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Signal Processing for Multiple Receive Antenna Systems in the Presence of Multiple Access Interference

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Signal Processing for Multiple Receive Antenna Systems in the Presence of Multiple Access Interference

by

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Abstract

In wireless communications systems employing DS-CDMA, the base stations with multiple antennas receive signals from different mobiles that are affected by fading and noise as well as interference from other users, thus diminishing the capacity. This dissertation considers a base station with multiple receive antennas which are weighted spatially by using maximal ratio combining. Moreover, at the transmitter side, the thesis examines the use of pilot signals along with the data stream for channel estimation purpose. For each user, the pilot signals are transmitted in parallel with the data signal using orthogonal signature waveforms to those of the data signal. The proposed receiver in this thesis uses a decorrelator detector to not only to decorrelate the data signal from the MAI components, but also to decorrelate the pilot signal to achieve good channel gain estimation. The study shows that the proposed system has a good resistance toward the near-far problem, give good space receiver performance efficiency comparing it with conventional space receivers and it has less power assigned to the pilot power stream to obtain best channel estimation within a certain fading rate.

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First of all, I want to express deep gratitude to my supervisor, Professor Claude D'Amours, for all the time and patience he has given to me through this research. He has provided a solid basis to my studies. Without the mercy of God and then without the supervisor's careful guidance, continuous support and critical insight, I would not have reached this point.

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Finally, I am indebted to my wonderful parents, my four lovely sisters and my friends for their steadfast love, support and encouragement. I dedicate this work as always to God then to my father, mother and for the whole engineering community.

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List of Abbreviations

AAA	Adaptive Antenna Array
AOA	Angle of Arrival
AMUD	Adaptive Multi-user Detector
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CAA	Conventional Antenna Array
CDMA	Code Division Multiple Access
CCI	Co-Channel Interference
dB	Decibel
DOA	Direction of Arrival
DS-CDMA	Direct Sequence-Code Division Multiple Access
DS/SS	Direct Sequence/ Spread Spectrum
FDMA	Frequency Division Multiple Access
FH/SS	Frequency Hopping / Spread Spectrum
FIR	Finite Impulse Response
ISI	Intersymbol Interference
LMS	Least Mean Square
MRC	Maximal Ratio Combining
MAI	Multiple Access Interference
MMSE	Minimum Mean Square Error

MUD	**Multi-User Detection**
PG	**Processing Gain**
PIC	**Parallel Interference Cancellation**
PN	**Pseudo-Random Noise**
RLS	**Recursive Least Square**
SIC	**Successive Interference Cancellation**
SNIR	**Signal to Interference plus Noise Ratio**
SNR	**Signal to Noise Ratio**
SDMA	**Space Division Multiple Access**
PSAM	**Pilot Symbol Assisted Modulation**
PT	**Pulse Tone**
SS	**Spread Spectrum**
ST	**Space-Time**
TDMA	**Time Division Multiple Access**
QOS	**Quality of Service**
ULA	**Uniform Linear Array**

List of Symbols

P_r	**Received power**
P_t	**Transmitted power**
Gr	**Gain of received antenna**
Gt	**Gain of transmitted antenna**
PL	**Path loss**
d_r	**The distance between the transmitter and the receiver.**
d_0	**The close in reference distance which is determined from measurements.**
$PL(d_o)$	**The ensemble average scale path loss at distance** d_o
n_o	**The path loss exponent**
ω	**Angular frequency**
λ	**The wavelength of the signal**
y(t)	**Output of the combiner**
x(t)	**Input data to antenna diversity**
r(t)	**The reference signal**
e(t)	**The error signal which is the difference between the reference and the output signal**
M	**The number of antenna elements**
$[.]^T$	**Denotes the transpose of the vector**
$[.]^H$	**Denotes the Hermitian transpose of the vector**
$[.]^*$	**Denotes the conjugate of the value**

d	The information signal
i	index for the user
θ_i	Direction of arrival (DOA) angle for the signal coming from user i
T_b	Signaling period
f	frequency in hertz
y(k)	Denotes the sample signal of y(t) at time instance k
x(k)	Denotes the sample signal of x(t) at time instance k
r(k)	Denotes the sample signal of r(t) at time instance k
E[.]	The expectation operator of [.]
Sgn[.]	The sign of [.]
R	The correlation matrix for the input data
ρ_{ij}	The cross correlation between the code user i and code user j
k	iteration index
I	Identity matrix
a(k)	The AR coefficient
c(t)	The spreading waveform
b(t)	The binary data signal composed of a sequence of symbols b_k
N_s	The number of samples in the chip
N	The processing gain
n	The bit index
$\hat{b}_{i,n}$	The estimated n^{th} bit for user i
f_c	Carrier frequency
P	The power of the signal

b_n	n^{th} data bit
ψ_T	A unit rectangular pulse of duration T
T_b	Bit period
v_i	The CDMA signal of user i
τ_i	The time delay for user i
ϕ_i	The random phase for the signal of user i
c	Speed of the light
m	Index for the antenna element
$a(\theta)$	The steering vector
q	Number of narrow band signals arriving at the antenna array
n(t)	Noise vector
L(t)	Correlated noise
u	Total transmitted signal for user i
A_{cp}	Amplitude of pilot signal
A_{cd}	Amplitude of data signal

Chapter 1

Introduction

1.1 Motivation

In the next generation of wireless communications, the industry intends to offer a high-speed wireless access and wireless multimedia services. It is desired to integrate both voice and data services such as voice over IP (VOIP) and high speed Internet access. However, in order to provide these services to the subscribers, high data rates and acceptable signal to interference and noise ratios (SINR) are required. The limitations in bandwidth as well as the fluctuations in the received power are some of the most important challenges that face wireless designers aiming to provide those services. Another important factor that must be taken into consideration is that the demand for more capacity is increasing rapidly these days. Statistics show that wireless services are currently growing at about 40% per year, and this trend is likely to continue for several years. The progress in radio technology is creating new and improved services without huge increase in cost to the consumer. This results in increasing the airtime usage and an increase in the number of subscribers [1].

An option for increasing the range of a wireless communications system and/or the overall amount of data that can be transmitted on it per unit time is by increasing the system's bandwidth, which will lead to higher data rates and/or more transmitting users. Another option is to increase the transmitted power of all transmitters, which will allow for greater data rates and increased transmission range. Unfortunately, in wireless communication systems the radio frequency spectrum and transmission power is limited. As much as clients try to benefit from wireless communication technology, because of the shortage of available spectrum within the cell, increasing transmitted power may cause unacceptable levels of multiple access interference (MAI). Thus, researchers are challenged to develop efficient transmission techniques and multiple access techniques that efficiently utilize the limited spectrum available, to increase the system capacity and to provide a good structure in canceling the multiple access interference.

Space time detectors which use multiple antennas in conjunction with multi-user detection has a key role in the next generation of mobile communications. It is one of the most promising technologies used for reducing the MAI and hence increasing the capacity of wireless cellular systems. The basic concept of space-time joint detection is to deliver high gain in the direction of the desired user and to null out the interference [49].

Most wireless applications such as cellular radios modems use temporal signal processing and ignore spatial signal processing. Using the temporal signal processing alone cannot effectively address the problem of multiple access interference (MAI) that arises from different users sharing the same channel, thereby limiting the range and capacity of the wireless network. In order to mitigate the MAI in the wireless networks, a spatial dimension element has to be added to the temporal signal processing by having multiple far away from the base station. These near users will create high interference levels and limit the capacity of the system. This problem is due to the MAI, where the conventional receiver ignores or makes no efforts to estimate the MAI when detecting the desired user. The interference caused by the stronger user to the weaker users may significantly degrade those users' performances. To combat the near-far problem, power control is required to keep all users' received power within reasonable limits at the receiver.

Since the power control requires a complex algorithm to achieve equality in power among users, multi-user detection may be used to overcome the near-far problem. In multi-user detection, the information of all users' (the desired plus the MAI user) is used to detect each user individually. Verdu in [48] and [49] derives the optimum multi-user detector for DS-CDMA systems and shows that the near far problem is not an inherent problem for DS-CDMA. Other multi-user detectors are presented in the literature, which we will discuss in more detail in chapter 3.

Most wireless applications such as cellular radios modems use temporal signal processing and ignore the spatial signal processing. Using the temporal signal processing, such as multi-user detectors alone, cannot effectively address the problem of multiple access interference (MAI) that arises from different users sharing the same channel as in CDMA [5]. In order to mitigate the MAI in the wireless networks, a spatial dimension can be added to the temporal signal processing by having multiple antennas at the front end of the modem.

Using multiple antennas, allowing for spatial processing techniques, is another useful technique added to temporal processing technique to mitigate the cochannel interference, fading and multipath in wireless channel [26]-[30]. Multiple antenna arrays may be combined with a multi-user detector or interference canceller to enhance the performance of the wireless communication system. Problems associated with conventional receivers and conventional antenna arrays, such as low signal to interference ratio, the near-far problem, interference with same angle of arrival (AOA) as the desired user as well as when the number of users is much larger than number of antenna elements can be solved by using the combined techniques. Therefore, the proposed space-time processing modem can improve network capacity, coverage and quality by not only reducing MAI, but also by enhancing diversity and array gain.

In the presence of multiple access interference, several space-time algorithms have been discussed in the open literature. The known joint detectors can be classified in terms of complexity and efficiency in performance. A space-time RAKE (ST-RAKE) receiver has a simple structure compared with other receivers [38], but the ST-RAKE can't estimate the MAI and performs poorly in the presence of severe MAI. On the other hand, adaptive array processing (or smart antennas) can cancel the MAI by placing nulls in the antenna beam pattern in the direction of the interference [40]-[42]. Moreover, the adaptive antenna array can be combined with the multi-user detector or interference canceller to enhance the performance of the wireless communication system. This technique can eliminate the MAI, even if the MAI has the same angle of arrival (AOA) as the desired user [59]. However, this class of space-time detector requires a lot of information at the receiver side, such as array response vector (also called the steering factor), and angle of arrival (AOA) of desired user. The blind algorithm requires only to know about the spreading code of the desired user [46]. Furthermore, because of the time-varying channel, the adaptive antenna-multi-user detector needs a long training sequence to find the near-optimal weights. The rapid fluctuations of the channel may cause the antenna array's response to jump from one local minima to another. Overall, although the combined smart antenna with multi-user detector will show significant improvement of capacity, it needs a large number of digital signal processing (DSP) operations.

In this thesis we focus on developing a new simple space-time joint detection system, used with DS-CDMA systems to increase the wireless system capacity and improve the performance. The proposed system employs a linear decorrelator detector in each branch of a multiple antenna receiver that will almost effectively combat the MAI while producing a system that is stable and low in complexity. The decorrelator is not only used to eliminate the MAI in the data signals, but also it is used to decorrelate the pilot signals needed for channel estimation as well. As we will see in the simulation results in Chapter 4, our proposed system will show improvement when compared with the conventional receiver. Here are some of these features;

1. The power allocated to pilot stream needs only 15% of the total transmitted power in order to have a good tradeoff between channel estimation and reliable data recovery when the channel's normalized fading rate is 0.1.
2. The proposed system has good immunity to the near far problem even in sever conditions. Its near-far resistance is superior to that of the conventional system.

3. Since our proposed system uses more than one antenna elements at the receiver, space diversity is achieved.
4. At higher signal to noise ratio, the propose system will continue showing its improvement in performance in the presence of MAI, while an error floor occurs in the conventional receiver.

1.3 Thesis Structure

This thesis is outlined as follows:

Chapter 2 introduces the fundamental characteristics of channels in the digital communication field. Then an overview of wireless channel will be presented. This is followed by a description of our proposed channel model. Afterwards, channel estimation will be discussed by showing some methods that are used to estimate the channel response. Next we will jump to study the performance of binary modulation in additive white Gaussian noise (AWGN) and in a Rayleigh fading channel, with a single receive antenna and then with multiple antennas to show space diversity. Our last topic in this chapter will be spread spectrum (SS). We will show different SS types, as well as its functionality and the advantages and disadvantages of using SS techniques.

Chapter 3 present analyses for the space-time receiver for wireless communication. In this chapter, we will show the spatial diversity advantages that will be gained by having multiple antenna elements at the receiver. Moreover, we will examine how the spatial processing will able to separate multiple users. An overview of adaptive antennas will be presented as an application example to show the gain of using spatial diversity. The advantages of using more than one antenna by the receiver will be presented at the end of this section. Then we will discuss several types and classes of multi-user detectors. Finally, previous literature research work will be discussed to show most of the joint detection techniques (combining multiple antennas with multi-user detection) that already has been studied in space-time processing at the receiver modem.

Chapter 4 presents the simulation models and results. Here we will describe our proposed space-time CDMA system by presenting the mathematical expressions of the proposed algorithm for the transmitter, channel and receiver model. Then we will summarize our proposed system by discussing the system parameters that are used in our simulations. We evaluate the performance of the proposed algorithm by showing the bit error rate (BER) performance for different performance evaluation criteria and the constellation diagram between the conventional receivers and our proposed system. A discussion of all simulation results will conclude this chapter.

Chapter 5 will present our conclusions based on the results in Chapter 4 as well as suggestions for future work to be done in this field.

Chapter 2

Digital Communications Over Wireless Channels

2.1 Communication Channel

The purpose of any communication system is to transmit the information signal from its source to the desired destination [1]. The medium over which information is transmitted is called the communication channel. During transmission, the modulated signal is subject to a variety of changes. Some of these changes are deterministic in nature. For example, attenuation, linear and nonlinear distortion is deterministic while additive noise and multipath fading are probabilistic. Since deterministic changes can be considered to be special cases of random changes, in the general sense the mathematical model for a communication channel is stochastic [1][2].

Most communication channels, including wire line and wireless channels, may be generally characterized as a band limited linear filter [2]. Hence, we start by exploring the ways in which the channels can affect signals. This is followed by a detailed discussion of the radio channel.

2.2 Distortion

Any change to a signal caused by the channel other than delaying it or multiplying it by a constant is considered to be distortion. Distortion affects the spectrum of the transmitted signal. Linear distortion does not affect the bandwidth of the transmitted signal while nonlinear distortion does. Some nonlinear forms of distortion are significant at the higher transmission frequencies.

