INDOOR CDMA CAPACITY USING SMART ANTENNA BASE STATION

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Indoor CDMA Capacity Using Smart Antenna Base station

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Abstract

The performance of smart antenna in an indoor environment, with the emphasis on improving the system capacity of direct sequence code division multiple access (DS-CDMA) base station is studied. A model for the uplink CDMA is presented which assumes that the base station alone uses an antenna array to transmit and receive signals. In the channel model we assume that the signal bandwidth is much larger than the coherence bandwidth to assure the existence of resolvable paths. The base station uses Differential Phase Shift Keying (DSPK) for modulation and a RAKE receiver.

Evaluation of different performance measures such as bit error probability and outage probability is performed. A significant improvement in capacity for the 8 sensors smart antenna over the conventional one antenna element is achieved.

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Table of Contents

List of Figures

List of Tables

Notations and Abbreviations

AC	Access channel
AMPS	Advanced mobile phone system
BER	Bit error ratio
CDF	Cumulative distribution function
CDMA	Code division multiple access
CM	Constant modulus
DECT	Digital European cordless telephone
DF	Direction finding
DPSK	Differential phase shift keying
DS-SS	Direct sequence spread spectrum
FAF	Floor attenuation factor
FDD	Frequency division duplexing
FDMA	Frequency division multiple access
FEC	Forward error correction
FF	Fast fading
F-FH	Fast Frequency-hopping
FH	Frequency-hopping
FM	Frequency modulation

GSK	Gaussian shift keying
GSM	Global system for mobile communications
IS-54	Interim standard 54
IS-95	Interim standard 95
ISI	Inter-symbol interference
JDC	Japanese digital cellular
LOS	Line of sight
OBS	Obstructed
O-QPSK	Offset-quadrant phase shift keying
PDF	Probability distribution function
PN	pseudo noise
PSK	Phase shift keying
QPSK	Quadrant phase shift keying
RF	Radio frequency
RMS	Root mean square
RTC	Reverse traffic channel
SDMA	Space division multiple access
SF	Slow fading
S-FH	Slow Frequency-hopping
SINR	Signal to interference-plus-noise ratio
SIR	Signal to interference ratio
SS	Spread spectrum
TC2	Cordless telephone
TDD	Time division duplexing

TDMA	Time division multiple access
TH	Time-hopping
TIA	Telecommunication industry association
USDC	United states digital cellular

1 INTRODUCTION

1.1 Smart Antennas

In wireless communication, especially in mobile phone systems, two major aspects are always considered, the system capacity and the quality of the communication services. Recent trends in mobile communication have shown that intelligent smart antenna system at the base stations would enhance the communication capacity and the quality of the communication services. Therefore, smart antenna systems have been introduced for base stations to locate and track mobile users and provide them with an improved quality of services.

Smart antennas are composed of phase array antennas and beam formers which combine the signals from the antenna array elements. Adjusting the amplitude and phase with which the individual antenna signals are combined can control a composite antenna pattern. This enables the array to act as a spatial filter, which can enhance or reject signals, based on their direction of arrival. Smart antenna systems, which are applicable to the multi-access communication systems, allow customize beams to be generated for each mobile or group of mobiles. This allows channel reuse with the same cellular domain.

The main drawbacks to high performance wireless communications are interference from other users (co-channel interference), which limits the system capacity (number of users, which can be served by the system), the intersymbol interference (ISI) and the signal fading (time varying amplitude) caused by the multipath signals.

Since the desired signal and the co-channel interference arrive at the receiver from different directions, smart antenna can use these differences to reduce the co-cannel interference and hence increase the system capacity. In addition, the reflected multipath components of the transmitted signal also arrive at the receiver from different directions. Spatial processing can use these differences to attenuate the multipath, thereby reducing ISI and fading, which will lead to higher data rate and better bit error rate BER performance, (Simon et. al, 1998).

1.2 Channel Models

The use of smart antennas in small, lightweight and low-power hand-held devices is unlikely in the next generation systems. However, the base station for these applications can use antenna array with space–time processing at the transmitter to reduce the co-channel interference and multipath, providing similar performance advantage as smart antenna in the receiver. The design of smart antennas and the analysis of their performance requires a new class of channel models, which incorporate both spatial and temporal characteristics. These models can be divided into the following groups:

1.2.1 General statistically based models

These models contain Lee's model, Discrete Uniform Distribution model, Geometrically Based Signal Bounce Statistical model and Gaussian Angle of Arrival model. These models are useful for general system performance analysis (Ertel and Cardieri, 1998).

1.2.2 More site-specific models

These are Extended Tap Delay Line model and Measurement-based Channel model. They can be expected to yield greater accuracy but require measurement data as an input (Ertel and Cardieri, 1998).

1.2.3 Entirely site-specific models

An example from this group is Ray Tracing model, which has the potential to be extremely accurate but require a comprehensive description of the physical propagation environment as well as measurement to validate the models, (Ertel and Cardieri, 1998).
The objective of modeling is to substantially reduce the amount of physical measurement required in the system planning process.

1.3 Smart Antennas in Mobile Communications

Cellular systems use 120° sectorization at each base station. Each base station uses three separate sets of antenna for each 120° sector, with dual receiver diversity in each sector. But each sector uses different frequency to reduce co-channel interference, therefore handoffs between sector are required. For higher performance, narrower sectors could be used, but this will result in too many handoffs. The introduction of multibeam or adaptive array antenna (diversity antenna) without handoffs between beams (smart antennas) overcomes this problem (Ertel and Cardieri, 1998).

This phenomenon leads to the use of antenna arrays in many applications like IS-136 digital TDMA systems, GSM system and IS-95 Digital CDMA systems to enhance the range and capacity increase.

The IS-95 CDMA system (Feher, 1995) has multiple simultaneous users in each 1.25MHz channel with 8 kb/s per user and a spreading gain of 128. A RAKE receiver, which combines delayed version of the CDMA signals overcomes the delayed spread problem and provides a diversity gain. The CDMA spreading codes can provide the reference signal for adaptive array weight calculation. The advantage and disadvantage of the adaptive array antennas and the multibeam antennas on range, capacity and data rate for the CDMA systems can be summarized as follows:

The RAKE receiver generally provides three-fold diversity, and different beams can be used for each figure of the RAKE receiver. The net effect is that the additional diversity gain of the adaptive array is much smaller, and the antenna gain limitation is much less (Winters, 1998). Therefore an adaptive array provides only a slightly larger range increase. Whereas multibeam antennas are more preferable for CDMA, because they require less complexity with respect to weight/beam tracking and to the downlink traffic.

CDMA capacity depends on the spreading gain and the corresponding number of equal-power co-channel interference. A multibeam antenna with M beams reduces the number of interference per beam by a factor M, and thereby increases the capacity by M-folds. As for an adaptive array, it can provide only limited additional interference suppression, since the number of intereferers is much greater than the number of antennas. Because multibeam antennas are less complex than adaptive arrays, particularly since beams need to be switched at most every few seconds versus tracking 179 Hz fading signals in adaptive arrays, multibeam antennas are generally preferred in CDMA systems.

In general an adaptive array antenna can separate signals from closely spaced antennas. This enables multiple spatial channels to be used to greatly increase the data rate between a mobile and a base station.

1.4 Multiple Access Systems

Multiple access schemes are used to allow many mobile users to share simultaneously a finite amount of radio spectrum and to achieve high capacity services by simultaneously allocating the available amount of channels to multiple users.

In wireless communications systems, it is often desirable to allow the subscriber to send simultaneous information to the base station while receiving information from the base station.

There are three major access techniques used to share the available bandwidth in wireless communication systems. These techniques can be categorized into two basic groups, narrowband channelize systems and wideband systems, depending upon how the available bandwidth is allocated to the users.

1.4.1 Narrowband Systems

In narrowband systems the available radio spectrum is divided into a large number of narrowband channels. Two multi-access techniques are categorized under this system. In Frequency Division Multiple Access (FDMA), a user is assigned a particular channel, which is not shared by other users in the vicinity. Time Division Multiple Access (TDMA) on the other hand allows users to share the same channel but allocates a unique time slot to each user in a cyclical fashion on the channel. Therefore, a small number of users are separated in time on a single channel. Figure 1.1 shows these systems.

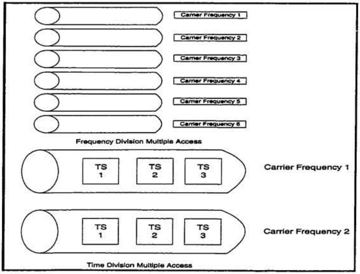

Figure 1. 1 Narrowband multiple access

1.4.2 Wideband Systems

In wideband systems, the entire system bandwidth is made available to each user and it is much larger than the bandwidth required for transmitting information. Such systems are referred to as Spread Spectrum (SS) systems. These systems can use codes or time slots to select users as shown in Figure 1. 2.

TDMA allocates time slots to the many users on the channel and allows only one user to access the channel at any instant of time.

Systems that use codes as a basis of selection are called Code Division Multiple Access (CDMA). It allows all users to access the channel at the same time. This system has been adopted as a standard called IS-95 CDMA. It promises improved capacity of either the analog AMPS system or the digital TDMA system. The major attributes of IS-95 CDMA systems are:

System Capacity. The capacity of CDMA system is higher than that of the existing analog system due to improved code gain/modulation density, voice activity, three sector sectorization and reuse of the same spectrum in every cell, (Kohno, 1998).

Economies. CDMA system is a cost-effective technology that requires fewer, less-expensive cells and no costly frequency reuse pattern.

Quality of Service. It can be improved by providing robust operation in fading environments and soft handoffs. CDMA takes the advantages of multipath fading to enhance communication and voice quality. By using a RAKE receiver and other improved signal-processing techniques, each mobile station selects the three strongest multipath signals and coherently combines them to produce an enhanced signal. Thus, the fading multipath nature is used to an advantage in CDMA.

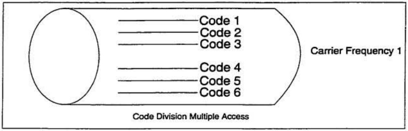

Figure 1. 2 Wideband Multiple Access

1.5 Duplexing

The effect of sending and receiving information simultaneously is called duplexing. It is generally required in wireless systems. Duplexing may be performed using frequency or time domain techniques.

Frequency Division Duplexing (FDD) provides two distinct bands of frequency for each user. The forward band provides traffic from the base station to the user while, the reverse band provides traffic from the user to the base station. A device called duplexer is used to allow simultaneous transmission and reception on the duplex channel.

Time Division Duplexing (TDD) uses time instead of frequency to provide both forward and reverse links.

Table 1.1 shows the use of multiple access techniques in wireless communication systems.

Table 1. 1 Multi Access Technical Applications

Wireless System	Multiple Access Technique
Advanced Mobile Phone System (AMPS)	FDMA/FDD
Global System for Mobile (GSM)	TDMA/FDD
U. S. Digital Cellular (USDC)	TDMA/FDD
Japanese Digital Cellular (JDC)	TDMA/FDD
Cordless Telephone (TC2)	FDMA/TDD
Digital European Cordless Telephone (DECT)	FDMA/TDD
U. S. Narrowband Spread Spectrum (IS-95)	CDMA/FDD

1.6 Thesis Overview

The application of the CDMA systems can be classified into three major categories, Indoor Wireless Communications, Outdoor Cellular CDMA Systems and Mobile Satellite CDMA Systems. Indoor wireless communication has significant advantages over the conventional cabling. Therefore, DS-SS modulation for indoor wireless multiple access communications over multipath fading channels is widely employed. DS-SS provides both multiple access and resistance to multipath fading. DS-SS will be discussed in chapter three.

The general propagation characteristics of an indoor environment can be assumed as the channel, which is characterized as a frequency selective, slow fading channel. The most important effects are path loss due to distance between transmitter and receiver and multipath propagation due to reflection and absorption of radio waves. The multipath effect can be modeled by describing the channel as a filter with a discrete impulse response. This response consists of a number of resolvable paths, each having an independent gain, phase and delay described by random variables, these parameters will be coverd in chapter two and four. In chapter five we will study the performance of the system configuration from the bit error probability, outage probability and diversity technique point of view.

The CDMA protocols achieve their multiple access properties by assigning each user a different code. This code is used to transform a user's signal in wide bandwidth signals. If a receiver receives multiple wide bandwidth signals, it will use the code assigned to a particular user to transform the wideband signals received from that user back to the original signal.

During this process the desired signal power is compressed into the original signal to appear as noise when compared to the desired signal (Garg et. al, 1997).

If the number of the interfering users is not too large, the signal-to-noise ratio will be large enough to extract the desired signal without error. In this case the protocol behaves as a contentionless protocol. However, if the number of the users exceeds a certain limit, the interference becomes too large for the desired signal to be extracted and contention occurs. Therefore the protocol is basically contentionless unless too many users access the cannel at the same time.

The concept of division in the CDMA protocols will be discussed in more detail in chapter three.

