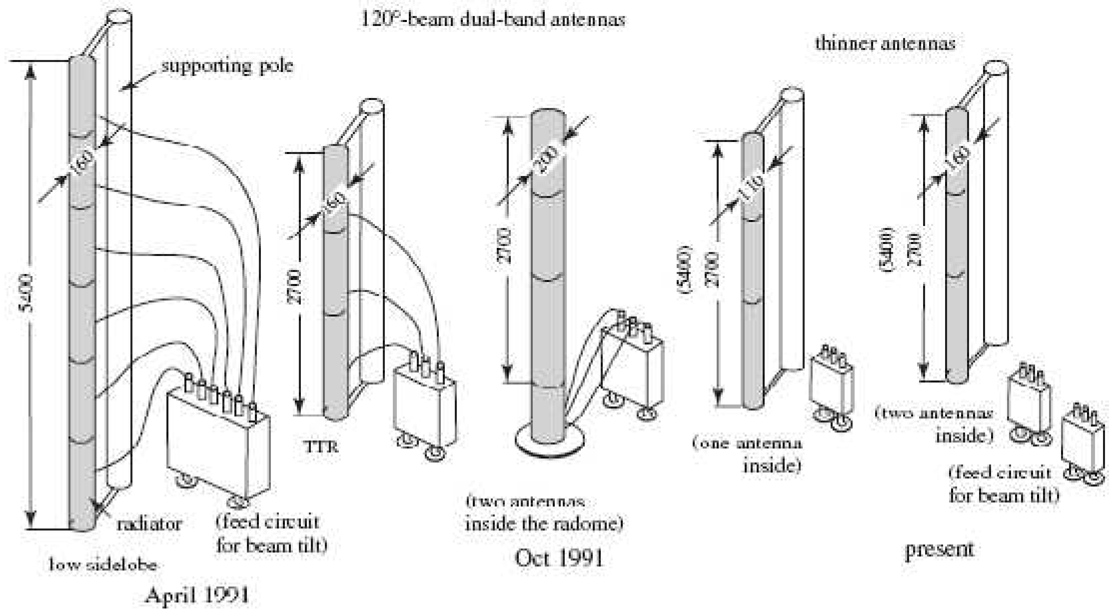

Figure Base station antenna.

Dual-Frequency Antennas

The inside view of a practical base station antenna is shown in Figure. As dual-band array elements, the dual radiator configuration is employed. Printed dipole configurations are used as the 900- and 1500-MHz elements. The 1500-MHz element is placed in front of the 900-MHz element. Stubs are inserted between array elements in order to suppress mutual coupling due to the close spacing of the elements at 0.6 wavelength. A reflector placed behind these radiators achieves a 120° beam in the horizontal plane. The radome diameter is 100 mm.

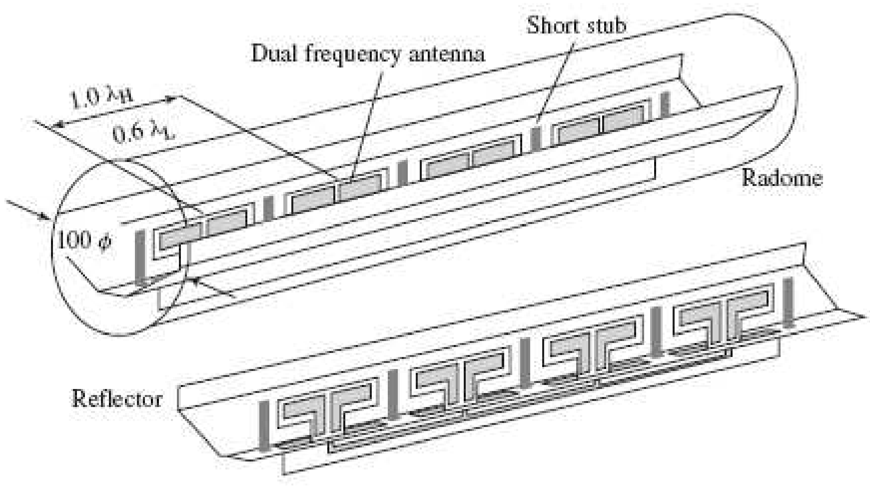

Figure Internal view of a base station antenna.

Triple-Band Antenna

IMT-2000 system uses the 2-GHz band and antennas additional to existing dual-band antennas or exchange of dual-band antennas have been required. Since there is little space to install additional antennas, multiband antennas have been developed: two types of dual-band antennas, which operate at 900 MHz/2 GHz and 1.5/2 GHz and one triple-band antenna, covering 0.9-, 1.5-, and 2-GHz bands. In IMT-2000 systems, new sector-beam antennas having 60◦ beamwidth are required in addition to 90◦ and 120◦ sector-beam antennas. By adding a parasitic element to dual-band antennas, triple-band antennas have been attained. Two horizontal parasitic elements are used for increasing the bandwidth at 900-MHz bands

Dual-Beam Antenna

In cellular systems, six-sectored cells are employed to increase the system capacity. Then a maximum of 18 antennas are needed when a diversity system is applied. Thus reduction in the number of antennas is desired to ease the antenna installation on the tower.

Diversity Antenna

Space Diversity

Shared use of an antenna element for reception and transmission is mandatory for reducing the number of antennas. The approach to accommodate two sector elements inside a cylindrical radome is being applied to most base station antennas in order to reduce the apparent number of antennas for multisector cell sites. An example of this scheme is illustrated in Figure, which is a cross-sectional view of a three-sector antenna system consisting of two antennas, each of which covers a corresponding sector. In the 3G system in Japan, a six-sector zone system has been adopted in the 2-GHz band in addition to the three-sector zone system in 900- and 1500-MHz bands. Thus two types of antennas are required, one with 120°beamwidth (BW) for all three bands, and another with 60◦ BW for 2-GHz and 120°BW for both 900-MHz and 1.5-GHz bands. Three more 60° BW antennas at 2GHz are needed in order to cover the whole area. Two 60◦ BW antennas are placed side by side near the center of two 120° BW antennas for the lower frequency bands. The six-sector space diversity scheme is shown in Figure

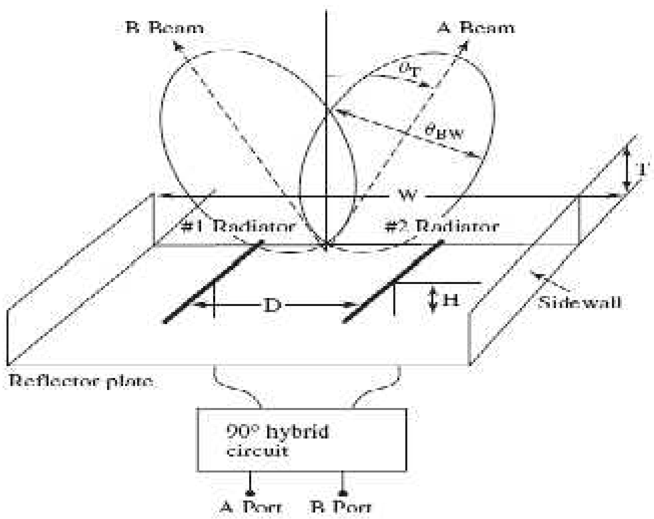

Figure Configuration of dual-beam antenna

Polarization Diversity

Polarization diversity, on the other hand, does not require two spatially separated antennas. Multiple dipole elements with orthogonal polarization can be alternately mounted on a piece of dielectric substrate in a vertical radome. Elements for the orthogonal polarization may be vertical/horizontal or $+45^{\circ}/-45^{\circ}$ crossed dipoles, depending on the particular design. Polarization diversity is a well-known diversity technique; however, it has not been used in cellular phone systems. This was because it was not necessarily useful in the propagation environment where mobile terminals were mostly automobiles and vertical-polarization components dominated, as the mobile stations employed were a vertical trunk-lid element. However, as the number of hand-held phone subscribers has increased, the propagation condition has changed. Horizontal-polarization components sometimes exceed vertical-polarization components, as users hold their phones in a tilted position, typically 60° from zenith in a talk position. The polarization diversity is then recognized as effective in the field where both vertical-and horizontal-polarization components exist.

Antennas for Micro/Pico Cellular Systems

Cellular phone service has been extended to inside tunnels and subway stations, inside large buildings, in underground shopping malls, and so forth. In order to provide services for these areas, the relay station is installed between the outside base station and the areas to be covered. The relay station has a booster, which receives a downlink signal from the outside station, amplifies it, and reradiates it. As for the uplink signal, the operation is reversed in the same way. A flat antenna is used for the booster.

For example, for tunnel systems, a low profile, two-element half-wave dipole antenna is used. The antenna is installed on the wall of a tunnel with very low height. The radiation pattern is a figure eight in both *E*-and *H*-planes, having nulls in the direction normal to the antenna element. In the tunnel system, an optical fiber cable is used to connect the outside station with the relay station, since the optical fiber has very low transmission loss and wide bandwidth. The optical signal is modulated with both 900-MHz and 1.5-GHz signals directly and conveyed by the optical fiber. In the downlink, from the outside station to the relay station, the received signal is reradiated by a low gain antenna to the relay station. The antennas are flat types and installed on the sidewall of a building or some other construction. The antennas are arranged one by one with separation of about a hundred meters. The reverse system is used for the uplink. The subway link uses a similar system, by which mobile phone service can be made available in the subway stations and even on the trains.

Antennas for Personal Handy-phone System (PHS)

In the PHS system, a TDD (time division duplex) transmission scheme is employed. This system provides service at the 1900-MHz band. A block diagram of the transmission systems with transmission diversity is shown in Figure. The base station (BS) has two antennas and receivers. The portable station (PS) has only one antenna and receiver. In a TDD system, a single carrier frequency is used to provide two-way communication (upward channel—PS to BS; and downward channel—BS to PS). The BS is able to predict the received-signal strength at the PS because of the reciprocity between upward and downward channels.

The BS receives an upward link signal from the PS using the diversity reception method and measures the received signal strength during a receiving period. In addition, it predicts which antenna gives the highest received signal strength at the PS. Then the selected antenna is used for transmission.

Adaptive Array Antenna System

AAAS in PHS

Among the endeavors to reduce the interference and improve the spectrum efficiency, as transmitting data and the number of users are increasing, study of the adaptive array antenna has been one of the most significant subjects to which antenna engineers devoted attention.

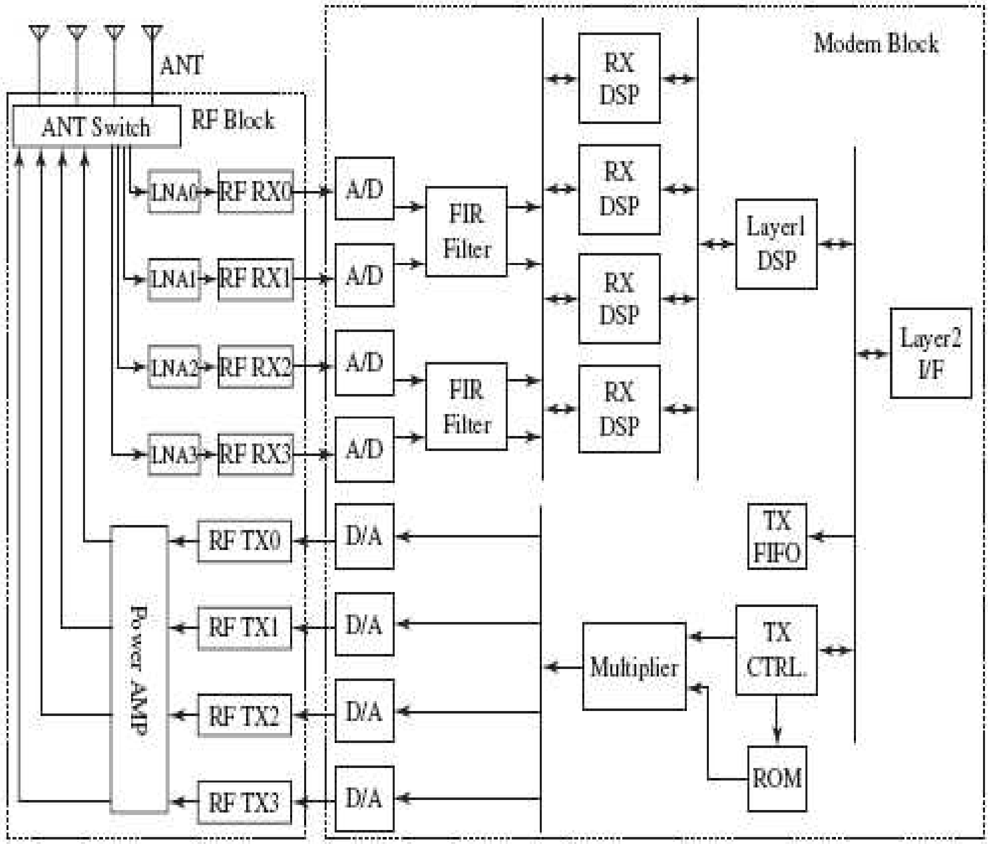

Figure Block diagram of adaptive antenna system for base station.

The adaptive array system was first introduced in 1998. The block diagram of an adaptive base station antenna applied to PHS is shown in Figure. Four antennas are used for transmission and reception. Adequate weight is determined in the CPU module for the output signal of each antenna to achieve the best bit error rate (BER) value by using a constant modulus algorithm (CMA) concept.

Antenna radiation pattern maximum is achieved in the desired signal direction, and pattern nulls are achieved in the undesired signal directions. Each antenna has gain of 10 dBi. Antenna spacing is 5 wavelengths. A radio unit is installed at the foot of the antenna.

AAAS in Wireless Local Loop (WLL) Systems

An adaptive array antenna system (AAAS) applied to WLL systems is called Super PHS–WLL, because the PHS standards are applied to the WLL, and wideband radios and spatial channel processing are employed to enhance system performance. By means of adaptive beam forming, the channel capacity is increased, the number of multipath signals is reduced, the coverage is expanded, and the flexible configuration of the coverage for each base station to match the local propagation environment is made feasible.

The system concept is briefly illustrated in Figure. The figure shows an example of a system configuration, which is composed of a BS (base station), single-and four-line SU (subscriber unit), and connection to the local exchange. The operating frequency of the system is the same as that of the PHS, 1880–1930 MHz, and the TDMA–SDMA system for the channel access and the duplex system, TDD (time division duplex), are employed.

The AAAS, which produces multiple beams by digital beam forming (DBF) technology, performs the function of SDMA. The beams are automatically directed to multiple SUs, wherever they are located. The received signal at the BS in the multipath environment is processed adaptively to enhance the processing gain, thus improving the quality of the link. It also contributes to interference rejection by directing a null against it, and enhancing the reuse of the same frequency, thereby increasing the number of channels.

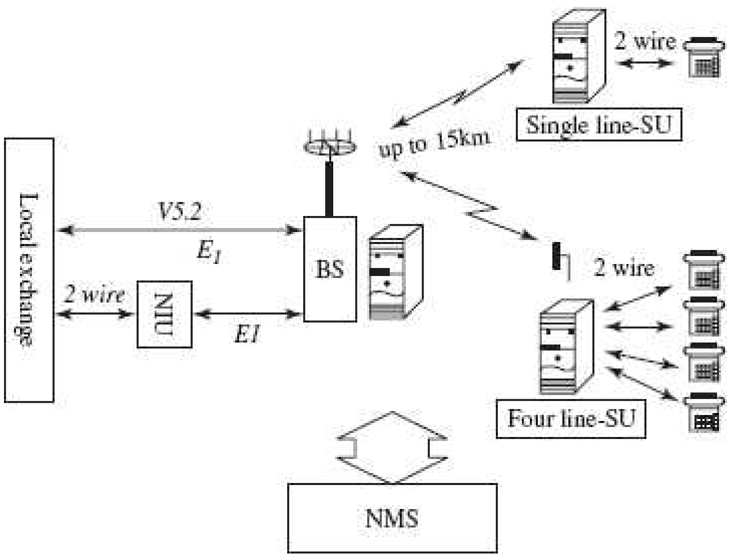

Figure System concept of PHS WLL.

System performance that can dynamically adapt to changeable propagation environments is achieved by DBF technology. The DBF software processes, in real time, the phase and time differences among the incoming signals in the multipath environment so that the signal transmission can be directed to each of the desired SUs. By this means, a stable and robust link can be established. Other factors that contribute to enhancing the channel capacity are the multiple frequency operation by using multiple wideband radios in addition to the AAAS, and use of the PHS standard TDD, which has four time slots in one data frame that can be used for both transmitting and receiving on four channels simultaneously on one frequency.

A base station of the super PHS–WLL system, which is in operation, is composed of 12 antennas and 12 radios and uses 16 radiofrequencies for the 16-way multiplex operation. By combining the spatial channel processing with the AAAS, a spatial channel efficiency of about 2.5 is achieved. Consequently, the voice-channel capacity achieved is up to 155; that is, 4*(*time slot*)* × 16*(*frequency*)* × 2.5 (spatial channel efficiency) −5 (control channel). It can serve 2730 subscribers as the total traffic capacity and is capable of covering a range up to about 15 km.

AAAS in the iBurst Systems

The iBurst system is one of the latest broadband wireless access systems, which features efficient frequency utilization achieved by means of an adaptive modulation scheme and application of an AAAS. The iBurst system was first introduced in Australia and began commercial service in 2000. It is presently deployed in some countries in southern Africa and Asia, while many other countries such as the United States, Canada, and Japan are closely investigating introduction of the system. Kyocera Corporation is playing a major role in the development and commercialization of the system. The system has been standardized by the ANSI as one of the high capacity (HC) SDMA systems and also will be taken up by the IEEE 802.20 Committee to be included in the standard as a wideband mobile broadband system (WMBS).

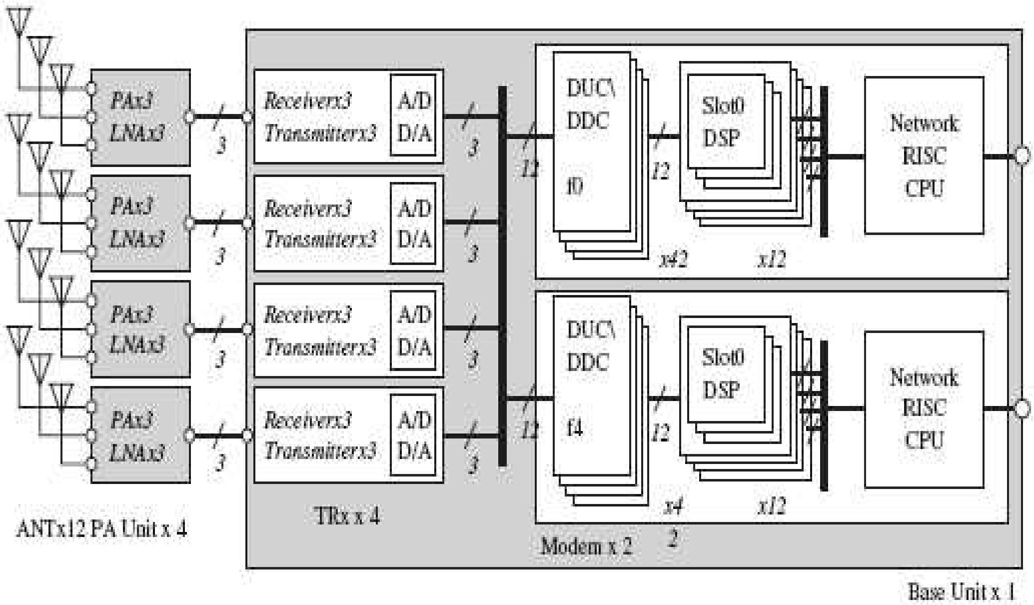

Figure Block diagram of iBurst base station unit.

The iBurst system technology is based on PHS (Personal Handy-phone System), which employs a TDD/TDMA system. The TDD system is advantageous for realizing the AAAS, as the channel estimation, which is necessary for adaptive control of the antenna pattern, can easily be made, since the channel uses the same frequency for both transmitting and receiving. In addition, the system employs very advanced technologies such as an adaptive modulation scheme and advanced SDMA system. The block diagram of the system is shown in Figure and shows a base station installed on the top of a building, where a 12-element array and base station units, including T/R circuits, are shown.