Linear distortion can cause problems in pulse transmission systems that are used in pulse modulation or in digital communication [3]. This distortion is characterized by time dispersion (spreading), due either to multipath or to the limited bandwidth of the channel. For now, we look at the effect that can be readily characterized by the transfer function of the channel. The channel can be characterized by a transfer function of the form [3];

$$H(f) = A(f)e^{-j\theta(f)} \tag{2.1}$$

where $A(f)$ is the frequency-dependent amplitude response of the channel, and the frequency-dependent phase response is $\theta(f)$.

Distortion arises from these two frequency dependent quantities as follows: If $A(f)$ is not a constant, we have what is known as amplitude distortion; if $\theta(f)$ is not linear in f, we have phase distortion.

2.3 Wireless Communication Channel

In broad terms, a radio channel is any channel through which RF energy passes to complete a transmit / receive circuit. In the channel, or environment, characteristics of a transmitted electromagnetic wave are altered by such phenomena as fading, interference, propagation (path) loss, etc. In addition, there are three types of spreading dimensions that results the multipath propagation. These are the delay spread (time scale), Doppler spread (frequency scale) and angle spread (space scale) [3][4]. The mean path loss, long-term fading, short-term fading as well as Doppler frequency shift, delay spread, and angle spread are the main channel effects and they will be discussed in the next section. Moreover, the sections below will also show the graphical representation of the channel model.

2.3.1 Path Loss

A channel can be modeled as a free space or a multipath environment. A free space environment is an environment in which no multipath and no obstructions exist. The mean path loss is a range dependent and changes very slowly even for fast mobiles. The equation (2.2) governing the propagation path loss in a free space environment is known as the Friis Transmission Formula, and is given by

$$P_r = P_t\left(\frac{\lambda}{4\pi d}\right)^2 G_t G_r \qquad (2.2)$$

where P_t and P_r are the transmitted and received powers, respectively, λ is the wavelength, G_t, G_r are the power gains of the transmit and receive antennas, respectively, and d is the propagation distance (range).

2.3.2 Fading

In addition to the path loss, if the received signal exhibits fluctuations in signal level, then this is called fading [3][4][5][6]. In a fading channel, the received strength of the transmitted signal fluctuates around some mean that is range dependent. The received power varies and this variation is made up of two types of fading: long-term fading and short-term fading, which are described as follows:

$$\alpha(t) = \alpha_s(t)\alpha_r(t) \tag{2.1}$$

Where $\alpha_s(t)$ represents the long-term fading and $\alpha_r(t)$ represents the short term fading.

A) Long Term Fading

Long term fading is caused by shadowing effects of buildings or natural features and sometimes it is called a shadowing fading. The long-term fading is determined by the local mean of the short term fading [4][5] and it represents the envelope of the signal level $\alpha_s(t)$ in equation (2.1). This distribution is influenced by the antenna heights, operating frequency, and specific type of environment.

B) Short Term Fading

Short-term fading corresponds to the rapid fluctuations of the received signal in space compared with the long-term fading [5]. It is caused by the scattering of the signal off objects near the moving mobile. If there is no direct path presented, the envelope of the received signal has a Rayleigh density function. Otherwise, it has a Rician distribution.

C) Frequency Non-Selective Fading, Frequency Selective Fading and Space Selective Fading (Angle Spread)

In frequency nonselective fading or flat fading, a channel is said to experience flat fading if all frequencies of interest propagate in the same manner through the channel. That is, all frequencies of interest are equally faded. It represents the envelope or the complex amplitude of the fading channel having a Rayleigh density function distribution with one path delay spread ($\tau_m \approx 0$), where τ_m is the channel's delay spread. In other words, if the bandwidth of the information signal is much less than the coherence bandwidth of the faded channels, the channel is said to be a flat fading or frequency-nonselective channel. On the other hand, if the delay spread is significant ($\tau_m \neq 0$), the signal can be resolved into several time-shifted and scaled versions of the transmitted signal at the receiver. Thus, the channel can be modeled as a discrete filter and the magnitude and phase of the fading depends on the input frequency. Therefore this is called frequency-selective fading [4][6]. Again it can be characterized in terms of the coherence bandwidth, where if we have small ratio of coherence channel bandwidth to signal bandwidth, we have frequency selective fading.

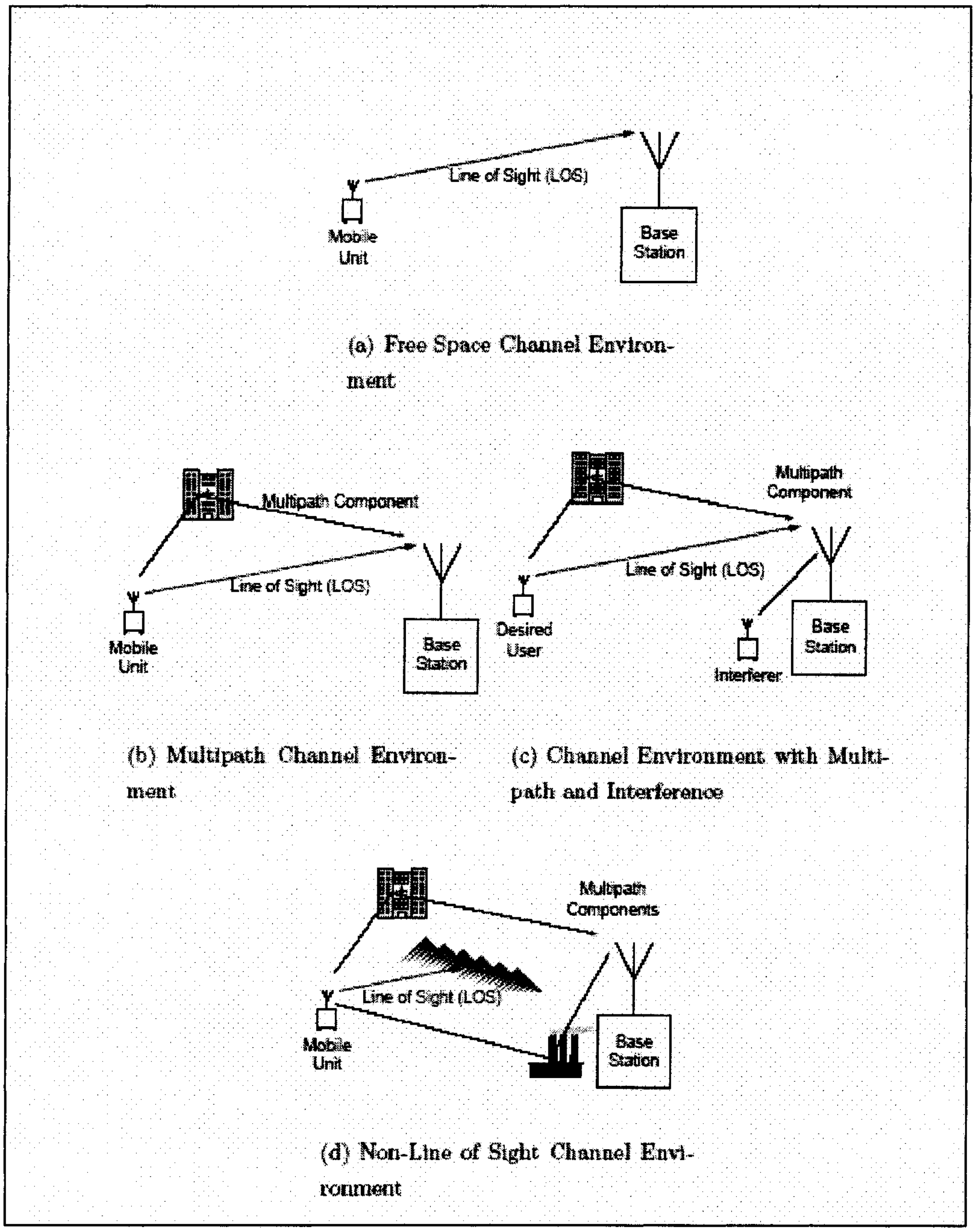
Space selective fading [5] arises from the angle spread, which means that the signal amplitude depends on the spatial location of the antenna. Angle spread at the receiver refers to the spread of angles of arrival of the multipath components at the antenna array and it is characterized by the coherence distance. The coherence distance is defined as the maximum spatial separation for which the channel responses at two antennas remain strongly correlated. Thereby, the shorter coherence distance, the larger the angle spread.

D) Slow Fading and Fast Fading (Doppler Spread)

The fading rate is a measure of how fast the channel changes in time [4][6]. This measurement is based on the coherence time of the channel. The coherence time is the time separation over which the time varying channel impulse responses remains strongly correlated. Thus, the larger the coherence time, the slower the channel changes. By taking the Fourier transform of the autocorrelation of the impulse channel response, we will find the Doppler spectrum of the channel. The coherence time is inversely proportional to the Doppler spread; the Doppler spread is defined as the range of frequencies of the Doppler power spectrum. The Doppler power spectrum is given by:

$$S(f) = \frac{3\sigma}{2\pi f_m}\left[1-\left(\frac{f-f_c}{f_m}\right)^2\right]^{-1/2}, f_c - f_m < f < f_c + f_m \qquad (2.2)$$

Where $f_m = v / \lambda$ is the maximum Doppler shift, where v is the mobile velocity, λ is the wavelength, and f is the carrier frequency. In figure 2.1, we see the different scenarios that can exist between the base station and mobile station which give rise to fading conditions. Figure 2.1(a) shows only one path between the source and destination therefore can be classified as free space. Multipath propagation is shown in Figure 2.1(b) which yields either flat fading or frequency selective fading (multipath) depending on the delay between the paths. Since there is a strong line of sight component, the envelope of the received signal will have a Ricean distribution. Figure 2.1(c) shows the same channel as 2.1(b) with an interference component, which is typical of wireless communication networks. Figure 2.1(d) shows Rayleigh fading since we don't have a line of sight component arriving at the receiver.



(a) Free Space Channel Environment

(b) Multipath Channel Environment

(c) Channel Environment with Multipath and Interference

(d) Non-Line of Sight Channel Environment

Figure 2.1: Channel Models

2.4 Channel Modeling

To enhance deployment, optimization and maintenance of wireless systems, channel models have been developed to analyze propagation and transmit/receive effects in a channel. Channel models are analytical tools that give system designers insight to the performance of a proposed system in a given transmission environment. In this section we will present our proposed channel model.

2.4.1 Proposed Channel Model

The early classic channel models, which were developed for narrowband transmission systems, provide only information about signal amplitude level distribution and Doppler shifts of the receive signals. These models have their origins in the early days of cellular radio, when wideband digital modulation techniques were not readily available. As wireless system becomes more complex and more accurate models are required, additional concepts, such as time delay spread, were incorporated into channel models.

In our thesis, the channel model that we will demonstrate is slow flat fading with a Rayleigh distribution. In terms of flat fading, we are only concerned with a signal whose transmission bandwidth is small compared to the coherence bandwidth of the channel. We have chosen the flat fading for the it's ease of implementation in the simulation of our proposed system.

The general multipath channel model is shown in figure 2.2 [10]. The model for the fading channel can be characterized by the impulse response:

$$h(t) = \sum_i a_i(t)\delta(t - T_i)) \qquad (2.3)$$

Where a_i (t) is a set of time varying independent complex Gaussian random variables and T_i (t) is a set of time-varying random delays is called the multipath delay spread of the channel for i=1,2…N, where N is number of multipath components. In our case, since there is only one resolvable multipath component in flat fading channel, (2.3) becomes:

$$h(t) = a(t)\delta(t) \tag{2.4}$$

The spectrum of the set of path weights, a_i (t), is approximately limited to the frequencies between $\pm B_d$, where B_d is the single-sided bandwidth of the fading process of the channel and it is called the Doppler spread of the channel [15]. As we will see later, the $a_i(t)$ is determined by using an autoregressive model.

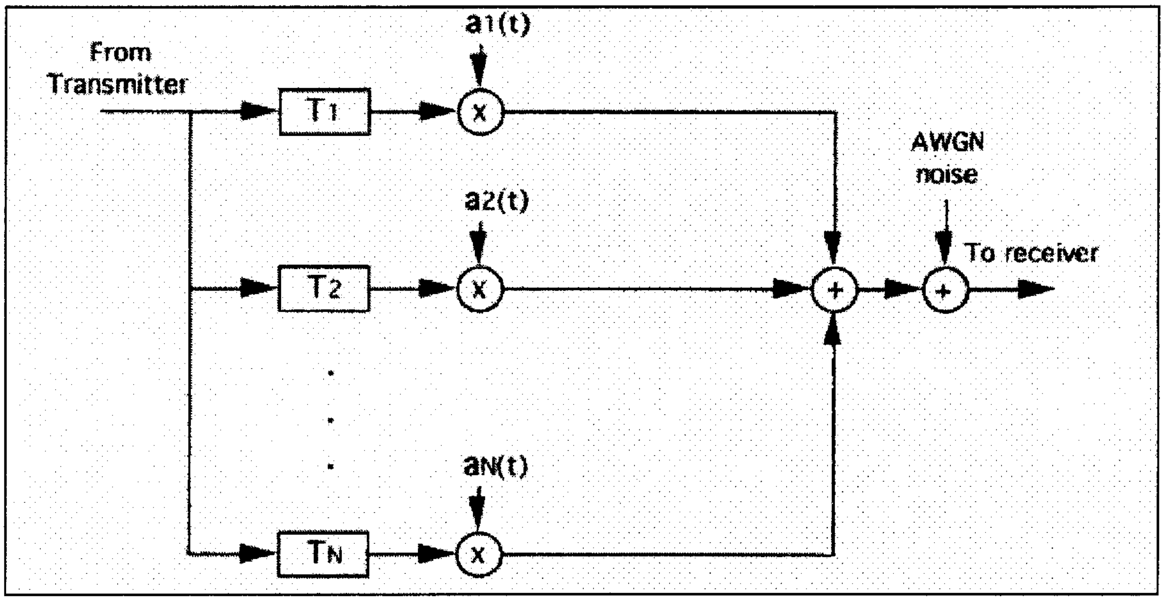


Figure 2.2: Multipath channel model [9]

In the next section, we will show the channel model that is used in the simulations in Chapter 4.