1.6.1 Motivation

The capacity of a set of portable stations sharing a single indoor radio channel was conducted in a previous work by (Shad, et, al, 1997). The stations communicate with a base station, which is equipped with a smart antenna operating in multibeam space division multiple access and time division multiple access, (SDMA/TDMA) mode. Both theoretical and measured data showed that dynamic slot allocation is capable of increasing the static capacity of SDMA/TDMA system. (Winters, 1987) showed that optimum combining can increase the capacity of the system. Moreover, (Prasad and Misser, 1995) used the diversity techniques to improve the system performance from the bit error point view. The latest trends indicate that CDMA systems for wireless personal communications operating at wideband RF shows a tremendous increase in usage in the coming years. The topic of system performance will be a big issue in the design of this system. Moreover, introducing smart antennas gives the challenge to use them in improving the performance of CDMA systems. In this thesis, both the CDMA system and the smart antenna are studied to achieve an improvement in the system capacity.

1.6.2 Thesis objective

The objective of this thesis is to investigate the impact of using smart antennas to increase the capacity of indoor CDMA wireless communication systems.

1.6.3 Thesis outline

This thesis examines the capacity of an indoor CDMA system, using a smart antenna base station. A number of portable stations sharing a single indoor radio channel are considered. The base station is equipped with a smart antenna operating in multibeam CDMA mode. First the concept of smart antenna in wireless personal communications will be reviewed. The fundamentals of the technology and statistically based models for the spatial channels for antenna array systems will be outlined. The use of these models in the analysis and simulation of antenna array systems will be presented. Finally, the performance improvement of the antenna array including diversity, interference suppression and capacity increase will be investigated as follows:

The space-time models for CDMA system is reviewed, which include array vector estimation technique and techniques for combined equalization and multi-user detection. The application of these modems to improve the performance of CDMA cellular systems is outlined. Finally the techniques and the fundamentals of the spread spectrum system are discussed, followed by the concept of division in the CDMA protocols.

The CDMA standards and the design concepts for mobile communications and techniques used to increase the capacity of the CDMA are discussed. This includes the performance of the system configuration from the bit error probability and outage probability.

A wideband RF channel system is evaluated. After constructing the system models, the propagation characteristic in the indoor environment is investigated. Finally the capacity of the system is determined.

Simulation results using MATLAB and SIMULINK models of the systems are presented

2 PROPAGATION CHARACTERISTICS

2.1 Fundamentals of radio propagation

Knowing the propagation characteristics of a channel is an important aspect in radio communications. The transmission path between the transmitter and the receiver can vary from direct line of sight (LOS) to one that is severely obstructed by buildings and foliages. To establish a good communication between a mobile and base station, the communication link must be well defined. In general, there are three basic propagation mechanisms which impact propagation in a mobile propagation system: Reflection, Diffraction, and Scattering. The first occurs when propagating electromagnetic wave impinges upon an object, which has very large dimension when compared to the wavelength of the propagating wave (Rappaport, 1996). The second occurs when the radio path between the transmitter and the receiver is obstructed by a surface that has sharp edges. The resulting diffraction depends on the geometry of the object as well as the amplitude, phase and polarization of the incident wave at the point of diffraction (Rappaport, 1996). The third occurs when the medium through which the wave travels consists of objects with dimensions that are small compare to the wavelength. Scattered waves are produced by rough surfaces, small objects or any other irregularities. In an indoor environment there are two aspects associated with building losses. First is the loss through the outer structure, known as penetration loss. The second refers to the losses between locations within the building resulting from the layout of the building, the nature of material used in the construction of the outside walls, floors, the partitions, and the building type. All are known as the building losses. Many measurements have been done to obtain information concerning propagation loss (Rappaport, 1996).

2.2 Penetration Loss

The best definition to building penetration loss as given by (Doble, 1996), is " Building penetration loss is the difference between the mean signal level measured right around the outside of the building at ground level and the mean level over the floor of interest. (The research results obtained were subjected to a variety of factors such as the amount of glass in the walls, the different ways in which the floors are divided, and the material used). In an experiment conducted in the Electrical Engineering Department at Liverpool University, (Doble, 1996), the penetration loss measured on the ground floor was 12.1 dB and on the first floor 1.6 dB. The difference is due to a large area of glass used in the first floor. Moreover, RF penetration was found to be a function of frequency as well as height within a building. Penetration loss decreases with the increase of frequency. The experimental studies also showed that penetration loss decreased at a rate of about 2 dB per floor from ground level up to the 10^{th} floor and then started to increase due to shadowing effects of adjacent buildings.

2.3 Indoor Losses

Indoor radio propagation is dominated by the same mechanisms as outdoor propagation. However, conditions are much more variable due to people movement and whether interior doors and widows are open or closed. The field of indoor radio propagation is relatively new with the first wave of research occurring in the early 1980's. The indoor channels are subjected to a variety of characteristics such as path loss, multipath propagation and channel fading and shadowing. Multipath fading is described by its envelope fading (nonfrequency-selective amplitude distribution), Doppler spread (time-selective or time variable random phase noise), and a time-delay spread (variable propagation distance of reflected signals causing time variation in the reflected signals). In general indoor channels can be classified either as line-of-sight (LOS) or obstructed (OBS), with simple or heavy clutter.

2.3.1 Path loss

Path loss in radio propagation is a measure of interest. It is defined as the ratio between the received power P_r and the transmitted power P_t.

$$L = \frac{P_r}{P_t} \tag{2.1}$$

The average power at a distance d from the transmitter is a decreasing function of distance d, which is represented by a path loss power law

$$P_0 \approx d^{\gamma} \tag{2.2}$$

Measurements have been done to obtain information about the value of the path loss law exponent γ in indoor environment. In free space this exponent is equal to two, therefore the power law follows an inverse square law (Prasad 1996).

The value of γ is reported in different research papers. In Bultitude (1987) and Saleh and Valenzuela (1987), the location of the transmitter and receiver is considered. Table. 2.1 lists these values

Table 2. 1 Measured values for path loss exponent

Location of transmitter	Location of receiver	Frequency	Value of γ
Hallway	Hallway	910 MHz	1.8
Hallway	Rooms off the hallway	,,	3.0
Hallway	Rooms off and perpendicular to the transmitter	,,	4.0 to 6.0

Other factors such as partition material, type of building, LOS path, OBS path, and light and heavy clutters are also considered and reported in (Bultitude, 1987), (Paroakis, 1995) and (Prasad, 1996). Table 2.2 shows these measurements.

It is reported that the upnormal condition takes place when the receiver is located in rooms off the hallway and perpendicular to the transmitter hallway.

The average power received by the base station from any mobile terminal is a function of the distance. If the terminal is close to the base station then the average power is much higher than that which is located in a larger distance. This is called the near-far zone effect and it is a result of the path-loss law.

Table 2. 2 Measured values for path loss exponenet building type

Building type	Frequency GHz	Value Of γ				
		LOS		OBS		
		Light clutter	Heavy clutter	Light clutter	Heavy clutter	Upnormal condition
Metalized partition						6.0
Factories		1.79	1.79	2.38	2.81	
	2.4	1.2 to 2		3.3		
	4.75	1.2 to 2		3.8		
	11.5	1.2 to 2		4.5		

Propagation between floors is another important aspect in dealing with wireless system of multifloor buildings that need to share frequencies within the building. Frequencies are reused on different floors to avoid the co-channel interference. Measurements have shown that loss between floors does not increase linearly in dB. The greatest floor attenuation in dB occurs when the receiver and transmitter are separated by one floor. Typical values of attenuation between floors are 15 dB for a single floor separation and an additional 6-10 dB per floor separation up to four floors. For more than four floor the path-loss will increase by only few dB for each additional floor. Moreover the signal strength or average power received inside a building increases with height of the building.

There are different models to calculate the path-loss. Since the mean path loss is a function of distance and the γ th power, then the path-loss in dB can be modeled as, (Garg, et. al, 1997):

$$L(R)=L(R_0)+10\times\gamma\log\left(\frac{R}{R_0}\right)dB \tag{2.3}$$

where, L(R) is the mean path-loss in dB, $L(R_0)$ is a path-loss in dB from the transmitter to reference distance R_0, γ is the path-loss exponent, R is the distance from the transmitter(m) and R_0 is a reference distance from the transmitter (m). L (R_0) is found to be 31.7 dB at 914 MHz. Equation 2.3 is modified to emphasize the mean path-loss exponent as a function of the number of floors between the receiver and transmitter.

$$L(R)=L(R_0)+10\times\gamma(multifloor)\log\left(\frac{R}{R_0}\right)dB \tag{2.4}$$

The floor attenuation factor (FAF) also can be used as another path-loss prediction model. A constant FAF in dB which is a function of the number of floors and building type is added to equation 2.3 to model the path-loss in the same floor.

$$L(R)=L(R_0)+10\times\gamma\ \log(\frac{R}{R_0})+FAF\ dB \tag{2.5}$$

2.3.2 Multipath

Multipath is the phenomenon of signal reflections, which arrive at the receiver antenna at different times. These reflections are due to building structure and surrounding inventory. In other words, in most application, no complete direct LOS propagation exists between the base station antenna and the mobile antennas because of natural and constructed obstacles. In such cases the radio link may be modeled as a randomly varying propagation path. But in general, there may exist more than one propagation path, which in this case is referred to as multipath propagation. Figure 2. 1shows these propagations.

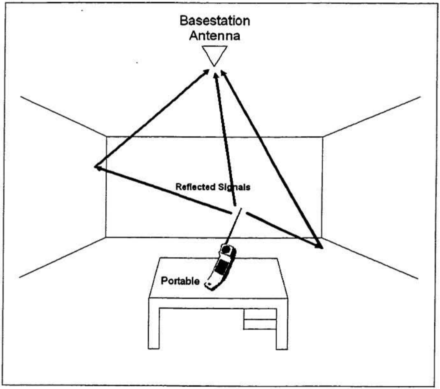

Figure 2. 1 Multipath propagation

The reflected signals add up together, which cause signal peaks or dips, depending on the phases of the reflected signals.

Indoor multipath channel can be represented by certain factors. The baseband complex response of the λ^{th} path is given by, (Prasad, 1996):

$$h_l(t)=\beta_l \exp\{j\gamma_l\}\delta_l(t-\tau_l) \tag{2.6}$$

where β_λ is a positive gain, τ_λ is a propagation delay, γ_λ is a phase shift and λ indicates the desired path. These parameters are randomly changing functions of time because of people and other environmental factors. Therefore, these parameters cause a variation in the received signal strength. Another important distortion is also considered, which is due to frequency selectivity of the channel. Each of the two distortions is related to a different aspect. The former depends on the coherence bandwidth of the channel (multipath spread). The later depends on the time variation of the channel characterized by the coherence time (Doppler spread).

2.3.2.1 Delay spread

Good information about the multipath characteristics of the channel can be obtained from the impulse response of the paths. This impulse response can be considered as a power delay density function, which can be characterized by using the RMS multipath delay spread T_m. Delay spread is the standard deviation of the power delay profile. It can be calculated by using the following equation, (Prasad, 1996):

$$T_m \triangleq \sqrt{E[\tau^2]-E^2[\tau]} \tag{2.7a}$$

with

$$E[\tau] \underline{\Delta} \frac{\sum_{\lambda} \tau_{\lambda} \beta_{\lambda}^{2}}{\sum_{\lambda} \beta_{\lambda}^{2}} \tag{2.7b}$$

and

$$E[\tau^{2}] \underline{\Delta} \frac{\sum_{\lambda} \tau_{\lambda}^{2} \beta_{\lambda}^{2}}{\sum_{\lambda} \beta_{\lambda}^{2}} \tag{2.7c}$$

where β_{λ} is the received power of the λ^{th} pulse and τ_{λ} is rms time delay spread. Typical values for τ_{λ} in an indoor environment are of the order of nanoseconds. Some of these values which are measured in different indoor locations and reported in the literature, (Prasad, 1996) are listed in Table 2.3

Table 2. 3 Time delay spread

Frequency range (MHz)	Delay spread range (ns)
910	50 to 250
1500	10 to 50
1900	70 to 94
2400 to11500	10 to 20
850 to 4000	270 to 300

The coherence bandwidth of the channel is the bandwidth over which the signal propagation characteristics are correlated. If the coherence bandwidth is smaller than the transmitted signal bandwidth then the channel is frequency selective and the different frequency components in the signal are subjected to different gains and phase shifts. On the other hand if the coherence bandwidth is larger than the bandwidth of the transmitted signal, the channel is frequency nonselective and all frequency components are subjected to the same gain and phase shift. In general the bandwidth of a spread spectrum system is larger than the coherence bandwidth, which implies that the channel is frequency selective. The coherence bandwidth is expressed as the reciprocal of the rms time delay spread. In other words the rms delay spread and the coherence bandwidth are inversely proportional to one another. The characteristic of the channel is frequency selective and leads to an existence of the so-called resolvable paths. The delayed version of the initial signal can be grouped in clusters (paths), which can be resolved independently. The maximum number of the resolvable paths L can be determined using the multipath delay spread and the chip duration. For the signals to be resolved, they should be separated by one chip time T_c. Therefore the maximum paths can be estimated as

$$L = \left\lfloor \frac{T_{max}}{T_c} \right\rfloor + 1 \tag{2.8}$$

Choosing the largest integer that is less than or equal to T_{max} / T_c will determine L. Equation 2.8 can be used only if the number of resolvable paths is fixed, but in practice this number is a random variable.

2.3.2.2 Doppler spread

Doppler spread is the range of values of frequency over which the Doppler power spectrum is essentially nonzero. It is related to the coherence time of the channel which, is the duration over which the channel characteristics do not change significantly. Therefore the coherence time of the channel can be calculated as the reciprocal of the Doppler spread. Since the coherence time is the measure of the width of the time correlation function, a slow changing channel has a large coherence time or a small Doppler spread.