Antennas for Small Mobile Terminals

Design of Small Mobile Terminal Antennas

In designing antennas for small mobile terminals typically handsets, factors that should be taken into consideration are as follows:

- Small size
- Light weight
- Compact structure
- Low profile or flush mount
- Robustness
- Flexibility
- Low cost

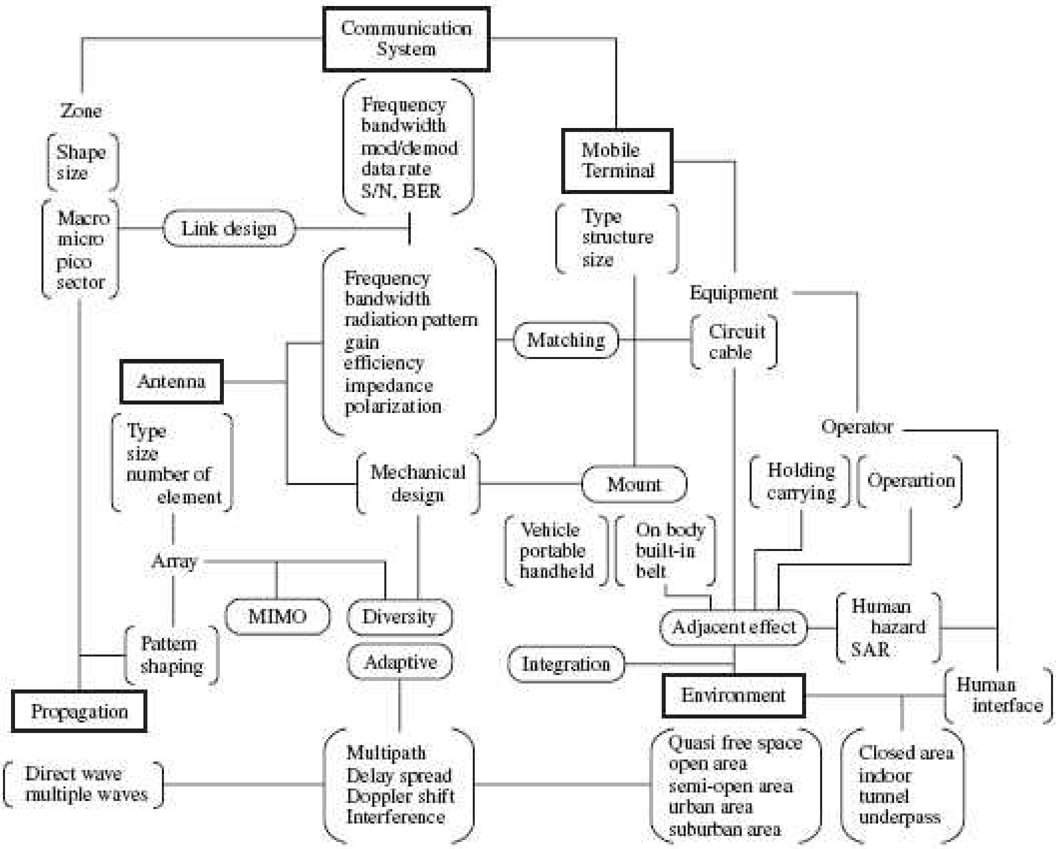

Figure Design parameters necessary for mobile terminal antennas

For specific applications, there may be other important requirements for the design. Parameters pertaining to not only antennas but also communication systems, propagation problems, and environmental conditions are summarized in Figure. In fact, all of these parameters, more or less, should be taken into consideration when handset antennas are designed.

When systems require particular specifications, antenna design should be made specifically to satisfy the requirements. Examples of such requirements are multiband operation and built-in structure. In GSM systems, various types of built-in multiband antennas have been used in small mobile terminals. In addition, after introduction of 3G systems, there have been areas where both 2G and 3G systems operating with different frequency bands are in service and hence multiband antennas have been installed in mobile terminals operating in these areas. In addition, there have been various small mobile terminals to which nonmobile phone systems such as Bluetooth, RFID, NFC (near-field communication), and mobile WiMAX are added and function not only for the purpose of communication but also for control, data transmission, and identification.

Accordingly, antennas that operate at frequencies corresponding to such systems should also be mounted. Propagation problems are presently diversified to include complicated environments such as inside buildings and subway stations and in underpasses. In addition, the problem now is to include higher frequency regions such as microwaves and millimeter waves.

Typical Antenna Elements Used for Mobile Terminals

Typical antenna elements used for small mobile terminals are MP (monopole), NMHA (normal mode helical antenna), PIFA (planar inverted-F antenna), ceramic chip, meander line, and MSA (microstrip antenna). A majority of mobile terminals such as PDC handsets employ an MP element and a built-in PIFA as a pair of elements for a diversity antenna. The NMHA element is also used in small mobile terminals, where the NMHA is placed on the top of an MP element and operates when the MP is retracted. The NMHA element is used as the main antenna of PHS mobile terminals, since the operating frequency is 1.5 GHz, allowing the antenna length to be short enough to be fixed on the unit body. Present trends are to employ a built-in antenna. The most popular built-in antenna is a type of PIFA, but with various modifications from the original structure. Many GSM handsets have used such modified types of built-in, multiband PIFAs. Another antenna element has a meander line structure, which is useful in smaller-sized antennas. The smallest antenna presently used is a type of ceramic-chip antenna, having a size of several cubic millimeters; however, it is necessary to use some appropriate size of ground plane, on which the chip antenna is placed.

Monopole and Dipole

A monopole is the most simple, thin, lightweight, low cost antenna element useful for small mobile terminals. It has long been used as the typical antenna element for small mobile terminals. The dipole antenna also has features of being simple, thin, lightweight, and low cost; however, it wasn’t used until 3G systems, because the dipole length is too long to mount on the mobile terminals for frequency bands below 1 GHz. Antenna performance depends not only on the antenna element but also on dimensions of the ground plane on which the antenna element is mounted. Figure illustrates the gain of a MP with respect to the length of the unit, which is assumed to be a rectangular conducting box, simulating a handset unit.

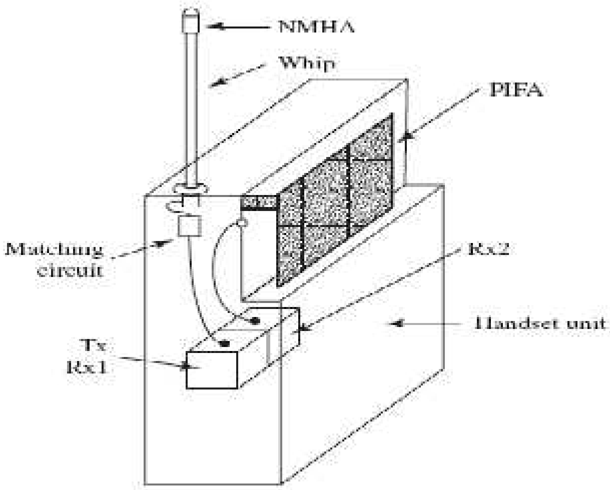

Figure Antenna system in a PDC mobile terminal

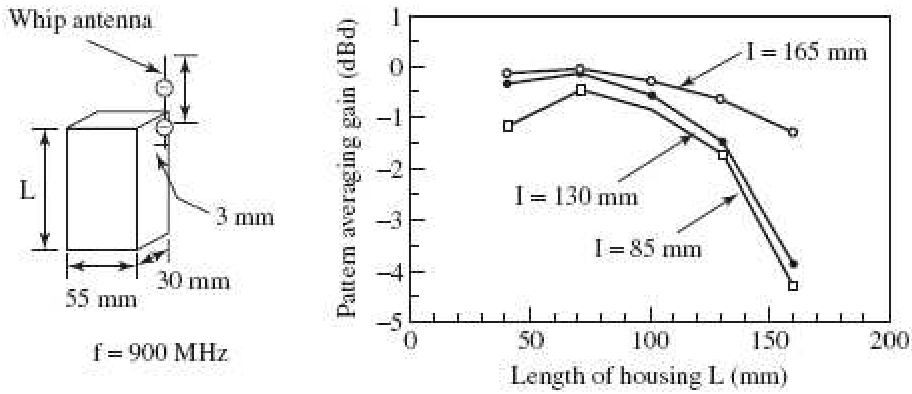

Figure Gain of a monopole with respect to the length

The dimensions of the box are shown in the figure. Radiation currents flow heavily on the handset unit as well as the antenna element when the length of the unit is about 14 λ; however, the handset-unit current is reduced remarkably when the monopole length is about 12 λ. Nevertheless, that is still not advantageous, because the input impedance is too high to match the load impedance. When the element length is either 38 λ or 58 λ, handset current flows are relatively small and yet the input impedance characteristics are adequate to match the load impedance. This concept has been applied to the design of PDC handset antennas.

Inverted-F Antenna (IFA)

The IFA can be treated equivalently as an antenna composed of two parts: an ILA with its image, and 2 two-wire shorted transmission lines. The two-wire transmission lines compensate capacitive impedance of the ILA and at the same time step up the input impedance so that matching to the 50-ohm load is made feasible. The input impedance of an IFA is expressed by

$$Z\text{in} = 4Z\text{a} \cap 2Zb$$

where Za denotes the input impedance of the ILA with its image, Zb is the impedance of the shorted two-wire transmission line, and ∩ expresses parallel combination of these two impedances.

Planar Inverted-F Antenna (PIFA)

The PIFA seems to be an antenna modified from an IFA by replacing the linear horizontal element of the IFA with a planar element. That is misleading, however. The radiation principle of the PIFA differs from that of the IFA. The source of radiation of a PIFA is the magnetic current on the peripheral aperture of the planar element, whereas that of the IFA is the electric currents on the linear F-shaped vertical and horizontal elements. In addition, the length of the peripheral aperture of the PIFA is about a half-wavelength, while that of the IFA is about a quarter-wavelength.

Study of the PIFA structure initially starts with a half-wave slot placed on the side of a rectangular conducting body as shown in Figure. The slot is fed at a point near a shorted end in order to attain adequate impedance to match the load impedance. The idea was born in the process of developing handset antennas. The basic concept was to develop a small, low profile antenna, suitable for a handset antenna.

In order to have a low profile or flush-mounted structure, use of a magnetic current as a source of radiation was considered. The slot antenna satisfies this requirement. Meanwhile, the microstrip antenna (MSA) had been known as a useful planar antenna, having a small and low profile structure. Then an attempt was made to place that slot antenna on a ground plane to form a planar structure like an MSA and the performance was confirmed to be satisfactory for handset use. Finally, the antenna configuration was arranged to have a structure as illustrated in Figure, in which the shorted part is reduced to make the antenna size smaller. In a PIFA, its peripheral aperture of a half-wavelength contributes to radiation, whereas in the MSA a patch-end aperture of about a quarter-wavelength contributes to the radiation.

In fact, an attempt was made initially to increase the bandwidth of a linear IFA by replacing the linear horizontal element with a planar element; however, it was not successful, as the bandwidth increase obtained was only 1–2%, unless the height was otherwise increased.

An example of the input impedance of a similar antenna model is shown in Figure. The equivalent circuit of this antenna model is shown in Figure. It is interesting to note that in the equivalent circuit there is a reactance jX . The bandwidth enhancement of an antenna system, being composed of a PIFA element placed on the handset unit, may be attributed to this reactance. This reactance has constant impedance over a much wider bandwidth than that of the antenna system. In practice, there are many variations in a PIFA to be found. In fact, some can hardly be identified as a PIFA, because the modified antennas have entirely different appearances from the original structure.

APPLICATIONS

Example of Practical antenna and base station units

Example of Monopole antenna in mobile application

POST TEST-MCQ TYPE

1. Friis free space equation
   1. Is an expression for noise power
   2. Is a function of transmitting and receiving antenna gain
   3. Depends upon the distance between transmitting and receiving antenna

a. 1 and 2 are correct
b. 1 and 3 are correct
c. 2 and 3 are correct
d. All are correct

2. According to Friis free space equation
   1. Received power falls with square of the distance between the transmitter and receiver
   2. Increases with square of the distance between the transmitter and receiver
   3. Received power increases with gains of transmitting and receiving antennas

a. 1 and 2 are correct
b. 1 and 3 are correct
c. All the three are correct
d. 2 and 3 are correct

3. Centre excited hexagonal cells use
a. Sectored directional antennas
b. Omni directional antennas
c. Yagi uda antennas
d. None of the above

4. The irreducible error floor in a frequency selective channel is primarily caused by
a. Errors due to ISI
b. Adjacent Channel Interference
c. Co-channel Interference
d. delay spread

5. Shannon's coding theorem proves that a ___________exists that achieves data rates arbitrarily close to capacity with arbitrarily small probability of bit error.
a. data
b. noise
c. codeword
d. code

6. PIFA stands for
a. Polarized Inverted-F Antenna
b. Polarized Iconic-F Antenna
c. Planar Inverted-F Antenna
d. Planar Iconic-F Antenna

7. Currently, the smallest antenna used for mobile terminals is a type of
a. dipole Antenna
b. microstrip antenna
c. inverted-F Antenna
d. ceramic-chip antenna

8. In the design of small mobile terminal antennas, one of the following factor should NOT be considered. Identify it.
a. Small size
b. Robustness
c. Propagation medium
d. Light weight

9. The AAAS in the iBurst Systems was first introduced in which of the following country.
a. United states
b. Japan
c. South Africa
d. Australia

10, The AAAS that produces multiple beams by digital beam forming (DBF) technology, performs the function of
a. FDMA
b. SDMA
c. CDMA
d. TDMA

11. In the PHS system, a TDD (time division duplex) transmission scheme is employed providing service at the
a. 1500-MHz band
b. 1700-MHz band
c. 1800-MHz band
d. 1900-MHz band

12. The number of horizontal parasitic elements used for increasing the bandwidth at 900-MHz bands in triple band is
a. 1
b. 2
c. 3
d. 4

13. The capacity of frequency-selective fading channels for ________the capacity is known.
a. frequency variant channels
b. frequency-invariant channels
c. time-invariant channels
d. time variant channels

14. For the AWGN channel, the maximizing input distribution is ______that results in the channel capacity.
a. zero variance
b. zero mean
c. Shannon
d. Gaussian

15. CDI stands for
a. Channel Distribution Information
b. Channel Distribution Interference
c. Code Distribution Information
d. Channel Data Information

16. In Time-varying channels, the frequency-selective fading channel consists of a set of ___________ in parallel with SNR.
a. AWGN channels
b. Noiseless channels
c. Rayleigh fading
d. Noisy channels

17. MSA stands for
a. Multi Strip Antenna
b. Micro Strap Antenna
c. Macro Strip Antenna
d. Micro Strip Antenna

18. Antennas with a gain in the range of ____are normally used for base station antennas in cellular systems
a. 10 –20 dB
b. 100 -200 dB
c. 1- 6 dB
d. 50-500dB

19. The capacity of time variant parallel set of channels is the _____________associated with each channel with power optimally allocated over all channels.
a. sum of the rates
b. double of the rates
c. difference of rates
d. zero

20. One of the following is NOT a key item in design of base station antenna. Identify it.
a. Diversity
b. Element
c. Feed circuit
d. Propagation

21. MLA stands for
a. Meander Line Antenna
b. Micro Line Antenna
c. Macro Line Antenna
d. Microstrip Line Antenna

22. Mobile communication path loss in free-space propagation do not depend on
a. operating frequency
b. antenna height
c. particular environmental conditions
d. battery size

23. The radome diameter for a dual frequency antenna is
a. 100m
b. 100 cm
c. 100 mm
d. 100nm

CONCLUSION

In this unit the capacity of flat and frequency selective channels were first discussed in detail. In the second part the antennas, its types, designing them for mobile terminal were discussed. PIFA and base station antennas and arrays used for mobile communication was elaborated in detail.

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ASSIGNMENT

1. Differentiate between Flat Fading and Frequency Selective Fading.
2. Explain the capacity of a Flat fading channel with neat diagrams.
3. Explain the capacity of a Frequency Selective fading channel with neat diagrams.
4. Explain Capacity of a channel with receiver diversity.
5. Describe about Planar Inverted F Antenna (PIFA).
6. Explain the monopole antenna system.
7. What is a base station antenna? Explain its arrays.

MOBILE COMMUNICATION AND NETWORKS

UNIT IV MULTI-ANTENNA COMMUNICATION

Prepared by

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Assistant Professor
ECE, SCSVMV

AIM & OBJECTIVES

- To understand the issues involved in mobile communication system design and analysis.
- To understand the characteristics of wireless channels.
- To know the fundamental limits on the capacity of wireless channels.
- To acquire knowledge about different types of Diversity receivers.

PRE TEST-MCQ TYPE

1. Equalizer is usually implemented in
a) Transmitter
b) Baseband or at IF in a receiver
c) Radio channel
d) Modulator stage

2. Which of the following is not a characteristic of FIR filter?
a) Many zeroes
b) Poles only at z=0
c) Transfer function is a polynomial of z-1
d) Many poles

3. Which of the following is not a non-linear equalization technique?
a) Decision feedback equalization
b) Maximum likelihood symbol detection
c) Minimum square error detection
d) Maximum likelihood sequence detection

4. Small scale fades are characterized by ____________ amplitude fluctuations.
a) Large
b) Small
c) Rapid
d) Slow

5. Large scale fading can be mitigated with the help of
a) Modulation
b) Demodulation
c) Macroscopic diversity technique
d) Microscopic diversity technique

UNIT IV MULTI-ANTENNA COMMUNICATION

Receiver structure- Diversity receivers- selection and MRC receivers, RAKE receiver, equalization: linear-ZFE and adaptive, DFE. Transmit Diversity-Altamonte scheme.

THEORY

Introduction

Apart from the better transmitter and receiver technology, mobile communications require signal processing techniques that improve the link performance. Equalization, Diversity and channel coding are channel impairment improvement techniques. Equalization compensates for Inter Symbol Interference (ISI) created by multipath within time dispersive channels. An equalizer within a receiver compensates for the average range of expected channel amplitude and delay characteristics. In other words, an equalizer is a filter at themobile receiver whose impulse response is inverse of the channel impulse response. As such equalizers finds their use in frequency selective fading channels. Diversity is another technique used to compensate fast fading and is usually implemented using two or more receiving antennas. It is usually employed to reduce the depths and duration of the fades experienced by a receiver in a flat fading channel. Channel coding improves mobile communication link performance by adding redundant data bits in the trans mitted message. At the baseband portion of the transmitter, a channel coder maps a digital message sequence in to another specific code sequence containing greater number of bits than original contained in the message.

One of the most powerful techniques to mitigate the effects of fading is to use diversity-combining of independently fading signal paths. Diversity-combining uses the fact that independent signal paths have a low probability of experiencing deep fades simultaneously. Thus, the idea behind diversity is to send the same data over independent fading paths. These independent paths are combined in some way such that the fading of the resultant signal is reduced. For example, consider a system with two antennas at either the transmitter or receiver that experience independent fading. If the antennas are spaced sufficiently far apart, it is unlikely that they both experience deep fades at the same time. By selecting the antenna with the strongest signal, called selection combining, Now obtain a much better signal than can just have one antenna. Diversity techniques that mitigate the effect of multipath fading are called microdiversity, and that is the focus of this chapter. Diversity to mitigate the effects of shadowing from buildings and objects is called macrodiversity. Macrodiversity is generally implemented by combining signals received by several base stations or access points. This requires coordination among the different base stations or access points. Such coordination is implemented as part of the networking protocols in infrastructure-based wireless networks.

Realization of Independent Fading Paths

There are many ways of achieving independent fading paths in a wireless system. One method is to use multiple transmit or receive antennas, also called an antenna array, where the elements of the array are separated in distance. This type of diversity is referred to as *space diversity*. Note that with receiver space diversity, independent fading paths are realized without an increase in transmit signal power or bandwidth. Moreover, coherent combining of the diversity signals leads to an increase in SNR at the receiver over the SNR that would be obtained with just a single receive antenna. Conversely, to obtain independent paths through transmitter space diversity, the transmit power must be divided among multiple antennas.

Thus, with coherent combining of the transmit signals the received SNR is the same as if there were just a single transmit antenna. Space diversity also requires that the separation between antennas be such that the fading amplitudes corresponding to each antenna are approximately independent. For example, in a uniform scattering environment with isotropic transmit and receive antennas the minimum antenna separation required for independent fading on each antenna is approximately one half wavelength (0.38λ to be exact). If the transmit or receive antennas are directional (which is common at the base station if the system has cell sectorization), then the multipath is confined to a small angle relative to the LOS ray, which means that a larger antenna separation is required to get independent fading samples.

A second method of achieving diversity is by using either two transmit antennas or two receive antennas with different polarization (e.g. vertically and horizontally polarized waves). The two transmitted waves follow the same path. However, since the multiple random reflections distribute the power nearly equally relative to both polarizations, the average receive power corresponding to either polarized antenna is approximately the same.

Since the scattering angle relative to each polarization is random, it is highly improbable that signals received on the two differently polarized antennas would be simultaneously in deep fades. There are two disadvantages of polarization diversity. First the atmost two diversity branches, corresponding to the two types of polarization. The second disadvantage is that polarization diversity loses effectively half the power (3 dB) since the transmit or receive power is divided between the two differently polarized antennas.

Directional antennas provide angle, or directional, diversity by restricting the receive antenna beamwidth to a given angle. In the extreme, if the angle is very small then at most one of the multipath rays will fall within the receive beamwidth, so there is no multipath fading from multiple rays. However, this diversity technique requires either a sufficient number of directional antennas to span all possible directions of arrival or a single antenna whose directivity can be steered to the angle of arrival of one of the multipath components (preferably the strongest one).

Note also that with this technique the SNR may decrease due to the loss of multipath components that fall outside the receive antenna beamwidth, unless the directional gain of the antenna is sufficiently large to compensate for this lost power.

Smart antennas are antenna arrays with adjustable phase at each antenna element: such arrays form directional antennas that can be steered to the incoming angle of the strongest multipath component. Frequency diversity is achieved by transmitting the same narrowband signal at different carrier frequencies, where the carriers are separated by the coherence bandwidth of the channel. This technique requires additional transmit power to send the signal over multiple frequency bands. Spread spectrum with RAKE reception does provide independently fading paths of the information signal and thus is a form of frequency diversity. Time diversity is achieved by transmitting the same signal at different times, where the time difference is greater than the channel coherence time (the inverse of the channel Doppler spread). Time diversity does not require increased transmit power, but it does decrease the data rate since data is repeated in the diversity time slots rather than sending new data in these time slots.

Time diversity can also be achieved through coding and interleaving, Clearly time diversity can't be used for stationary applications, since the channel coherence time is infinite and thus fading is highly correlated over time.