2.4.2 Doppler Spread Structure

From the impulse response of flat fading channel of (2.4), we need to find the complex channel gain, $a(t)$. To determine this, an auto regressive (AR) model is used as in [10]. The auto regressive model is IIR filter structure with an M taps. In order to uniquely define the AR model for order M, depicted in figure 2.3, we need to specify two sets of model parameters:

1. The AR coefficients $w_1, w_2, .., w_M$ (The output of the AR model)
2. The variance σ_v^2 of the Gaussian input v (n) used as excitation.

Our first task is to know what those AR taps values are. If we apply the Yule-Walker formula [9], we can obtain easily the taps of the filters. The Yule formula state as in equation (2.5)

$$\begin{bmatrix} r(0) & r(1) & \cdots & r(M-1) \\ r^*(1) & r(0) & \cdots & r(M-1) \\ \vdots & \vdots & \ddots & \vdots \\ r^*(M-1) & r^*(M-2) & \cdots & r(0) \end{bmatrix} \begin{bmatrix} w_1 \\ w_2 \\ \vdots \\ w_M \end{bmatrix} = \begin{bmatrix} r^*(1) \\ r^*(2) \\ \vdots \\ r^*(M) \end{bmatrix} \tag{2.5}$$

Where M is number of taps filters, r(i) is an autocorrelation functions of the stochastic input signals, and W_i is the taps coefficients of the filter.

We may express the Yule-Walker equations in compact matrix form as in equation (2.6),

$$\text{Rw} = \text{r} \tag{2.6}$$

Therefore, the taps (W) can be easily be determined by inversing the matrix R (assuming that the correlation matrix R is nonsingular) as in equation (2.7)

$$w = R^{-1}r \tag{2.7}$$

Where R^{-1} is the inverse of the matrix R, and the vector W is defined by (2.8)

$$w = [w_1 \quad w_2 \quad \cdots \quad w_M]^T \tag{2.8}$$

In order to find the filter taps (w_i), we require need to find the equation for the autocorrelation function of the fading coefficients r(k). According to [9], the auto correlation function can be found using the Bessel function as a common assumption of a U-shaped fading spectrum corresponding to isotropic scattering. The autocorrelation equation is given in (2.9) as follows,

$$R(\tau) = \sigma^2 J_0(2\pi B_D \tau) \tag{2.9}$$

where B_D is the Doppler spread and J_o is the 0^{th} order Bessel function of the first kind. The i^{th} and j^{th} elements in the R matrix are given by in equation (2.10)

$$R_{ij} = \sigma^2 J_0(2\pi f_D TM|i-j|) \tag{2.10}$$

The variance of the Gaussian input with zero mean, v(n), is calculated by finding the autocorrelation of the input value of the AR filter. Thus, the formula for the variance of the Gaussian input is as follows:

$$\sigma_v^2 = \sum_{k=0}^{M} w_k^* r(k) \tag{2.11}$$

Finally, the overall structure of our channel model is given in figure 2.3.

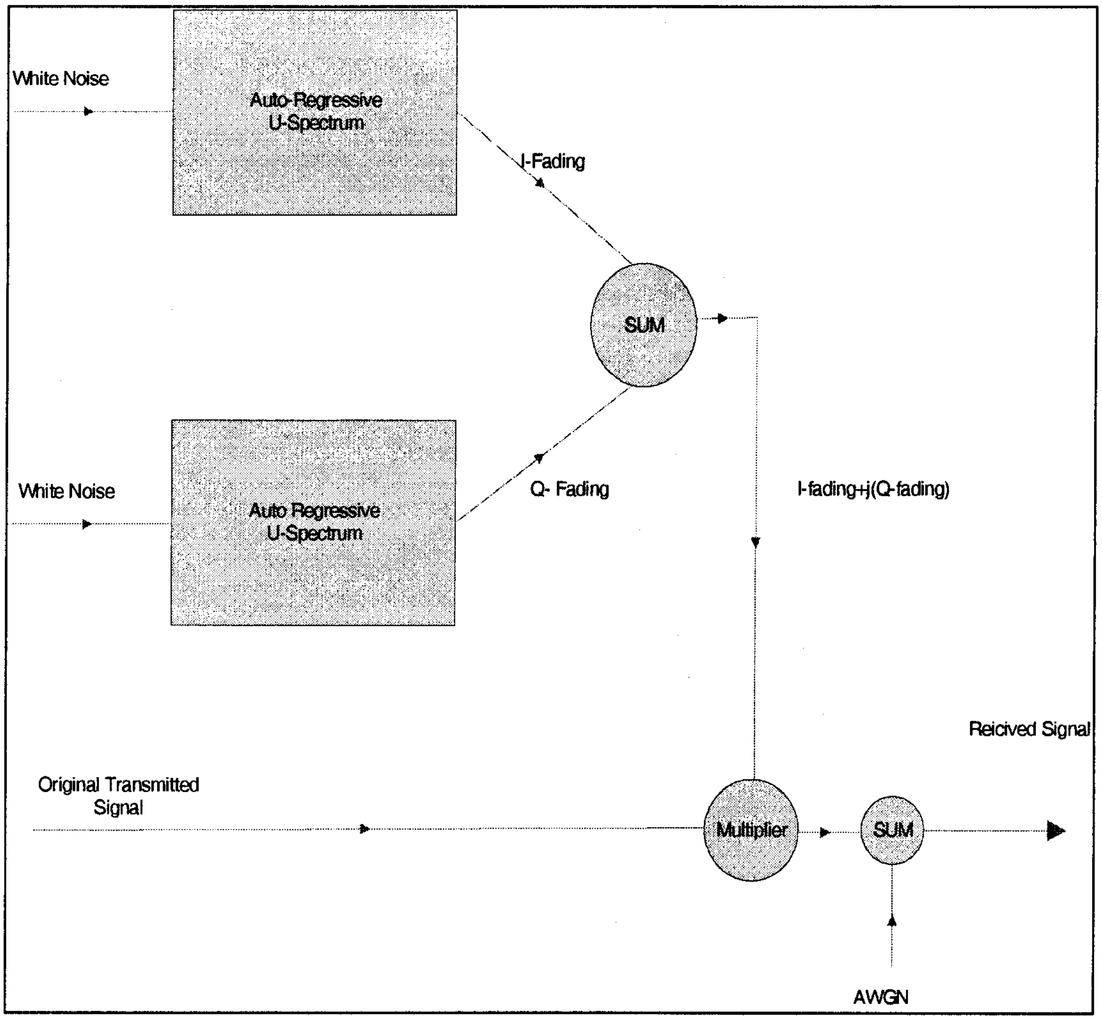


Figure 2.3: Proposed Channel Model

Furthermore, we have used the interpolation technique to interpolate both the in-phase and the quadrate components of the Doppler coefficients before multiplying to the transmitted signal. The reason for doing this is to be able to produce slow flat fading without increasing the number of filter taps in the AR model. This will improve the simulation speed.

2.5 Channel Estimation

Channel estimation is necessary in digital mobile communication to combat rapid fading [18][19]. Fading negatively affects the overall performance by degrading the bit error rate (BER), producing an irreducible BER or error floor as well as making difficult the use of multilevel formats, with their greater spectral efficiency. Channel estimation can improve the degradation and remove limitations on multilevel formats. In order to estimate the channel gain, a known pilot signal can be used to observe how it is affected by its transmission over the channel [11][12][13][15]. There are two major pilot signals systems used in the digital communication, pilot tone (PT) [11] [12] and pilot symbol assisted modulation (PSAM) [13][14][21].

In PT, the tone provides the receiver with an explicit amplitude and phase reference for detection and thereby suppresses the error floor. But the problem with the PT is to find the proper location within the channel spectrum so that it provides a good channel estimate for the entire spectrum without interfering with the desired signal. To overcome this problem, transparent tone in band TTIB is used. However, this technique requires complex signal processing and increases the peak to average power ratio.

The PSAM transmitter generates known pilot symbols. The receiver extracts the pilot symbols and uses them to derive its amplitude and phase reference. Using PSAM will not change the transmitted pulse shape as in PT and the processing at the transmitter and receiver is less complex than using the TTIB. Figure 2.4 shows the block diagram of the

PSAM system. It is clear from the figure 2.4, after getting the output from the matched filter, the receiver reference branch (lower branch) interpolates the channel measurement provided by the pilot symbols to obtain an amplitude and phase reference to the data branch (upper branch). As in the PT scheme, it will suppress noise floor. However, the drawback of using this system is the delay associated with filling the buffer of the interpolator, which try to the fill the gaps between the pilot samples with a certain amount at the output of the decimator device. In addition, as the size of the interpolator increases, which is sometimes necessary to produce a good channel estimate, the delay increases.

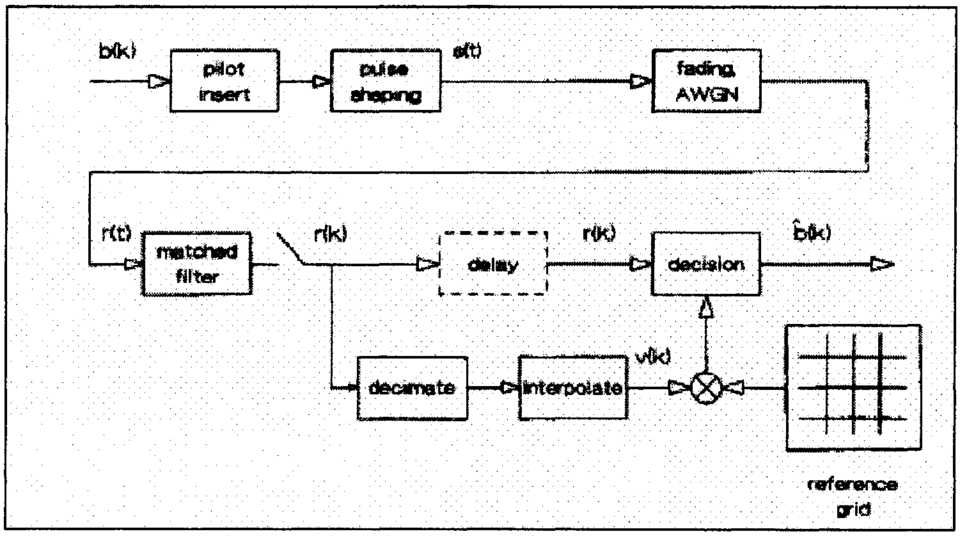


Figure 2.4: The PSAM channel estimation

In our proposed system, we will use the PSAM system with some modification, rather using the interpolator that introduce a time delay to have some performance on noise, an estimated channel filter will be used instead as in figure 2.5. After receiving the pilot signal signature waveform that is orthogonal to the data signal signature waveform from each antenna, we will compare the received pilot symbol samples with known pilot symbols and therefore one can obtain an estimate the gain and the phase changes of the channel. Furthermore, to reduce the background noise in the received pilot symbols, we use a filter that has a cutoff equal to Doppler frequency. The advantages of our channel estimation system are summarized as follows;

- No transmitting pulse shaping required (no increase in the peak-to-average power ratio).
- Simpler signal processing than PT method.
- Its structure has less complexity than the conventional pilot channel detection for spatial processing purpose.

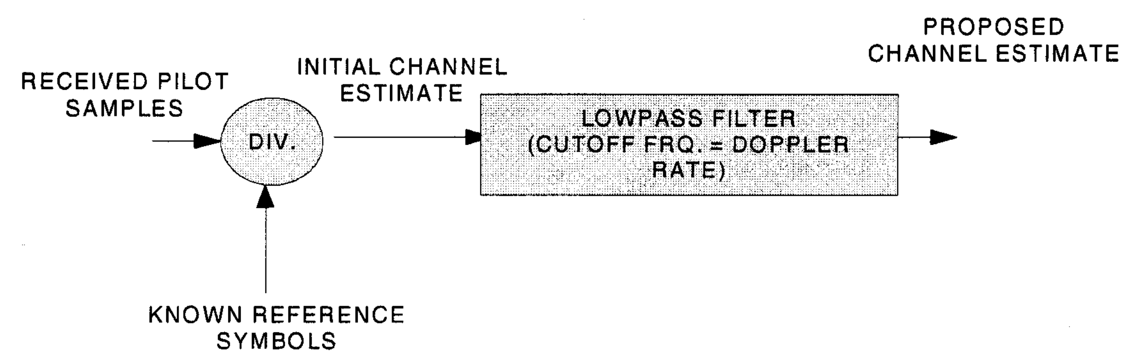


Figure 2.6: Pilot channel estimation

Figure 2.5: Pilot channel estimation

2.6 Binary Modulation Schemes

Binary shift keying (PSK) was developed during the early days of the deep-space program; PSK is now widely used in both military and commercial communications systems [1][4]. The general expression for PSK over one signaling interval is in equation (2.12),

$$s_i(t) = \sqrt{\frac{2P}{T}} \cos[\omega_o t + \phi_i(t)] \qquad \begin{matrix} 0 \le t \le T \\ i = 1, \cdots, M \end{matrix} \qquad (2.12)$$

Where T is the signaling interval, p is the energy per symbol, w_0 is the angular frequency in rads/sec and $\phi_i(t)$, is the information phase which has M discrete values, typically given by

$$\phi_i(t) = \frac{2\pi i}{M} \qquad i = 1, \cdots, M$$

For the binary PSK (BPSK), M=2. In BPSK, modulation, the modulating data signal shifts the phase of the waveform, $s_i(t)$, to one of two states,

$$\phi_i = \begin{cases} 0 & bit = "1" \\ \pi & bit = "0" \end{cases}$$

2.6.1 Binary Phase Shift Keying in AWGN Channels

In an additive white Gaussian noise (AWGN) channel model, we assume no distortion or other effects other than the addition of white Gaussian noise. White Gaussian noise is a model for the thermal noise generated by random electron movement in the receiver.

The equation for the received signal in an AWGN channel is

$$r(t) = s_i(t) + n(t) \qquad (2.13)$$

Where $s_i(t)$ is the transmitted signal (i=1,2), r(t) is the received signal, and n(t) is a zero-mean wide-sense stationary random process with power spectral density $S(j\omega) = No/2$. The probability of receiving a bit in error is given by (2.14);

$$P_b = Q(\sqrt{\frac{2E_b}{N_o}}) \tag{2.14}$$

Where E_b/N_o is the bit energy to noise spectral density ratio.