In an indoor environment measurements, in any fixed location, temporal variations in the received signal envelope caused by movement of personnel and machinery are considered to be slow, and have a maximum Doppler spread of 6.1 Hz.

2.3.3 Fading characteristics

The rapid fluctuation of the amplitude of a radio signal over a short period of time or travel distance is called fading. It is caused by interference between two or more versions of the transmitted signals, which arrive at the receiver at slightly different times, (multipaths). These signals combine at the receiver to give a resultant signal which can vary widely in amplitude and phase. The fading depends on the nature of the transmitted signal with respect to the characteristics of the channel. The relation between the signal parameters and the channel parameters determine the type of fading that the signal will undergo. Small Scale fading based on Multipath Time Delay Spread has two types of fading, Flat Fading and Frequency Selective Fading. While Small-Scale fading based on Doppler Spread has another two types of fading, Fast Fading and Slow Fading.

The channel may be classified to be either fast fading or slow fading depending on ratio of the rate of change of the transmitted signal, to the rate of change of the channel. Fast fading occurs when the channel impulse response changes rapidly within the symbol duration, that is the coherence time of the channel is smaller than the symbol period of the transmitted signal. This characteristic causes the signal to be distorted. Fast fading occurs only for very slow data rate.

In slow fading the channel impulse response changes at a rate much slower than the transmitted baseband signal. Therefore the channel may be assumed to be static over one or several reciprocal bandwidth intervals. In the frequency domain, this implies that the Doppler spread of the channel is much less than the bandwidth of the baseband signal. We may notice that the velocity of the mobile or the objects in the channel determine whether a signal undergoes fast fading or slow fading.

2.3.4 Fading distribution

Two types of distributions are always considered when talking about the fading statistics of the channel, Rayleigh distribution and Ricean distribution. Depending on the nature of the reflected signals arriving at the receiver antenna, the channel distribution is determined.

2.3.4.1 Rayleigh distribution

The Rayleigh distribution is commonly used to describe the statistical time varying nature of the received envelope of a flat fading signal, or the received envelope of an individual multipath component. The Rayleigh distribution has a probability density function (pdf) given by

$$P(r)=\frac{r}{\sigma^2}\exp\left(-\frac{r^2}{2\sigma^2}\right) \quad (0\le r\le\infty) \tag{2.9}$$

where σ is the rms value of the received voltage signal, σ^2 is the time average power of the received signal and r is the distance. The probability that the envelope of the received signal does not exceed a specified threshold R, is given by the following cumulative distribution function (cdf)

$$P(r)=\Pr(r\le R)=\int_0^R P(r)dr=1-\exp\left(-\frac{R^2}{2\sigma^2}\right) \tag{2.10}$$

2.3.4.2 Ricean distribution

The existence of a dominant stationary signal component, such as a Line-of-Sight (LOS) propagation path makes the fading statistics of the signal envelope to be considered as a Ricean distribution. The random multipath components that arrive at different angles are superimposed on a stationary dominant signal. Therefore two distributions related to the Ricean distribution can be distinct. First the spatial signal strength distribution; if the environment is static, there is a fixed spatial pattern of signal maxima and minima. The measurements confirmed that since a significant part of the received signal envelope is due to a constant path, the spatial signal strength distribution could be described by a Ricean distribution. Second the signal strength distribution, which is due to the existence of a LOS component at the input of a fixed receiver antenna, is also a Ricean distribution. The Ricean distribution is characterized by a parameter R, which is the ratio of the peak power and the power received over specular paths.

$$R=\frac{s^2}{2\sigma^2} \tag{2.11}$$

According to the recent measurement done in office building (Bultitude, 1987 and Rappaport, 1996), the parameter R was found to be equal to 6.8 dB, which corresponds to a brick building with reinforced concrete and plaster, as well as some ceramic block interior partitions. The parameter R is equal to 11 dB corresponding to a building having the same construction, but with an open-office interior floor plan and nonmetallic ceiling tiles throughout. The Ricean distribution has a probability density function (pdf) given by (Prasad, 1996)

$$P(r) = \frac{r}{\sigma^2}\exp\left(-\frac{r^2+s^2}{2\sigma^2}\right) I_0\left[\frac{sr}{\sigma^2}\right] \quad (0 \le r \le \infty), (s \ge 0) \tag{2.12}$$

where s is the peak value of the specular radio signal due to the superposion of the dominate LOS signal and the time invariant scattered signals reflected from walls, ceiling and stationary inventory. I_0 ( ) is the modified Bessel function of the first kind and zero order. Figure 2. 2 shows a Ricean probability density function for R equal to 6.8 and 11dB.

In the case where there is no dominant component and the received signal consists of reflected versions of the original signal, then the parameter R approaches zero and the Ricean distribution changes into Rayleigh distribution, (Prasad, 1996). This shows that the Rayleigh distribution is similar to the Ricean distribution with direct to specular ratio R=0.

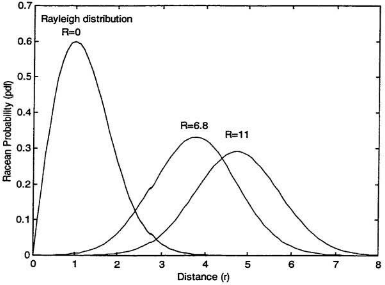

Figure 2. 2 Ricean distribution

3 CODE DIVISION MULTIPLE ACCESS

3.1 CDMA Protocols

Code division multiple access system uses coding to achieve its multiple access property. The property of division is achieved by assigning each subscriber a unique code sequence. This code is used to transform the original signal in a wideband signal before transmission. This phenomenon is called spread-spectrum. On the other hand the receiver uses the code which is assigned to a particular user to transform back the received wideband signal to the original signal. In this processing the desired signal power is compressed into the original signal bandwidth while the other signal bandwidths remain unchanged and appear as interfering signals to the desired user's signal.

The CDMA protocol is considered to be placed between the contentionless and contention protocols. Depending on the interfering signals the behavior of the CDMA system describes its protocol. If the number of the interfering signals is not too large, the signal-to-noise ratio will be large enough to extract the desired signal without error and the system will behave as contentionless protocol. However if the number is too large, it will be too difficult for the desired signal to be extracted and the system will behave as contention protocol.

3.1.1 Direct-Sequence CDMA

Modulation is used as a basis of division in CDMA system to obtain the wideband signal. This leads to different types of protocol which in turn describes the techniques used in generating the spread spectrum signals The most well known types of spread spectrum are: Direct Sequence CDMA (DS-CDMA), Frequency-Hopping CDMA (FH-CDMA), Time-Hopping CDMA (TH-CDMA) and hybrid CDMA (Prasad, 1996). The first two types, are the most common used ones in wireless communications.

The spread spectrum modulation transforms the information data into transmission signal with a much larger bandwidth. The ratio of the transmitted bandwidth to the information bandwidth is called the processing gain of the spread spectrum system α_p.

In DS-CDMA, the information data signal is directly modulated by a binary code sequence with a bandwidth much larger than the original bandwidth to transform it to wideband signal. The resultant signal modulates the carrier, as shown in Figure 3. 1. The modulated signal is then transmitted through the channel.

The modulated carrier then is modulated by the code signal. This code signal consists of a number of code bits (chips) that can be either +1 or –1. To obtain the desired spreading of the signal, the chip rate must be much higher than the data bit rate.

For the modulation, different techniques can be used. The most known techniques are those of the form of phase shift keying (PSK).

At the receiving point the receiver uses coherent demodulation to despread the signal using a locally generated code sequence. The desired modulated signal will be accompanied with other users signals. The receiver correlates the desired signal with the code sequence of the original signal. To be able to perform the despreading operation, the receiver must not only recognize the code sequence used to spread the original sequence, but also the codes of the received signal and the locally generated code must be synchronized. This synchronization must be accomplished at the beginning of the reception and maintained until the whole signal is received.

For the original data to be demodulated and recovered without error, the desired power signal to interfering signals ratio should be large and the cross–correlations of the code sequences of the interfering signals and the code sequence of the desired signal should be small.

DS-CDMA protocol has some properties, which make it more preferable among other protocols. These properties mainly relate to multiple access, multipath, narrowband intrferences and low probability of interception.

If the multiple users use the channel at the same time, then there will be multiple direct sequence signals overlapping in time and frequencies. At the receiver, coherent demodulation is used to remove the code modulation. This operation concentrates the power of the desired user in the information bandwidth. If the cross-correlation between the code of desired user signal and the code of the interfering signal is small, coherent detection will put only small part of the interfering signal into the information bandwidth. This property makes the DS-CDMA more capable for multiple access systems.

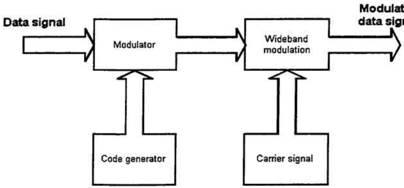

Figure 3. 1 Direct Sequence CDMA

Another property is related to multipath interference rejection. If the code sequence has an ideal auto-correlation function, as it will be mentioned later, then the correlation function is zero outside the interval $[-T_c, T_c]$ where T_c is the chip duration. This means that if the desired signal and a version which is delayed for more than $2T_c$ are received, coherent demodulation will treat the delayed version as an interfering signal putting only a small part of the power in the information bandwidth.

The transmitter involves a multiplication of narrowband signal with a wideband code sequence. The code sequence spreads the spectrum of the narrowband signal so that its power in the information bandwidth decreases by a factor equal to the processing gain.

This property resists the existence of narrowband interference. The DS signal uses the whole signal spectrum all the time; therefore it will have a very low transmitted power per hertz. This makes it very difficult to detect the signal.

In general DC-CDMA protocol has the following specific advantages: first, the generation of the coded signal is easy. It can be done by a simple multiplication. Second, the carrier generator is simple since only one carrier frequency has to be generated. Coherent demodulation of the spread-spectrum is possible. Finally, no synchronization among the users is required.

The main problem associated with the DS-CDMA protocol is the Near-Far effect. The power received from a user close to the base station is much higher than that received from further away. Since users continuously transmit over the whole bandwidth, users close to the base station will constantly create a lot of interference for users far from the base station, making their reception difficult and sometimes impossible.

3.1.2 Frequency hopping CDMA

Frequency hopping is the one of the most important protocols of CDMA systems. In frequency hopping, the carrier frequencies assigned to the individual users are varied in pseudorandom fashion within the limitation of a wideband channel. The digital data (information bits) is broken into an equal sized bursts, which are transmitted, on different carrier frequencies. The instantaneous bandwidth of any transmitted burst is much smaller than the total spread bandwidth. In frequency hopped receiver, a locally generated Pseudo noise (PN) code is used to synchronize the receiver instantaneous frequency with that of the transmitter.

Consider the transmitter/receiver diagram of a frequency hopping system shown in Figure 3. 2. The data signal is baseband modulated on a carrier. In most cases FM modulation is used for analog signals and GSK modulation is used for digital signals. The carrier frequency is converted to the transmission frequency by using fast frequency synthesizers that are controlled by the PN code signal. The binary PN code generation drives the frequency synthesizer to hop to one of the many available frequencies chosen by the PN sequence generator.

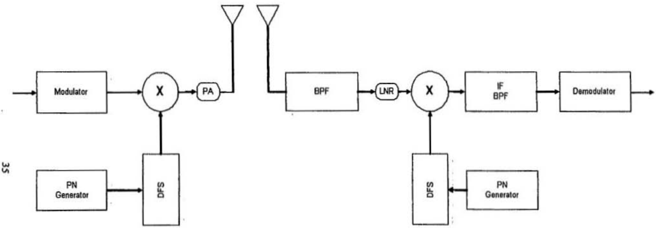

Figure 3. 2 Transmitter and Receiver of FH-CDMA Protocol

At the receiver, a locally generated code sequence is used. The received signal then is converted down to the baseband-modulated carrier. The data is recovered after baseband demodulation. The synthsization / tracking circuit ensures that the hopping of the locally generated carrier will make a possible despreading of the original signal.

There are two types of frequency hopping CDMA protocol. The distinction between these two types can be made according to the rate of change of both the carrier and the information data. Each type has its own properties, which distinguish it from the other. First, if the rate of change of the carrier frequency (hop) is greater than the information data bit rate, then the system is referred to as a fast frequency hopping CDMA (F-FH). In this case the carrier frequency changes a number of times during the transmission of one bit, in other words, one bit is transmitted in different frequencies. This phenomenon can be useful in the multiple access system. If the desired user is the only one to transmit in most of the frequency band, the received power of the desired user signal will be much higher than the interfering signal power and the signal will be received correctly. Another property for the F-FH protocol is the ability to reduce the multipath interference. Any particular signal frequency will be modulated and transmitted on a number of carrier frequencies. The multipath effect is different at different carrier frequencies. Therefore signal frequencies that are amplified at one carrier frequency will be attenuated at another carrier frequency. Then at the receiver the responses at the different hopping frequencies are averaged, thus reducing the multipath interference.If the processing gain G_p is taken into consideration, then the desired signal will use the hopping frequency where the interferer is located $1/G_p$ percent of the time

The interference is therefore reduced by a factor G_p. Another advantage to F-FH is that the synchronization is much easier than with DS-CDMA. The different frequency bands that F-FH can occupy do not have to be contiguous. Therefore the probability of multiple users transmitting in the same time is small.