Receiver Diversity

System Model

In receiver diversity the independent fading paths associated with multiple receive antennas are combined to obtain a resultant signal that is then passed through a standard demodulator. The combining can be done in several ways which vary in complexity and overall performance. Most combining techniques are linear: the output of the combiner is just a weighted sum of the different fading paths or branches, as shown in Figure for *M*-branch diversity. Specifically, when all but one of the complex α_is are zero, only one path is passed to the combiner output. When more than one of the α_i's is nonzero, the combiner adds together multiple paths, where each path may be weighted by different value. Combining more than one branch signal requires co-phasing, where the phase θ_i of the *i*th branch is removed through the multiplication by $\alpha_i = a_i\, e^{-j\theta i}$ for some real-valued a_i. This phase removal requires coherent detection of each branch to determine its phase θ_i. Without co-phasing, the branch signals would not add up coherently in the combiner, so the resulting output could still exhibit significant fading due to constructive and destructive addition of the signals in all the branches.

The multiplication by α_i can be performed either before detection (pre detection) or after detection (post detection) with essentially no difference in performance. Combining is typically performed post-detection, since the branch signal power and/or phase is required to determine the appropriate α_i value. Post-detection combining of multiple branches requires a dedicated receiver for each branch to determine the branch phase, which increases the hardware complexity and power consumption, particularly for a large number of branches.

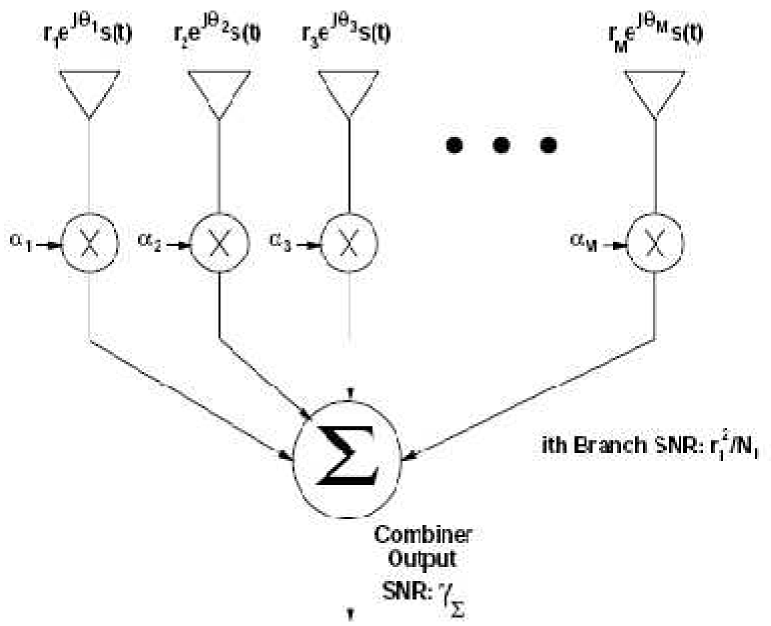


Figure Linear combiner

The main purpose of diversity is to coherently combine the independent fading paths so that the effects of fading are mitigated. The signal output from the combiner equals the original transmitted signal $s(t)$ multiplied by a random complex amplitude term $\alpha\Sigma = \Sigma i\ a_i r_i$. This complex amplitude term results in a random SNR $\gamma\Sigma$ at the combiner output, where the distribution of $\gamma\Sigma$ is a function of the number of diversity paths, the fading distribution on each path, and the combining technique, as shown in more detail below

There are two types of performance gain associated with receiver space diversity: array gain and diversity gain. The array gain results from coherent combining of multiple receive signals. Even in the absence of fading, this can lead to an increase in average received SNR. For example, suppose there is no fading so that $ri = \sqrt{Es}$ for Es the energy per symbol of the transmitted signal. Assume identical noise PSD $N0$ on each branch and pulse shaping such that $BTs = 1$. Then each branch has the same SNR $\gamma i = Es/N0$. Set $ai = ri/\sqrt{N0}$. Then the received SNR is

$$\gamma_\Sigma = \frac{\left(\sum_{i=1}^M a_i r_i\right)^2}{N_0 \sum_{i=1}^M a_i^2} = \frac{\left(\sum_{i=1}^M E_s\right)^2}{N_0 \sum_{i=1}^M E_s} = \frac{M E_s}{N_0}.$$

Thus, in the absence of fading, with appropriate weighting there is an M-fold increase in SNR due to the coherent combining of the M signals received from the different antennas. This SNR increase in the absence of fading is referred to as the array gain. More precisely, array gain Ag is defined as the increase in averaged combined SNR $\gamma\Sigma$ over the average branch SNR γ

$$A_g = \frac{\overline{\gamma}_\Sigma}{\overline{\gamma}}.$$

Array gain occurs for all diversity combining techniques, but is most pronounced in MRC. Both diversity and array gain occur in transmit diversity as well. The array gain allows a system with multiple transmit or receive antennas in a fading channel to achieve better performance than a system without diversity in an AWGN channel with the same average SNR.

In fading the combining of multiple independent fading paths leads to a more favorable distribution for $\gamma\Sigma$ than would be the case with just a single path. In particular, the performance of a diversity system, whether it uses space diversity or another form of diversity, in terms of Ps and $Pout$ is as defined

$$\overline{P}_s = \int_0^\infty P_s(\gamma) p_{\gamma_\Sigma}(\gamma) d\gamma,$$

where $Ps(\gamma)$ is the probability of symbol error for demodulation of $s(t)$ in AWGN with SNR $\gamma\Sigma$, and

$$P_{out} = p(\gamma_\Sigma \le \gamma_0) = \int_0^{\gamma_0} p_{\gamma_\Sigma}(\gamma) d\gamma,$$

for some target SNR value $\gamma 0$. The more favorable distribution for $\gamma\Sigma$ leads to a decrease in $P\ s$ and $Pout$ due to diversity combining, and the resulting performance advantage is called the diversity gain. In particular, for some diversity systems their average probability of error can be expressed in the form $P\ s = c\gamma{-}M$ where c is a constant that depends on the specific modulation and coding, γ is the average received *SNR* per branch, and M is called the diversity order of the system. The diversity order indicates how the *slope* of the average probability of error as a function of average SNR changes with diversity. This expression has a diversity order of one, consistent with a single receive antenna. The maximum diversity order of a system with M antennas is M, and when the diversity order equals M the system is said to achieve full diversity order.

Selection Combining

In selection combining (SC), the combiner outputs the signal on the branch with the highest SNR $r^2\ i\ /Ni$. This is equivalent to choosing the branch with the highest $r^2\ i + Ni$ if the noise power $Ni = N$ is the same on all branches 1. Since only one branch is used at a time, SC often requires just one receiver that is switched into the active antenna branch. However, a dedicated receiver on each antenna branch may be needed for systems that transmit continuously in order to simultaneously and continuously monitor SNR on each branch. With SC the path output from the combiner has an SNR equal to the maximum SNR of all the branches. Moreover, since only one branch output is used, co-phasing of multiple branches is not required, so this technique can be used with either coherent or differential modulation.

For M branch diversity, the CDF of $\gamma\Sigma$ is given by

$$P_{\gamma_\Sigma}(\gamma) = p(\gamma_\Sigma < \gamma) = p(\max[\gamma_1, \gamma_2, \ldots, \gamma_M] < \gamma) = \prod_{i=1}^{M} p(\gamma_i < \gamma).$$

the pdf of $\gamma\Sigma$ is obtained by differentiating $P\gamma\Sigma(\gamma)$ relative to γ, and the outage probability by evaluating $P\gamma\Sigma(\gamma)$ at $\gamma = \gamma 0$. Assume M branches with uncorrelated Rayleigh fading amplitudes ri. The instantaneous SNR on the ith branch is therefore given by $\gamma i = r^2\ i\ /N$. Defining the average SNR on the ith branch as $\gamma i = E[\gamma i]$, the SNR distribution will be exponential:

$$p(\gamma_i) = \frac{1}{\overline{\gamma}_i} e^{-\gamma_i/\overline{\gamma}_i}.$$

The outage probability for a target $\gamma 0$ on the ith branch in Rayleigh fading is

$$P_{out}(\gamma_0) = 1 - e^{-\gamma_0/\overline{\gamma}_i}.$$

The outage probability of the selection-combiner for the target $\gamma 0$ is then

$$P_{out}(\gamma_0) = \prod_{i=1}^{M} p(\gamma_i < \gamma_0) = \prod_{i=1}^{M} \left[1 - e^{-\gamma_0/\overline{\gamma}_i}\right].$$

$$P_{out}(\gamma_0) = p(\gamma_\Sigma < \gamma_0) = \left[1 - e^{-\gamma_0/\overline{\gamma}}\right]^M$$

Differentiating above with relative to $\gamma 0$ yields the pdf for $\gamma\Sigma$:

$$p_{\gamma_\Sigma}(\gamma) = \frac{M}{\overline{\gamma}}\left[1 - e^{-\gamma/\overline{\gamma}}\right]^{M-1} e^{-\gamma/\overline{\gamma}}$$

The average SNR of the combiner output in i.i.d. Rayleigh fading is

$$\begin{aligned}\overline{\gamma}_\Sigma &= \int_0^\infty \gamma p_{\gamma_\Sigma}(\gamma) d\gamma \\ &= \int_0^\infty \frac{\gamma M}{\overline{\gamma}}\left[1 - e^{-\gamma/\overline{\gamma}}\right]^{M-1} e^{-\gamma/\overline{\gamma}} d\gamma \\ &= \overline{\gamma}\sum_{i=1}^{M}\frac{1}{i}.\end{aligned}$$

Thus, the average SNR gain increases with *M*, but not linearly. The biggest gain is obtained by going from no diversity to two-branch diversity. Increasing the number of diversity branches from two to three will give much less gain than going from one to two, and in general increasing *M* yields diminishing returns in terms of the SNR gain.

Threshold Combining

SC for systems that transmit continuously may require a dedicated receiver on each branch to continuously monitor branch SNR. A simpler type of combining, called threshold combining, avoids the need for a dedicated receiver on each branch by scanning each of the branches in sequential order and outputting the first signal with SNR above a given threshold γT . As in SC, since only one branch output is used at a time, co-phasing is not required. Thus, this technique can be used with either coherent or differential modulation. Once a branch is chosen, as long as the SNR on that branch remains above the desired threshold, the combiner outputs that signal. If the SNR on the selected branch falls below the threshold, the combiner switches to another branch.

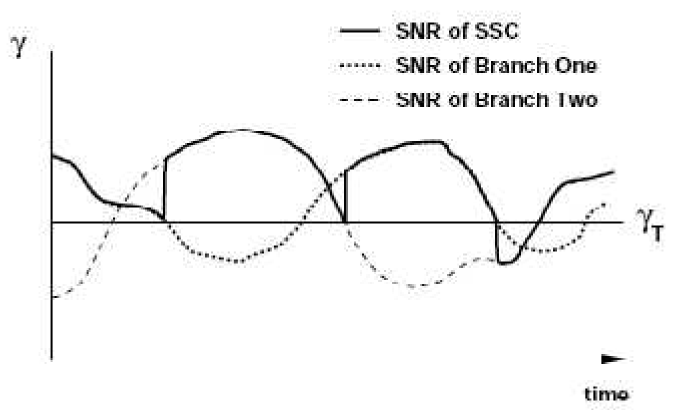


Figure SNR of SSC Technique

There are several criteria the combiner can use to decide which branch to switch. The simplest criterion is to switch randomly to another branch. With only two-branch diversity this is equivalent to switching to the other branch when the SNR on the active branch falls below γT . This method is called switch and stay combining (SSC). The switching process and SNR associated with SSC is illustrated in Figure. Since the SSC does not select the branch with the highest SNR, its performance is between that of no diversity and ideal SC.

Denote the SNR on the *i*th branch by γi and the SNR of the combiner output by $\gamma\Sigma$. The CDF of $\gamma\Sigma$ will depend on the threshold level γT and the CDF of γi. For two-branch diversity with i.i.d. branch statistics the CDF of the combiner output $P\gamma\Sigma(\gamma) = p(\gamma\Sigma \leq \gamma)$ can be expressed in terms of the CDF $P\gamma i(\gamma) = p(\gamma i \leq \gamma)$ and pdf $p\gamma i(\gamma)$ of the individual branch SNRs as

$$P_{\gamma_\Sigma}(\gamma) = \begin{cases} P_{\gamma_1}(\gamma_T)P_{\gamma_2}(\gamma) & \gamma < \gamma_T \\ p(\gamma_T \leq \gamma_1 \leq \gamma) + P_{\gamma_1}(\gamma_T)P_{\gamma_2}(\gamma) & \gamma \geq \gamma_T. \end{cases}$$

For Rayleigh fading in each branch with $\gamma i = \gamma$, $i = 1, 2$ this yields

$$P_{\gamma_\Sigma}(\gamma) = \begin{cases} 1 - e^{-\gamma_T/\overline{\gamma}} - e^{-\gamma/\overline{\gamma}} + e^{-(\gamma_T+\gamma)/\overline{\gamma}} & \gamma < \gamma_T \\ 1 - 2e^{-\gamma/\overline{\gamma}} + e^{-(\gamma_T+\gamma)/\overline{\gamma}} & \gamma \geq \gamma_T. \end{cases}$$

The outage probability *Pout* associated with a given $\gamma 0$ is obtained by evaluating $P\gamma\Sigma(\gamma)$ at $\gamma = \gamma 0$

$$P_{out}(\gamma_0) = P_{\gamma_\Sigma}(\gamma_0) = \begin{cases} 1 - e^{-\gamma_T/\overline{\gamma}} - e^{-\gamma_0/\overline{\gamma}} + e^{-(\gamma_T+\gamma_0)/\overline{\gamma}} & \gamma_0 < \gamma_T \\ 1 - 2e^{-\gamma_0/\overline{\gamma}} + e^{-(\gamma_T+\gamma_0)/\overline{\gamma}} & \gamma_0 \geq \gamma_T. \end{cases}$$

Maximal Ratio Combining

In SC and SSC, the output of the combiner equals the signal on one of the branches. In maximal ratio combining (MRC) the output is a weighted sum of all branches, so the $\alpha$is in Figure are all nonzero. Since the signals are cophased, $\alpha_i = a_i e^{-j\theta i}$, where θ_i is the phase of the incoming signal on the *i*th branch. Thus, the envelope of the combiner output will be $r = \sum M\ i=1\ a_i r_i$. Assuming the same noise PSD $N0$ in each branch yields a total noise PSD N_{tot} at the combiner output of

$$N_{tot} = \sum_{i=1}^{M} a_i^2 N_0$$

Thus, the output SNR of the combiner is

$$\gamma\Sigma = \frac{r^2}{N_{tot}} = \frac{1}{N_0} \frac{\left(\sum_{i=1}^{M} a_i r_i\right)^2}{\sum_{i=1}^{M} a_i^2}$$

The goal is to chose the α_is to maximize $\gamma\Sigma$. Intuitively, branches with a high SNR should be weighted more than branches with a low SNR, so the weights a^2_i should be proportional to the branch SNRs $r^2{}_i / N_0$. Now find the a_is that maximize $\gamma\Sigma$ by taking partial derivatives or using the

Swartz inequality. Solving for the optimal weights yields $a^2_i = r^2_i / N_0$, and the resulting combiner SNR becomes

$$\gamma_\Sigma = \sum_{i=1}^{M} r_i^2 / N_0 = \sum_{i=1}^{M} \gamma_i.$$

Thus, the SNR of the combiner output is the sum of SNRs on each branch. The average combiner SNR increases linearly with the number of diversity branches M, in contrast to the diminishing returns associated with the average combiner SNR in SC. As with SC, even with Rayleigh fading on all branches, the distribution of the combiner output SNR is no longer Rayleigh.

To obtain the distribution of $\gamma\Sigma$ take the product of the exponential moment generating or characteristic functions. Assuming i.i.d. Rayleigh fading on each branch with equal average branch SNR γ, the distribution of $\gamma\Sigma$ is chi-squared with $2M$ degrees of freedom, expected value $\gamma\Sigma = M\gamma$, and variance $2M\gamma$:

$$p_{\gamma_\Sigma}(\gamma) = \frac{\gamma^{M-1} e^{-\gamma/\bar{\gamma}}}{\bar{\gamma}^M (M-1)!}, \quad \gamma \geq 0$$

The corresponding outage probability for a given threshold $\gamma 0$ is given by

$$P_{out} = p(\gamma_\Sigma < \gamma_0) = \int_0^{\gamma_0} p_{\gamma_\Sigma}(\gamma) d\gamma = 1 - e^{-\gamma_0/\bar{\gamma}} \sum_{k=1}^{M} \frac{(\gamma_0/\bar{\gamma})^{k-1}}{(k-1)!}.$$

Equal-Gain Combining

MRC requires knowledge of the time-varying SNR on each branch, which can be very difficult to measure. A simpler technique is equal-gain combining, which co-phases the signals on each branch and then combines them with equal weighting, $\alpha_i = e^{-\theta i}$. The SNR of the combiner output, assuming equal noise PSD N_0 in each branch, is then given by

$$\gamma_\Sigma = \frac{1}{N_0 M} \left(\sum_{i=1}^{M} r_i \right)^2$$

The pdf and CDF of $\gamma\Sigma$ do not exist in closed-form. For i.i.d. Rayleigh fading and two-branch diversity and average branch SNR γ, an expression for the CDF in terms of the Q function can be derived as

$$P_{\gamma_\Sigma}(\gamma) = 1 - e^{-2\gamma/\bar{\gamma}} \sqrt{\frac{\pi\gamma}{\gamma}} e^{-\gamma/\bar{\gamma}} \left(1 - 2Q\left(\sqrt{2\gamma/\bar{\gamma}}\right)\right)$$

The resulting outage probability is given by

$$P_{out}(\gamma_0) = 1 - e^{-2\gamma_R} - \sqrt{\pi\gamma_R} e^{-\gamma_R} \left(1 - 2Q\left(\sqrt{2\gamma_R}\right)\right)$$

RAKE Receiver

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

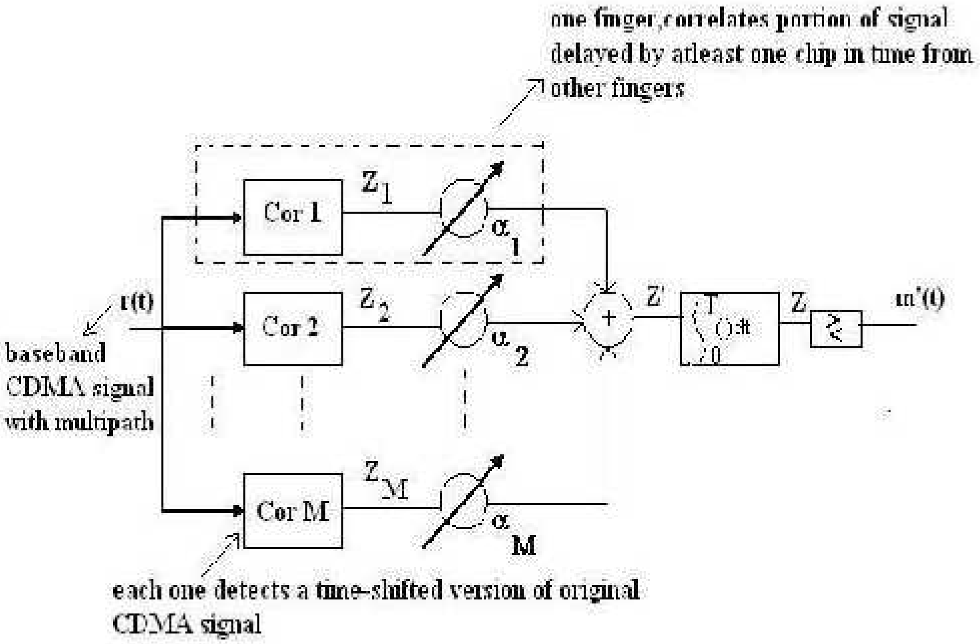


Figure RAKE receiver.

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

Interleaver

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

A block interleaver formats the encoded data into a rectangular array of m rows and n columns, and interleaves nm bits at a time. Each row contains a word of source data having n bits. an interleaver of degree m consists of m rows. source bits are placed into the interleaver by sequentially increasing the row number for each successive bit, and forming the columns.

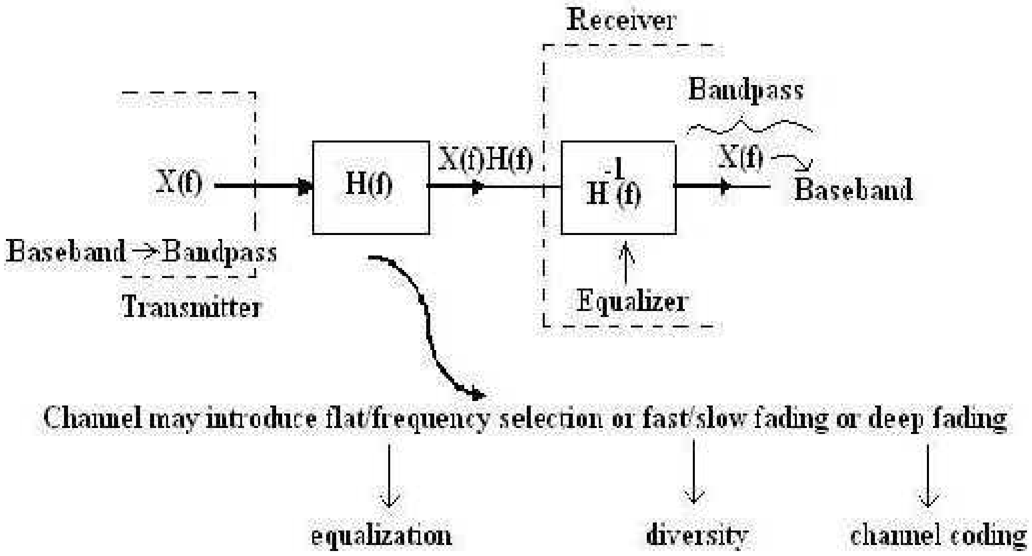


Figure A general framework of fading effects and their mitigation techniques

The interleaved source data is then read out row-wise and transmitted over the channel. This has the effect of separating the original source bits by m bit periods. At the receiver, de-interleaver stores the received data by sequentially increasing the row number of each successive bit, and then clocks out the data row-wise, one word at a time. Convolutional interleavers are ideally suited for use with convolutional codes. A general framework of the fading effects and their mitigation techniques is shown in Figure.