2.6.2 Binary Phase Shift Keying in Rayleigh Fading Channels

In this section, we analyze the performance of a binary phase shift keying (BPSK) in terms of probability of error in flat Rayleigh fading (frequency nonselective) channel [4][5]. The signal bandwidth W is assumed to be much smaller than the coherence bandwidth of the faded channel.

The channel impulse response is given in equation (2.15),

$$h(\tau;t) = a(t)\delta(t-\tau_0(t)) \tag{2.15}$$

Where $|a(t)|$ has a Rayleigh distribution at any instant in time.

Since our channel is a slowing faded channel, the time variation of $a(t)$ and $\tau_o(t)$ are very slow compared to the symbol interval. So our channel impulse response over one signaling interval can be appointed by:

$$h(\tau;t) = h(\tau) = \alpha\delta(t-\tau_0)$$

where $|a|$ is the magnitude of the complex channel gain and is Rayleigh distributed, i.e.,

$$p(\alpha) = \begin{Bmatrix} \frac{\alpha}{\sigma_\alpha^2} e^{-\alpha/2\sigma_\alpha^2}, & \alpha > 0 \\ 0, & otherwise \end{Bmatrix}, \ldots \sigma_\alpha^2 = E(\alpha^2)/2 \qquad (2.16)$$

Now let the received signal in the interval 0 <t < T be

$$r(t) = \alpha\, s(t) + n(t) \qquad (2.17)$$

Since the transmitted signal is BPSK, two possible signals can be transmitted (m=0,1) over given signaling interval. The received signal in (2.17) can be rewritten as:

$$r = \alpha\sqrt{E_b}\cos m\pi + n, \qquad m=0,1 \qquad (2.18)$$

The probability of error is given by:

$$P_2(\alpha) = Q\left(\sqrt{\frac{2\alpha^2 E_b}{N_o}}\right) \qquad (2.19)$$

Let gamma has a chi-squared distribution and it is given by:

$$\gamma = \frac{\alpha^2 E_b}{N_o}$$

But this probability is for a fixed value of α. To find the probability of error of averaged over all possible values of α, we should perform the following:

$$P_2 = \int_0^{\infty} P_2(\gamma)p(\gamma)d\gamma \qquad (2.20)$$

Where $p(\gamma)$ is the distribution of the bit energy to noise spectral density ratio. Since α is a Rayleigh distributed, α^2 has a chi-square probability distribution with 2 degrees of freedom [4]. If the envelope has a Rayleigh distribution, the bit energy (which is proportional to the envelope squared) has a chi-squared distribution. Therefore we can find the performance of BPSK in Rayleigh fading channel, assuming no phase recovery errors to be:

$$P_2 = \frac{1}{2}\left[1 - \sqrt{\frac{\tilde{\rho}}{1+\tilde{\rho}}}\right] \qquad (2.21)$$

Given that $\tilde{\rho} = \frac{E_b}{N_0}E(\alpha^2)$, where a is Rayleigh-distributed and E(α^2) is simply the average value of a^2. For large values of $\tilde{\rho}$, the probability of error for the binary of antipodal signal will be then

$$P_2 \approx \frac{1}{4\tilde{\rho}} \qquad (2.22)$$

2.6.3 Binary Phase Shift Keying in Rayleigh Fading Channels with Diversity

There are several types of diversity such as frequency, time, polarization and space diversity [4][5][15][16][17]. In our project, we are interested in space diversity. The space diversity can be obtained by using a multiple receive antennas, and/or multiple antenna. The receive antennas must be spaced sufficiently far apart so that the multipath components in the signal have significantly different propagation paths. Usually the space between the received antennas depends at least half wavelength in order to obtain signals that fade independently and sometimes depends on the location of the mobile user from the base station.

Moreover, there are different techniques to combine the desired received signals from different antennas. The most well known technique called the maximal ratio combining, which is the modification of the equal-gain combining. Those complex methods combine the D independent fading receiving signals. The equal-gain combiner is to sum the D received signal after they have been compensated with the phase offset of the D independent paths. If the received signal was estimated in terms of their power level for each D independent received path and multiply by the phase correction (square-root of the power level) before it goes to the detector, this is called maximal ratio combiner. More explaining of space combing will be discussed in chapter 3.

The probability of bit error of the BPSK in Rayleigh distribution with diversity is relatively similar without diversity. With diversity the slope BER curve becomes steeper as we increase the number of receive antennas.

The bit error rate will be of the form

$$\frac{K_D}{\tilde{\rho}_b^D}$$

where K_D is a constant that depends on D and $\tilde{\rho}_b$ is obtained from equation (2.21). Therefore the error probability will have a decreasing exponential shape by increasing the D paths. For the BPSK modulation, the probability of error will be with diversity,

$$P_2 = \frac{K_D}{(4\tilde{\rho}_b)^D}, \qquad \tilde{\rho}_b >> 10\text{dB} \tag{2.23}$$

and K_D is defined as

$$K_D = \frac{(2D-1)}{D!(D-1)!}$$

2.7 Spread Spectrum Communication and Multiple Access Interference

Spread spectrum systems were originally developed and initially used for military communication purpose for the following reasons; (1) hiding the signal at lower power level, (2) its ant-jam properties, (3) allowing multiple users to transmit in the same bandwidth simultaneously. These days, spread spectrum signals are being used to provide reliable communication in a variety of commercial applications, including mobile vehicular communications and interoffice wireless communications [20][22][23][24].

Spread spectrum allows several users to transmit their message over the same frequency range. This is achieved by assigning a different codes to each user, therefore this system is called code division multiple access (CDMA) [23].

In direct sequence code division multiple access (DS-CDMA), each user is assigned a unique spreading code that has low cross-correlation values with other spreading codes. This is the reason why a receiver, which has knowledge about the code of the intended transmitter, is capable of selecting the desired signal.

We could summarize the benefits of using the code division multiple access (CDMA). As the signal is spread over a large frequency-band, the power spectral density is getting very small, so other communications systems do not suffer from this kind of communications. Other than that, in all situations the whole frequency-spectrum is used, so there is low possibility for a narrowband interferes to degrade the system. In addition, it gives high security due to unknown random codes. This means that it is difficult to detect the message of other users. Moreover, it allows random access. Users can start their transmission at any arbitrary time as in FDMA.

The main parameter in spread spectrum systems is the processing gain: the ratio of transmission and information bandwidth:

$$G_p = \frac{BW_c}{BW_b} \tag{2.24}$$

which is also called the "spreading factor". The processing gain determines the number of users that can be allowed in a system, the difficulty to jam or detect a signal. Therefore, to increase the number of users that have low cross correlation among them, it is advantageous of spread spectrum systems to have a processing gain as high as possible.

There are two basic types of spread spectrum signal for digital communications-namely, direct-sequence spread spectrum (DSSS) [24][22] and frequency-hopped spread spectrum (FHSS) [21][22]. Since in our thesis paper we are using the BPSK as a modulation stream, direct-sequence spread spectrum is a better choice to be used in our project.

Direct sequence is the best-known spread spectrum technique [24]. The data signal is multiplied by a pseudonoise code (PN code). A PN code is a sequence of chips valued -1 and 1 (polar) or 0 and 1 (non-polar) and has noise-like properties. Since each user has it unique code as well as users has low cross-correlation values among their codes, jamming or detecting a data message will be difficult to achieve. A usual way to create a PN code is by means of at least one shift-register.

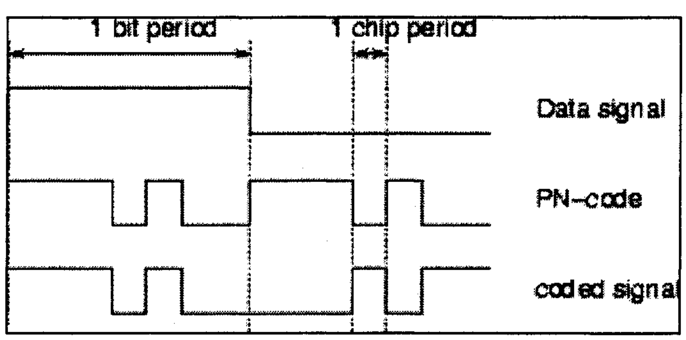


Figure 2.6 Direct-sequence spreading in time domain

As can be seen in the figure 2.6, in time domain, the original data signal for each user is now multiplied by a factor chip period. As a result, each coded signal carrying a single data signal.

Figure 2.7 in the frequency domain, the bandwidth of the data signal is now multiplied by a factor (1/chip period).

1. The power content however stays the same, with the result that the power spectral density is reduced
2. Since the code existed of +1s and -1s, multiplying the spread signal by the PN code completely removes the code from the signal and the original data-signal is left.
3. At the receiver, the spread signal is once again multiplied by the same PN sequence. This despreads the desired signal while at the same time spreads a narrowband interference signal.

However, there are two main problems with applying direct sequence spreading. One of them is called the multiple access interference [13][20][21][22]. As we mention earlier in this section, the number of users will have low code correlation among them. By fixing the spreading gain for a certain value and by increasing the number of users within cell, the orthogonally will be lost between users codes and each user will behave as an interference to the other user. Let us explain it in mathematical expression. Assume we have N users in our system trying to transmit $s_i(t)$ information signal using a signature waveform $c_i(t)$ for each user i. Then at the receiver side, the received signal before despreading is given by:

$$r(t) = \sum_{i=1}^{N} s_i(t)c_i(t) + n(t) \tag{2.25}$$

where n(t) is AWGN with spectral density equal $N_o/2$.

In order to reconstruct the desired user, we have again to multiply the same code signal that we assigned to the desired user at the transmitter to the received signal. Then information signal of the desired user is given by;

$$u_i = b_i A_i T_b + \sum_{j \neq i}^{N} b_j A_j T_b \rho_{ij} + N_i \tag{2.26}$$

where b_i is the modulation bit, A_i is amplitude assigned to each user, T_b is the signaling interval bit, Ni is noise signal for each user i and cross correlation between signature waveform between users (ρ_{ij}) is given by;

$$\rho_{i,j} = \begin{cases} 1 & i = j \\ \frac{1}{T_b} \int_{Tb} ci(t)cj(t)dt & i \neq j \end{cases} \tag{2.27}$$

As it can be seen from equation 2.26, the first part of the equation represents the desired data signal that we are interested to obtain and next part represents the multiple access interference. The multiple access interference depends on the cross correlation of the user code signals. If codes are orthogonal between users (ρ_{ij}=0), then we will not suffer from MAI. But in reality, by increasing the number of users sharing the radio spectrum, the value of the cross correlation (ρ_{ij}) will increase and contribute to make the detector to take an incorrect decision.

Another problem encountered by direct sequence spread spectrum systems is the Near-Far effect [22], which is illustrated in figure 2.8. This effect is present when an interfering transmitter is much closer to the receiver than the intended transmitter. The MAI component from the receiver is significant in power compared to the desired signal component. The result is that proper data detection is less reliable. This problem could be solved by using power control. In our work, we assume that near-far problem is eliminated using perfect power control.

This research will try to find a method to suppress as much as possible the multiple access interference and to reduce the error rate by spatial and temporal signal processing structure as will be shown in Chapter 3.

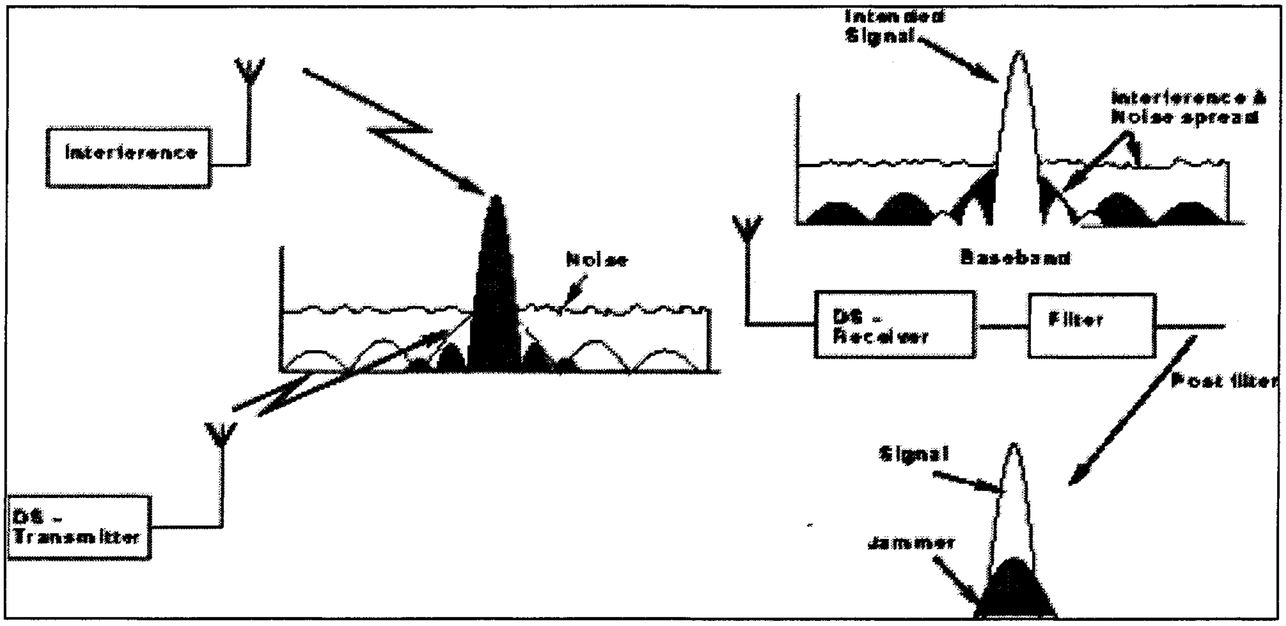


Figure 2.7: DS-concept, before and after dispreading in frequency domain

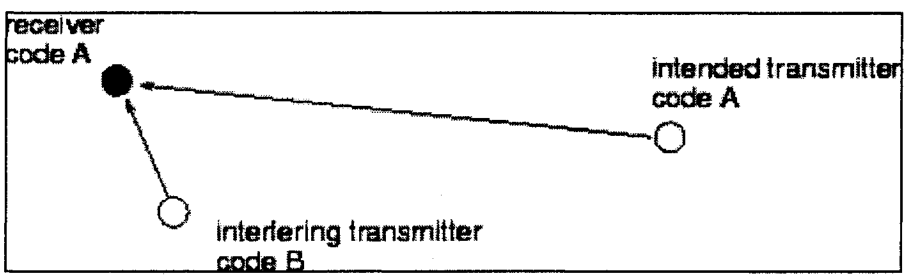


Figure 2.8: Near-far effect

Chapter 3

Space-Time Receivers for Wireless Communications

One of the biggest challenges of using CDMA in wireless communications is multiple access interference (MAI) caused by co-channel users. The signature waveforms of other users are not always orthogonal to the desired user's signature waveforms. Different techniques can be used to overcome these impediments; these include multiuser detection and antenna array processing. To combat and exploit the inherent diversity caused by multipath, RAKE matched filtering (or a bank of correlators or integrators) is used to coherently combine resolvable multipath replicas of the desired signal. Multiuser detectors (MUDs) [23] reduce the effects of MAI by exploiting its well-known structure. Those time domain signal processing techniques can be enhanced using antenna array processing [24]-[26]. Space domain techniques can further combat multipath fading through diversity combining and provide signal to interference- and noise ratio (SINR) gain by coherently combining the signals across antennas.