The second type of frequency hopping CDMA is the slow frequency hopping protocol (S-FH). This type is accomplished if the rate of change of hops is less than the rate of change of the information data bit. In this case multiple bits are transmitted at the same frequency. This phenomenon can be useful to some degree in multiple access systems. If the probability of other users transmitting in the same frequency band is low enough, the desired user will be received correctly most of the time. During transmission the FH signal uses, as much power per hertz as a continuous transmission would, but the frequency at which the signal is going to be transmitted is unknown and the duration of the transmission at a particular frequency is quite small. This gives the FH signal more readily intercept. Moreover, a frequency hopping system provides a level of security, especially when a large number of channels are used, since an intercepting receiver that does not know the pseudorandom sequence of frequency slots retune rapidly to search for the signal it wishes to intercept. The frequency hopping signal is somewhat immune to fading, since error control coding and interleaving can be used to protect the frequency hopped signal against deep fades which may occasionally occur during the hopping sequence. Both error control coding and interleaving can also be combined to guard against erasures which can occur when two or more users transmit on the same channel at the same time, (Rappaport, 1996).

3.2 PN Sequence

To spread the bandwidth of the modulated signal to a larger transmission bandwidth and to distinguish among the different users using the same bandwidth, pseudo noise (PN) sequences are used in both DS-CDMA and FH-CDMA systems. These sequences are deterministic and periodic.

The idea in generating the PN sequences is to use a shift register in the operation. The feedback outputs of the shift registers are combined to form the sequence. The shift register binary sequence is shifted in response to clock pulses. The contents of the stages are logically combined to form the input of the first stage. The contents of the last stage form the code sequence. A feedback shift register and its output are called linear when the feedback logic contents are entirely of module-2 adders.

Consider the shift register shown in Figure 3. 3, the operation of the shift register is controlled by a sequence of clock pulses. At each clock pulse, the contents of each stage in the register is shifted by one stage to the right. In the same time the contents of the stages S and S_1 are module-2 added and the result is fed back to stage S_1. The shift register sequence is defined to be the output of stage S.

It is found that the contents of the register repeats after $2^n - 1$, where n is the number of stages, (Dixon, 1976).

The output sequences of the shift register are classified as either maximal or non-maximal length. Maximal length sequences are the longest sequences that can be generated by a given shift register of a given length L. All other sequences are considered to be non-maximal.

Maximal length sequences are quite useful in spread spectrum techniques. They have several properties, which make them very popular.

The number of binary zeros differs from the number of ones by at most one chip (i.e. Ones are 2^n-1 and Zeros are 2^n-2). Second, the relative positions of the turns vary from code sequence to another but the number of each turn length is the same (turn is defined as a sequence of a single type of binary digits). Most importantly is the existence of good auto-correlation and cross-correlation behavior.

The rejection of the interfering signals from users other than the desired user depends on the ratio of the maximum cross-correlation coefficient to auto-correlation coefficient. The smaller is this ratio, the better is the rejection.

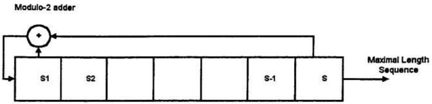

Figure 3. 3 Seven-Stage Shift Register

Auto-correlation function is defined as the degree of correspondence between a sequence and its phase shifted version, and it is given by (Feher, 1995)

$$R_A(\tau)=\int_{-\infty}^{\infty} x(t)\,x(t+\tau)\,dt \qquad (3.1)$$

If x (t) represents the PN code sequence and it is periodic over a period of T_0, then the above equation becomes

$$R_A(\tau)=\frac{1}{R_0(0)}\frac{1}{T_0}\int_{-T0/2}^{T0/2} x(t)\,x(t+\tau)\,dt \tag{3.2}$$

where

$$R_A(0)=\frac{1}{T_0}\int_{-T0/2}^{T0/2} x^2(t)\,dt \tag{3.3}$$

The cross-correlation is the correlation between two different signals, and is given by:

$$R_c(\tau)=\int_{-T_0/2}^{T_0/2} x(t)\,y(t+\tau)\,dt \tag{3.4}$$

One important drawback of the maximal length sequence is the limitation of the number of sequences that can be produced. To overcome this disadvantage, a new type of code sequence has been introduced. This code does not satisfy all the properties of a maximal length code, but it has a satisfactory correlation behavior compared to non-maximum length codes.

Using a module-2 adder to the output of two different maximal length codes a new code, known as the gold code is generated (Dixon, 1976). Figure 3. 4 shows this configuration. The operation can be described by shifting the maximal length sequences with respect to one another, then a new code is produced giving 2^{n+1} sequences plus the original two maximal length codes. In other words, it will give $2^{n+1}+1$ different codes.

The maximum cross-correlation coefficient for the gold code is given by

$$K=\begin{cases}2^{(n+1)/2}+1 & n-\text{odd}\\ 2^{(n+1)/2}-1 & n=\text{even}\end{cases}$$

The auto-correlation coefficient is equal to the length of the sequence L.

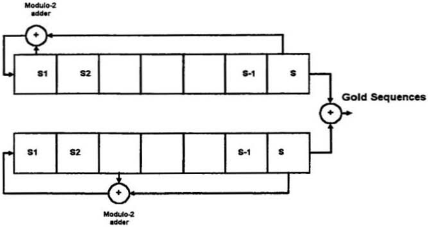

Figure 3. 4 Gold Codes Generation Scheme

3.3 CDMA Standard IS-95

CDMA technology offers several advantages over the traditional FDMA AMPS and the IS-54 TDMA. The USA digital cellular system based on CDMA which promises increased capacity has been standardized as Interim Standard 95 (IS-95) by the USA Telecommunication Industry Association (TIA).

Statistics of telephone conversations suggest that in a typical full-duplex two-way voice conversation, the duty cycle of each voice is less than 35%. Exploring the voice activity in each FDMA and TDMA systems might be hard to implement because of the time delay associated with reassigning channel resources during each speech pauses. In CDMA it is possible to reduce interference to other users. This reduction in interference power can be transferred to either an increase in system capacity or reduction in the average power transmitted by the mobile. Moreover CDMA allows each user within a cell to use the same radio channel, and users in adjacent cells also use the same radio channel, since this is a direct sequence spread spectrum CDMA system. CDMA completely eliminates the need for frequency planning within a market.

In the CDMA IS-95 standard each channel occupies 1.25 MHz of spectrum on each one way link, or 10% of the available cellular spectrum. Unlike the other cellular standards, the user data rate changes in real-time, depending on the voice activity and requirements in the network. Therefore the CDMA IS-95 standard uses time and path diversity to mitigate the effect of frequency selective multipath fading. Time diversity is obtained by use of forward error correction (FEC) and interleaving. Path diversity is inherently provided by the CDMA approach, by spreading the signal over 1.25MHz wide bandwidth. Such signal with wide bandwidth will resolve the multipath component and, thus provide the receiver with several independent fading signal paths. This path diversity is exploited by the use of RAKE receiver to combine different multipath components.

CDMA IS-95 uses different modulation and spread techniques for the forward and reverse links. On the forward link, the base station simultaneously transmits the user data for all mobiles in the cell by using different spreading sequence for each mobile. A pilot code is also transmitted simultaneously at a higher level, thereby allowing all mobiles to use coherent carrier detection while estimating the channel conditions. On the reverse link, all mobiles respond in an asynchronous fashion and have ideally constant signal level due to power control applied to the base station.

In CDMA every user is considered to be as a source of interfering to the other users. Therefore the mobile system transmits a power control acknowledgement in its current implementation. The CDMA standard uses several power control techniques (open loop and close loop control) to optimize the system performance. In addition, the base station uses a three-sector antenna each sector covering 120° of the azimuth, to reduce the multipath access interference and thereby increase system capacity.

3.3.1 CDMA forward link channel

The forward CDMA channel consists of a pilot channel, a synchronization channel, paging channels and forward traffic channels. In the CDMA standard, the forward link uses a combination of frequency division pseudorandom code division and orthogonal signal multiple access techniques. Dividing the available cellular spectrum into nominal 1.25 MHz bandwidth channel employs frequency division.

Data on the forward channel is grouped into 20 msec blocks. The user data is first convolutionally coded and then formatted and interleaved to adjust for the actual user data rate, which may vary. Then the signal is spreaded using Walsh code and a long PN sequence at a rate of 1.2288 Mchips/sec. The forward CDMA channel is shown in Figure 3. 5. The overall processing can be described briefly as follows:

The underlaying data rate for the system is 9600 bits/sec which represents the speech code rate of 8550 bits/sec. The speech encoder actually detects speech activity and changes data rate to lower values (1200 bits) during quite silent periods. The 9600 bits/sec stream is segmented into 20 msec frames and then further convolutionally encoded to provide the capability of error correction and detection at the receiver. The conventional encoder has constraint length K=9 and code rate of ½. This will bring the data rate to 19.2 kbits/sec. Whenever the user data rate is less than 9600 bits/sec, each symbol from the convolution encoder is repeated before block interleaving. This repetition results in a constant coded rate of 19200 bits/sec for all possible information data rate. The convolution encoding is followed by interleaving over 20 msec intervals for burst error protection, due to fast fading in the radio channel.

The 19.2 kbits/sec output of the interleaver is modified by long-codes which serve as data scrambling for security. The decimator keeps only the first chip out of every 64 consecutive PN chips.

The modified stream is encoded for spread spectrum transmission using 64 sequences each of length 64, binary orthogonal Walsh codes, (Feher, 1995) at a fixed chip rate of 1.2288 Mchips/sec. The structure of Walsh code provides 64 orthogonal sequences and one of the 64 sequences is assigned to a mobile unit during call set-up. In this way, six

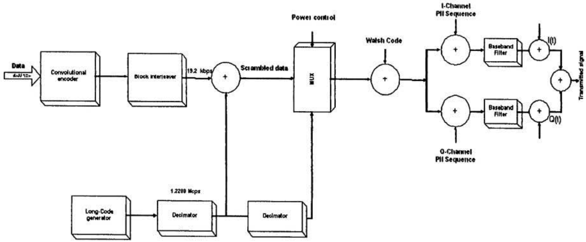

Figure 3. 5 CDMA Forward Traffic Link

four orthogonal channels can be established on the forward link.

Then the spreaded data stream is spread into I and Q streams each of which is modified by a unique short code of length 2^{15}-1 chips. The resulting spread spectrum stream is carried over the air interface with filttered QPSK modulation

All signals transmitted from a base station in a particular CDMA radio channel share the set of 64 Walsh codes and the same pair of short codes. However, signals from different base stations are distinguished by time offset from the basic short code which allows CDMA from each base station to be uniquely identified.

Different signals from a given base station in particular CDMA radio channel are distinguished at the mobile receiver by the orthogonal Walsh codes.

A pilot signal is transmitted in each base station, which used as coherent carrier frequency for demodulation by all mobile receivers. The pilot is transmitted at relatively higher level than other types of signals, which allows for tracking of the carrier phase. The pilot signal is unmodulated by information and uses the zero Walsh function. Thus the pilot signal simply consists of the quadrature pair of short codes.

The synchronization channel broadcasts synchronization messages to the mobile station and operates at 1200 bits/sec. The mobile receiver can obtain synchronization with the nearest base station without prior knowledge of the identity of the base station by searching for the entire length of the short code. The strongest signal time offset corresponds to the time offset of the short code of the base station which the mobile has the best propagation channel (nearest base station). The synchronization channel is assigned channel number 32. The paging channel is used to send control information and paging messages from the base station to the mobile stations and operates at 9600, 4800, 2400 or 1200 bits/sec, and it is assigned the lowest channel number, (Feher, 1995).

3.3.2 CDMA Reverse link channel

The CDMA reverse traffic link employs the same processing concepts as the forward link with slightly different methodology. Figure 3. 6 illustrates the reverse link channel. Two channels are assigned to the reverse traffic link, Access Channel (AC) and Reverse Traffic Channel (RTC). Both share the same frequency assignment. The access channel is used by the mobile to initiate communication with the base station and to respond to paging channel messages. The access channel is a random access channel with each channel user uniquely identified by their long code.

The CDMA reverse link also employs PN spread spectrum modulation using the same short codes as that used for the forward link. However all mobiles use the same mobile phase offset. Signals from different mobiles are distinguished at the base station by use of a very long (2^{42} -1) PN sequence with a user address determined time offset.

Because every possible time offset is a valid address, an extremely large address space is provided. The data rate system is also 9600 bits/sec. The transmitted digital information stream is grouped into 20 msec frames and then further convolutionally encoded using code of rate 1/3 and constraint K=9. This will bring the data rate to 28.8 Kbits/sec. The encoded information bits are then interleaved over the 20 msec frame. The interleaved information are grouped into symbol groups or code words of 6 bits. These code words are used to select one of the 64 different orthogonal Walsh functions for transmission. At the output of the Walsh modulator, the chip rate is 307.2 Kchips/sec. We may note that the Walsh codes are used for different purposes on the forward and reverse links. On the

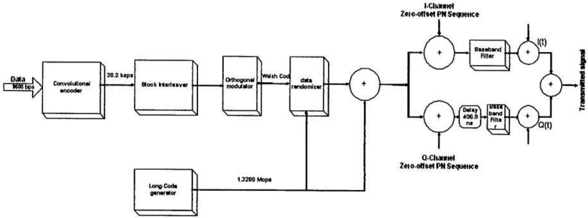

Figure 3. 6 CDMA Reverse Traffic Link

forward link, Walsh codes are used for spreading to denote a particular user, while on the reverse link they are used for data modulation.