Equalization

ISI can cause an irreducible error floor when the modulation symbol time is on the same order as the channel delay spread. Signal processing provides a powerful mechanism to counteract ISI. In a broad sense, equalization defines any signal processing technique used at the receiver to alleviate the ISI problem caused by delay spread. Signal processing can also be used at the transmitter to make the signal less susceptible to delay spread: spread spectrum and multicarrier modulation fall in this category of transmitter signal processing techniques.

ISI mitigation is required when the modulation symbol time Ts is on the order of the channel's rms delay spread σTm. For example, cordless phones typically operate indoors, where the delay spread is small. Since voice is also a relatively low date rate application, equalization is generally not needed in cordless phones. However, the IS-54 digital cellular standard is designed for outdoor use, where $\sigma Tm \approx Ts$, so equalization is part of this standard.

Higher data rate applications are more sensitive to delay spread, and generally require high-performance equalizers or other ISI mitigation techniques. In fact, mitigating the impact of delay spread is one of the most challenging hurdles for high-speed wireless data systems. Equalizer design must typically balance ISI mitigation with noise enhancement, since both the signal and the noise pass through the equalizer, which can increase the noise power. Nonlinear equalizers suffer less from noise enhancement than linear equalizers, but typically entail higher complexity, as discussed in more detail below.

Moreover, equalizers must typically have an estimate of the channel impulse or frequency response to mitigate the resulting ISI. Since the wireless channel varies over time, the equalizer must learn the frequency or impulse response of the channel (training) and then update its estimate of the frequency response as the channel changes (tracking). The process of equalizer training and tracking is often referred to as adaptive equalization, since the equalizer adapts to the changing channel. Equalizer training and tracking can be quite difficult if the channel is changing rapidly. An equalizer can be implemented at baseband, RF, or IF. Most equalizers are implemented digitally after A/D conversion, since such filters are small, cheap, easily tuneable, and very power efficient.

Equalizer Types

Equalization techniques fall into two broad categories: linear and nonlinear. The linear techniques are generally the simplest to implement and to understand conceptually. However, linear equalization techniques typically suffer from more noise enhancement than nonlinear equalizers, and are therefore not used in most wireless applications. Among nonlinear equalization techniques, decision-feedback equalization (DFE) is the most common, since it is fairly simple to implement and generally performs well. However, on channels with low SNR, the DFE suffers from error propagation when bits are decoded in error, leading to poor performance. The optimal equalization technique is maximum likelihood sequence estimation (MLSE). Unfortunately, the complexity of this technique grows exponentially with the length of the delay spread, and is therefore impractical on most channels of interest.

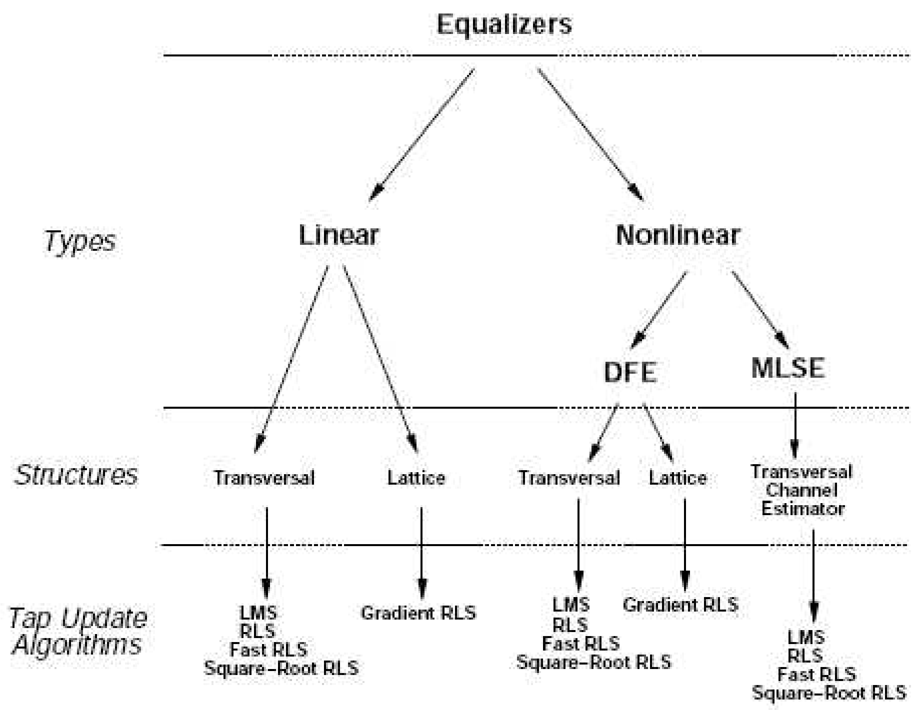


Figure Equalizer Types, Structures, and Algorithms

However, the performance of the MLSE is often used as an upper bound on performance for other equalization techniques. Linear and nonlinear equalizers are typically implemented using a transversal or lattice structure. The transversal structure is a filter with $N - 1$ delay elements and N taps with tunable complex weights.

The lattice filter uses a more complex recursive structure. In exchange for this increased complexity relative to transversal structures, lattice structures often have better numerical stability and convergence properties and greater flexibility in changing their length. In addition to the equalizer type and structure, adaptive equalizers require algorithms for updating the filter tap coefficients during training and tracking. Many algorithms have been developed over the years for this purpose. These algorithms generally entail tradeoffs between complexity, convergence rate, and numerical stability.

Linear Equalizers

If *F*Σ(*f*) is not flat, use the equalizer *Heq*(*z*) in Figure to reduce ISI. In this section, assume a linear equalizer implemented via an $N = 2L + 1$ tap transversal filter

$$H_{eq}(z) = \sum_{i=-L}^{L} w_i z^{-i}.$$

The length of the equalizer *N* is typically dictated by implementation considerations, since a large *N* usually entails higher complexity. Causal linear equalizers have $wi = 0$, $i < 0$. For a given equalizer size *N* the equalizer design must specify the tap weights *{wi}L i*=−*L* for a given channel frequency response, and the algorithm for updating these tap weights as the channel varies.

Recall that the performance metric in wireless systems is probability of error (or outage probability), so for a given channel the optimal choice of equalizer coefficients would be the coefficients that minimize probability of error. Unfortunately it is extremely difficult to optimize the *{wi}*s with respect to this criterion. Since it cannot directly optimize for the desired performance metric, It must instead use an indirect optimization that balances ISI mitigation with the prevention of noise enhancement, as discussed relative to the simple analog example above.

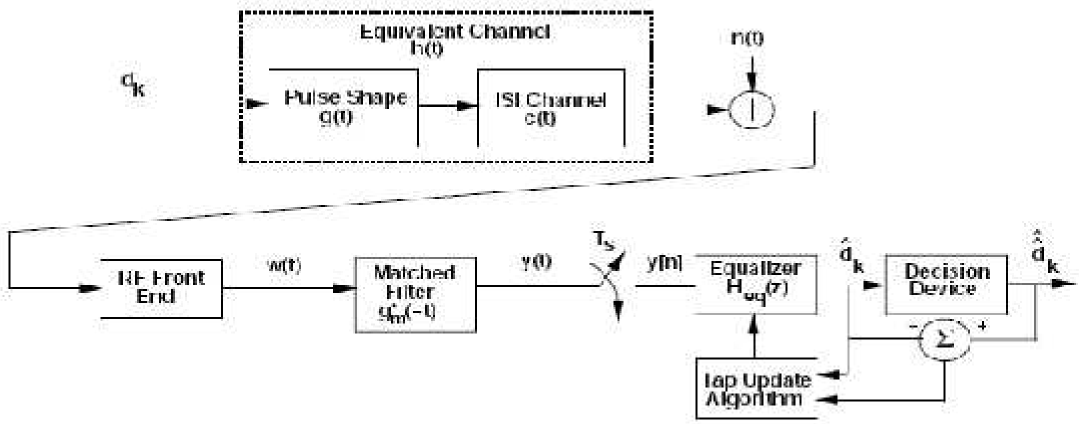


Figure End to End system

The two linear equalizers: the Zero Forcing (ZF) equalizer and the Minimum Mean Square Error (MMSE) equalizer. The former equalizer cancels all ISI, but can lead to considerable noise enhancement. The latter technique minimizes the expected mean squared error between the transmitted symbol and the symbol detected at the equalizer output, thereby providing a better balance between ISI mitigation and noise enhancement. Because of this more favorable balance, MMSE equalizers tend to have better BER performance than equalizers using the ZF algorithm.

A Mathematical Framework

The signal received by the equalizer is given by

$$x(t) = d(t) * h(t) + nb\ (t)$$

where $d(t)$ is the trans mitted 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

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

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

$$g\ (t) = h\ (t) * heq\ (t) = \delta\ (t)$$

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

$$Heq\ (f)\ H\ (f) = 1$$

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 (ZF) Equalizers

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

$$Hch\ (f)\ Heq\ (f) = 1,\ |f\,| < 1/2T$$

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.

A Generic Adaptive Equalizer

The basic structure of an adaptive filter is shown in Figure. This filter is called the transversal filter, and in this case has N delay elements, N+1 taps and N+1 tunable complex multipliers, called weights. These weights are updated continuously by an adaptive algorithm. In the figure the subscript k represents discrete time index. The adaptive algorithm is controlled by the error signal *ek*. The error signal is derived by comparing the output of the equalizer, with some signal *dk* which is replica of trans mitted signal. The adaptive algorithm uses *ek* to minimize the cost function and uses the equalizer weights in such a manner that it minimizes the cost function iteratively.

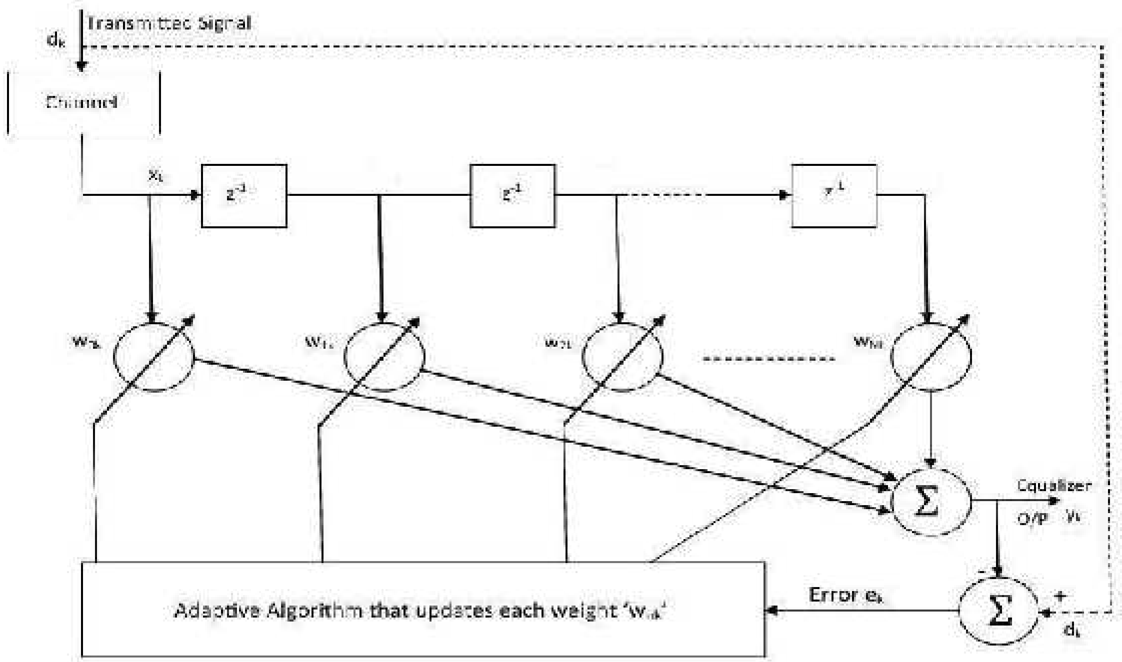


Figure A generic adaptive equalizer

All of the equalizers described so far are designed based on a known value of the composite channel response $h(t) = g(t) * c(t)$. Since the channel $c(t)$ in generally not known when the receiver is designed, the equalizer must be tunable so it can adjust to different values of $c(t)$. Moreover, since in wireless channels $c(t) = c(\tau, t)$ will change over time, the system must periodically estimate the channel $c(t)$ and update the equalizer coefficients accordingly. This process is called equalizer training or adaptive equalization.

The equalizer can also use the detected data to adjust the equalizer coefficients. This process is called equalizer tracking. Blind equalizers do not use training: they learn the channel response via the detected data only. During training, the coefficients of the equalizer are updated at time k based on a known training sequence $[dk-M, \ldots, dk]$ that has been sent over the channel.

The length M of the training sequence depends on the number of equalizer coefficients that must be determined and the convergence speed of the training algorithm. Note that the equalizer must be retrained when the channel decorrelates, i.e. at least every Tc seconds where Tc is the channel coherence time.

Thus, if the training algorithm is slow relative to the channel coherence time then the channel may change before the equalizer can learn the channel. Specifically, if $MT_s > T_c$ then the channel will decorrelate before the equalizer has finished training. In this case equalization is not an effective countermeasure for ISI, and some other technique (e.g. multicarrier modulation or CDMA) is needed.

Let $\{\hat{d}_k\}$ denote the bit decisions output from the equalizer given a transmitted training sequence $\{d_k\}$. The goal is to update the N equalizer coefficients at time $k + 1$ based on the training sequence have received up to time k. Now denote these updated coefficients as $\{w_{-L}(k + 1), \ldots, w_L(k + 1)\}$. MMSE is used as criterion to update these coefficients, i.e. chose $\{w_{-L}(k + 1), \ldots, w_L(k + 1)\}$ as the coefficients that minimize the MSE between d_k and $\hat{d}_k$. Recall that $\hat{d}_k = \sum_{i=-L}^{L} w_i(k) y_{k-i}$, where $y_k = y[k]$ is the output of the sampler in at time k with the known training sequence as input. The $\{w_{-L}(k + 1), \ldots, w_L(k + 1)\}$ that minimize MSE are obtained via a Weiner filter.

Specifically $\mathbf{w}(k + 1) = \{w_{-L}(k + 1), \ldots, w_L(k + 1)\} = \mathbf{R}^{-1}\mathbf{p}$,

where $\mathbf{p} = d_k[y_{k+L} \ldots y_{k-L}]^T$ and

$$R = \begin{bmatrix} |y_{k+L}|^2 & y_{k+L}y^*_{k+L-1} & \cdots & y_{k+L}y^*_{k-L} \\ y_{k+L-1}y^*_{k+L} & |y_{k+L-1}|^2 & \cdots & y_{k+L-1}y^*_{k-L} \\ \vdots & \ddots & \ddots & \vdots \\ y_{k-L}y^*_{k+L} & \cdots & \cdots & |y_{k-L}|^2 \end{bmatrix}$$

Note that the optimal tap updates in this case requires a matrix inversion, which requires N^2 to N^3 multiply operations on each iteration (each symbol time T_s). However, the convergence of this algorithm is very fast: it typically converges in around N symbol times for N the number of equalizer tap weights.

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 response 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 environment 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.

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 compensates 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 algorithm of the equalizer tracks the changing channel.

• As a result, the adaptive equalizer continuously changes the filter characteristics over time.

Decision-Feedback Equalization

The DFE consists of a feed forward filter $B(z)$ with the received sequence as input (similar to the linear equalizer) followed by a feedback filter $D(z)$ with the previously detected sequence as input. The DFE structure is shown Figure. Effectively, the DFE determines the ISI contribution from the detected symbols *{dn}* by passing them through the feedback filter that approximates the combined discrete equivalent baseband channel $F(z)$.

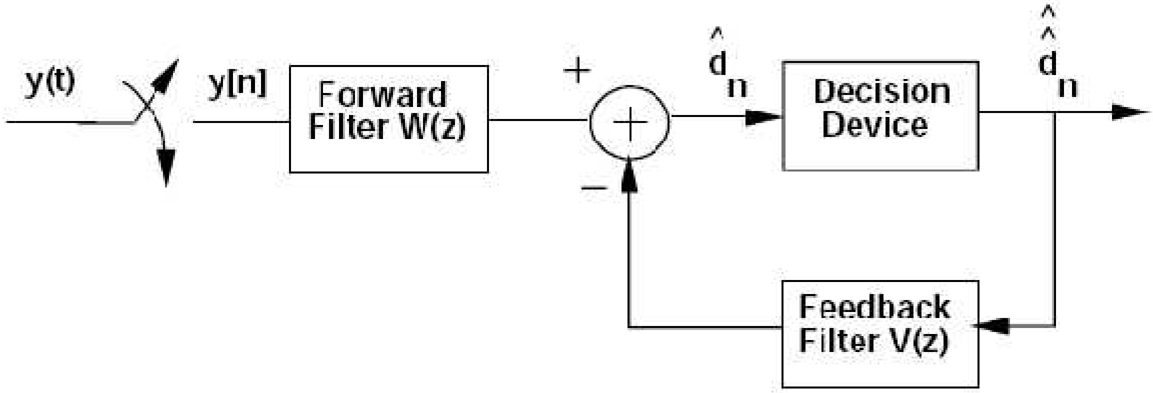


Figure Decision-Feedback Equalizer Structure

The resulting ISI is then subtracted from the incoming symbols. Since the feedback filter $D(z)$ in Figure sits in a feedback loop, it must be strictly causal, or else the system is unstable. The feedback filter of the DFE does not suffer from noise enhancement because it estimates the channel frequency response rather than its inverse. For channels with deep spectral nulls, DFEs generally perform much better than linear equalizers

Assuming W(z) has N1 taps and V (z) has N2 taps, the DFE output is given as

$$\hat{d}_k = \sum_{i=-N_1}^{0} w_i y[k-i] - \sum_{i=1}^{N_2} v_i \hat{\hat{d}}_{k-i}.$$

The typical criteria for selecting the coefficients for $W(z)$ and $V(z)$ are either zero-forcing (remove all ISI) or MMSE (minimize the expected MSE between the DFE output and the original symbol). When both $W(z)$ and $V(z)$ have infinite duration, it was shown by Price that the optimal feed forward filter for a zero-forcing DFE is $1/G*m(1/z*)$, the same noise whitening filter as in the linear MMSE equalizer. In this case the feedback filter $V(z)$ should be essentially the same as the combined baseband channel $F(z)$.

For the MMSE criterion, minimize $E[dk - \hat{}\, dk]^2$. Let $fn = f[n]$ denote the samples of $f(t)$. Then this minimization implies that the coefficients of the feed forward filter must satisfy the following set of linear equations:

$$\sum_{i=-N_1}^{0} q_{li} w_i = f^*_{-l},$$

for $qli = \sum f*j\, fj{+}l{-}i + N0\delta[l - i]$, $l, i = -N1, \ldots, 0$. The coefficients of the feedback filter are then determined from the feed forward coefficients by

$$v_k = -\sum_{i=-N_1}^{0} w_i f_{k-i}$$

These coefficients completely eliminate ISI when there are no decision errors, i.e. when $\hat{}\hat{}d_k$ =

d_k. The resulting minimum MSE is

$$J_{min} = \exp\left[T_s \int_{-.5/T_s}^{.5/T_s} \ln\left[\frac{N_0}{F\Sigma(f) + N_0}\right] df\right]$$

In general the MMSE associated with a DFE is much lower than that of a linear equalizer, if the impact of feedback errors is ignored. DFEs exhibit feedback errors if ˆˆd_k _= d_k, since the ISI subtracted by the feedback path is not the true ISI corresponding to d_n. This error therefore propagates to later bit decisions. Moreover, this error propagation cannot be improved through channel coding, since the feedback path operates on coded channel symbols before decoding. That is because the ISI must be subtracted immediately, which doesn't allow for any decoding delay. The error propagation therefore seriously degrades performance on channels with low SNR.

MIMO Diversity -Alamouti

Consider the instance at which, the transmission is from a mobile to the base-station. The base-stations can be equipped with multiple antennas with sufficient separation easily, the signal transmitted by the mobile unit can be picked up by multiple receive antennas and they can be combined using a diversity-combining technique, e.g., maximal-ratio combining, selection combining, equal-gain combining, etc., to obtain receive diversity. However, if the situation is reversed (i.e., for downlink transmission) achieving diversity gain is not that simple due to the fact that the mobile units are typically limited in size, and it is usually difficult to place multiple antennas that are separated by sufficiently large distances for reception of multiple copies of the transmitted signal through independent channels. Therefore, It is desirable to have a scheme where the benefits of spatial diversity are exploited through "transmit diversity". With this motivation, Alamouti (1998) introduced a way of obtaining transmit diversity when there are two transmit antennas.

Alamouti -Transmission Scheme

The Alamouti scheme is a simple transmit diversity scheme suitable for two transmit antennas. Two symbols are considered at a time, say x1 and x2, and they are transmitted in two consecutive time slots. In the first time slot, *x*1 is transmitted from the first antenna and *x*2 is transmitted from the second one. In the second time slot, −*x*2* is transmitted from the first antenna, while x1* is transmitted from the second antenna. This process is illustrated in Figure The signals *x*1 and *x*2 are picked from an arbitrary (*M*-ary) constellation. Since two symbols are transmitted in two time slots, the overall transmission rate is 1 symbol per channel use, or log2*M* bits per channel use.