For CDMA systems, a superior alternative to time-only multiuser detection or single antenna array is joint space-time (ST) multiuser detection, which considers both the spatial and temporal aspect of the information signal. ST-MUDs have been shown to provide significant performance improvements over single-user detectors, which detect the desired user with regard to the MAI over multiuser detectors with a multiple antennas [56]. Optimum maximum likelihood space-time multiuser detection was suggested for flat fading [57]. Because the optimum detector's implementation is prohibitively complex for a large number of users, linear space-time multiuser detectors with lower complexity have been explored. These linear detectors are all based on the decorrelating detector, which is combined with the multiple antenna arrays, the linear detectors null out the interfering users.

In this chapter, we begin with an introduction of multiple antenna receivers, while examining the benefits obtained by using multiple antennas at the receiver. Next, we will explain in more detail several multi-user detectors, which employ temporal signal processing. Finally, previous research in (ST-MUD) receivers for wireless communications will be discussed. Overall, it has been shown that ST-MUDs provide significant performance improvements over conventional (space-only, time-only, and/or single antenna) detectors [58].

3.1 Multiple Antenna Receivers

Multiple antennas at the receiver provide space diversity. Other diversity methods include frequency diversity, time diversity, polarization diversity and angle-of-arrival AOA diversity [24], [25]. In this section we will concentrate on space diversity receivers that use more than one antenna at the receiver.

The key benefit of any diversity type is to receive several replicas of the same information signal transmitted over independently fading channels, so that the probability that all signal components fade simultaneously is reduced considerably. In the uplink, to obtain independently faded signals at different antennas of the receiver, a separation of at least 10 wavelengths is required between adjacent antennas [27] [32].

Multiple antenna systems can provide different kinds of diversity. The most popular are spatial diversity and polarization diversity. In this thesis, we will explore the spatial diversity aspect of multiple antenna receivers. Spatial diversity means that the signal is received by the receiver's antennas that are spatially separated and that the fading on the different paths taken by the signal to reach these antennas is lowly correlated. Furthermore, the correlation of the fading process of different signals received by adjacent antennas depends on the angular spread between the adjacent antennas.

There are three types of combiners used in antenna arrays to improve the signal to interference ratio: selection combiner (SC), maximal ratio combiner (MRC) and adaptive combining (AC).

Selection diversity is based on the power level of all antenna outputs for each user's signal and tries to select the antenna element that has the highest received signal power [30] as shown in Figure 3.1. According to [30], Brennan shows the relationship between the average power as function of diversity order, $P_{av}(L)$, to be given by:

$$P_{av}(L) = P\sum_{k=1}^{L}\frac{1}{k} \tag{3.1}$$

where P is the average power of a single diversity path.

The disadvantage of using this combiner is that it has poor resistance to the near far problem since if an interfering user that has a high average power, SC will select it and ignore the desired user. On the other hand, simplicity of its structure makes it more attractive to researchers. Switched smart antennas are based on this technique [31].

The second combiner combines weighted signals so as to maximize the signal to noise ratio. This is called a *maximal ratio combiner* (MRC) [30]. A MRC system is shown in figure 3.2. The weights are the complex conjugate of the complex channel gain (assuming that all paths have the same transmitted energy). The advantage of MRC is that it performs best when only one transmitter exists with multiple antenna elements (assuming zero interfering signals). According to [30], the maximal ratio diversity combining (MRC) is the optimum diversity combing in the case of coherent detection. In general, when the channel gains are all equal to 1 in the MRC, in other word, each path is received and summed with the other paths without being weighted; the combiner is called equal-gain diversity combining (EGC). But equal gain combining can be used even when the channel gains are not equal. It is sub-optimal in the signal to noise ratio sense. However MRC is optimal in the signal to noise ratio sense, even in the presence of MAI, as long as the MAI is accounted for. The performance of the MRC is measured by the equivalent energy per bit to noise density ratio, $(E_b/N_o)_{eq}$ for the Lth order diversity combining system,

$$\left(\frac{E_b}{N_0}\right) = \sum_{k=1}^{L} |a_k|^2 \frac{E_b}{N_o} \tag{3.2}$$

where $|a_k|$ is the magnitude of the kth path and is Rayleigh distributed and $|a_k|^2$ has a Chi-Square distribution. Hence, the probability density function of $(E_b/N_o)_{eq}$ is shown in [30] to be given by:

$$p\left(\frac{E_b}{N_o}\right) = \frac{1}{(L-1)!\gamma_c}\left(\frac{E_b}{N_o}\right)^{L-1} e^{-\frac{(E_b/N_0)_{eq}}{\gamma_c}} \tag{3.3}$$

where γ_c is the average E_b/N_o per channel and it equals to $\gamma_c = (E_b/N_o)_{eq}/L$. Also the bit error probability $P_b(L)$ can be found to be [30],

$$P_b(L) = (\frac{1-\mu}{2})^L \sum_{k=0}^{L} \binom{L-1+k}{k} \left(\frac{1+\mu}{2}\right)^k \qquad \text{where } \mu = \sqrt{\frac{\gamma_c}{1+\gamma_c}} \tag{3.4}$$

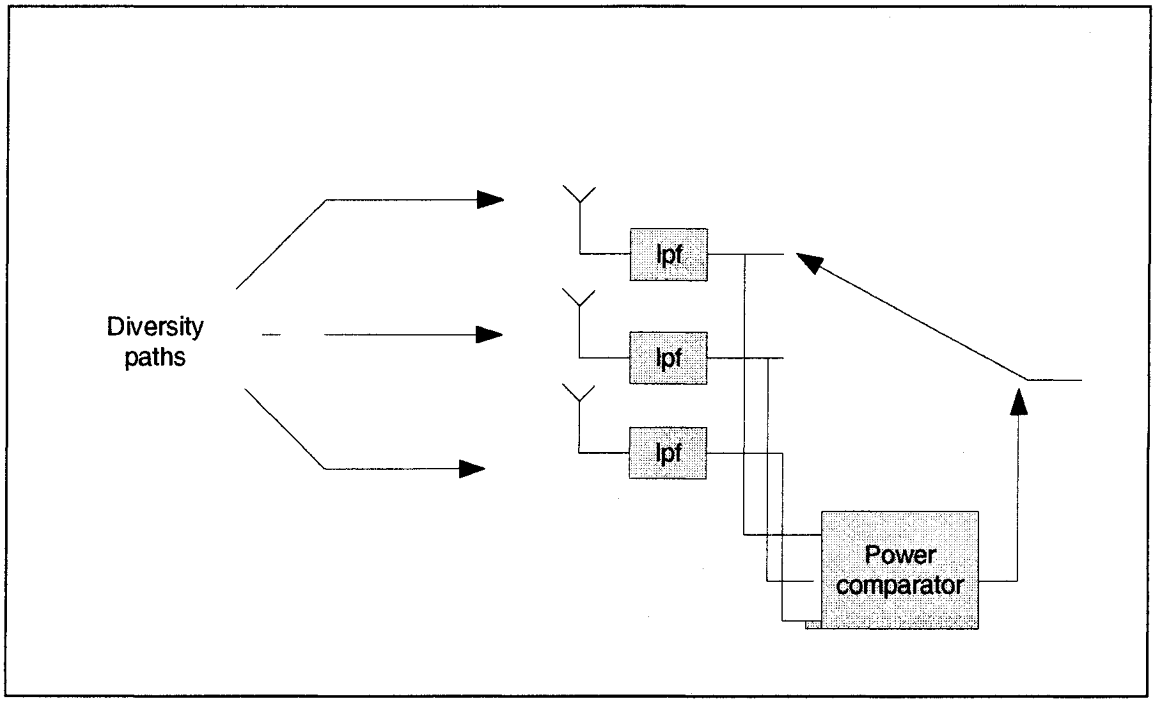


Figure 3.1 Selective combining

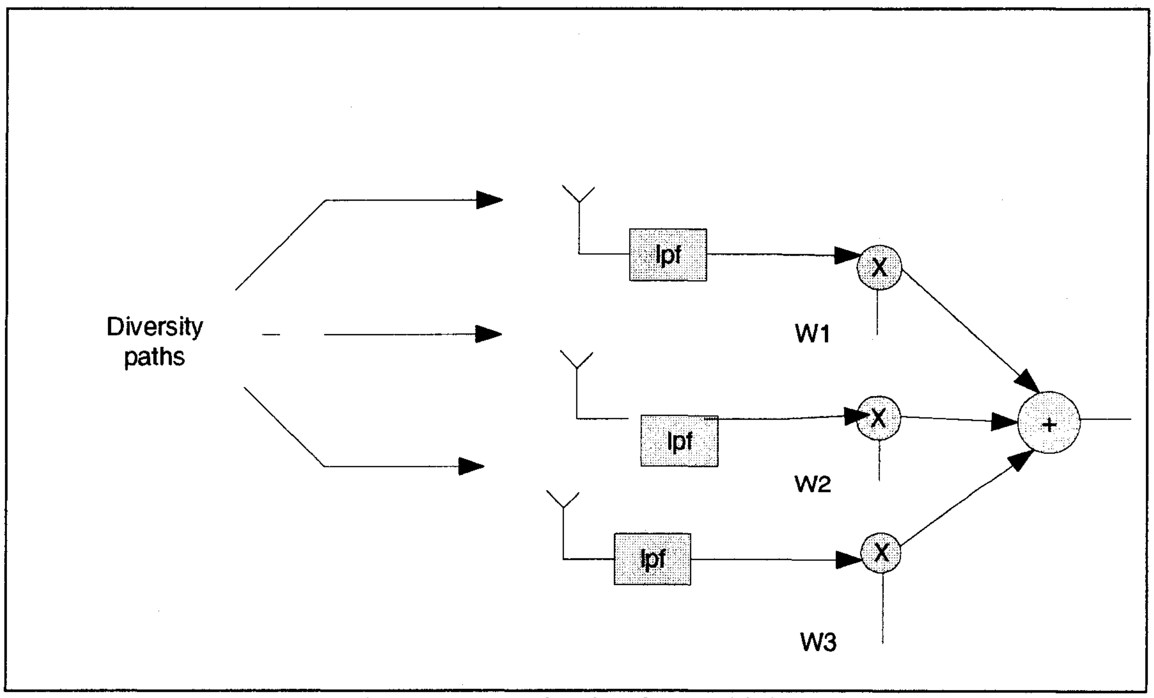


Figure 3.2 Maximal ratio combining

Adaptive combining systems [27][28] provide a higher degree of freedom, since they have the ability to adapt the radiation pattern to the RF signal environment in real time. Adaptive antennas are used to enhance received signals and may also be used to form beams for transmission. A basic block diagram of adaptive combining is shown in figure 3.3.

Since our system uses the channel gain to weight the different signal components, our proposed system uses maximal ratio diversity combining antennas.

In general, antenna arrays change their antenna patterns dynamically either to null out noise and analog interference or to collect multipath vectors. Adaptive array antennas can adjust their pattern to track portable users. In other words, they can direct the main beam toward the pilot signal, while suppressing the beam pattern in the direction of the interferers. In other words, adaptive antenna array systems can customize an appropriate radiation pattern for each individual user. The switch beam that uses the selective diversity systems allows placing the desired user at the maximum of the main lobe, but it also shows this system inability to reject the unwanted signal. Adaptive systems use variants of the steepest descent algorithm i.e. RLS, LMS etc.

In the presence of a low level interference, both the SC and MRC system will provide significant gain over conventional sectored antennas [33]. However, in the high-level interference case, the interference rejection capability of the MRC systems provides significantly improvement over the SC system. Recent studies show that a hybrid combination of SC and MRC will give better results than implementing each combiner alone [34][36].

In order for the adaptive system to maximize the signal to noise plus interference signals ratio (SNIR), they use many adaptive systems using different signal processing algorithms. One of those algorithms is the directional of arrival (DOA) algorithm, which continuously locates and tracks the angle of arrival of the interest signal, by dynamically updating the weights of the antenna elements. The other algorithm is the reference (pilot) signal algorithm.