The final-processing element performs the direct sequence spreading functions. First the modulation symbols or Walsh function are spread by using the mobile specific long code at a rate of 1.2288 Mchips/sec (256 chips per modulation symbol). Then the data stream is split into Q and I streams where it is modified with short code pair. The resulting spread spectrum signal is then carried over the air interface with filtered O-QPSK (Offset QPSK) modulation.

4 ADAPTIVE ARRAY ANTENNA

4.1 Smart Antennas

Smart-antennas are composed of phase array antennas and beam formers which combine the signals from the antenna array elements. A composite antenna pattern is produced, which can be controlled by adjusting the amplitude and phase with which the individual antenna signals are combined. This enables the array to act as a spatial filter, which can enhance or reject signals based on their direction of arrival. Smart antenna systems, which are applicable to the multi-access communication systems, allow customized beams to be generated for each mobile or group of mobiles. This allows channel reuse with the same cellular domain. Cellular systems usually use 120° sectorization at each base station. Each base station uses three separate sets of antenna for each 120°sector, with dual receive diversity in each sector. But each sector uses different frequency to reduce co-channel interference; therefore handoffs between sectors are required. For higher performance, narrower sectors could be used, but this will result in too many handoffs, (Goldsmith, 1998).

Multi-beam or adaptive array antenna (diversity antenna) without handoffs between beams (smart antenna) overcomes this problem.

4.2 Adaptive array antennas

Let us first consider the adaptive array antenna. Here, simple multiple antennas are usually used. The signals from the antennas are weighted and combined to form the array output to obtain a maximum signal-to- interference plus noise ratio (SINR).

Figure 4. 1 shows a general layout for a phase array antenna. The antenna elements in the adaptive array should all have similar antenna patterns.

Adjusting the phase between different antennas controls the direction where the maximum gain would appear. The phases are adjusted such that the signals due to a source in the direction where maximum gain is required are added in phase. This results in the gain of the combined antenna being equal to the sum of the gains of the individual antenna, (Ertel, 1998).

Phase array antennas have the following advantages: first, the arrays can theoretically cancel N interferes with M antennas ($M>N$) and achieve M-N fold diversity gain. Second, an M-fold diversity gain which produces a reduction in the required average output signal-to-noise ratio (SNR) for a given bit error rate (BER) with fading. Finally, the antenna can also have an M-fold gain, which is the reduction in the required received signal power for a given average of output signal-to-noise ratio. This gain is independent of the environment.

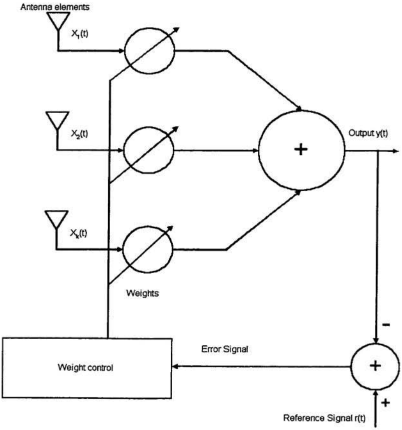

Figure 4. 1 Phase Array Antenna

4.3 Multi-beam Antennas

In the multi-beam antennas the signals induced on different elements are combined to form a single output. The response as a function of angle is referred to as beam pattern. The process of combining the signals from different antenna elements and adjusting their phase to point a beam in the desired direction is known as beam forming. The direction where the array has maximum gain is called beam direction. All patterns on either side of the beam direction have a low value, which is referred to as a null. Therefore beam forming exploits the differential phase between different antennas to modify the antenna pattern of the whole array into single antenna pattern. Moreover, multiple fixed beams can be used in a sector. An M-beam antenna can provide an M-fold antenna gain and some diversity gain by combining the received signals from different beams. This is referred to as angle diversity or dual diversity by using second antenna array that uses an orthogonal polarization (Goldsmith, 1998). The same beam can be used for the down-link as well as for the up-link, but it will provide only antenna gain on the down-link. Multi-beam antennas have the following disadvantages: non-uniform gain with respect to angle, second possibility of locking onto the wrong beam due to multipath or interference, and finally, they cannot suppress interference if it is in the same beam as the desired signal. The property of the flexibility of adjusting the array weighting to specify the array pattern has a significant application in canceling directional sources operating at the same frequency as the desired source but not in the same direction. In these situations, where the directions of the interferences are known, it is possible to cancel these interferences by placing the nulls in the pattern corresponding to these directions and steering the main beam in the direction of the desired signal. Beam forming in this situation is normally known as null beam forming, Figure 4. 2 shows this pattern. More material regarding beam forming and null beam forming can be found in (Winters, 1998).

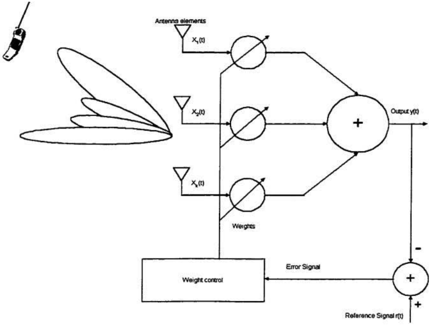

Figure 4. 2 Beam forming Pattern

4.4 Antenna-array Techniques

We referred in (Elfarawi, 1998) to certain problems concerning the wireless communication systems. Array technology has been established to overcome these drawbacks. In array technology the signals received by different antenna elements are weighted and combined to construct the desired output signals. This arrangement of the antenna array provides a diversity gain due to multi-path fading in addition to antenna gain. Diversity is used to overcome the problem of fading and clarifies the fact that signals arriving at different locations fade at different rates. To utilize diversity in practice, three different techniques can be used: Spatial, polarization and angle diversity (Garg, 1997).

Diversity combining is different from antenna array processing or beam forming. In diversity combining the signals are combined to increase the signal level without affecting the individual antenna pattern. A system employing a diversity combiner uses signals induced on other antenna separated far enough and combines these signals in one of the following ways:

Equal gain combiner adjusts the phases of the desired signals and combines them in-phase after equal weighting.

Maximal ratio combiner applies weight in proportion to the SNR and combines the weighted signal in-phase.

Selection diversity combiner selects the signal from one of the antennas for processing. The selection may be according to the power of the desired signal, the total power or the SIR available at each antenna.

The selection based upon the SIR is most effective in combating channel interference, whereas the equal gain combiner provides the lowest outage probabilities, (Ertel, 1998).
The above techniques are applied to the base station. The same techniques can be used for the handset with a limitation of the cost and the power consumption of the receiver electronics for each antenna.

4.5 Adaptive Beam forming for Wireless CDMA

The use of adaptive array antennas (smart antennas) in small, lightweight and low-power handled devices is unlikely in the next generation systems. However, the base station for these applications can use antenna arrays with space-time processing at the transmitter to reduce the co-channel interference and multipath, providing similar performance advantages as smart antenna in the receiver.
The phenomena of sectoring the base station into separate sets, (Elfarawi, et. al, 1998), leads to the use of antenna arrays in many applications like IS-136 TDMA systems, GSM systems and IS-95 CDMA systems to enhance the range and capacity increase.

4.6 Beam forming Techniques

The main purpose of implementing beam forming in communications systems is to combat multipath, extend coverage by the base station and reuse frequency channels within the cell.
In CDMA wireless systems, each user within a cell modulates the information signal with a unique coding sequence that identifies the sender. Although all users use the same frequency band at the same time, the intended receiver can detect each of the received

signals by using an appropriate decoding technique to detect the user of the subject matter. The users actually interfere with each other within the same cell and with other users in adjacent cells. Both interfering signal powers are reduced by the spreading gain of the code, (Viterbi, 1995).

There are several techniques (algorithms) used to design proper beam formers to select the desired signal from a multiple co-channel signals based on the available information. The traditional techniques combine high resolution direction-finding (DF) techniques such as MUSIC, ESPRIT and weighted subspace fitting (WSF), (Paulraj and et, al, 1985) and (Viberg, 1991) with optimum beam forming to estimate the signal waveforms, (Anderson, et. al 1991) and (Ottersten et. al, 1989). Other techniques used reference or training signal to find the optimum beam former, (Winters et. al, 1994).

In recent years, several property-restoral techniques have been developed which, exploit the temporal or spectral structure of communication signals while assuming no prior spatial information. These techniques take advantage of signal properties such as constant modulus (CM), (Mayrargue, 1993), discrete alphabet, (Talwar et. al, 1994), (Swindlhurst et. al, 1993), self-coherence (Agee at. al, 1994), and high order statistical properties (Tong et. al, 1993).

4.7 Signal Model

The above beam forming techniques are not suitable for CDMA systems. One reason for that can be explained as following: since all users in CDMA wireless system are co-channels, and their number may easily exceed the number of the antennas in addition to multipath propagation, and the fact that each transmission path may contain direct, reflected and diffracted paths at different time delays, the array manifold may be poorly defined. Hence, DF-based beam forming techniques are not applicable. Another reason is that there are no training or reference signals present in the mobile to base link, (Litva et. al, 1996). Hence reference signal based techniques cannot be used.

(Nugib, 1996) proposed a technique which is based on space-time processing framework. In this technique a coding filtering at each antenna for each user in the system is performed. The eigenstructure of the pre-and post-correlation array covariance matrices to estimate the channel vector and derive the corresponding beam formers is also exploited.

We used this technique to construct the beam former for the indoor transmission for the following reason: first, this technique does not require any training signal, and second, it does not require any assumptions on the signal propagation, which makes it suitable for different propagation settings.

In the signal model, we consider an array of M omnidirectional elements immersed in a homogeneous media in the far field signal sources. Let us consider the origin of the coordinate system to be the time reference as shown in Figure 4. 3, then the time taken by the plane wave arriving from the ith source in the direction (ϕ_i, θ_i) and measured from the λ_{th} element to the origin is given by (Godara, 1997)

$$\tau_\lambda(\phi_i, \theta_i) = \frac{r_\lambda . \upsilon(\phi_i, \theta_i)}{c} \tag{4.1}$$

where r_λ is the position vector of the λ_{th} element, $\nu(\phi_i, \theta_i)$ is the unit vector in the direction of (ϕ_i, θ_i), c is the speed of propagation of the plane wave, and (.) represents the inner product.

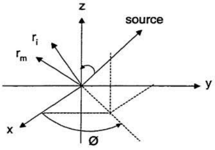

Figure 4. 3 System Coordinate

If we have a linear array of equispaced elements with element spacing d aligned with the x-axis such that the first element is situated at the origin, then the time taken by the plane wave becomes

$$\tau_{\lambda}(\theta_i) = \frac{d(\lambda - 1)\cos\theta_i}{c} \tag{4.2}$$

We can express the signal induced in the reference element due to the ith source as:

$$f_i(t)\, e^{j2\pi f_0 t} \tag{4.3}$$

where $f_i(t)$ denotes the complex modulating function. This function reflects the particular modulation used in the communication system.

For our particular system CDMA this function is given by (Godara, 1997)

$$f_i(\mathrm{t}) = \mathrm{b}_i(\mathrm{t})\ \mathrm{n}_i(\mathrm{t}) \tag{4.4}$$

where $b_i(t)$ denotes the message sequence and $n_i(t)$ is a pseudo random noise binary sequence. The modulating function in fact represents the complex low pass process of the system with zero mean and variance equal to the source power P_i as measured at the reference element.

The message sequence can be represented by considering the case of system block diagram represented in Figure 4. 4 where we have S number of stations (mobiles) with K users each of which simultaneously communicate with a base station that uses CDMA system with a Differential Phase Shift Keying (DPSK). DPSK is used here as a modulation scheme to avoid the need for synchronous carrier recovery at the receiver, which is a difficult task in a multipath propagation environment. Moreover, since no phase estimation is required, DPSK is often considered to be noncoherent communication technique. To elaborate the concept of DPSK in general, the received signal in any given signaling interval is compared to the phase of the received signal from the proceeding-signaling interval.

The data waveform from station S of user K is denoted as

$$b_{sk}(t) = \sum_{k=-\infty}^{\infty} b_{sk} P_{T_b}(t - kT_b) \quad , b_{sk} \in (1,-1) \tag{4.5}$$

where b_{sk} is a differentially encoded information data bit, P_{Tb} is a rectangular pulse of unit height and duration T_b.

Each user has a unique spread spectrum PN code of N chips that fit into one data bit (i.e. $T_b = NT_c$ or $T_b >> T_c$), where T_c is the chip duration with a processing gain G. The spread spectrum code of user K can be expressed as

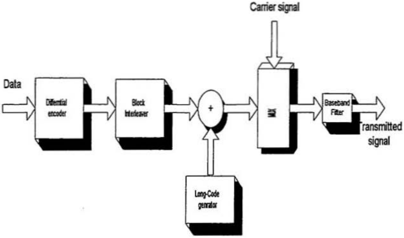

Figure 4. 4 Block Diagram of CDMA Up Link Transmitter

$$C_{sk}(t) = \sum_{k=-\infty}^{\infty} C_{sk} P_c\left(t - kT_c\right) \ , \quad C_{sk} \in \left(\mathbf{1}, \mathbf{-1}\right) \tag{4.6}$$

where $C_{sk,}$ the spread spectrum code of user k, it is a random variable taking values of ±1 with equal probability, P_c is a rectangular pulse of duration T_c. The processing gain G is defined as

$$G = T_b / T_c \tag{4.7}$$

In our work we used the configuration in figure 3.4 to generate the code sequence of the system.