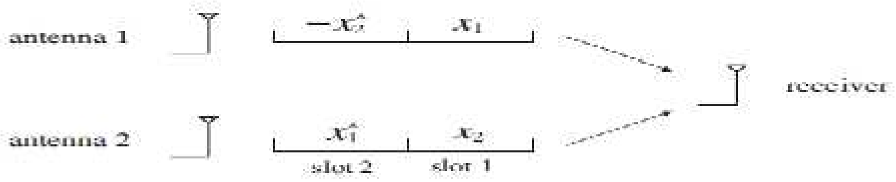


Figure Alamouti Scheme

Optimal Receiver for the Alamouti Scheme

The optimal receiver for the Alamouti scheme is of two types.

a) Single receive antenna

b) Multiple receive antenna

a) Single receive antenna System

Consider the case of a single receive antenna. The received signal in the first time slot is then

$$y_1(1) = \sqrt{\rho}(h_{1,1}x_1 + h_{2,1}x_2) + n_1(1)$$

And in the second time slot is

$$y_1(2) = \sqrt{\rho}(-h_{1,1}x_2^* + h_{2,1}x_1^*) + n_1(2).$$

Define the vector of the received signals (where the second signal is conjugated) as

$$y = \begin{bmatrix} y_1(1) \\ y_1^*(2) \end{bmatrix}$$

Which can be written as

$$y = \sqrt{\rho}\begin{bmatrix} h_{1,1} & h_{2,1} \\ h_{2,1}^* & -h_{1,1}^* \end{bmatrix}\begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} n_1(1) \\ n_1^*(2) \end{bmatrix}$$

b) Multiple Receive Antenna System

The Alamouti scheme can easily be extended to systems with multiple receive antennas as well, resulting in receive diversity in addition to the existing transmit diversity. For this case, the available diversity order is twice the number of receive antennas, and it can also be achieved by a simple linear receiver.

Assume that the received signal during the kth time slot at the j th receive antenna is yj (k), where k = 1, 2, j = 1, 2, . . .,Nr . The channel coefficient from the ith transmit antenna to the jth receive antenna is denoted by hi,j. then it is written as

$$y_j(1) = \sqrt{\rho}(h_{1,j}x_1 + h_{2,j}x_2) + n_j(1)$$

in the first time slot, and in the second time slot

$$y_j(2) = \sqrt{\rho}(-h_{1,j}x_2^* + h_{2,j}x_1^*) + n_j(2)$$

After linear combining the received signals of all the receive antennas and scaling by the factor

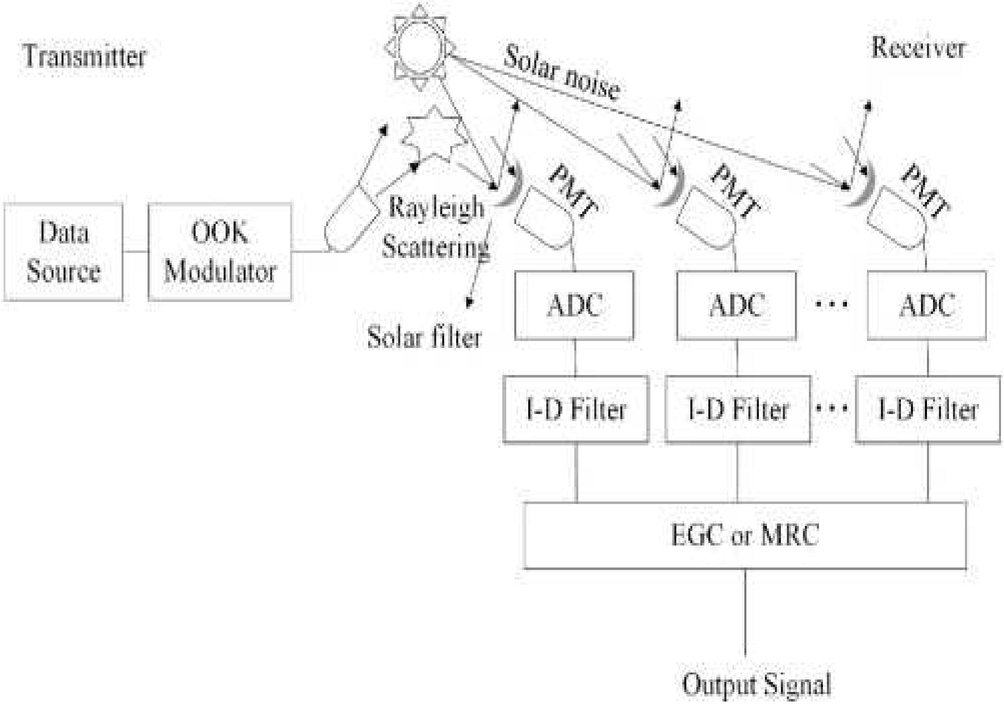
$$1\Big/\sqrt{|h_{1,j}|^2 + |h_{2,j}|^2},$$

the sufficient statistics for optimal receiver as

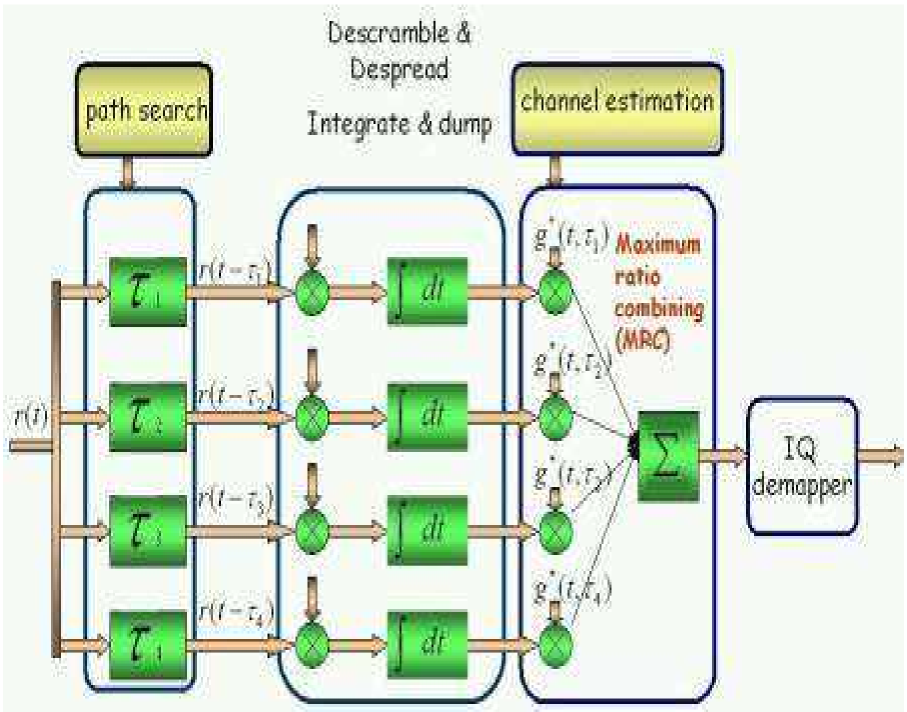
$$y_j(1) = \sqrt{\rho}\sqrt{|h_{1,j}|^2 + |h_{2,j}|^2}\;\; x_1 + n_j''(1),$$

$$y_j(2) = \sqrt{\rho}\sqrt{|h_{1,j}|^2 + |h_{2,j}|^2}\;\; x_2 + n_j''(2),$$

APPLICATIONS



Example of MRC diversity reception algorithm



Example of W-CDMA RAKE receiver

POST TEST-MCQ TYPE

1. In maximal ratio combining, the output SNR is equal to
a) Mean of all individual SNRs
b) Maximum of all SNRs
c) Sum of individual SNRs
d) Minimum of all SNRs

2. Time diversity repeatedly transmits information at time spacings that exceed
a) Coherence bandwidth
b) Dwell time
c) Run time
d) Coherence time

3. Which of the following is not used to improve received signal quality over small scale times and distance?
a) Modulation
b) Equalization
c) Diversity
d) Channel coding

4. Equalization is used to compensate
a) Peak signal to noise ratio
b) Intersymbol interference
c) Channel fading
d) Noises present in the signal

5. Training and tracking are the operating modes of __________
a) Diversity techniques
b) Channel coding techniques
c) Equalization techniques
d) Demodulation techniques

6. An equalizer is said to be converged when it is properly ________
a) Trained
b) Tracked
c) Installed
d) Used

7. Time for convergence of an equalizer is not a function of ________
a) Equalizer algorithm
b) Equalizer structure
c) Time rate of change of multipath radio channel
d) Transmitter characteristics

8. Equalizer is _________ of the channel.
a) Opposite
b) Same characteristics
c) Inverse filter
d) Add on

9. Which of the following controls the adaptive algorithm in an equalizer?
a) Error signal
b) Transmitted signal
c) Received signal
d) Channel impulse response

10. The adaptive algorithms in equalizer that do not require training sequence are called
a) Linear adaptive algorithms
b) Blind algorithms
c) Non-linear adaptive algorithms
d) Spatially adaptive algorithms

11. Equalization techniques can be categorised into _______ and ______ techniques.
a) Linear, non linear
b) Active, passive
c) Direct, indirect
d) Slow, fast

12. Which of the following is not an advantage of lattice equalizer?
a) Simple structure
b) Numerical stability
c) Faster convergence
d) Dynamic assignment

13. Which of the following does not hold true for MLSE?
a) Minimizes probability of sequence error
b) Require knowledge of channel characteristics
c) Requires the statistical distribution of noise
d) Operates on continuous time signal

14. Which of the following factor could not determine the performance of algorithm?
a) Structural properties
b) Rate of convergence
c) Computational complexity
d) Numerical properties

15. Rate of convergence is defined by __________ of algorithm.
a) Time span
b) Number of iterations
c) Accuracy
d) Complexity

16. Computational complexity is a measure of ________
a) Time
b) Number of iterations
c) Number of operations
d) Accuracy

17. Choice of equalizer structure and its algorithm is not dependent on ________
a) Cost of computing platform
b) Power budget
c) Radio propagation characteristics
d) Statistical distribution of transmitted power

18. Which of the following is a drawback of zero forcing algorithm?
a) Long training sequence
b) Amplification of noise
c) Not suitable for static channels
d) Non zero ISI

19. Diversity decisions are made by
a) Receiver
b) Transmitter
c) Channel
d) Adaptive algorithms

20. Which of the following is used to prevent deep fade for rapidly varying channel?
a) Modulation
b) Demodulation
c) Macroscopic diversity technique
d) Microscopic diversity technique

21. Which of the following is not a category of space diversity technique?
a) Selection diversity
b) Time diversity
c) Feedback diversity
d) Equal gain diversity

22. Frequency diversity is implemented by transmitting information on more than one
a) Carrier frequency
b) Amplitude
c) Phase
d) Modulation scheme

23. A RAKE receiver collects the __________ versions of the original signal.
a) Time shifted
b) Amplitude shifted
c) Frequency shifted
d) Phase shifted

24. RAKE receiver uses separate _________ to provide the time shifted version of the signal.
a) IF receiver
b) Equalizer
c) Correlation receiver
d) Channel

25. Each correlation receiver in RAKE receiver is adjusted in
a) Frequency shift
b) Amplitude change
c) Phase shift
d) Time delay

26. The range of time delays that a particular correlator can search is called
a) Search window
b) Sliding window
c) Time span
d) Dwell time

27. RAKE receiver is used for _______ technique.
a) CDMA
b) TDMA
c) FDMA
d) OFDM

28. A RAKE receiver uses __________ to separately detect the M strongest signals.
a) Single correlator
b) Multiple correlator
c) Single IF receiver
d) Multiple IF receivers

29. A RAKE receiver uses
a) Equalization
b) Channel coding
c) Diversity
d) Encryption

30. Interleaving is used to obtain ___________ diversity.
a) Time
b) Frequency
c) Polarization
d) Antenna

CONCLUSION

In this unit, the Realization of Independent Fading Paths with Receiver Diversity was discussed in detailed. The different types of diversity receivers like Selection Combining Threshold Combining, Maximal Ratio and Equal-Gain Combining schemes were elaborated. The RAKE is essentially another form of diversity combining, since the spreading code induces a path diversity on the transmitted signal so that independent multipath components separated by more than a chip time can be resolved. The name RAKE comes from the notion that the multibranch receiver resembles a garden rake, and has the effect of raking up the energy associated with the multipath components on each of its branches. The RAKE was invented in the 1950s to deal with the ionospheric multipath on a spread spectrum HF transcontinental link. The different equalizer types-linear-ZFE and adaptive, DFE and their structures were discussed. The Alamouti scheme that is a simple transmit diversity scheme suitable for two transmit antennas is decrsibed.

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ASSIGNMENT

1. What is the need for Selection Combining in diversity receivers?
2. Why does an equalizer that tracks the channel during data transmission still need to train periodically? Name the benefits of tracking.
3. Describe the concept of RAKE receiver.
4. Explain in detail about Decision-Feedback Equalization.
5. Explain in detail about the Alamouti scheme of MIMO diversity

MOBILE COMMUNICATION AND NETWORKS

UNIT V MIMO AND MULTIPLEXING

Prepared by

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AIM & OBJECTIVES

- To understand the issues involved in mobile communication system design and analysis.
- To understand the characteristics of wireless channels.
- To know the fundamental limits on the capacity of wireless channels.
- To acquire knowledge about different types of Diversity receivers.

PRE TEST-MCQ TYPE

1. Which of the following is the world's first cellular system to specify digital modulation and network level architecture?
a) GSM
b) AMPS
c) CDMA
d) IS-54

2. Previously in 1980s, GSM stands for
a) Global system for mobile
b) Groupe special mobile
c) Global special mobile
d) Groupe system mobile

3. US digital cellular system based on CDMA was standardized as
a) IS-54
b) IS-136
c) IS-95
d) IS-76

4. Which is used to resolve and combine multipath components?
a) Equalizer
b) Registers
c) RAKE receiver
d) Frequency divider

5. MIMO was initially developed in the year
a) 1980
b) 1990
c) 1980
d) 1975

UNIT V MIMO AND MULTIPLEXING

MIMO and space time signal processing, spatial multiplexing, diversity/multiplexing tradeoff. Performance measures- Outage, average SNR, average symbol/bit error rate. System examples- GSM, EDGE, GPRS, IS-95, CDMA 2000 and WCDMA

THEORY

Introduction

In this unit, consideration of systems with multiple antennas at the transmitter and receiver, which are commonly referred to as multiple input multiple output (MIMO) systems is done. The multiple antennas can be used to increase data rates through multiplexing or to improve performance through diversity. In MIMO systems the transmit and receive antennas can both be used for diversity gain. Multiplexing is obtained by exploiting the structure of the channel gain matrix to obtain independent signalling paths that can be used to send independent data. These spectral efficiency gains often require accurate knowledge of the channel at the receiver, and sometimes at the transmitter as well. In addition to spectral efficiency gains, ISI and interference from other users can be reduced using smart antenna techniques. The cost of the performance enhancements obtained through MIMO techniques is the added cost of deploying multiple antennas, the space and power requirements of these extra antennas (especially on small handheld units), and the added complexity required for multi-dimensional signal processing.

Narrowband MIMO Model

In this section a narrowband MIMO channel is considered. A narrowband point-to-point communication system of *Mt* transmit and *Mr* receive antennas is shown in Figure. This system can be represented by the following discrete time model:

$$\begin{bmatrix} y_1 \\ \vdots \\ y_{M_r} \end{bmatrix} = \begin{bmatrix} h_{11} & \cdots & h_{1M_t} \\ \vdots & \ddots & \vdots \\ h_{M_r1} & \cdots & h_{M_rM_t} \end{bmatrix} \begin{bmatrix} x_1 \\ \vdots \\ x_{M_t} \end{bmatrix} + \begin{bmatrix} n_1 \\ \vdots \\ n_{M_r} \end{bmatrix}$$

or simply as y = Hx + n. Here x represents the *Mt*-dimensional transmitted symbol, n is the *Mr*-dimensional noise vector, and H is the *Mr* × *Mt* matrix of channel gains *hij* representing the gain from transmit antenna *j* to receive antenna *i*.

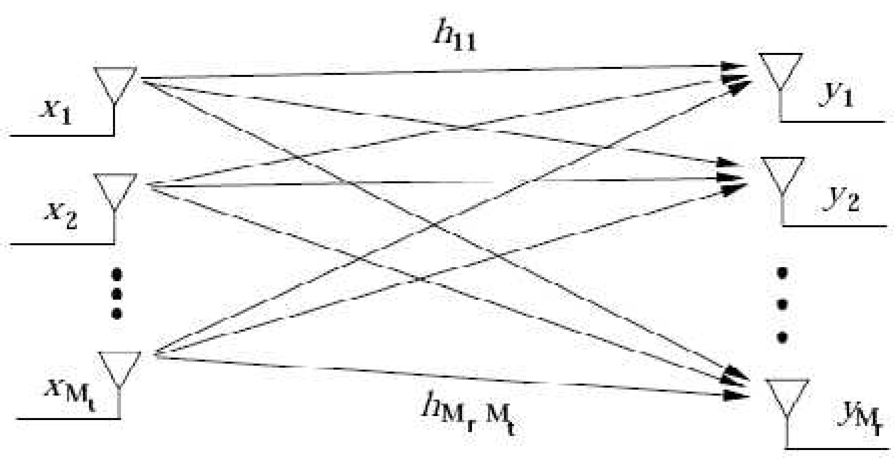


Figure MIMO systems

Assume a channel bandwidth of B and complex Gaussian noise with zero mean and covariance matrix $\sigma 2n$ IMr , where typically $\sigma 2n = N0B$. For simplicity, given a transmit power constraint P will assume an equivalent model with a noise power of unity and transmit power $P/\sigma 2n = \rho$, where ρ can be interpreted as the average SNR per receive antenna under unity channel gain. This power constraint implies that the input symbols satisfy

$$\sum_{i=1}^{M_t} \mathrm{E}[x_i x_i^*] = \rho,$$

or, equivalently, that Tr(Rx) = ρ, where Tr(Rx) is the trace of the input covariance matrix Rx = E[xxT]. Different assumptions can be made about the knowledge of the channel gain matrix H at the transmitter and receiver, referred to as channel side information at the transmitter (CSIT) and channel side information at the receiver (CSIR), respectively. For a static channel CSIR is typically assumed, since the channel gains can be obtained fairly easily by sending a pilot sequence for channel estimation.

If a feedback path is available then CSIR from the receiver can be sent back to the transmitter to provide CSIT: CSIT may also be available in time-division duplexing systems without a feedback path by exploiting reciprocal properties of propagation. When the channel is not known at either the transmitter or receiver then some distribution on the channel gain matrix must be assumed. The most common model for this distribution is a zero-mean spatially white (ZMSW) model, where the entries of H are assumed to be i.i.d. zero mean, unit variance, complex circularly symmetric Gaussian random variables1. This model is adopted unless stated otherwise. Alternatively, these entries may be complex circularly symmetric Gaussian random variables with a non-zero mean or with a covariance matrix not equal to the identity matrix. In general, different assumptions about CSI and about the distribution of the H entries lead to different channel capacities and different approaches to space-time signalling.

Optimal decoding of the received signal requires ML demodulation. If the symbols modulated onto each of the Mt transmit antennas are chosen from an alphabet of size $|X|$, then because of the cross-coupling between transmitted symbols at the receiver antennas, ML demodulation requires an exhaustive search over all $|X|Mt$ possible input vector of Mt symbols. For general channel matrices, when the transmitter does not know H this complexity cannot be reduced further. This decoding complexity is typically prohibitive for even a small number of transmit antennas. However, decoding complexity is significantly reduced if the channel is known at the transmitter

Parallel Decomposition of the MIMO Channel

The multiple antennas at the transmitter or receiver can be used for diversity gain. When *both* the transmitter and receiver have multiple antennas, there is another mechanism for performance gain called multiplexing gain. The multiplexing gain of a MIMO system results from the fact that a MIMO channel can be decomposed into a number R of parallel independent channels. By multiplexing independent data onto these independent channels to get an R-fold increase in data rate in comparison to a system with just one antenna at the transmitter and receiver.

This increased data rate is called the multiplexing gain. In this section, how to obtain independent channels from a MIMO system is described.

Consider a MIMO channel with $Mr \times Mt$ channel gain matrix H known to both the transmitter and the receiver. Let *RH* denote the rank of H. From matrix theory, for any matrix H, Now obtain its singular value decomposition (SVD) as

$$H = U\Sigma VH,$$

where the $Mr\times Mr$ matrix U and the $Mt\times Mt$ matrix V are unitary matrices2 and Σ is an $Mr\times Mt$ diagonal matrix of singular values $\{\sigma i\}$ of H. These singular values have the property that $\sigma i = \sqrt{\lambda i}$ for λi the *i*th eigen value of *HHH*, and *RH* of these singular values are nonzero, where *RH* is the rank of the matrix H. Since *RH* cannot exceed the number of columns or rows of H, $RH \le \min(Mt,Mr)$. If H is full rank, which is sometimes referred to as a rich scattering environment, then $RH = \min(Mt,Mr)$. Other environments may lead to a low rank H: a channel with high correlation among the gains in H may have rank 1.

The parallel decomposition of the channel is obtained by defining a transformation on the channel input and output x and y through transmit precoding and receiver shaping. In transmit precoding the input to the antennas x is generated through a linear transformation on input vector ˜x as x = VH˜x. Receiver shaping performs a similar operation at the receiver by multiplying the channel output y with *U H*, as shown in Figure.