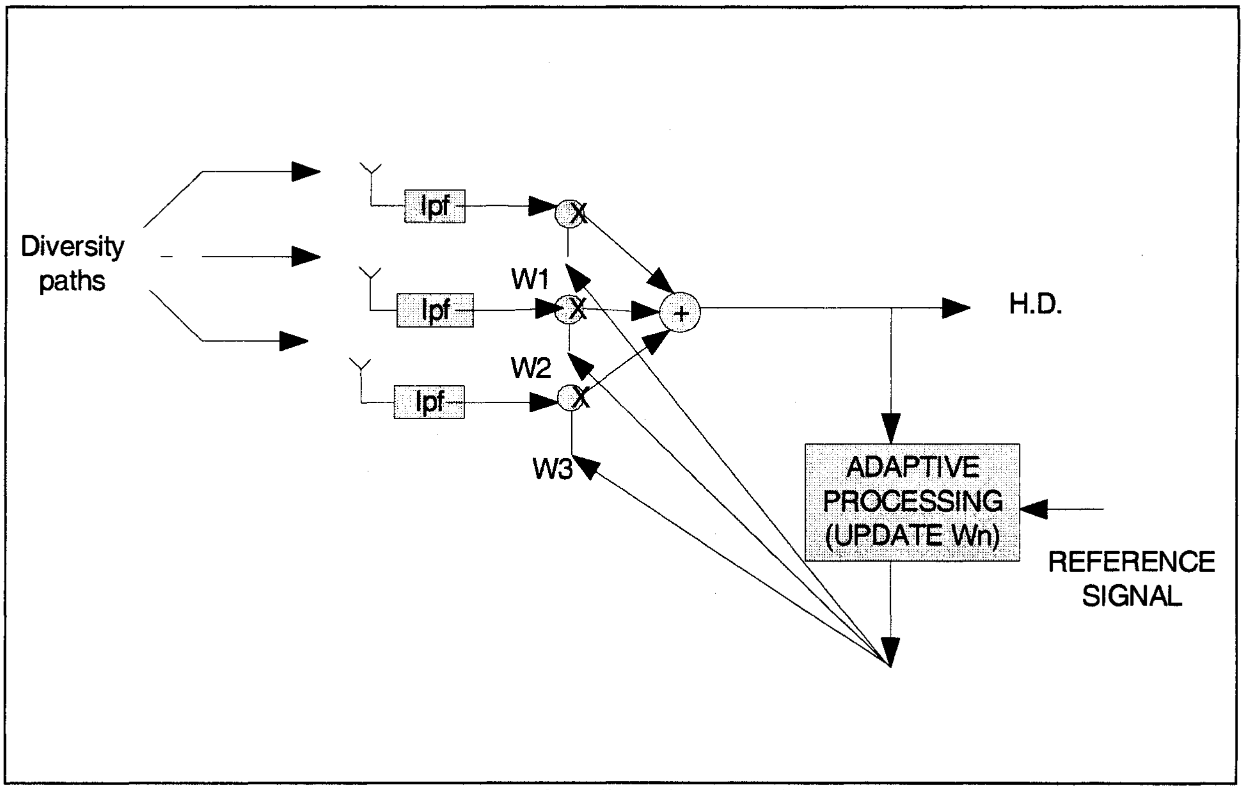


Figure 3.3: Adaptive diversity combining

3.2 Benefits of Multiple Antennas

Multiple antennas offer a broad range of ways to improve wireless system performance by providing greater gain at the base station. Those major benefits can be summarized as following:

A. **Range Extension:** Multiple antennas can increase range by increasing the gain of the base station antenna while keeping the received power at a certain level. Let the received uplink power from the mobile unit at the base station be:

$$P_r = P_t + G_s + G_b - PL \tag{3.8}$$

Where P_r is the power received at the base station, P_t is the power transmitted by the subscriber, G_s is the gain of the subscriber unit antenna, PL is the path loss and G_b is the gain of the base station antenna.

And the path loss is give by

$$PL = \overline{PL} + 10 n_0 \log\left(\frac{d_r}{d_0}\right) + X_\sigma \tag{3.9}$$

Where d_r is the distance between the transmitter and the receiver, d_o is a reference distance which is determined from measurement.

As we observe from the two equations (3.8) & (3.9), by increasing the tolerable path loss, we can increase the reception range, d, of the base station.

B. **Reducing the transmitted power:** For a given range, an increase in receiver gain means a signal will be received at higher power for the same transmitted power. For a system with higher receive antenna gain, we can decrease the transmit power to achieve the same performance as a system with lower receive antenna gain. Also, it will reduce the power management in the mobile unit such as the size of the batteries size and extending their charging period life.

C. **Improve System Capacity:** CDMA systems are power limited. In other words, as more users access the system, the overall bit error rate performance of the system degrades due to increased MAI. The maximum number of simultaneous users is determined by the highest acceptable bit error rate. The points below will help to increase the capacity for more than one antenna:

- CDMA systems are interference limited. Multiple antennas can be used to allow the subscriber to transmit less power for each link. Thus, the multiple user interference is reduced which allow increasing the number of simultaneous targeted users that operates within the same cell.
- Multiple antennas can be used to separate different subscribers partially while they are using the same spectral resource, compared with the conventional antennas. This technique is called Space Division Multiple access (SDMA).
- The crucial reason in the growing interest of multiple antenna systems is the increase of number of subscribers. In densely populated areas, mobile systems are normally interference limited, meaning that the interference

from the other users is the main source of noise in the system. This means that the signal to interference ratio (SIR) is much larger than the signal to noise ratio (SNR). In general, multiple antennas will increase the SIR by simultaneously increasing the useful received signal level and lowering the interference level.

3.3 Previous research in advanced studies in antenna array

The previous research in antenna array processing and still currently demanding in the wireless field is the smart antennas. When the smart antennas systems are called smart not because the design or the structure is smart [40][41][42][43], but for the digital signal processing that is attached with the antennas makes the system smart. In fact, the theoretical theory of using this technology is not new. For example; they used it in the defense related systems in World War II. In the recent years, the emergence of powerful, low cost, digital signal processing (DSPs), general-purpose processors (and application-specific integrated circuits) as well as innovation signal processing algorithms, smart antennas systems are been used for commercial application.

We can understand the smart antennas system by showing an example related to our anatomy. Let imagine two persons carrying on a conversation inside a pitch-dark room. The listener between two persons is capable of determining the location of the speakers, as he or she moves about the room because the voice of the speakers arrives at each acoustic sensor - the ear - at a different time. The human signal processor – the brain - computes the direction of the speaker from the time difference or delays received by the two ears. Afterwards, the brain adds the strength of the signals from each user, to focus on the sound of the computed direction. But if additional speakers join in the conversation, the brain can tune out unwanted interferers and concentrate on one conversation at a time. Conversely, the listener can respond back to the same direction of the desired speaker by orienting the transmitter towards the speaker.

Electrical smart antenna systems work the same way, using many antennas instead of the ears and a digital signal processor instead of the brain. Therefore, after the digital signal processing receives the time delays from each antenna element, it computes the direction of arrival (DOA) of the desired user. It then adjusts the excitation (the amplitudes and phases of the signal) to produce a radiation pattern that focuses on the desired user, while nulls any signal interfering the target user.

There are two main types of smart antennas used in the wireless communication, one is called the switch beam antenna that selects the beam that has the largest signal power and nulls other beams. The second kind of smart antenna is called adaptive beam antenna, which is able to change its antenna pattern dynamically to adjust to noise, interference and multipath. To adjust the tracking capability, they use an adaptive algorithm such as least mean square (LMS) or the Recursive Least Square (RLS) adaptation algorithms. Our proposed system, although similar in concept to the adaptive beam antennas, does not require feedback for weight updates.

A multiple input multiple output (MIMO) is another active study in the antenna array processing to increase the capacity of the wireless communication. There is a lot of research in this field that still needs to be studied since it can be the gateway for the third or fourth generation for the mobile communication.

3.4 Multi-User Detector

The study on adaptive antenna array systems has been investigated to combat multiple access interference for CDMA systems. Adaptive antenna array can cancel out undesired signal using beamforming. However, if the there are interfering signals from undesired users with the same arrival angle of a desired user, an adaptive antenna array cannot suppress it. Also if the number of interferers is higher than the number of receiver antenna elements, an adaptive antenna array can't suppress all of the interfering signals. Here multi-user detection will be used to overcome this problem [50] - [54].

In this section we will review the more widely used multiuser detection techniques. It can be combined with antenna arrays to cancel out the remaining interference at the same beamforming angle (arrival of angle) or the remaining interference users that are in the same beamforming region.

3.4.1 Conventional Receiver

The conventional receiver consists of K bank of matched filters for each of K interfering DS-CDMA signals followed by decision devices [50]. If we assume the channel has a single resolvable path, the bit decisions are given by:

signal in its appropriate correlator. However, the performance of the weak users will be greatly degraded by the strong users. This is the previously mentioned near-far problem. Designing spreading codes that have low cross correlation can help improve this problem, but this task is difficult to achieve.

A modified version of the matched filter receiver is the RAKE receiver that is a popular single antenna receiver used in the presence of multipath [44][45]. As it is known in DS-CDMA, multipath has both positive and negative effects. One on hand, the independently fading paths can be a valuable source of diversity. One the other hand, the multipath introduces interpath interference and in the case of orthogonal codes is used, it also introduces MAI.

The RAKE receiver uses multiple correlators, one for each path, and the outputs of the correlators (called fingers) are then combined into a single output to maximize the SNR. The received signal r(t) for q user that have L paths is given by:

$$r_q(k) = \sum_{l=0}^{L-1} \alpha_{lq}^* \int_{kT+tc}^{(k+1)T} x(t) c_q(t - kT - tc) dt \tag{3.12}$$

Where α_{lq}^* is a complex conjugate of channel gain for a certain path L transmitted form user q.

The performance of a coherent RAKE receiver with perfect channel estimation (complex weights are matched to the channel response at these fingers) tends towards that of transmission in an AWGN channel. A practical implementation of a coherent RAKE in a fading channel employs some form of channel estimation, such as pilot symbols. The most applied combiner in coherent RAKE receiver is the maximal ratio combiner where the weights of the MRC are proportional to the path gains. Thus, the combining weights in the RAKE receiver are used to maximize the SINR rather than SNR as in the conventional receiver.

3.4.2 Optimum Multi-User Detector

Verdu has developed the optimum multi-user detector or maximum likelihood detector to minimize the probability of error [51]. The optimum receiver consists of a bank of matched filters followed by a multi-user Viterbi processor.

The beauty of this detector is that Verdu shows that the performance degradation due to increased MAI or the near-far problem is minimized. However, in terms of signal processing complexity, the optimum multi-user detector's complexity grows exponentially as the number of users increases and this complexity becomes prohibitive for most cellular systems since the number of users is usually high. Besides the complexity problem, the receiver requires the exact channel gain for each received signal as well as its received power and spreading signature of all users. Estimation of these parameters requires additional signal processing. The need for simpler yet robust receivers led to search for sub-optimum receivers.

3.4.3 Sub Optimum Multi-User Detector

To lessen the effect of the MAI – which also results in improved near-far problem compared to the conventional receiver, which depends on the number of users, suboptimum multi-user detectors are designed which have comparable performance to the optimum multiuser detector but with less mathematical operations. One of those sub optimum multiuser detector is called linear multiuser detector. There are two types of linear sub optimum detectors; one called decorrelating detector [49][48][52] and other is called minimum mean square error multi-user detector (MMSE) [53].

3.4.3.1 Decorrelator

The decorrelator is one of the most important multi-user detectors. The decorrelator consists of bank-matched filters that produce a vector of decision variables that are then multiplied by the inverse of the cross correlation matrix R of the spreading codes. Assuming synchronous DS-CDMA the correlation matrix R as is given by:

$$R = \begin{bmatrix} 1 & \rho_{21} & \cdots & \rho_{N1} \\ \rho_{12} & 1 & \cdots & \rho_{N2} \\ \vdots & \vdots & \ddots & \vdots \\ \rho_{1N} & \rho_{2N} & \cdots & 1 \end{bmatrix} \qquad (3.13)$$

ρ_{ij} is given by

$$\rho_{ij} = \begin{cases} 1 & i = j \\ \frac{1}{T_b}\int_{Tb} ci(t)cj(t)dt & i \neq j \end{cases} \qquad (3.14)$$

Where T_b is the symbol period. Assume the output signal of the entire matched filter is

$$r(t) = R * d + n \tag{3.15}$$

Where d is the decision variable vector obtained from the bank of matched filters and n is AWGN samples at the output of the matched filters. Multiplying (3.15) by the inverse of the correlation matrix R^{-1}, we get:

$$r = d + R^{-1}n \tag{3.16}$$

As we observe from the equation (3.16), the decorrelator completely eliminates the MAI. The most interesting thing about the decorrelator is that there is no need to know the power of the individual users and it is shown to provide large capacity gains. However, the drawback of the decorrelating receiver is noise enhancement. From the second term of equation (3.16) (R^{-1} n), the noise (n) will be boosted by the R^{-1} factor. Moreover, if the noise level is higher than the interference level, the decorrelator receiver may be worse than the conventional receiver. In this condition, the minimum mean square error (MMSE) multi-user detectors that will suppress the noise development will be the proper choice.

3.4.3.2 Minimum Mean Square Error Detector

This detector structure is very close to the decorrelator detector. The only difference between this detector and the decorrelator detector is the background noise level in the inverse matrix equation. The MMSE can eliminate the background noise when the noise level becomes too high.

The linear transformation H of the MMSE detector is

$$H = [R + \left(\frac{N_0}{2}\right) A^{-2}]^{-1} \tag{3.17}$$

Where R is the cross correlation matrix as same as the cross correlation of the decorrelator matrix inverter, A is the diagonal matrix containing the received amplitudes of the users and $N_0/2$ is the two sided power spectral density of the background noise.

As it can be seen, if the noise level goes to zero, the MMSE detector will convert to the decorrelator detector. The linear transformation H required from the detector to know about all of amplitudes of the users, therefore its performance depends on the power of the MAI users. Thus the MMSE will some near-far resistance compared to the decorrelator detector which is completely near-far resistant.

3.4.4 Multi-Stage Interference Cancellation

There are two main types of interference cancellation used as a multi-user detector, the successive interference cancellation (SIC) [47][53] and parallel interference cancellation (PIC) [46][54].

3.4.4.1 Successive Interference Cancellation (SIC)

The structure of successive interference cancellation is almost like multistage conventional receiver, which in return, will give a simple structure advantage. First, the conventional receiver will detect the strongest interference, make a hard decision to get the useful date, respread the data again and then cancel it from the received signal. We will repeat this step until we detect all interferences. The problem that arise in using the SIC is as follows:

- Since each stage depends on the previous stage of its estimation, if the first stage estimation was incorrectly detected, then the probability of error detection in the forward stages will be high.
- The first detected user that has the strongest signal will contain a lot of MAI components that in turn will make variable risky decision.
- If the MAI power profile changes, we will require reordering the signal detection of the detector again.