Let us assume that X_{sk} represents the signal transmitted by user K. Then the signal transmitted by this user can be expressed as a function of amplitude of the carrier A, the spreading code, the differentially encoded bit, the angular carrier frequency ω_c, and the carrier phase shift for the K_{th} user θ_{sk}.

The amplitude of the carrier A can be evaluated as the product of a binary random variable and the square root of the power P transmitted by user K. We assume the existence of perfect power control, (i.e. the user adjusts its transmitted power such that the power received at the base station is kept constant).

Then we can write the transmitted signal from the K_{th} user as

$$X_{sk}(t) = A\, b_{sk}(t)\, C_{sk}(t)\, \mathrm{Cos}\,(\omega_c t + \theta_{sk}) \tag{4.8}$$

It is important to have resolvable paths existing in the system. This can be accomplished if the signal bandwidth is much larger than the coherence bandwidth of the radio channel. The multipath lowpass (complex) equivalent expression of the channel for the link of K_{th} user of station S and the base station can be written as

$$h_{sk}(t) = \rho_{sk} \sum_{\lambda=1}^{L} S_{sk}^{j}\; \alpha_{\lambda}^{sk}(t)\; \beta\left(t - \tau_{\lambda}^{sk}\right) e^{j\phi_{\lambda}^{sk}} \tag{4.9}$$

where ρ_{sk} is the effect of path loss and lognormal shadowing, which can be expressed as

$$10 \log_{10} \rho_{sk} \sim N\,(-\gamma\,.\;\; 10 \log_{10} r_{sk}\,.\; \sigma^2\,) \tag{4.10}$$

where γ is the path loss exponent, r_{sk} is the distance between the K_{th} user and the base station and σ^2 is the variance which varies between 6-12 depending on the degree of shadowing.

α_l^{sk} is the λ_{th} path gain or attenuation factor, ϕ_l^{sk} is the λ_{th} path phase shift, and τ_l^{sk} is the λ_{th} path time delay. L is the number of resolvable paths. The number of resolvable paths could be fixed or randomly changed. If fixed, it can be calculated as

$$L = \lfloor T_m / T_c \rfloor + 1 \tag{4.11}$$

where, T_m is the delay spread and T_c is the chip duration.

The channel parameter α, ϕ and τ in reality randomly change with time. However, the rate of their change is too slow compared to the rate of change of the signal so that they are constant over several bit durations.

The phase ϕ_l^{sk} is assumed to be uniformly distributed over $(0,2\pi)$, the path time delay is assumed to be uniformly distributed over ($0,T_{max}$ or T_b), where T_m is the maximum delay spread of the channel and T_b is the bit duration.

According to measurements performed in offices and factories (Bultitude, 1987 and Saleh, 1987), the path gain α_l^{sk} is assumed to follow independent Ricean distribution random variables.

In IS-95 standard, the spread signal bandwidth is very small compared to the carrier frequency. Therefore the narrowband assumption for antenna array is valid, (Naguib, 1994). This enables the modeling of the time delay due to propagation across the array as phase shifts.

If we include the effect of the antenna array on the signal, then the channel expression can be modified by inserting the response vector of the antenna arrays a_λ^{sk} into the original expression. Therefore, the vector multipath channel as seen by the K_{th} user with respect to the base station antenna array can be written as

$$h_{sk}(t) = \rho_{sk} \sum_{\lambda=1}^{L} S_{sk}^{j} \ \alpha_{\lambda}^{sk}(t) \ \beta\left(t - \tau_{\lambda}^{sk}\right) e^{j\varphi_{\lambda}^{sk}} a_{\lambda}^{sk} \tag{4.12}$$

Now without loss of generality we consider the case of a single cell. The complex received signal at the base station can be classified as two opposite signals. A desired signal that belongs to the K_{th} user of the subject matter, (in this case reference user number 1), and unwanted signals, which represent the interfering signals plus the noise. The received signal then can be written as

$$\mathrm{X}\ (\mathrm{t}) = \mathrm{X}_{\mathrm{d}}\ (\mathrm{t}) + \mathrm{X}_{\mathrm{u}}\ (\mathrm{t}) \tag{4.13}$$

where:

$$\begin{aligned} x_d(t) = x_{sk}(t) = \psi_{sk}\sqrt{P_{sk}}\ \rho_{sk} \sum_{\lambda=1}^{L} b_{\lambda}^{sk}\left(t - \tau_{\lambda}^{sk}\right) \\ .C_{\lambda}^{sk}\left(t - \tau_{\lambda}^{sk}\right)\alpha_{\lambda}^{sk}(t) e^{j\theta_{\lambda}^{sk}} a_{\lambda}^{sk} \end{aligned} \tag{4.14}$$

and

$$\begin{aligned} x_u(t) = x_{sk}(t) = \sum_{k=2}^{K} \sum_{\lambda=1}^{L} \psi_{sk}\sqrt{P_{sk}}\ \rho_{sk}\, b_{\lambda}^{sk}\left(t - \tau_{\lambda}^{sk}\right) \\ .C_{\lambda}^{sk}\left(t - \tau_{\lambda}^{sk}\right)\alpha_{\lambda}^{sk}(t) e^{j\theta_{\lambda}^{sk}} a_{\lambda}^{sk} + n(t) \end{aligned} \tag{4.15}$$

where n(t) is the white Gaussian noise and it has zero mean and covariance. Equation 4.14 represents the signal from the desired user in the presence of multipath, while equation 4.15 shows the signals received from other users within the desired user cell site.

Matlab codes and simulink models were used to evaluate the above partameters and signals.

4.7.1 Receiver Signal

For the λ_{th} path of the desired user K, the antenna outputs are correlated using a code denoted as C_1 $(t-\tau_{\lambda,1})$, where 1 indicates user number 1. Then the output expression will be

$$y_{\lambda 1} = \int_{\tau_{\lambda,1}}^{T+\tau_{\lambda,1}} x(t)\, C_1\left(t-\tau_{\lambda,1}\right) dt \tag{4.16}$$

substituting for the input signal we get

$$y_{\lambda 1} = \int_{\tau_{\lambda,1}}^{T+\tau_{\lambda,1}} [x_{s1}(t)+n(t)]\, C_1\left(t-\tau_{\lambda,1}\right) dt \tag{4.17}$$

$$y_{\lambda,1}(t) = \int_{\tau_{\lambda,1}}^{T+\tau_{\lambda,1}} \begin{bmatrix} A_{s1}\, \rho_{s1} \sum_{\lambda=1}^{L} b_{\lambda}^{s1}\left(t-\tau_{\lambda}^{s1}\right). \\ C_{\lambda}^{s1}\left(t-\tau_{\lambda}^{s1}\right) \alpha_{\lambda}^{s1}(t)\, e^{j\theta_{\lambda}^{s1}}\, a_{\lambda}^{s1} \end{bmatrix} C_1\left(t-\tau_{\lambda,1}\right) dt \tag{4.18}$$

Let,

$$I_{\lambda,1}^{s1} = \int_{\tau_{\lambda,1}}^{T+\tau_{\lambda,1}} b_{s1}\left(t-\tau_{\lambda}^{s1}\right) c\left(t-\tau_{\lambda}^{s1}\right) C_1\left(t-\tau_{\lambda,1}\right) dt \tag{4.19}$$

and

$$\tilde{n}(t) = \int_{\tau_{\lambda,1}}^{T+\tau_{\lambda,1}} n(t)\, C_1\left(t-\tau_{\lambda,1}\right) dt \tag{4.20}$$

Substituting the above two equations into the output expression we get

$$y_{\lambda,1}(t) = \sum_{sk=1}^{K} A_{sk}\,\rho_{sk} \sum_{\lambda=1}^{L} \alpha_{\lambda}^{sk}(t)\; e^{j\theta_{\lambda}^{sk}}\; a_{\lambda}^{sk}\; I_{\lambda,1}^{sk} + \tilde{n}(t) \qquad (4.21)$$

The post correlation noise covariance is given by

$$E\left\{\tilde{n}(t)\,\tilde{n}^{*}(t)\right\} = \sigma^{2}T\,I \qquad (4.22)$$

when we sample the output signal at sample instant T, the output will be

$$y_{\lambda,1}(T) = TA_{1}\alpha_{\lambda,1}\,e^{j\theta_{\lambda,1}}\,b_{1}\,a_{\lambda,1} + q_{\lambda,1} \qquad (4.23)$$

where $q_{\lambda,1}$ is the noise component due to thermal noise and multiple access interference.

The pre and post-correlation array covariance are given by

$$R_{xx} = E\left\{x\,x^{*}\right\} = A_{1}^{2}\,\alpha_{\lambda,1}^{2}\,a_{\lambda,1}\,a_{\lambda,1}^{*} + Q_{\lambda,1} \qquad (4.23)$$

$$R_{yy} = \frac{1}{TT_{c}}E\left\{y_{\lambda,1}\,y_{\lambda,1}^{*}\right\} = L\,A_{1}^{2}\,\alpha_{\lambda,1}^{2}\,a_{\lambda,1}\,a_{\lambda,1}^{*} + Q_{\lambda,1} \qquad (4.24)$$

$Q_{\lambda,1}$ is the array covariance due to all antenna signals other than the desired λ_{th} path signal of user K. $a_{\lambda,1}$ can be estimated as the principal eigenvector of the pair matrix (R_{xx} , R_{yy}). $Q_{\lambda,1}$ also can be estimated as

$$Q_{\lambda,1} = \frac{L}{L-1}\left(R_{xx} - \frac{1}{L}R_{yy}\right) \qquad (4.25)$$

The above estimation can be used to calculate the optimum beam forming weights for the system

$$W_{\lambda,1} = \frac{Q_{\lambda,1}^{-1}\,a_{\lambda,1}}{a_{\lambda,1}^{*}\,Q_{\lambda,1}^{-1}\,a_{\lambda,1}} \qquad (4.26)$$

Using these weights the output at sampling instant T is

$$Z_{\lambda,1} = W^*_{\lambda,1}\, y_{\lambda,1} \tag{4.27}$$

$$Z_{\lambda,1} = A_1\, \alpha_{\lambda,1}\, e^{j\theta_{\lambda,1}}\, b_1 + W^*_{\lambda,1}\, q_{\lambda,1} \tag{4.28}$$

If we consider the current and the previous sampling outputs, we will have the following outputs

$$Z^0_{\lambda,1} = A_1\, \alpha_{\lambda,1}\, e^{j\theta_{\lambda,1}}\, b^0_1 + W^*_{\lambda,1}\, q_{\lambda,1} \tag{4.29}$$

$$Z^{-1}_{\lambda,1} = A_1\, \alpha_{\lambda,1}\, e^{j\theta_{\lambda,1}}\, b^{-1}_1 + W^*_{\lambda,1}\, q_{\lambda,1} \tag{4.30}$$

The output decision of the DSPK demodulator and the hard decision then will be (Feher, 1995)

$$\xi_{\lambda,1} = \mathrm{Re}\left\{Z^0_{\lambda,1}\, Z^{*-1}_{\lambda,1}\right\} \tag{4.31}$$

$$\xi_1 = \mathrm{Re}\left\{\sum_{\lambda}^{L} Z^0_{\lambda,1}\, Z^{*-1}_{\lambda,1}\right\} \tag{4.32}$$

4.8 Matched Filter

Consider the output of the antenna array for a single user case and adaptive white Gaussian noise

$$X(t) = \sum_{\lambda=1}^{L} \sqrt{p_\lambda}\, b(t-\tau_\lambda)\, \tilde{C}(t-\tau_\lambda)\, e^{j\phi_\lambda} a_\lambda + n(t) \tag{4.33}$$

In order to derive the optimal space-time matched filter for the received signal, we consider the likelihood function of the received signal, conditioned on the knowledge of all parameters. First, let

$$X(t)=\sqrt{P_\lambda}\, b(t-\tau_\lambda)\tilde{C}(t-\tau_\lambda)e^{j\phi_\lambda}a_\lambda \tag{4.34}$$

The likelihood function of the received multipath signal can be written via the Cameron-Martin formula as

$$L(\{S(t);-\infty<t<\infty\})=C.\exp\{\Omega(b(t))/\sigma_n^2\} \tag{4.35}$$

Where C is an arbitrary constant and $\Omega(\,b(t)\,)$ is defined as

$$\Omega(b(t))=2.R\left\{\int_{-\infty}^{\infty}\sum_{\lambda=1}^{L}X_\lambda^*(t)S(t)\,dt\right\}-\int_{-\infty}^{\infty}\left|\sum_{\lambda=1}^{L}X_\lambda(t)\right|^2 dt \tag{4.36}$$

The objective is to select the bits b(t) that will maximize the likelihood function. Assuming that the user sends M information bits, the first integral yields

$$\int_{-\infty}^{\infty}\sum_{\lambda=1}^{L}X_\lambda^*(t)S(t)\,dt=\sum_{\lambda=1}^{L}\sqrt{p_\lambda}\,e^{j\phi_\lambda}\int_{-\infty}^{\infty}b(t-\tau_\lambda)\tilde{C}^*(t-\tau_\lambda)a_\lambda^*\,S(t)\,dt \tag{4.37}$$

$$=\sum_{n=1}^{M}b(n)\sum_{\lambda=1}^{L}\sqrt{p_\lambda}e^{j\phi_\lambda}\,Z_\lambda(n) \tag{4.38}$$

Where $z_1(n)$, $z_2(n)$,......, $z_L(n)$, are the matched filter outputs synchronously sampled with respect to each path signal and

$$Z_\lambda(n)=\int_{(n-1)T_b+\tau_\lambda}^{nT_b+\tau_\lambda}\tilde{C}^*(t-\tau_\lambda)a_\lambda^*\,S(t)\,dt,\quad \lambda=1,\ldots\ldots,L \tag{4.39}$$

The second integral in Ω (equation 4.36) does not depend on the received signal at the array, therefore, the only adequate statistic for the detection of bits b(n), where n=1 ... M, is $Z_\lambda(n)$, where $\lambda = 1 \ldots L$, b(n) is obtained by a linear operation on the received vector S(t).