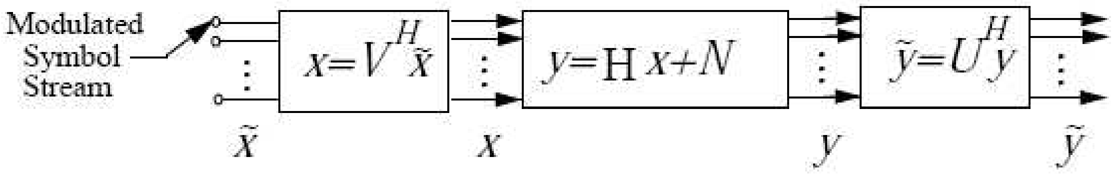


Figure Transmit Precoding and Receiver Shaping

The transmit precoding and receiver shaping transform the MIMO channel into RH parallel single-input single-output (SISO) channels with input ˜x and output ˜y, since from the SVD,

$$\begin{aligned}
\tilde{\mathbf{y}} &= \mathbf{U}^H(\mathbf{H}\mathbf{x} + \mathbf{n}) \\
&= \mathbf{U}^H(\mathbf{U}\Sigma\mathbf{V}\mathbf{x} + \mathbf{n}) \\
&= \mathbf{U}^H(\mathbf{U}\Sigma\mathbf{V}\mathbf{V}^H\tilde{\mathbf{x}} + \mathbf{n}) \\
&= \mathbf{U}^H\mathbf{U}\Sigma\mathbf{V}\mathbf{V}^H\tilde{\mathbf{x}} + \mathbf{U}^H\mathbf{n} \\
&= \Sigma\tilde{\mathbf{x}} + \tilde{\mathbf{n}},
\end{aligned}$$

where ˜n = U*H*n and Σ is the diagonal matrix of singular values of H with *σi* on the *i*th diagonal. Note that multiplication by a unitary matrix does not change the distribution of the noise, i.e. n and ˜n are identically distributed. Thus, the transmit precoding and receiver shaping transform the MIMO channel into RH parallel independent channels where the ith channel has input ˜xi, output ˜yi, noise ˜ni, and channel gain σi. Note that the σis are related since they are all functions of H, but since the resulting parallel channels do not interfere with each other, and say that the channels with these gains are independent, linked only through the total power constraint.

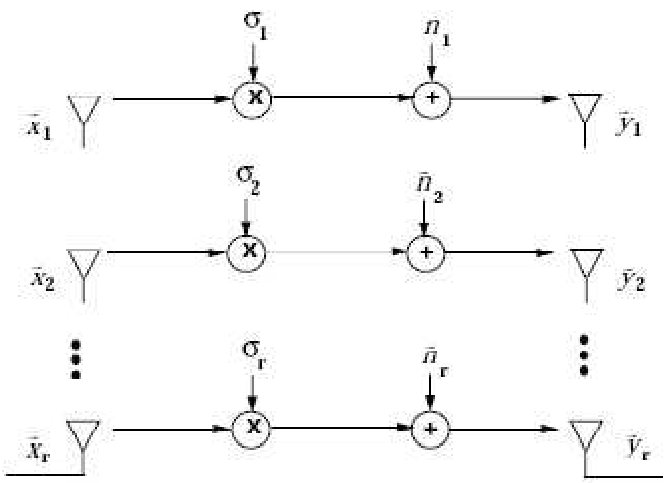


Figure Parallel Decomposition of the MIMO Channel

This parallel decomposition is shown in Figure. Since the parallel channels do not interfere with each other, the optimal ML demodulation complexity is linear in RH, the number of independent paths that need to be decoded. Moreover, by sending independent data across each of the parallel channels, the MIMO channel can support RH times the data rate of a system with just one transmit and receive antenna, leading to a multiplexing gain of RH. Note, however, that the performance on each of the channels will depend on its gain σi. The next section will more precisely characterize the multiplexing gain associated with the Shannon capacity of the MIMO channel.

Spatial Multiplexing and BLAST Architectures

The basic premise of spatial multiplexing is to send *Mt* independent symbols per symbol period using the dimensions of space and time. In order to get full diversity order an encoded bit stream must be transmitted over all *Mt* transmit antennas. This can be done through a serial encoding, illustrated in Figure. With serial encoding the bit stream is temporally encoded over the channel block length *T*, interleaved, and mapped to a constellation point, then demultiplexed onto the different antennas. If each codeword is sufficiently long, it can be transmitted over all *Mt* transmit antennas and received by all *Mr* receive antennas, resulting in full diversity gain. However, the codeword length *T* required to achieve this full diversity is *MtMr*, and decoding complexity grows exponentially with this codeword length. This high level of complexity makes serial encoding impractical.

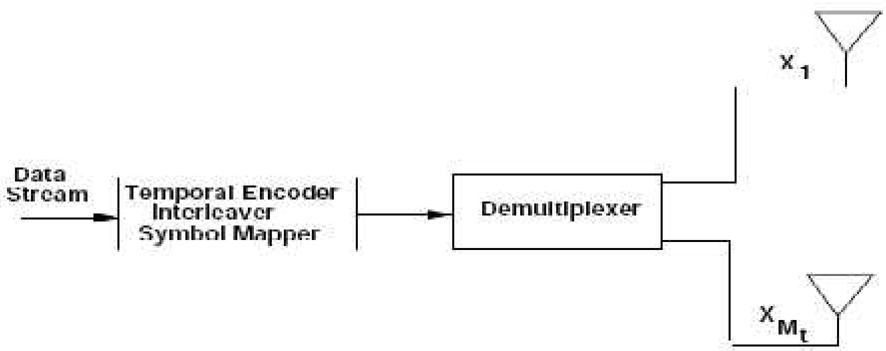


Figure Spatial Multiplexing with Serial Encoding

A simpler method to achieve spatial multiplexing, pioneered at Bell Laboratories as one of the Bell Labs Layered Space Time (BLAST) architectures for MIMO channels, is parallel encoding, illustrated in Figure. With parallel encoding the data stream is demultiplexed into *Mt* independent streams. Each of the resulting substreams is passed through a SISO temporal encoder with blocklenth *T*, interleaved, mapped to a signal constellation point, and transmitted over its corresponding transmit antenna. This process can be considered to be the encoding of the serial data into a vertical vector, and hence is also referred to as vertical encoding or V-BLAST .

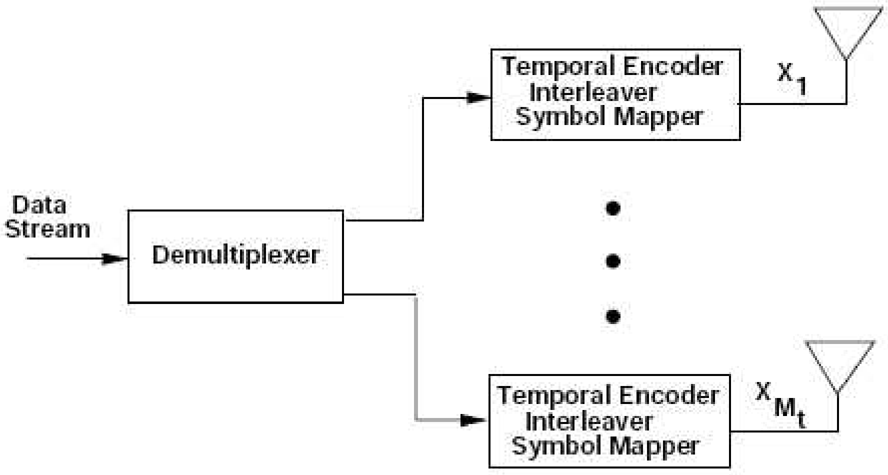


Figure Spatial Multiplexing with Parallel Encoding: VBLAST

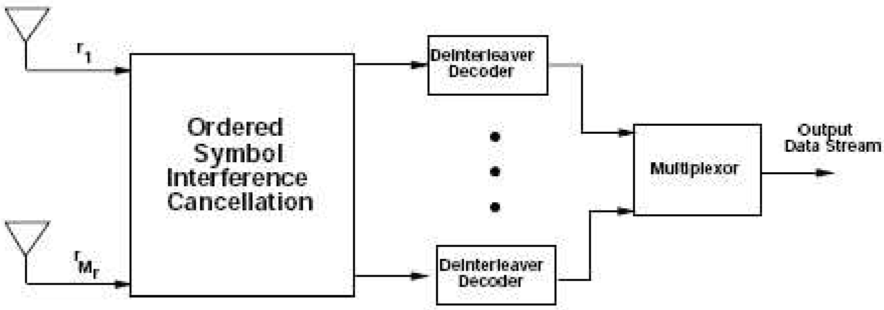


Figure VBLAST Receiver with Linear Complexity

Vertical encoding can achieve at most a diversity order of*Mr*, since each coded symbol is transmitted from one antenna and received by *Mr* antennas. This system has a simple encoding complexity that is linear in the number of antennas. However, optimal decoding still requires joint detection of the codewords from each of the transmit antennas, since all transmitted symbols are received by all the receive antennas. It was shown , that the receiver complexity can be significantly reduced through the use of symbol interference cancellation, as shown in Figure. The symbol interference cancellation, which exploits the synchronicity of the symbols transmitted from each antenna, works as follows. First the*Mt* transmitted symbols are ordered in terms of their received SNR. An estimate of the received symbol with the highest SNR is made while treating all other symbols as noise. This estimated symbol is subtracted out, and the symbol with the next highest SNR estimated while treating the remaining symbols as noise. This process repeats until all *Mt* transmitted symbols have been estimated. After cancelling out interfering symbols, the coded substream associated with each transmit antenna can be individually decoded, resulting in a receiver complexity that is linear in the number of transmit antennas. In fact, coding is not even needed with this architecture, and data rates of 20-40 bps/Hz with reasonable error rates were reported in using uncoded V-BLAST.

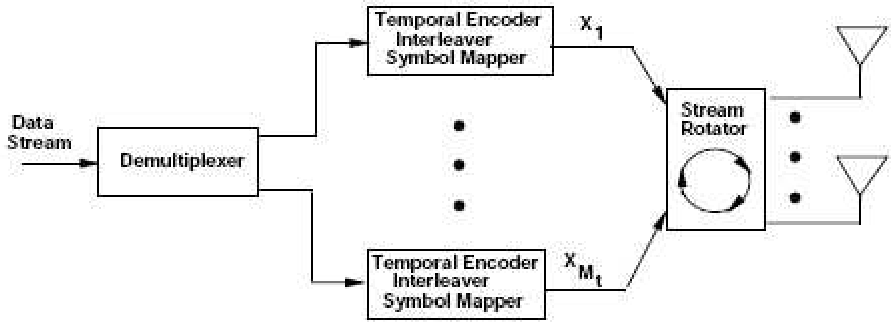


Figure Diagonal Encoding with Stream Rotation

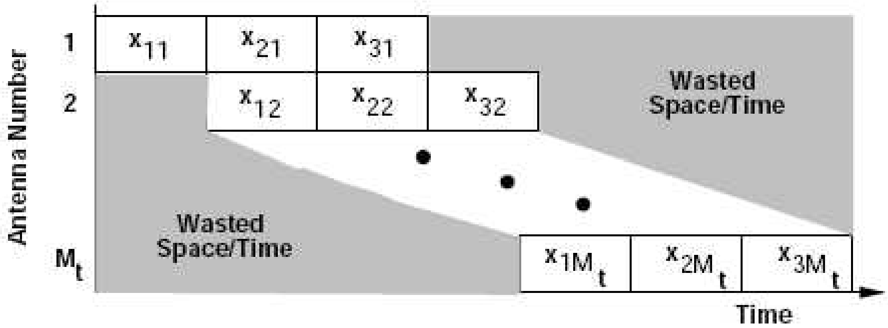


Figure Stream Rotation

The simplicity of parallel encoding and the diversity benefits of serial encoding can be obtained using a creative combination of the two techniques called diagonal encoding or D-BLAST illustrated in Figure. In D-BLAST, the data stream is first horizontally encoded. However, rather than transmitting the independent codewords on separate antennas, the codeword symbols are rotated across antennas, so that a codeword is spread over all *Mt* antennas. The operation of the stream rotation is shown in Figure. Suppose the *i*th encoder generates the codeword x*i* = *xi*1, . . . , *xiMt* . The stream rotator transmits each coded symbol on a different antenna, so *xi*1 is sent on antenna 1, *xi*2 is sent on antenna 2, and so forth. If the code block length *T* exceeds *Mt* then the rotation begins again on the 1st antenna. As a result, the codeword is spread across all spatial dimensions. Transmission schemes based on D-BLAST can achieve the full *MtMr* diversity gain if the temporal coding with stream rotation is capacity-achieving. Moreover, the D-BLAST system can achieve the maximum capacity with outage if the wasted space-time dimensions along the diagonals are neglected. Receiver complexity is also linear in the number of transmit antennas, since the receiver decodes each diagonal code independently. However, this simplicity comes as a price, as the efficiency loss of the wasted space-time dimensions illustrated in Figure can be large if the frame size is not appropriately chosen.

Diversity/Multiplexing Tradeoffs

The previous sections suggest two mechanisms for utilizing multiple antennas to improve wireless system performance. One option is to obtain capacity gain by decomposing the MIMO channel into parallel channels and multiplexing different data streams onto these channels. This capacity gain is also referred to as a multiplexing gain. However, the SNR associated with each of these channels depends on the singular values of the channel matrix. In capacity analysis this is taken into account by assigning a relatively low rate to these channels. However, practical signaling strategies for these channels will typically have poor performance, unless powerful channel coding techniques are employed. Alternatively, beamforming can be used, where the channel gains are coherently combined to obtain a very robust channel with high diversity gain. It is not necessary to use the antennas purely for multiplexing or diversity. Some of the space-time dimensions can be used for diversity gain, and the remaining dimensions used for multiplexing gain.

This gives rise to a fundamental design question in MIMO systems: should the antennas be used for diversity gain, multiplexing gain, or both? The diversity/multiplexing tradeoff or, more generally, the tradeoff between data rate, probability of error, and complexity for MIMO systems has been extensively studied in the literature, from both a theoretical perspective and in terms of practical space-time code designs.

This work has primarily focused on block fading channels with receiver CSI only since when both transmitter and receiver know the channel the tradeoff is relatively straightforward: antenna subsets can first be grouped for diversity gain and then the multiplexing gain corresponds to the new channel with reduced dimension due to the grouping. For the block fading model with receiver CSI only, as the block length grows asymptotically large, full diversity gain and full multiplexing gain (in terms of capacity with outage) can be obtained simultaneously with reasonable complexity by encoding diagonally across antennas. For finite block lengths it is not possible to achieve full diversity and full multiplexing gain simultaneously, in which case there is a tradeoff between these gains. A simple characterization of this tradeoff is given for block fading channels with block length $T \geq Mt + Mr - 1$ in the limit of asymptotically high SNR. In this analysis a transmission scheme is said to achieve multiplexing gain r and diversity gain d if the data rate (bps) per unit Hertz R(SNR) and probability of error Pe(SNR) as functions of SNR satisfy

$$\lim_{\log_2 \mathrm{SNR} \to \infty} \frac{R(\mathrm{SNR})}{\log_2 \mathrm{SNR}} = r,$$

$$\lim_{\log \mathrm{SNR} \to \infty} \frac{\log P_e(\mathrm{SNR})}{\log \mathrm{SNR}} = -d,$$

where the log in second can be in any base5. For each r the optimal diversity gain $dopt(r)$ is the maximum the diversity gain that can be achieved by any scheme. It is shown, that if the fading block length exceeds the total number of antennas at the transmitter and receiver, then

$$d_{opt}(r) = (M_t - r)(M_r - r), \ 0 \leq r \leq \min(M_t, M_r).$$

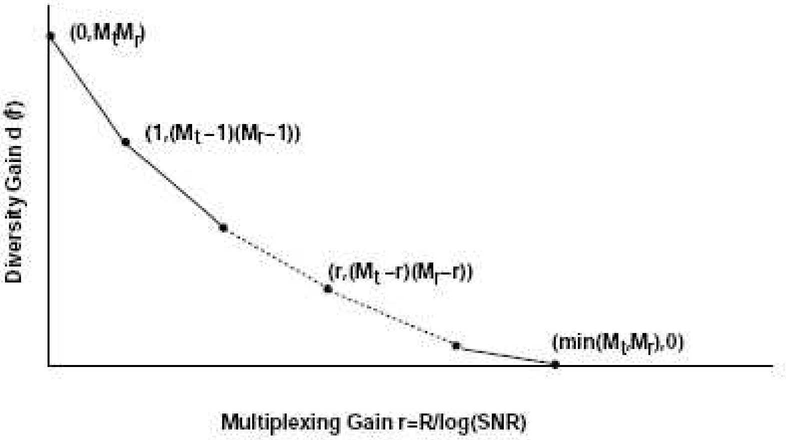


Figure Diversity-Multiplexing Tradeoff for High SNR Block Fading

It is also possible to adapt the diversity and multiplexing gains relative to channel conditions. Specifically, in poor channel states more antennas can be used for diversity gain, whereas in good states more antennas can be used for multiplexing.

GSM Architecture

GSM is a cellular network, which means that mobile phones connect to it by searching for cells in the immediate vicinity. GSM networks operate in four different frequency ranges. Most GSM networks operate in the 900 MHz or 1800 MHz bands..In the 900 MHz band the uplink frequency band is 890-915 MHz, and the downlink frequency band is 935-960 MHz. This 25 MHz bandwidth is subdivided into 124 carrier frequency channels, each spaced 200 kHz apart. Time division multiplexing is used to allow eight full-rate or sixteen half-rate speech channels per radio frequency channel. There are eight radio timeslots (giving eight burst periods) grouped into what is called a TDMA frame. Half rate channels use alternate frames in the same timeslot. The channel data rate is 270.833 kbit/s, and the frame duration is 4.615 ms.The transmission power in the handset is limited to a maximum of 2 watts in GSM850/900 and 1 watt in GSM1800/1900.

The structure of a GSM network GSM has used a variety of voice codecs to squeeze 3.1kHz audio into between 6 and 13kbps. Originally, two codecs, named after the types of data channel they were allocated, were used, called "Full Rate" (13kbps) and "Half Rate" (6kbps). These used a system based upon linear predictive coding (LPC). In addition to being efficient with bitrates, these codecs also made it easier to identify more important parts of the audio, allowing the air interface layer to prioritize and better protect these parts of the signal.GSM was further enhanced in the mid-nineties with the GSM-EFR codec, a 12.2kbps codec that uses a full rate channel.

Finally, with the development of UMTS, EFR was refactored into a variable-rate codec called AMR-Narrowband, which is high quality and robust against interference when used on full rate channels, and less robust but still relatively high quality when used in good radio conditions on half-rate channels.There are four different cell sizes in a GSM network - macro, micro, pico and umbrella cells. The coverage area of each cell varies according to the implementation environment. Macro cells can be regarded as cells where the base station antenna is installed on a mast or a building above average roof top level. Micro cells are cells whose antenna height is under average roof top level; they are typically used in urban areas. Picocells are small cells whose diameter is a few dozen meters; they are mainly used indoors. On the other hand, umbrella cells are used to cover shadowed regions of smaller cells and fill in gaps in coverage between those cells.

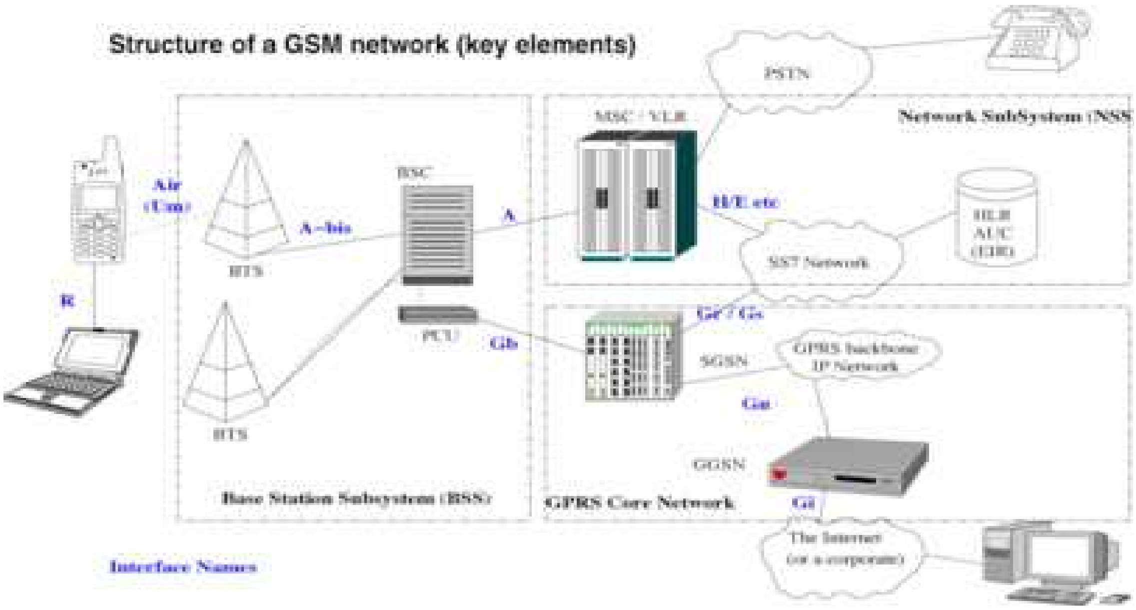


Figure The structure of a GSM network

Cell radius varies depending on antenna height, antenna gain and propagation conditions from a couple of hundred meters to several tens of kilometers. The longest distance the GSM specification supports in practical use is 35 km or 22 miles. There are also several implementations of the concept of an extended cell, where the cell radius could be double or even more, depending on the antenna system, the type of terrain and the timing advance.Indoor coverage is also supported by GSM and may be achieved by using an indoor picocell base station, or an indoor repeater with distributed indoor antennas fed through power splitters, to deliver the radio signals from an antenna outdoors to the separate indoor distributed antenna system. These are typically deployed when a lot of call capacity is needed indoors, for example in shopping centers or airports. However, this is not a prerequisite, since indoor coverage is also provided by in-building penetration of the radio signals from nearby cells.