3.4.4.2 Parallel Interference Cancellation (PIC)

In the PIC detector, all the MAI components for each user are estimated, summed and subtracted in parallel structure as in figure 3.4. The initial stage begins from the output of the conventional receiver. Then the output of the bank of match filter are multiplied by the estimated amplitude of received signals (if the MAI has different received power), respreading the data by spreading the codes for each user, summing all the MAI for each interesting user and subtracting it from the total received signal, then applying the conventional receiver to move to the next stage. The problems that may face the researchers are in both estimating signal amplitudes for each interfere and knowing the spreading code for each user in the detector. However, most applications in these days recommend using the PIC than SIC because of the disadvantages SIC causes.

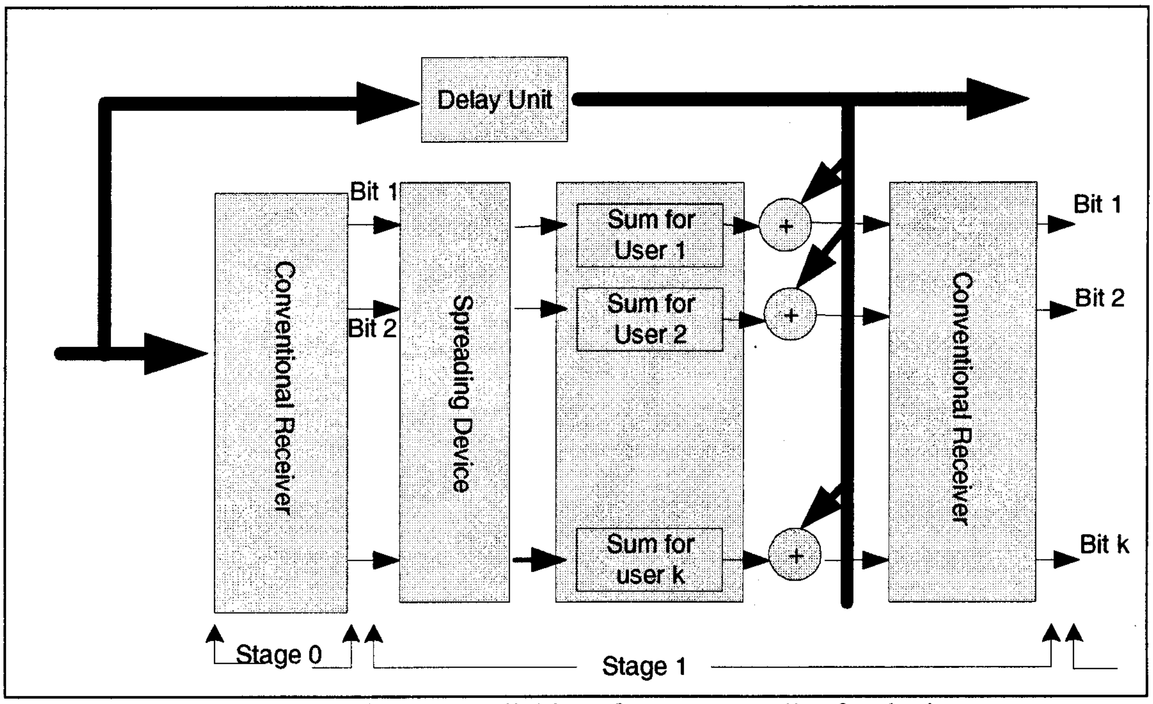


Figure 3.4: Multistage parallel interference canceller for the j stage.

3.4.5 Adaptive Multi-User Detector (AMUD)

This detector tries to adapt the weight of the interest signal from MAI by forcing a training sequence as a reference signal [60][61]. This structure is very close to the adaptive antenna array as in figure 3.3, which tries to update the weight coefficients according to the reference signal. The huge advantages of using adaptive multiuser detector can be summarized as follows;

1. AMUD do not require knowing about the spreading code, power and delays (multi path channel) for each user except the desired user as in the optimum and sub-optimum detectors.
2. AMUD only requirement is the spreading code for each user should be repeated in each transmission bit.
3. AMUD has less complexity in terms of signal processing than the optimum receiver and also its complexity doesn't depend on the growth number of users.
4. AMUD has less sensitivity from the impact of noise background than the linear multi-user detector (MMSE).
5. AMUD don't need to estimate the strength of all the MAI components as in SIC nor do they need to estimate the amplitude for each user as in PIC.
6. AMUD can be easily implemented because of the feature of linearity it owns.

However, in the adaptive antenna array, the performance will degrade as the number of MAI components increase or if the cross correlation between the signature waveforms of the MAI components and that of the desired component become very high. To overcome this problem, the adaptive multi-user detection should be combined with multistage interference cancellation or RAKE reception.

There are other types adaptive multi-user such as adaptive decision feedback multi-user detector, that works similar as the adaptive linear multi-user detector but it uses a decision variable as a reference signal.

3.5 Previous Research in Space-Time Receivers for Wireless Communications

In wireless code division multiple access (CDMA) systems (such as third generation wideband CDMA), the major impediments for signal detection at the physical layer are multipath fading and multiple access interference (MAI) caused by cochannel users, which their codes are not orthogonal to the desired user code signature. Moreover, if the MAI arrive from the direction other than that of the desired user, the spatial filtering introduced by the antenna array can significantly suppress the MAI's. However, an antenna array cannot suppress a high level interfering signal from undesired user with the same direction of arrival (DOA) as that of the desired user. One method to reduce the remaining interference is to use a multiuser detector in conjunction with the antenna array [55][56][57][58][59]. Several techniques can be used to overcome these impediments; these include the joint space-time detection, which combined either RAKE matched filtering or any multiuser detector with antenna array processing. In this section, we will present a global overview of the previous research in space-time receiver in the wireless communication.

3.5.1 Space-Time Rake Receiver

This is an extension of the time-only RAKE receiver discussed earlier. In space-time RAKE [55][56], a space-time correlator for each path as shown in Figure 3.5. This consists of beamformer weights, which is matched to the path array response vector, a_{lq}, followed by conventional code correlator matching to the path delay. The output fingers from the code correlator are summed in a RAKE combiner.

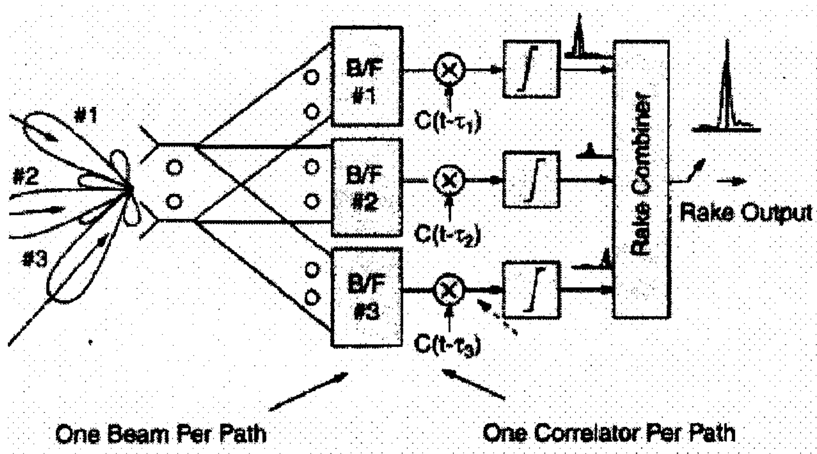


Figure 3.5: Space-Time RAKE receivers[56]

The output of the space-time RAKE for the qth user at the L paths is

$$\hat{S}(k) = dec\left\{\sum_{l=1}^{L} \alpha_{lq}^{*} \int_{kt+tc}^{(k+1)T+tc} a_{lq}^{H} x(t) c_q(t - kT - tc) dt\right\} \quad (3.18)$$

From (3.12), the space-time receiver gives more improvement than the conventional RAKE receiver. Adding the a_{lq} in each user was the reason that boosts the performance of the system.

3.5.2 Space-Time Blind Receiver

Before discussing this technique, it is worth defining the word "blind" in the CDMA context. A "blind" receiver requires no training or pilot signals [62], only the spreading code of the desired user. Some blind receivers do not even need to know the desired user's spreading code. In this section the thesis will refer to the first definition of the blind detection definition.

In space-time blind receivers, the algorithms are somehow similar. In other words, the only known parameter is the coding waveform for all users [62][63][64][65]. Recently, an interference canceller with a blind adaptive array has been proposed in [66], where the knowledge of the desired user's signature is exploited to design a blind adaptive receiver, without the need of transmitting a training sequence. The recent papers that deal with ST-blind multiuser detector are found in [62] and in [63]. In [62], a blind algorithm is formulated using a constrained optimization criteria and the adaptation is carried out using LMS algorithm for both the blind adaptive array and the adaptive parallel interference cancellation (APIC). The successive interference-canceling (SIC) receiver with a blind adaptive array is proposed in [63], having similar structure of the only time processing of SIC but adding a blind smart antenna as in figure 3.6. In general, those kinds of detectors are just combining the antenna array with the multiuser detectors in the blind adaptive fashion.

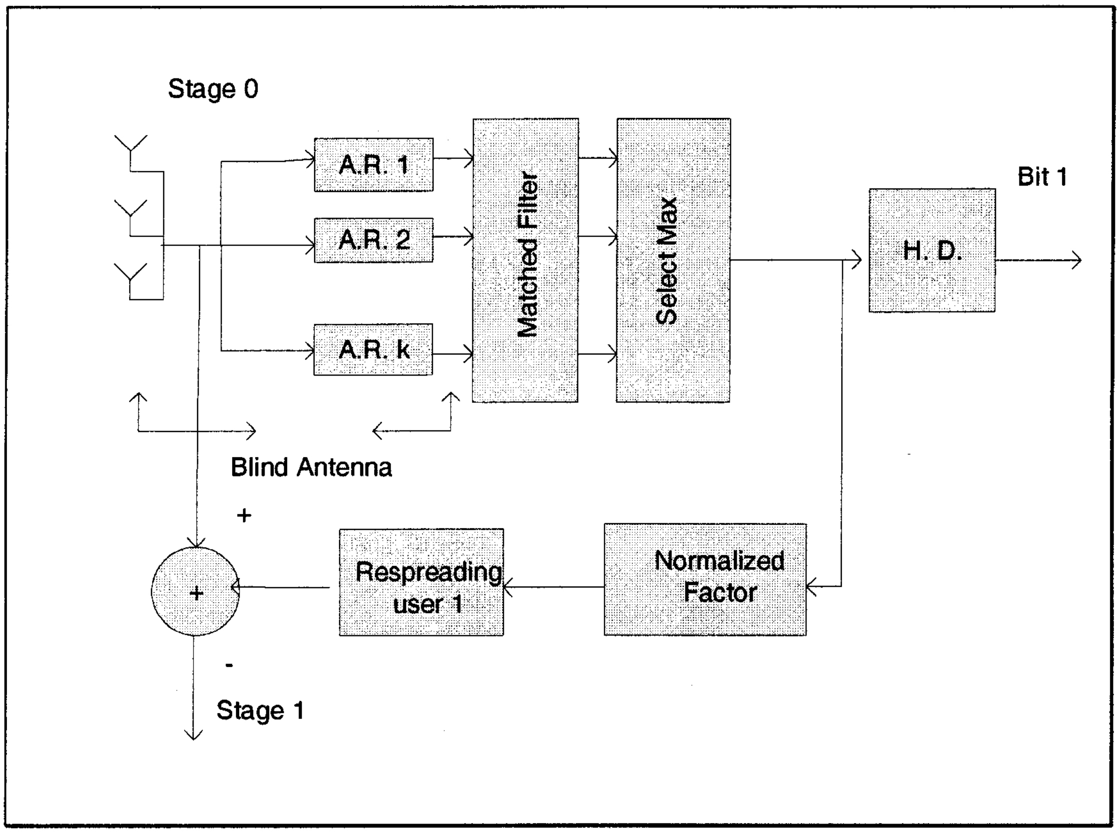


Figure 3.6: Blind space-time receivers using multistage serial interference canceller

Chapter 4

Proposed Receiver Structure and Simulation Models

In this chapter, the system model will be described and computer simulation methods and results will be presented. We will study, model and simulate the transmitter, channel and receiver. We will then present the simulation results of the proposed system in terms of bit error performance and constellation diagrams. A comparison of the performance between the conventional multiple antenna array algorithms and the proposed system will be presented also in this chapter.

The objective of this thesis is to reduce the effects of the MAI components from the desired user's signal. In the simulation, we will consider the case of a single cell system, which means we will only take into consideration the effect of the MAI coming from the same cell.

As previously mentioned, we assume perfect power control for ease of implementation of the simulations. This means that all the signals are received at the same power level, except for the near-far simulation case; we use different transmitted power for each user.

The beauty of our proposed system is that the receiver needs to know less parameters about the interfering users' received signals in order to produce reliable decision variables that are free of MAI for the desired user. In other words, the receiver (or base station) of the CDMA system requires only the spreading code sequences of all users.

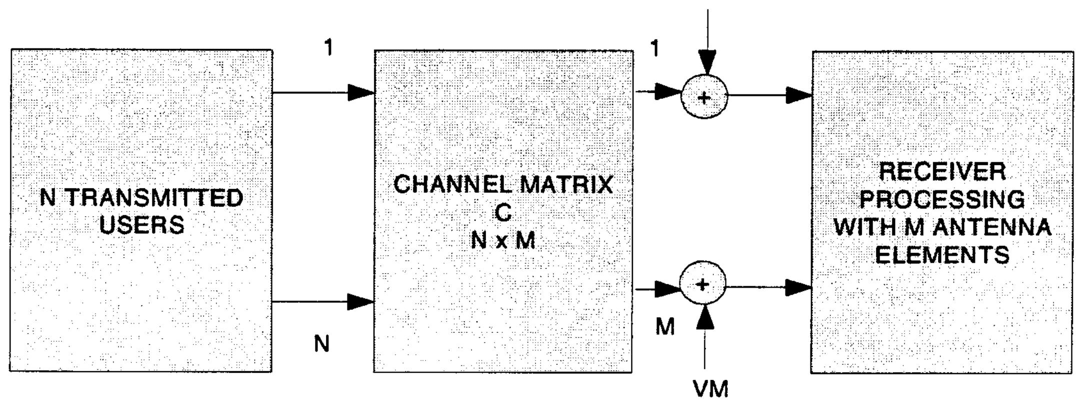


Figure 4.1: Matrix expression of end-to-end transceiver

4.1 Proposed System Model

We will study the model of the transmitter, the channel, and the receiver of the CDMA systems.