4.9 RAKE Receiver

The outputs of the space-time-matched filter can be combined. Then the output of the combiner is passed to a decision device. The overall structure is called a RAKE receiver. The RAKE receiver combines the information obtained from several resolvable multipath components. It consists of a bank of correlators as shown in Figure 4. 5, each of which correlate to a particular multipath component of the desired signal. These correlators are utilized to separately detect M strongest multipath components. The outputs of each correlator are weighted to provide better estimate of the transmitted signal than is provided by a single component. Demodulation and bit decision are then based on the weighted outputs of the M correlators.

In CDMA spread spectrum systems, the chip rate is typically much greater than the flat fading bandwidth of the channel. Whereas conventional techniques require an equalizer to undo the intersymbol interference between adjacent symbols, CDMA spreading codes are designed to provide very low correlation between successive chips. Thus, propagation delay spread in radio channel merely provides multiple version of the transmitted signal at the receiver. If these multipath components are delayed in time by more than chip duration, they appear like uncorrelated noise at CDMA receiver, and equalization is not required. However, since there is useful information in the multipath components, CDMA receivers may combine the time delayed versions of the original signal transmission in order to improve the signal to noise ratio at the receiver.

A RAKE receiver does just that. It attempts to collect the time-shifted version of the original signal by providing a separate correlation receiver for each multipath signal, (Rappaport, 1996).

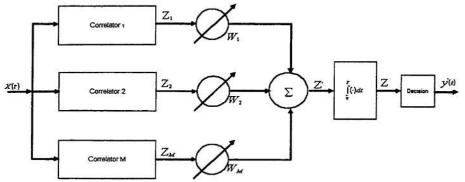

Figure 4. 5 RAKE Receiver

Since the path signals arrive at the receiver with different delays from different directions a more sophisticated means of exploiting the spatial structure of the received signals to obtain efficient combining of paths should be used. The proposed scheme for this service is called 2-D RAKE receiver which has been proposed by Khalaj, (Khalaj et. al, 1993). In the 2-D RAKE receiver,

Figure 4. 6, the weight vector W_N is used to improve PN code acquisition, and the weight vector W_k should be chosen in terms of an optimization criterion. As proposed in (Naguib, 1996) the weight vector can be obtained by using the array response vector of the λ_{th} path of the desired user signal and the covariance for the λ_{th} path interference plus noise. The array response vector can be estimated as the principal eigenvector of the matrix difference

In order to construct the beam former RAKE receiver, we use the code filtering approach proposed in (Naguib, 1995) for each resolvable multipath component.

First, we recall Eq (4.13), the total received signal vector. Since we considered user 1 as the reference user, then we assume that the time delays $\tau_{\lambda,1}$, $\lambda=1$ .... L are perfectly estimated and known. For the *n*th bit, the post-correlation signal vector for the multipath component is given by

$$y(n) = \frac{1}{\sqrt{T_b}} \int_{(n-1)T+\tau_{\lambda,1}}^{nT+\tau_{\lambda,1}} x(t) c(t-\tau_{\lambda,1}) \, dt \tag{4.40}$$

$$y(n) = 2\sqrt{T_b P_{\lambda,1}}\, b_1(n) e^{j\phi_{\lambda,1}} a_{\lambda,1} + i_{\lambda,1} + n_{\lambda,1} \tag{4.41}$$

where

$$n_{\lambda,1} = \frac{1}{\sqrt{T_b}} \int_{(n-1)T+\tau_{\lambda,1}}^{nT+\tau_{\lambda,1}} n(t) C_1^{*}(t-\tau_{\lambda,1}) \, dt \tag{4.42}$$

is the undesired component due to thermal noise, and

$$i_{\lambda,1} = \sum_{\substack{k=1 \\ k\neq\lambda}}^{L} \sqrt{P_{k,1}}\, I_{\lambda,1,k}\, e^{j\phi_{k,\lambda}} a_{k,\lambda} + \sum_{i=2}^{N} \sum_{k=1}^{L} \sqrt{P_{k,i}}\, I_{\lambda,1,k,i}\, e^{j\phi_{k,i}} a_{k,i} \tag{4.43}$$

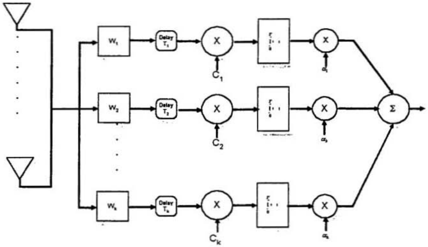

Figure 4. 6 2-D RAKE Receiver

is the undesired component due to multiple access interference plus self interference, where $I_{\lambda,1,k}$ is defined as

$$I_{\lambda,1,k,i} = \frac{1}{\sqrt{T_b}} \int_{(n-1)T+\tau_{\lambda,1}}^{nT+\tau_{\lambda,1}} b(t-\tau_{\lambda,1}) C_i^*(t-\tau_{k,i}) C_1^*(t-\tau_{\lambda,1})\, dt \tag{4.44}$$

The parameters R_{xx}, $R_{yy,\lambda,1}$ and $R_{uu,\lambda,1}$ are given by

$$R_{uu,\lambda,1} = \sum_{\substack{k=1 \\ k\neq\lambda}}^{L} 2P_{k,1}\; a_{k,\lambda} a_{k,\lambda}^* + \sum_{i=2}^{N} \sum_{\lambda=1}^{L} 2P_{\lambda,i}\, a_{\lambda,i}\, a_{\lambda,i}^* + \sigma_n^2 I \tag{4.45}$$

$$R_{xx} = 2P_{\lambda,1}\; a_{k,\lambda} a_{k,\lambda}^* + R_{uu,\lambda,1} \tag{4.46}$$

$$R_{yy,\lambda,1} = 2GP_{\lambda,1}\; a_{\lambda,1} a_{\lambda,1}^* + R_{uu,\lambda,1} \tag{4.47}$$

The same approach for calculating the weight $W_{\lambda,1}$ can be used here as well

$$W_{\lambda,1} = \xi R_{uu,\lambda,1}^{-1}\, a_{\lambda,1} \tag{4.48}$$

Therefore the beam former output for the λth path is

$$Z_{\lambda,1}(n) = W_{\lambda,1}^*\, y_{\lambda,1}(n) \tag{4.49}$$

$$Z_{\lambda,1}(n) = 2\sqrt{T_b P_{\lambda,1}}\, b_1(n) e^{j\phi_{\lambda,1}}\, W_{\lambda,1}^*\, a_{\lambda,1} + W_{\lambda,1}^*\, u_{\lambda,1} \tag{4.50}$$

and the corresponding path signal to interference-plus-noise ratio is

$$\gamma_{\lambda,1} = 2GP_{\lambda,1} \cdot a_{\lambda,1}^*\; R_{uu,\lambda,1}^{-1}\, a_{\lambda,1} \tag{4.51}$$

5 PERFORMANCE ANALYSIS

5.1 Diversity

In the radio propagation environment, diversity techniques can be used to improve the system performance. It implies that a number of signals carrying the same information can be combined to improve the performance. Two techniques are mostly considered: Selection diversity and maximal ratio combining.

5.1.1 Selection Diversity

Selection diversity is based on selecting the strongest signal carrying the same information. In DS-CDMA resolvable paths, this can be accomplished by selecting the path with largest auto-correlation peak. This implies that the highest order of diversity that can be achieved with one antenna is equal to the number of resolvable paths. Multiple antennas then can be used if the order of diversity is too low.

5.1.2 Maximal Ratio Combining

The idea of maximal ratio combining is to sum the demodulation results of a group of signals carrying the same information and then using the result as a decision variable. In the indoor environment DPSK modulation is much preferable to avoid the need for synchronous carrier recovery. The channel parameters are assumed to be constant over two consecutive signaling intervals, which makes it possible to avoid the need to estimate the channel parameters since the signal is automatically weighted. This techniques is used throughout the following analysis.

5.2 Bit Error Probability

The estimation of the covariance of the interference and the noise is given in equation 4.25. Then the post-correlation antenna outputs are combined via beam forming to estimate the signal from the desired user. For optimal combining, the optimum beam forming weight are given by Wiener solution, (Winters et. al, 1994)

$$\mathrm{W}_1 = \frac{\mathrm{Q}_1^{-1}\,\mathrm{a}_1}{\mathrm{a}_1^{*}\,\mathrm{Q}_1^{-1}\,\mathrm{a}_1} \tag{5.1}$$

and the corresponding bit energy to interference-plus-noise ratio is given by

$$\frac{E_b}{INR} = G\,P\,a_1^{*}\,Q_1^{-1}\,a_1 \tag{5.2}$$

As mentioned in (Winters et. al, 1994) if the array elements have half-wavelength spacing, and there are a large number of users uniformly distributed in space, then

$$\mathrm{Q}_1 = \mathrm{Scalar} \times \mathbf{I} \tag{5.3}$$

Which means that in this case $W_1 = a_1$ or simply we have a simple beam forming which is the optimal choice. The distribution function for the bit-energy to interference-plus-noise ratio can be obtained as

$$d_1(k) = a_1^* \, z_1(k) = S_1(k) + n_1(k) + n_2(k) + n_T(k) \tag{5.4}$$

$$d_1(k) = G\alpha_1 b_1 + \sum_{k=2}^{K} I_k(k) a_1^* a_k + a_1^* n_T(k) \tag{5.5}$$

The first term $S_1(k)$ is due to the signal from the desired user, the second term $n_1(k)$ is due to the multiple access interference from users within the cell which is zero mean, and the third term $n_T(k)$ is due to the additive thermal noise, which is normal with zero mean and variance equal to

$$Var\{n_T\} = \frac{G\sigma^2}{M} \tag{5.6}$$

The variance of n_1 is given by

$$Var\{n_1\} = GP \sum_{k=2}^{K} \psi_k \left\| a_1^* a_k \right\|^2 \tag{5.7}$$

This is a random variable that depends on the voice activity of the users, their distance from their cell site and shadowing and fading effects.

The signal to interference-plus-noise ratio SINR then can be written as

$$SINR = \frac{G}{\dfrac{\sigma^2}{MP} + I_1} \tag{5.8}$$

where I_1 is the interference to signal density ratio due to user cell and is given by

$$I_1 = \sum_{k=2}^{K} \psi_k \left\| a_1^* a_k \right\|^2 \tag{5.9}$$

For large K, I_1 can be approximated by Gaussian random variable with mean equal to μ_I (K-1), and variance equal to σ^2_I(K-1). The values of μ_I and σ_I depend on the voice activity factor v, the antenna array configuration and the cell geometry.

The probability of SINR then can be given by

$$\Pr\{SINR\}=Q\left(\frac{\delta-\mu_I}{\sigma_I}\right) \tag{5.10}$$

where

$$\delta=\frac{G}{\eta}-\frac{\sigma^2}{PM}, \mu_I=\mu_1(K-1), \sigma_I^2=\sigma_1^2(K-1)\, and\, Q(x)=\frac{1}{\sqrt{2\pi}}\int_x^{\infty} e^{-y^2/2}dy$$

If we compute equation 5.10 under certain condition it will give the outage probability of the system, which is defined, as the probability that the bit-energy to interference-plus-noise ratio will fall below the level η required for certain performance.

Therefore the outage probability of the system is given by

$$\Pr\{SINR\le\eta\}=Q\left(\frac{\delta-\mu_I}{\sigma_I}\right) \tag{5.11}$$

5.2.1 Computational Results

To study the performance of the system with the bit error probability and hence determine the capacity of the system in the direct signal propagation, we used the following parameters for the model, (Elfarawi and Sinha, 2000). We assumed a circular array configuration of the antenna array with 5 and 8 elements. Table 5.1 shows the parameters used in the computational procedure.

Table 5. 1 System parameters

Parameter	Value
BER	10^{-3}
Processing gain (G)	128 and 256
Variance σ_s	8dB
Voice activity (ν)	0.375
Array elements (M)	1, 5 and 8
Array spacing	$\lambda/2$

The BER is assumed to be 10^{-3}, which corresponds to SIN of 7dB. The mean and variance of the internal interference were evaluated using matlab code and the method described by (Gilhousen et. al, 1991). The probability of SINR is computed using equation 5.10.