The modulation used in GSM is Gaussian minimum shift keying (GMSK), a kind of continuous-phase frequency shift keying. In GMSK, the signal to be modulated onto the carrier is first smoothed with a Gaussian low-pass filter prior to being fed to a frequency modulator, which greatly reduces the interference to neighboring channels (adjacent channel interference).

Network structure

The network behind the GSM system seen by the customer is large and complicated in order to provide all of the services which are required. It is divided into a number of sections and these are each covered in separate articles.

• the Base Station Subsystem (the base stations and their controllers).

• the Network and Switching Subsystem (the part of the network most similar to a fixed network). This is sometimes also just called the core network.

• the GPRS Core Network (the optional part which allows packet based Internet connections).

• all of the elements in the system combine to produce many GSM services such as voice calls and SMS.

Subscriber identity module

One of the key features of GSM is the Subscriber Identity Module (SIM), commonly known as a SIM card. The SIM is a detachable smart card containing the user's subscription information and phonebook. This allows the user to retain his or her information after switching handsets.

GSM security

GSM was designed with a moderate level of security. The system was designed to authenticate the subscriber using shared-secret cryptography. Communications between the subscriber and the base station can be encrypted. GSM uses several cryptographic algorithms for security. The A5/1 and A5/2 stream ciphers are used for ensuring over-the-air voice privacy. A5/1 was developed first and is a stronger algorithm used within Europe and the United States; A5/2 is weaker and used in other countries.

A large security advantage of GSM over earlier systems is that the Ki, the crypto variable stored on the SIM card that is the key to any GSM ciphering algorithm, is never sent over the air interface. Serious weaknesses have been found in both algorithms, and it is possible to break A5/2 in real-time in a cipher text-only attack. The system supports multiple algorithms so operators may replace that cipher with a stronger one.

GPRS

General Packet Radio Services (GPRS) is a mobile data service available to users of GSM and IS-136 mobile phones. GPRS data transfer is typically charged per megabyte of transferred data, while data communication via traditional circuit switching is billed per minute of connection time, independently of if the user actually has transferred data or been in an idle state. GPRS can be utilized for services such as WAP access, SMS and MMS, but also for Internet communication services such as email and web access.2G cellular systems combined with GPRS is often described as "2.5G", that is, a technology between the second and third generations of mobile telephony. It provides moderate speed data transfer, by using unused TDMA channels.

GPRS is different from the older Circuit Switched Data (or CSD) connection included in GSM standards. In CSD, a data connection establishes a circuit, and reserves the full bandwidth of that circuit during the lifetime of the connection. GPRS is packetswitched which means that multiple users share the same transmission channel, only transmitting when they have data to send. This means that the total available bandwidth can be immediately dedicated to those users who are actually sending at any given moment, providing higher utilisation where users only send or receive data intermittently. Web browsing, receiving e-mails as they arrive and instant messaging are examples of uses that require intermittent data transfers, which benefit from sharing the available bandwidth.The multiple access methods used in GSM with GPRS is based on frequency division duplex (FDD) and FDMA. During a session, a user is assigned to one pair of uplink and downlink frequency channels. This is combined with time domain statistical multiplexing, i.e. packet mode communication, which makes it possible for several users to share the same frequency channel. The packets have constant length, corresponding to a GSM time slot. In the downlink, first-come first-served packet scheduling is used. In the uplink, a scheme that is very similar to reservation ALOHA is used. This means that slotted Aloha (S-ALOHA) is used for reservation inquiries during a contention phase, and then the actual data is transferred using first-come first-served scheduling.

GPRS speeds and profile

Packet-switched data under GPRS is achieved by allocating unused cell bandwidth to transmit data. As dedicated voice (or data) channels are setup by phones, the bandwidth available for packet switched data shrinks. A consequence of this is that packet switched data has a poor bit rate in busy cells. The theoretical limit for packet switched data is 171.2 kbit/s (using 8 time slots and CS-4 coding). A realistic bit rate is 30–80 kbit/s, because it is possible to use max 4 time slots for downlink. A change to the radio part of GPRS called EDGE (sometimes called *EGPRS* or *Enhanced GPRS* however it actually stands for Enhanced Data rates for GSM Evolution) allows higher bit rates of between 160 and 236.8 kbit/s. The maximum data rates are achieved only by allocation of more than one time slot in the TDMA frame. Also, the higher the data rate, the lower the error correction capability. Generally, the connection speed drops logarithmically with distance from the base station. This is not an issue in heavily populated areas with high cell density, but may become an issue in sparsely populated/rural areas.

GPRS coding scheme

Transfer speed depends also on the channel encoding used. The least robust (but fastest) coding scheme (CS-4) is available near the Base Transceiver Station (BTS) while the most robust coding scheme (CS-1) is used when the Mobile Station (MS) is further away from the BTS.Using the CS-4 it is possible to achieve a user speed of 20.0 kbit/s per time slot. However, using this scheme the cell coverage is 25% of normal. CS-1 can achieve a user speed of only 8.0 kbit/s per time slot, but has 98% of normal coverage. Newer network equipment can adapt the transfer speed automatically depending on the mobile location.

GPRS upgrades GSM data services providing:

• MMS - Multimedia Messaging Service

• Push To Talk over Cellular PoC / PTT - Push to talk

• Instant Messaging and Presence Wireless_Village

• Internet Applications for Smart Devices through WAP

• Point-to-point (PTP) service: internetworking with the Internet (IP protocols).

• Short Message Service (SMS): bearer for SMS.

• Future enhancements: flexible to add new functions, such as more capacity, more users, new accesses, new protocols, new radio networks.

IS-95 Cellular System

The IS-95 standard describes a Code Division Multiple Access (CDMA) system in which the audio band data signal is multiplied by a high rate spreading signal. This spreading signal is formed from a pseudo-noise code sequence, which is then multiplied by a Walsh code for maximum orthogonality to (ie. to have low cross-correlation with) the other codes in use in that cell. Typically, CDMA pseudo-noise sequences are very long, thereby giving excellent crosscorrelation characteristics. (IS-95 uses a 242-1 chip period, derived from a 42 bit mask.) The IS-95 system can be thought of as having many layers of protection against interference. It allows many users to co-exist, with minimal mutual interference. They can be described by the signal conditioning sequence that occurs on forward and reverse channels. The forward channel carries information from the base station to the mobile unit; the reverse channel carries information from the mobile unit to the base station. The transmission channels are shown; the reception of each channel follows the reverse sequence. The forward channels are between 869 and 894 MHz, while the reverse channels are between 824 and 849 Mhz. Within these bands, four sub-bands are available for CDMA, of widths 1, 0.1, 9 and 10 MHz; in the U.S., 1.25 MHz sub-bands near 849 and 894 MHz are employed.

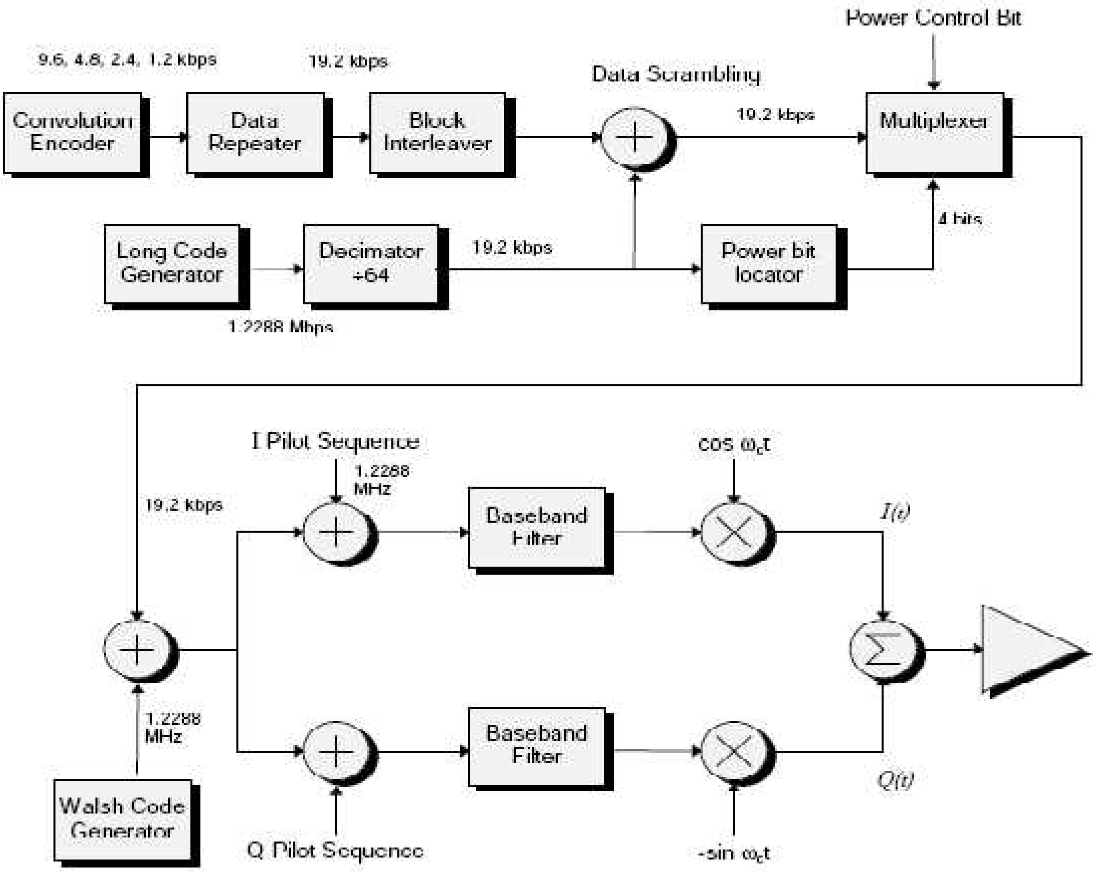


Figure Forward CDMA channel

All cells in the same area can employ the same spectral band, because the various signals are sorted out by the spread spectrum process rather than by frequency discrimination.

Forward channel transmission sequence

1. Convolution encoder- Encodes the data from one stream to two, doubling the nominal rate from 9.6 kpbs to 19.2 kbps, 4.8 kbps data to 9.6 kbps, etc.

2. Repetition circuit- Repeats coded symbols, so lower rate encoded data is increased from 9.6, 4.8 or 2.4 kbps to 19.2 kbps.

3. Block Interleaver- Reads data into the rows of a 24 x 16 array, and out of the columns; introduces a 20 msec delay, but spreads important bits (as produced by modern speech encoders) over time as proof against deep fades or noise bursts.

4. Data scrambling- The data are Modulo 2 added to every 64th bit of a pseudo-noise (PN) sequence created from a 42 bit shift register. (The resulting 242-1 bits repeat once per century after initiation.) The data rate at this point is still 19.2 kbps.

5. Power contro-l Every 1.25 msec, or 24 data symbols, a power control bit is inserted, in order to instruct the mobile unit to raise or lower its power (to equalize the power received from every mobile unit in the cell.) The location of the power control bit is determined from the PN sequence.

6. Orthogonal covering- The 19.2 kbps data are spread with a 1.2288 Mbps Walsh function, so that each one bit data symbol is spread by 64 Walsh chips. The Walsh function provides 64 mutually orthogonal binary sequences, each of length 64.

7. Quadrature spreading- The data are split into two bit streams, which are Modulo 2 added to two different but well defined “Pilot” pseudo-noise sequences generated from 15 bit shift registers. The code repeats 75 times every 2 seconds, or at 26.7 msec intervals.

8. Quadrature modulation- The binary I and Q outputs are mapped onto four phases of a quadrature modulator, at $\pm\pi/4$ and $\pm 3\pi/4$, using quadrature phase shift keying (QPSK).

9. RF modulation- The baseband quadrature data are raised to the forward cellular radio band, 869 to 894 MHz. The IS-95 channel occupies 1.25 MHz within this band, the rest of which is occupied by other cellular services such as AMPS.

Of the 64 available orthogonal channels (ie. channels which have minimum mutual interference), one is assigned to the pilot channel and one to the synchronization channel. Several low numbered channels are assigned to paging. The pilot channel corresponds to the all zeros Walsh code (Walsh channel 0), and contains the unmodulated quadrature PN spreading code. It is transmitted at higher power than the user channels, and is provided so that each subscriber within the cell can determine and react to the channel characteristics while employing coherent detection. Walsh channel 32 is assigned to the sync channel, which provides time and frame synchronization to the mobile unit. Time of day and station identification are continuously broadcast on this channel. As users are added to the system, they are assigned user channels from the available Walsh channels.

When over 60 users are present, the channels are assigned to multiple users, and protection from mutual interference within the same Walsh channel is provided by the private PN sequences that encode each user link. The number of users can therefore rise to large values, while reasonable quality is maintained.

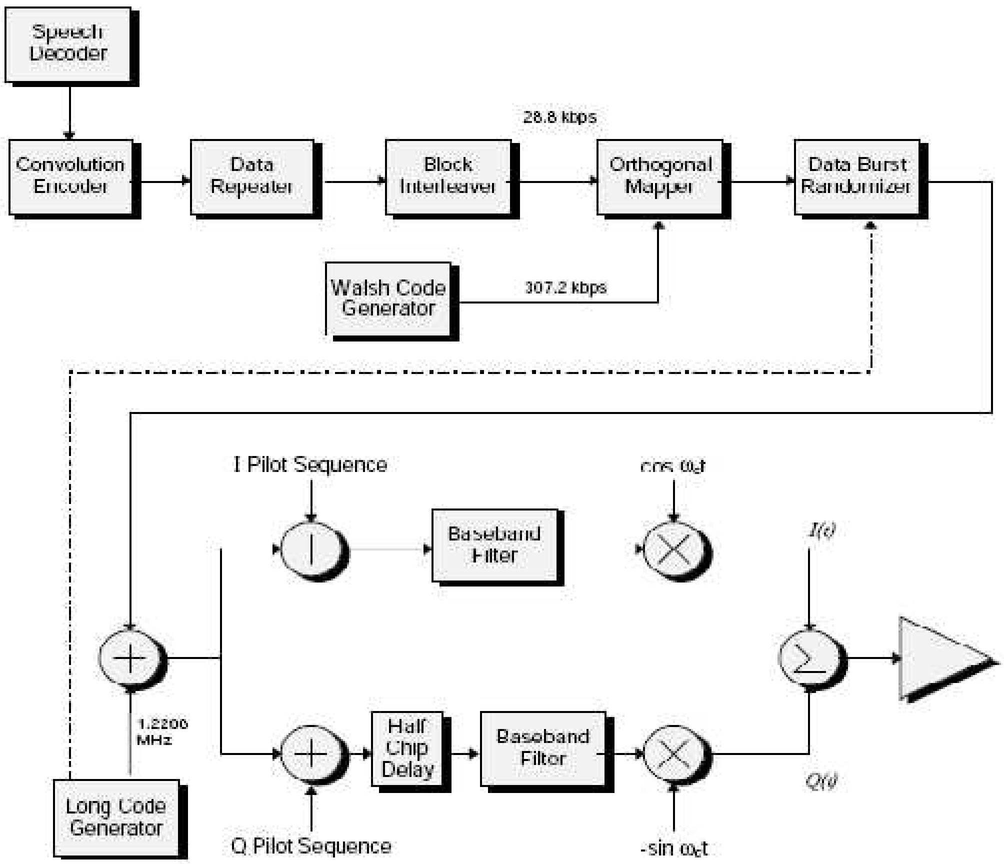


Figure Reverse CDMA channel

Reverse channel transmission sequence

1.Speech encoder - Produces nominal 9600 bps data stream, dynamically reduced to 4800, 2400, or 1200 bps during pauses and gaps in speech; quiet periods correspond to 1200 bps data.

2. Convolution encoder -Encodes the data from one stream to three, tripling the data rate from 9.6 kpbs to 28.8 kbps, 4.8 kbps data to 14.4 kbps, etc.

3. Repetition circuit- Repeats coded symbols, so lower rate encoded data is increased from 9.6, 4.8 or 2.4 kbps to 19.2 kbps.

4. Block Interleaver -Reads data into the columns of a 32 x 18 array, and out of the rows; introduces a 20 msec delay, but spreads important bits over time as proof against deep fades or noise bursts.

5. Orthogonal mapping- The 28.8 kbps data are split into sequential sets of six bits each, which are mapped to one of 64 Walsh functions. The data rate is therefore raised to 28.8 k x 64 chips/ 6 bits = 307.2 kpbs.

6. Burst Randomizing- The Walsh symbols are broken into groups of six, each group being 1.25 msec in duration. These are collected into frames of 16 power groups, or 1.25 msec x 16 = 20 msec. At 9600 bps, all 16 groups are transmitted; at 4800 bps, 8 randomly selected groups are transmitted; at 2400 bps, 4 groups; at 1200 bps, 2 groups. The transmitted groups are chosen randomly, according to a formula based on 14 bits of the PN sequence of the second last group in the previous frame.

7. Direct sequence spreading- The data are Modulo 2 added to every bit of a pseudo-noise (PN) sequence created from a 42 bit shift register. The PN sequence is generated at 1.2288 MHz, so each Walsh chip is spread by four long code PN chips.

8. Quadrature spreading- The data are split into two bit streams, which are Modulo 2 added to two different but well defined “Pilot” pseudo-noise sequences generated from 15 bit shift registers.

9. Quadrature modulation- The binary I and Q outputs are mapped onto four phases of a quadrature modulator, at ±π/4 and ±3π/4, using offset quadrature phase shift keying (OQPSK). (The Q channel is shifted by half a chip for improved spectral shaping.) 10. RF modulation The baseband quadrature data are raised to the reverse cellular radio band, 824 to 849 MHz. The IS-95 channel occupies 1.25 MHz within this band, the rest of which is occupied by other cellular services such as AMPS.

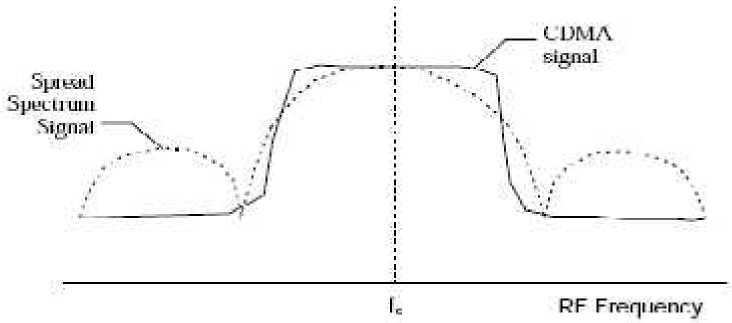


Figure **Spectral Shape of a generic Spread Spectrum signal and a CDMA signal**

Note that there is continuous transmission from a cell phone when a conversation is in progress. The lowest data rate is 1200 bps, with three 1.25 msec “power control” groups being transmitted in every 20 msec frame. A mobile phone is therefore not silent during conversations, and can be located by its telltale emissions. Spectral Considerations A generic spread spectrum occupies most of the available 1.25 MHz bandwidth. The spectral shape is best described by a sinx/x function, with many variations, such as pulse shaping, to curb out of band components.

The CDMA spectrum is nearly flat-topped, and does not have the prominent sidelobes. The final spectral shape of the CDMA forward link spectrum is given by the QPSK (quadrature phase shift keying) modulation process. The I and Q Pilot codes, at 1.2288 MHz, modulate the I and Q channels independently, and the QPSK process spreads out the spectral peaks left by the Walsh code. The 1.2288 MHz Walsh code modulates the 19.2 kbps data to produce an “orthogonal covering”. While separate Walsh codes have low cross-correlation, the Walsh code has a characteristic spectral signature.

CDMA 2000

The 2G mobile radio systems standardised by TIA in the United States are IS-95-A and IS-95-B. The radio access technology of both systems is based on narrowband DS-CDMA with a chip rate of 1.2288 Mcps, which gives a bandwidth of 1.25 MHz. IS-95-A was commercially launched in 1995, supporting circuit and packet mode transmissions at a maximum bit-rate of only 14.4 kbps . An enhancement to the IS-95-A standards, known as IS-95-B, was developed and introduced in 1998 in order to provide higher data rates, on the order of 115.2 kbps. This was feasible without changing the physical layer of IS-95-A. However, this still falls short of the 3G mobile radio system requirements. Hence the technical committee TR45.5 within TIA has proposed cdma2000, a 3G mobile radio system that is able to meet all the requirements laid down by ITU. One of the problems faced by TIA is that the frequency bands allocated for the 3G mobile radio system, identified during WARC'92 to be 1885-2025 MHz and 2110-2200 MHz, has already been allocated for Personal Communications Services (PCS) in the United States from 1.8 GHz to 2.2 GHz. In particular, the CDMA PCS based on the IS-95 standards has been allocated the frequency bands of 1850-1910 MHz and 1930-1990 GHz. Hence, the 3G mobile radio systems have to fit into the allocated bandwidth without imposing significant interference on the existing applications. Thus, the framework for cdma2000 was designed such that it can be overlaid on IS- 95 and it is backwards compatible with IS-95.