We examine the capacity of the system in the presence of direct propagation using different antenna elements and two different processing gains. We noticed that as the number of antenna array elements increases the number of users served by the base station increases for a certain SINR. Figure 5. 1 shows the capacity of the cell site in terms of the outage probability with no path signals available.

In Figure 5. 2 it is clear that a slight improvement in the capacity has been achieved for the base station using an antenna array, when the processing gain is increased to 256.

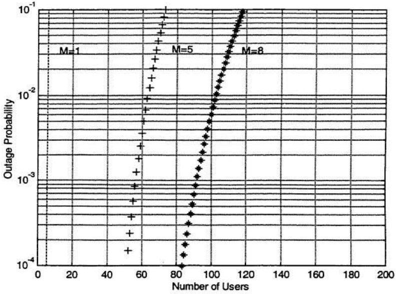

Figure 5. 1 Capacity of the system (Outage probability Vs Number of users) G=128

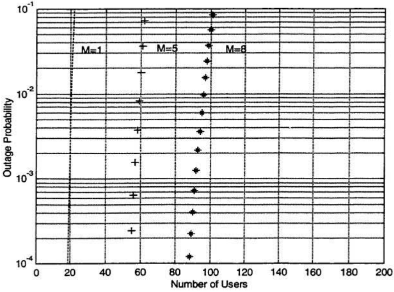

Figure 5. 2 Capacity of the system (Outage probability Vs Number of users) G=256

5.3 Bit Error Probability (Multipath Effect)

So far we considered the case of single user with direct signal propagation. In the case of multipath propagation, the transmitted signal arrives at the base station receiver with different time delays, attenuations and carrier phases as a result of reflections from the terrains. The structure of these individual paths can be identified and exploited to the degree that they can be resolved from one another. The wider the bandwidth of the signal compared to the coherence bandwidth of the channel, the more resolvable are the path (Paroakis, 1995). In spread spectrum signals using PN sequences with chip duration Tc =T/G, the individual paths can be resolved as long as their relative delays are more than one chip duration.

In general the path signals arrive at the receiver not only with different delays, but also from different directions in space. A single antenna receiver cannot utilize the spatial structure of the received signals. With antenna arrays, the spatial structure can be exploited to obtain a more efficient combining of the paths. This structure is called RAKE receiver and it has been introduced in chapter four.

Recalling equation 4.25, we can re-write it to estimate the covariance for the λ-th path interference-plus noise as

$$Q_{\lambda 1} = \frac{G}{G-1}\left(R_{xx} - \frac{1}{G} R_{\lambda,z1\lambda,z1} \right) \tag{5.12}$$

Then the corresponding optimum beam former is given by

$$\mathbf{W}_{\lambda 1} = \frac{\mathbf{Q}_{\lambda 1}^{-1}\, \mathbf{a}_{\lambda 1}}{\mathbf{a}_{\lambda 1}^{*}\, \mathbf{Q}_{\lambda 1}^{-1}\, \mathbf{a}_{\lambda 1}} \tag{5.13}$$

In the same way as before, we can write the bit-energy to interference-plus-noise ratio as

$$\frac{E_b}{INR} = SINR = GP \frac{\sum_{\lambda=1}^{L} \left\|\beta_{\lambda,1}\right\|^2}{\sum_{\lambda=1}^{L} \left\|\beta_{\lambda,1}\right\|^2 W_{\lambda,1}^{*}\, Q_{\lambda,1}\, W_{\lambda,1}} \tag{5.14}$$

If the conventional beam forming is used which means $W_{\lambda,1} = a_{\lambda,1}$, then the above equation can be written as

$$SINR = GP \frac{\sum_{\lambda=1}^{L} \left\|\beta_{\lambda,1}\right\|^2}{\sum_{\lambda=1}^{L} \left\|\beta_{\lambda,1}\right\|^2 a_{\lambda,1}^{*}\, Q_{\lambda,1} a_{\lambda,1}} \tag{5.15}$$

$$SINR = GP \frac{\sum_{\lambda=1}^{L} \left\|\beta_{\lambda,1}\right\|^2}{\sum_{\lambda=1}^{L} \left\|\beta_{\lambda,1}\right\|^2 I_{\lambda,1}} \tag{5.16}$$

where $I_{\lambda,1}$ is the interference-plus noise density for the λ_{th} path signal. $I_1 \ldots I_L$ are uncorrelated and can be approximated as Gaussian with the same mean and variance. The signal to noise-plus-interference ratio then is given by

$$SINR = \frac{G}{\max\{I_\lambda\}_{\lambda=1}^{L}} \tag{5.17}$$

and hence the probability of SINR not exceeding certain threshold is can be written as

$$\Pr\{SINR \leq \eta\} \leq \Pr\left\{\frac{G}{\max\{I_\lambda\}_{\lambda=1}^{L}} \leq \eta\right\} \tag{5.18}$$

Then the probability density of the maximum interference plus noise can be written as

$$f(\alpha) = G\left[1 - Q\left(\frac{\alpha - \mu_I}{\sigma_I}\right)\right]^{L} \frac{1}{\sqrt{2\pi\sigma_I^2}} \exp\left\{-\frac{(\alpha - \mu_I)^2}{2\sigma_I^2}\right\} \tag{5.19}$$

where μ_I is the mean and σ_I^2 is variance. Then we can write the outage probability of the system as

$$\Pr\{SINR \leq \eta\} \leq \int_{\chi}^{\infty} f(\alpha)\, d\alpha = 1 - \left[1 - Q\left(\frac{\delta - \mu_I}{\sigma_I}\right)\right]^L \tag{5.20}$$

where $\delta = \frac{G}{\eta} - \frac{\sigma^2}{PM}$, and $Q(y) = \frac{1}{\sqrt{2\pi}} \int_{y}^{\infty} e^{\frac{-y^2}{2}}\, dy$

5.3.1 Computational Results for Multipath Propagation

The number of paths is considered to be 3 and 4. The other parameters are mentioned in Table 5. 1.

The outage probability is computed using the above parameters, and the capacity of the cell site in terms of number of users is determined. First the capacity of the cell site is examined for 5 antenna array elements, processing gains of 128 and 256 with 3 resolvable paths. As multipath exist and a RAKE receiver is used as a filter, the capacity of the cell site for both multipath and no multipath is too close due to the good performance of the antenna arrays.

Figure 5. 3 and Figure 5. 4 show the performance of the system with 128 and 256 processing gains respectively.

We also examined the effect of the resolvable paths on the system with 5 antenna-array elements. In Figure 5. 5 and Figure 5. 6, the two configurations are performing almost the same for 4 resolvable paths.

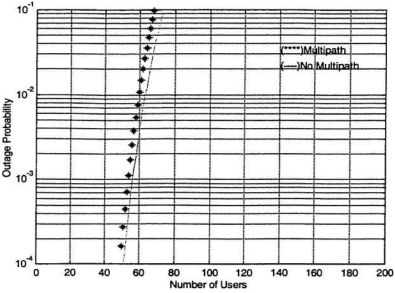

Figure 5. 3 Capacity of the system (Outage probability Vs Number of users) M=5, G=128 and L=3

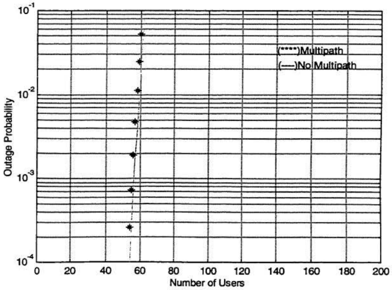

Figure 5. 4 Capacity of the system (Outage probability Vs Number of users) M=5, G=256 and L=3

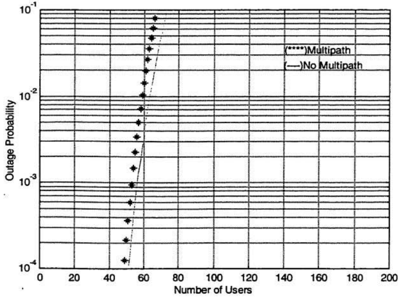

Figure 5. 5 Capacity of the system (Outage probability Vs Number of users) M=5, G=128 and L=4

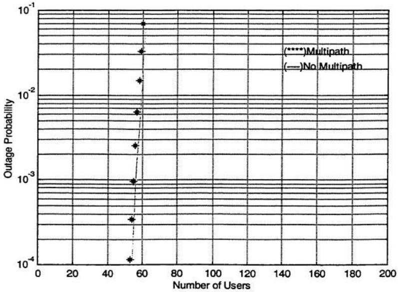

Figure 5. 6 Capacity of the system (Outage probability Vs Number of users) M=5, G=256 and L=4

We noticed that there is no major difference in the capacity with the increase of the resolvable paths due to the performance of the antenna arrays.

Now we consider the system outage probability when the number of sensors increases. In Figure 5. 7 and Figure 5. 8 we plot the system outage probability as a function of number of users when the number of antenna array sensors increases to 8 elements. From these figures we notice that the antenna array results in a many fold increase in the system capacity.

Figure 5. 9 and Figure 5. 10 show the performance with the number of resolvable paths increased to 4. Both figures show that the system is as good as when there is no multipath present.

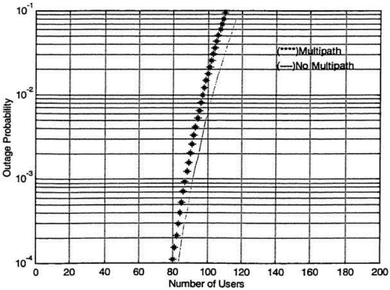

Figure 5. 7 Capacity of the system (Outage probability Vs Number of users) M=8, G=128 and L=3

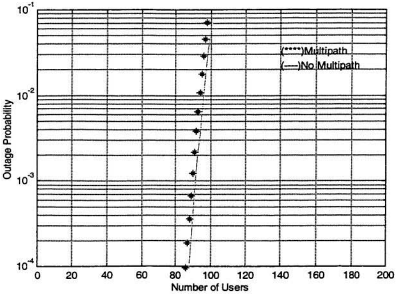

Figure 5. 8 Capacity of the system (Outage probability Vs Number of users) M=8, G=256 and L=3

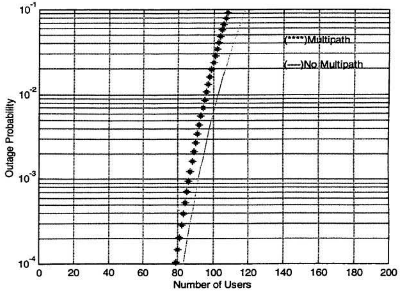

Figure 5. 9 Capacity of the system (Outage probability Vs Number of users) M=8, G=128 and L=4

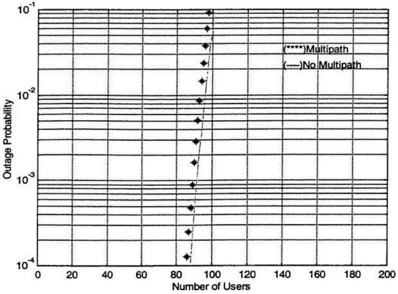

Figure 5. 10 Capacity of the system (Outage probability Vs Number of users) M=8, G=256 and L=4

6 CONCLUSIONS

6.1 Conclusions

The main focus of this work was the development and application of advanced antenna array signal processing techniques to indoor CDMA wireless communication systems. This work has been prompted by the current thrust in wireless communication technology and the need to support the projected capacity demands with the introduction of new personal communication services.

This work showed how could antenna array (smart antenna) can provide a significant improvement in the performance and increase the capacity of the second generation of wireless personal communication.

We overviewed the concept of smart antennas and discussed the techniques used in solving the problems associated with the wireless personal communication. We discussed how multibeam antennas are capable of increasing the capacity of CDMA systems. Chapters 4 and 5 presented the contributions of this thesis.

We focussed on the mobile to base station link. We assumed that only the base station was equipped with smart antenna, which is used to transmit and receive signals to and from the mobile unit. We also made the assumption of a perfect power control at the base station where all users transmitted to the base station with the same power.

An overview of wireless radio channel and statistical models for antenna channels was also provided. The assumption that indoor propagation could be modeled as Rician fading type is valid and it is dependent on the value of the parameter R that is found by measurement to be 6.8 or 11 dB.

The techniques used in spread spectrum technology were also investigated. We showed how codes could be used to an extent to improve the signal security and privacy.

Two types of propagation were considered. Direct signal propagation and multipath propagation. The performance of an indoor CDMA system was assessed in terms of error probability and outage probability. In general we can state that antenna array configuration at the base station improves the performance significantly.

We noticed that the performance is quite sensitive to the value of processing gain, the number of sensors in the array and the number of the propagation paths. For a certain BER value the number of subscribers can be improved as the number of sensors increases.

6.2 Future works

This research work by itself cannot answer all questions, which might arise when discussing the indoor wireless communications. Further studies not necessarily limited to the following areas could be of great assistance to complete this investigation.

- Through out this work the power is assumed to be perfect (i.e. all users transmit the same power). In reality this assumption is not quite correct. The mobile transmission criterion suffers from the so-called near-far effects. This phenomenon needs to be investigated.
- The Forward Error Correction FEC is an important concept. It can be used to further improve the performance of the system. FEC has the ability to reduce the channel error propability.
- An experimental work could give clearer picture and support the theoretical investigations.

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