Characteristics

The cdma2000 system has a basic chip rate of 3.6864 Mcps, which is accommodated in a bandwidth of 3.75 MHz. This chip rate is in fact three times the chip rate used in the IS-95 standards, which is 1.2288 Mcps. Accordingly, the bandwidth was also trebled. Hence, the existing IS-95 networks can also be used to support the operation of cdma2000. Higher chip rates on the order of N X 1:2288 Mcps, N = 6; 9; 12 are also supported. These are used to enable higher bit-rate transmission. The value of N is an important parameter in determining the channel coding rate and the channel bit-rate.

In order to transmit the high chip-rate signals (N > 1), two modulation techniques are employed. In the direct-spread modulation mode, the symbols are spread according to the chip-rate and transmitted using a single carrier, giving a bandwidth of N X 1:25 MHz. This method is used on both the uplink and downlink. In multicarrier (MC) modulation, the symbols to be transmitted are de-multiplexed into separate signals, each of which is then spread at a chip rate of 1.2288 Mcps. N different carrier frequencies are used to transmit these spread signals, each of which has a bandwidth of 1.25 MHz. This method is used for the downlink only, since in this case, transmit diversity can be achieved by transmitting the different carrier frequencies over spatially separated antennas.

By using multiple carriers, cdma2000 is capable of overlaying its signals on the existing IS-95 1.25 MHz channels and its own channels, while maintaining orthogonality. An example of an overlay scenario is shown in Figure. Higher chip rates are transmitted at a lower power, than lower chip rates. Hence, the interferences are kept to a minimum. Similarly to UTRA and IMT-2000, cdma2000 also supports TDD operation in unpaired frequency bands. In order to ease the implementation of a dual-mode FDD/TDD terminal, most of the techniques used for FDD operation can also be applied in TDD operation. The difference between these two modes is in the frame structure, whereby an additional guard time has to be included for TDD operation.

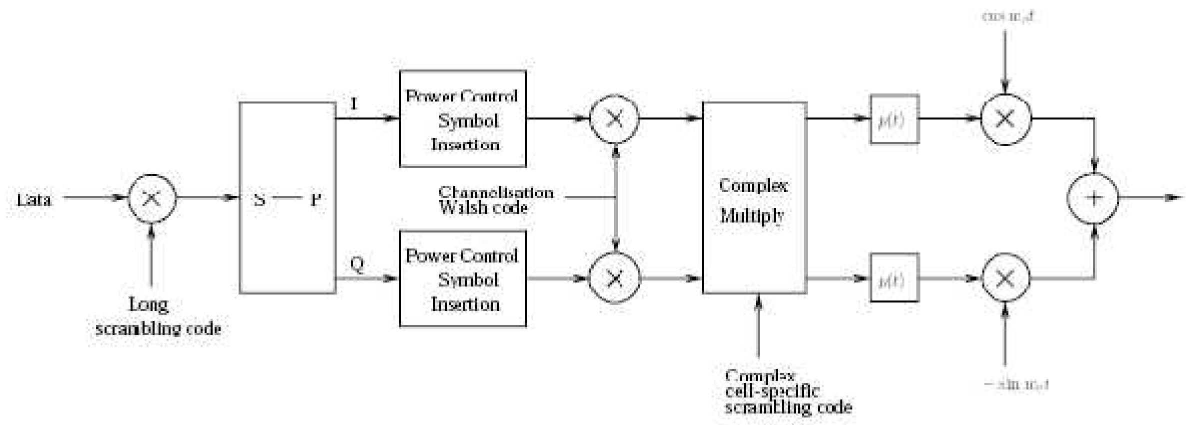


Figure The cdma2000 downlink transmitter

The Figure shows the structure of a downlink transmitter for a physical channel. In contrast to the IS-95 downlink transmitter shown earlier, the data in the cdma2000 downlink transmitter shown in Figure are first QPSK modulated before spreading using Walsh codes. As a result, the number of Walsh codes available is increased twofold due to the orthogonality of the I and Q carriers, as mentioned previously. The user data are first scrambled by the long scrambling code by assigning a different offset to different users for the purpose of improving user privacy, which is then mapped to the I and Q channels.

This long scrambling code is identical to the uplink user-specific scrambling code. The downlink pilot channels of PICH, CAPICH, DAPICH and the SYNC channel are not scrambled with a long code since there is no need for user-specificity. The uplink power control symbols are inserted into the FCH at a rate of 800 Hz, as it was shown in Figure. The I and Q channels are then spread using a Walsh code and complex multiplied with the cell-specific complex PN sequence, as portrayed in Figure. Each base station's downlink channel is assigned a different Walsh code in order to eliminate any intra-cell interference since all Walsh codes transmitted by the serving base station are received synchronously.

The length of the downlink channelization Walsh code is determined by the type of physical channel and its data rate. Typically for N = 1, downlink FCHs with data rates belonging to RS1, i.e. those transmitting at 9.6/4.8/2.7/1.5 kbps, use a 128-chip Walsh code and those in RS2, transmitting at 14.4/7.2/3.6/1.8 kbps use a 64-chip Walsh code. Walsh codes for downlink SCHs can range from 4-chip to 128-chip Walsh codes. The downlink PICH is an unmodulated sequence (all 0s) spread by Walsh code 0. Finally, the complex spread data in Figure is baseband filtered using the Nyquist’s filter impulse responses p(t) in Figure and modulated on a carrier frequency. For the case of multi-carrier modulation, the data are split into N branches immediately after the long code scrambling of which was omitted in the figure for the sake of simplicity. Each of the N branches is then treated as a separate transmitter and modulated using different carrier frequencies

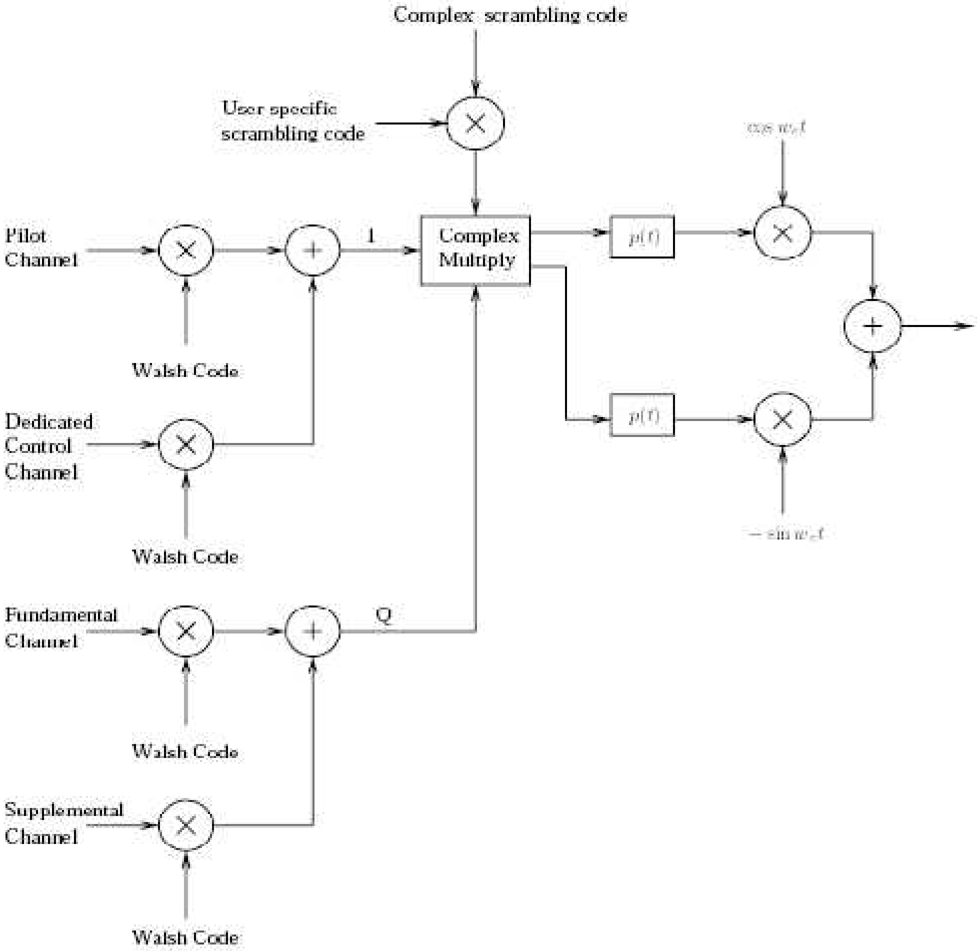
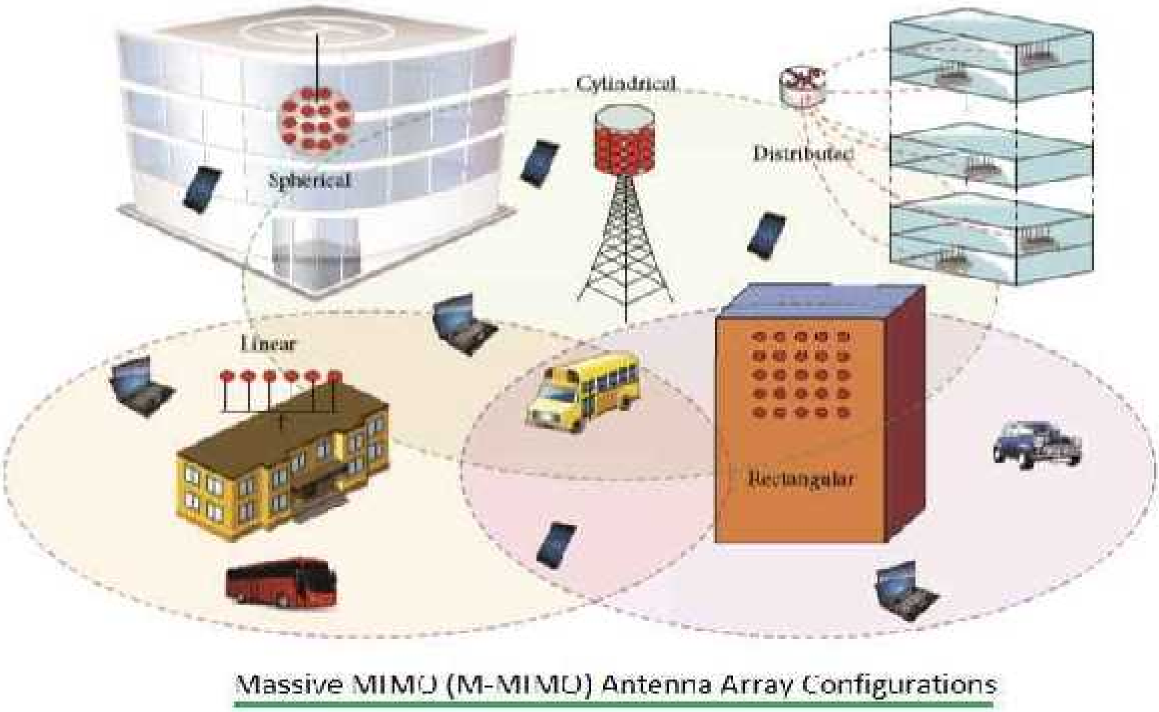


Figure The cdma2000 uplink transmitter.

The uplink cdma2000 transmitter is shown in Figure. The uplink PICH and DCCH of Table 10.9 are mapped to the I data channel, while the uplink FCH and SCH of Table 10.9 are mapped to the Q channel in Figure. Each of these uplink physical channels belonging to the same user is assigned different Walsh channelization codes in order to maintain orthogonality, with higher rate channels using shorter Walsh codes. The I and Q data channels are then spread by complex multiplication with the user-specifically offset real m-sequence based scrambling code and a complex scrambling code, which is the same for all the mobile stations in the system, as seen at the top of Figure.

However, this latter complex scrambling code is not explicitly since it is identical to the downlink cell-specific scrambling code. This complex scrambling code is only used for the purpose of quadrature spreading. Hence, in order to reduce the complexity of the base station receiver, this complex scrambling code is identical to the cell-specific scrambling code used on the downlink by all the base stations.

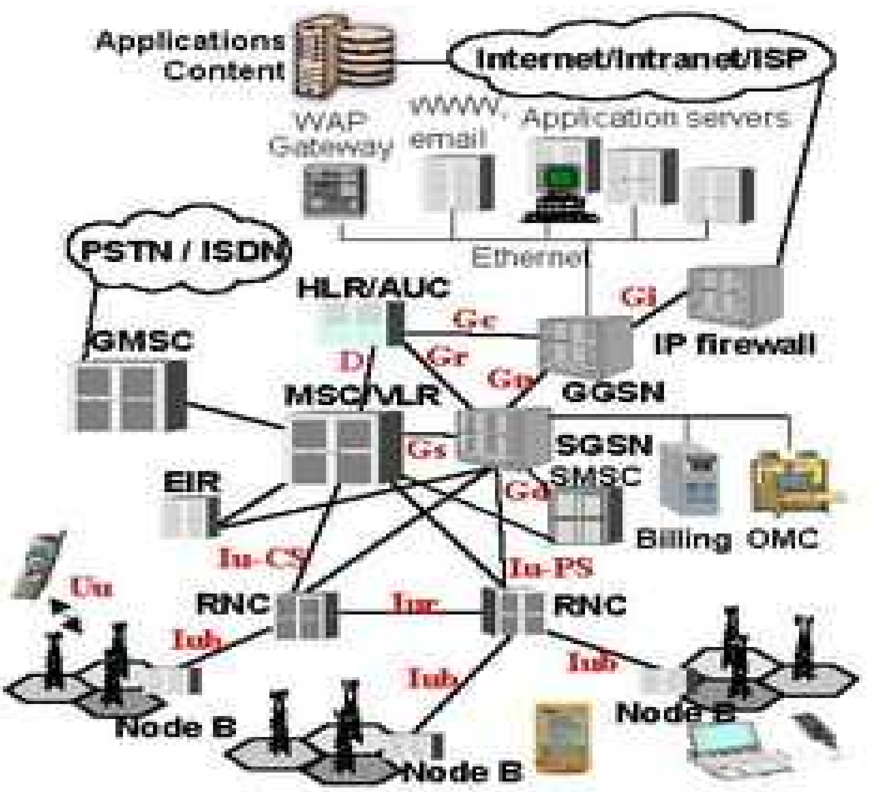
APPLICATIONS



Example of Massive MIMO

Example of MIMO application

Example of GSM/GPRS Remote Controller



Example of WCDMA & LTE

POST TEST-MCQ TYPE

1. MIMO stands for **_______**
a) Many input many output
b) Multiple input multiple output
c) Major input minor output
d) Minor input minor output

2. In MIMO, which factor has the greatest influence on data rates?
a) The size of antenna
b) The height of the antenna
c) The number of transmit antennas
d) The area of receive antennas

3. MIMO technology makes advantage of a natural radio wave phenomenon called
a) Reflection
b) Multipath
c) Refraction
d) Diffraction

4. Which of the following technology does not use MIMO?
a) 4G
b) Wifi
c) WiMax
d) AMPS

5. Which technique of transmit diversity used in UMTSS third-generation cellular systems?
a) STTD
b) SM
c) Collaborative Uplink MIMO
d) MU-MIMO

6. Which of the following is a transmission method used in MIMO wireless communications to transmit encoded data signals independently.
a) STTD
b) SM
c) Collaborative Uplink MIMO
d) MU-MIMO

7.Which is an additional open-loop MIMO technique considered by the WiMAX vendors?
a) STTD
b) SM
c) Collaborative Uplink MIMO
d) MU-MIMO

8. Which of the following does not come under the teleservices of GSM?
a) Standard mobile telephony
b) Mobile originated traffic
c) Base originated traffic
d) Packet switched traffic

9. Which of the following comes under supplementary ISDN services?
a) Emergency calling
b) Packet switched protocols
c) Call diversion
d) Standard mobile telephony

10. Which of the following memory device stores information such as subscriber's identification number in GSM?
a) Register
b) Flip flop
c) SIM
d) SMS

11.Which supports the operation and maintenance of GSM?
a) BSS
b) NSS
c) OSS
d) MSC

12. Which of the following does not come under subsystem of GSM architecture?
a) BSS
b) NSS
c) OSS
d) Channel

13. Which of the following feature makes impossible to eavesdrop on GSM radio transmission?
a) SIM
b) On the air privacy
c) SMS
d) Packet switched traffic

14. Which of the following is used by IS-95?
a) DSSS
b) FHSS
c) THSS
d) Hybrid

15. Each IS-95 channel occupies ____________ of spectrum on each one way link.
a) 1.25 MHz
b) 1.25 kHz
c) 200 kHz
d) 125 kHz

16. IS-95 is specified for reverse link operation in __________ band.
a) 869-894 MHz
b) 849-894 MHz
c) 849-869 MHz
d) 824-849 MHz

17. User data in IS-95 is spread to a channel chip rate of
a) 1.2288 Mchip/s
b) 9.6 Mchip/s
c) 12.288 Mchip/s
d) 0.96 Mchip/s

18. UMTS use which multiple access technique?
a) CDMA
b) TDMA
c) FDMA
d) SDMA

19. UMTS does not has backward compatibility with
a) GSM
b) IS-136
c) IS-95
d) GPRS

20. UMTS is also known as
a) IS-95
b) GPRS
c) CdmaOne
d) W-CDMA

21. What is the chip rate of W-CDMA?
a) 1.2288 Mcps
b) 3.84 Mcps
c) 270.833 Ksps
d) 100 Mcps

22. How much packet data rate per user is supported by W-CDMA if the user is stationary?
a) 2.048 Kbps
b) 100 Mbps
c) 2.048 Mbps
d) 1 Gbps

23. How much increase in spectral efficiency is provided by W-CDMA in comparison to GSM?
a) Two times
b) Three times
c) No increase
d) Six times

24. What is the minimum spectrum allocation required by W-CDMA?
a) 5 MHz
b) 20MHz
c) 1.25 MHz
d) 200 KHz

25. Which of the following has no backward compatibility with 3G Cdma2000?
a) IS-95
b) GPRS
c) IS-95A
d) IS-95B

26. Which of the following the first 3G CDMA air interface?
a) IS-95
b) IS-95B
c) Cdma2000 1xRTT
d) CdmaOne

27. How many users are supported by Cdma2000 1X in comparison to 2G CDMA standard?
a) Half
b) Twice
c) Six times
d) Ten times

28. Which of the following is not a characteristic of Cdma2000?
a) Adaptable baseband signalling rates
b) Adaptable baseband chipping rates
c) Multicarrier technologies
d) OFDMA

29. Cdma2000 1xEV was developed by **________**
a) Motorola
b) AT&T Laboratories
c) Qualcomm
d) NTT

30. How is bandwidth increased in Cdma2000?
a) Clubbing adjacent radio channels
b) Changing the hardware of base stations
c) Change of spectrum
d) Change of RF equipment

31. EDGE is sometimes also referred as
a) HSCSD
b) 3GPP
c) EGPRS
d) EGSCSD

32. What is one disadvantage of EDGE in comparison to HSCSD and GPRS?
a) Low data rates
b) Small coverage range
c) Low speed
d) No advancement

33. Which new modulation technique is used by EDGE?
a) BPSK
b) 8- PSK
c) DQPSK
d) AFSK

34. What changes GPRS need to acquire while upgrading itself from GSM?
a) A whole new base station
b) New transceiver at base station
c) New channel cards
d) New packet overlay including routers and gateways

35. GPRS and EDGE supports which 2G standard?
a) GSM only
b) IS-136 only
c) GSM and IS-136 both
d) PDC

36. Which of the following is a CDMA standard of second generation network?
a) IS-95
b) IS-136
c) ETACS
d) EDGE

37. Which is the protocol used for routing user data and control signaling within the GPRS backbone network?
a) Network Protocol(NP)
b) Internet Protocol(IP)
c) User Datagram Protocol(UDP)
d) GPRS Tunneling Protocol(GTP)

38. Which of the following is/are the limitations of GPRS?
i. Limited cell capacity for all users
ii. Speeds much lower in reality
iii. Transit delays
a) i and ii only
b) ii and iii only
c) i and iii only
d) All i, ii and iii

39. Which of the following needs enhancements to register GPRS user profiles.
a) Base Station Subsystem (BSS)
b) Base Transceiver Station (BTS)
c) Home Location Register (HLR)
d) Mobile Station (MS)

40. Which of the following is/are the applications of GPRS.
i. Communication
ii. E-Commerce
iii. Vertical applications
iv. Advertising
a) i, ii and iii only
b) ii, iii and iv only
c) i, iii and iv only
d) All i, ii, iii and iv

41. Which is the organization providing standards for GPRS network?
a) ANSI
b) ETSI
c) 3GPP
d) UMTS

42. A GPRS Network works same in
a) 2G
b) 3G
c) 2G and /or 3G
d) 4G

43. BLAST stands for
a) Bell Labs Layered Spectrum Time
b) Bell Labs Layered Space Time
c) Britain Labs Layered Space Time
d) Britain Labs Layered Spectrum Time

CONCLUSION

In this unit, MIMO and space time signal processing techniques used described in detail. The spatial multiplexing and its different architectures used for MIMO channels were discussed. The diversity/multiplexing tradeoff in MIMO channels were elaborated. The System examples like GSM, EDGE, GPRS, IS-95, CDMA 2000 and WCDMA were discussed.

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ASSIGNMENT

1. What is MIMO? What is the need for using it?
2. What is Spatial Multiplexing? Explain the different BLAST architectures.
3. Describe the GSM architecture in detail.
4. Describe about the forward and reverse channel characteristics of IS-95.
5. Explain, How CDMA2000 is superior to IS-95?
6. Explain the applications of GPRS.