4.1.1 Transmitter Model

BPSK Modulation

The type of modulation scheme used in the simulation is binary phase shift keying (BPSK). As mentioned in chapter 2, BPSK is now widely used in both military and commercial communication system. The expression for the n^{th} bit of the BPSK signal is:

$$s(t) = \sqrt{2P_i}\, b(t) \cos(2\pi f_c t) \tag{4.1}$$

$$b(t) = \sum_{n=-\infty}^{\infty} b_n \Psi(t - nT_b) \tag{4.2}$$

where P_i is the total transmitted power for user i, *fc* is the carrier frequency, b(t) is the binary data sequence composed of a sequence of symbols b_n, $b_n \in \{+1,-1\}$ and Ψ_{Tb} is a unit rectangular pulse of duration T_b and n is the data bit index.

Spread spectrum process

Direct sequence spread spectrum (DS-SS) will be used in the simulation as multiple access schemes by using a spreading signature waveform for each transmitted user.

The code signal that is assigned for each transmitter is used to spread the message signal. By using the code division multiple access (CDMA), the receiver code separates the desired user massage signal from the messages of the other users.

The spreading waveform c(t) may be expressed as

$$c(t) = \sum_{n=-\infty}^{\infty} \sum_{l=0}^{N} c_l \Psi_{Tc} \frac{(t-(l+nq)Tc)}{Tc} \tag{4.3}$$

where $c_l \in \{+1,-1\}$, q is the number of chips sent in the spreading code sequence, Tc is the chip period and Ψ_{TC} is a unit rectangular pulse of duration T_c.

The CDMA signal of user *i* may be expressed as,

$$v_i(t) = \sqrt{2P_i} c_i(t) b_i(t) \cos(2\pi f_c t) \tag{4.4}$$

and the baseband of equation (4.4) is given by,

$$v_i(t) = \sqrt{2P_i} c_i(t) b_i(t) \tag{4.5}$$

After gathering the basic information needed to transmit the signal, figure 4.2 will illustrate the transmitter block diagram for our proposed algorithm. As seen in the figure, each user transmits two signals, the data signal and the pilot signal. The data signal is the signal that carries the desired information that our propose system is trying to reconstruct at the receiver after passing through a faded channel along with other users' signals. The pilot signal is a stream of symbols with known values to both the transmitter and the receiver that is spread by a spreading code that is orthogonal to the spreading code of the same user's data signal. The objective of the pilot signal is to estimate the channel gain that in turn will be usde for phase correction and maximal ratio combination weighting.

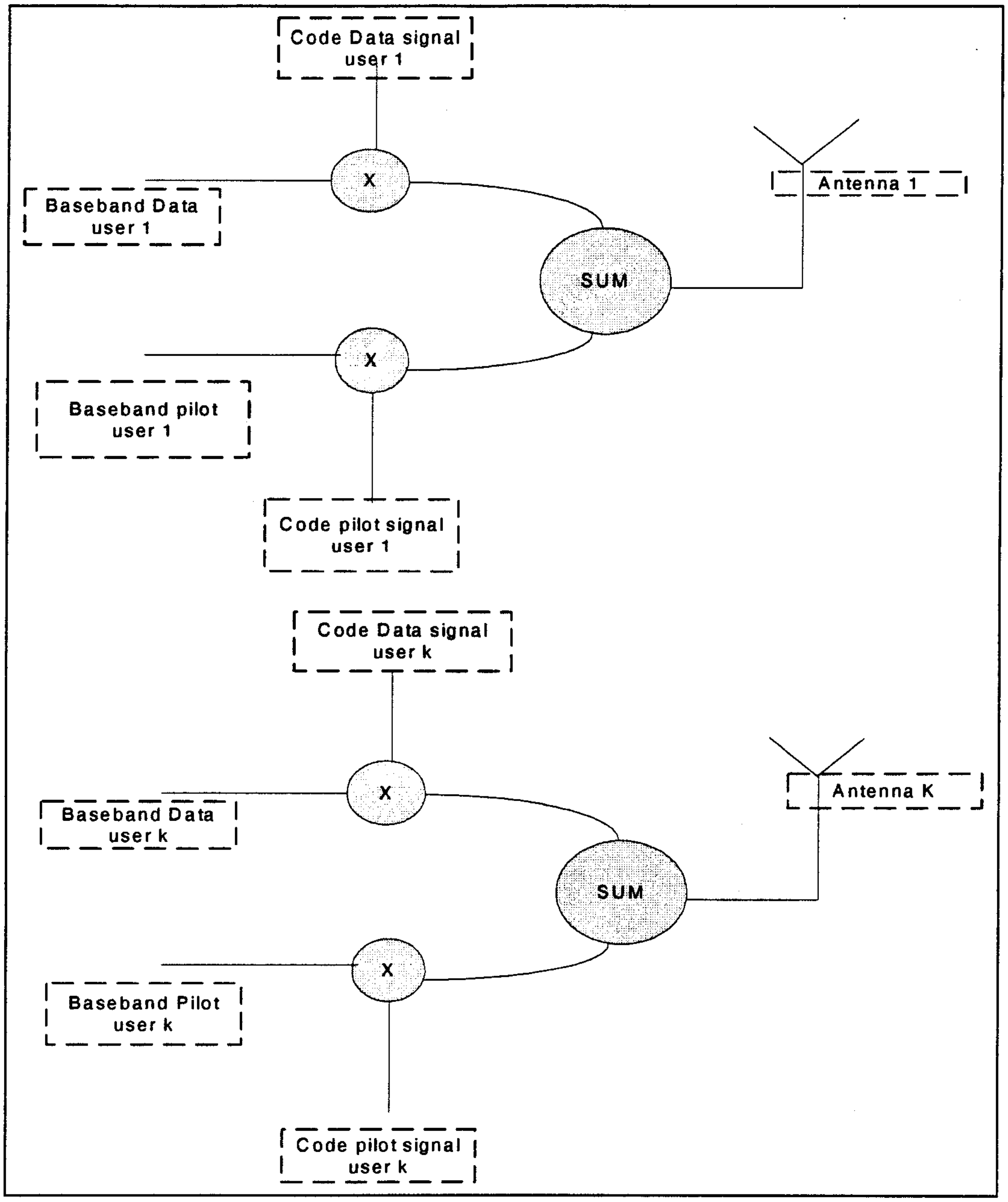


Figure 4.2: Proposed transmitter with N users

The transmitted signal from each user i will be given as,

$$v_i(t) = A_p c_{ip}(t) b_{ip}(t) + A_d c_{id}(t) b_{id}(t) \tag{4.6}$$

where A_p, A_d and c_p and c_d are the amplitude and code signature waveform signals of the pilot and information signal respectively of the i^{th} user. Where the cross correlation between the code signal of the data signal b_p and code signal of the pilot signal b_d at user i is

$$\rho_{cip,cjd} = \left\{ \frac{1}{T_b} \int_{Tb} cip(t) cjd(t) dt = 0 \right.$$

Moreover, if the total transmitted power is normalized to 1, A_p is given by:

$$A_p = \sqrt{(1 - A_d) A_r}$$

Where A_r is the fraction of the received energy that each antenna element in the array receives, therefore

$$A_r = \frac{1}{M}$$

Where M is number of received antenna elements.

4.1.2 Channel Model

The channel used in the simulation is flat slow fading. The flat fading channel means that there is no time delay spreading that maybe caused by inter symbol interference ISI. The Doppler rate is $B_d = 0.1/T_b$. The received signal at the base station for user *i*, after passing through the channel matrix is given by

$$r_i(t) = \alpha(t) v_i(t) \tag{4.7}$$

where $\alpha(t)$ is complex value referring to the channel response and $v_i(t)$ is defined in equation (4.6).

The channel in the wireless communication can be viewed as a matrix with size of NXM (seen in figure 4.3). Where N is the number of input (transmitted) signal that in our case will equal to number of transmitted users and M is number of antenna receiver array.

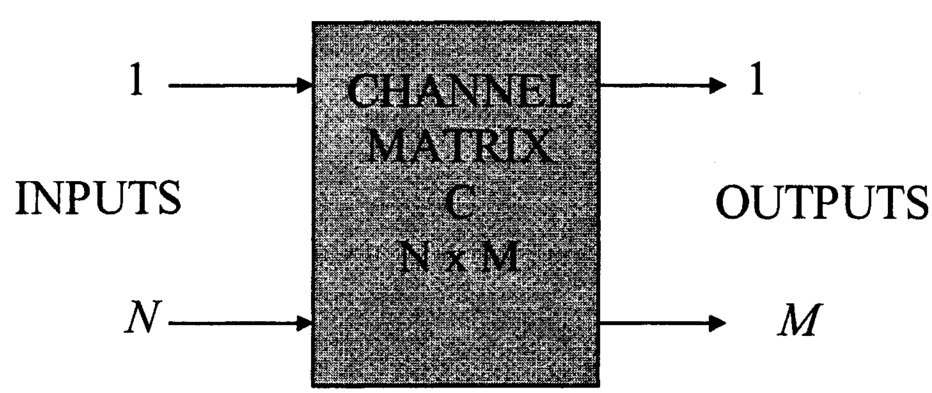


Figure 4.3: Channel matrix with N input and M outputs

As it is previously mentioned in chapter 2, we used the autoregressive model to find the Doppler taps of Rayleigh fading distribution characteristic and to find the variance of the Gaussian input.

Theoretically, if the distance between two adjacent receiver antennas is more than coherence distance [5], the fading on two incoming replicas of the same signal arriving at the two adjacent antennas will be uncorrelated. But practically the channel gains will have a non-zero correlation between, because of the fading environment criteria. Therefore in our simulation we take into consideration the correlation between the channel gains. Correlated channels gains are generated using the Cholesky factorization method.

4.1.3 Receiver Model

Our receiver model consists of four parts in space-time joint detection, beginning from the antenna array elements going through multiuser detectors, pilot filters, space weighting and the space combining and resulting in a decision variable. The block diagram of the receiver model is shown in figure 4.4.

Before we begin explaining our receiver model, we have to clarify an important issue related to the space gain produced by the antenna array elements, which is sometimes called array manifold or steering vectors.

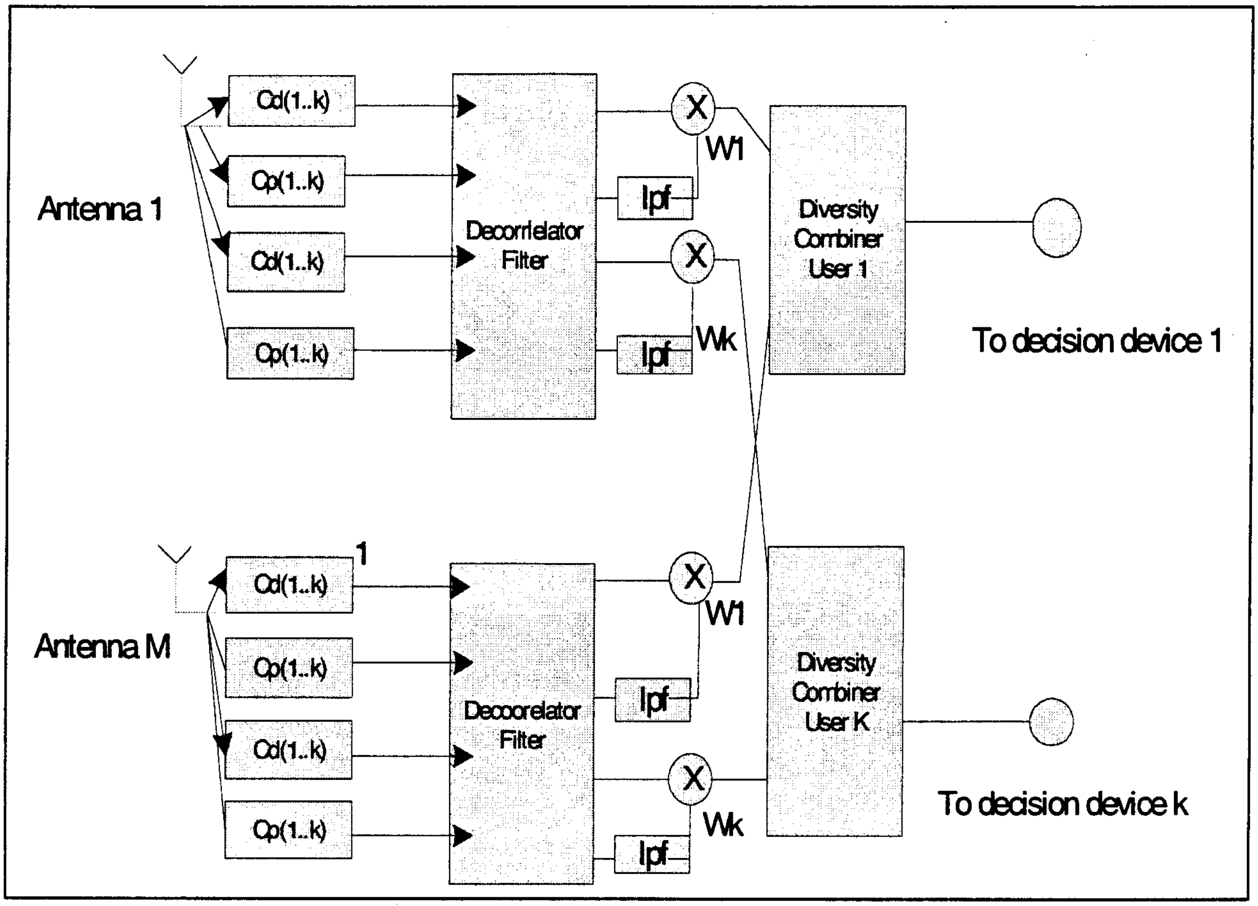


Figure 4.4: Proposed receiver block diagram

At the base station, antenna array elements have either a linear array or circular array. Circular arrays suffer from the disadvantage that high level side lobes are created in their beam patterns, while the main disadvantage of the linear array is that is less directional that the circular array [63]. However, the simplicity of the analysis makes the linear array attractive to cellular designers. Specifically, a linear antenna array that consists of identical elements (I) with equal inter-elements spacing (d) is called Uniform Linear Array (ULA) which is shown in Figure 4.5.

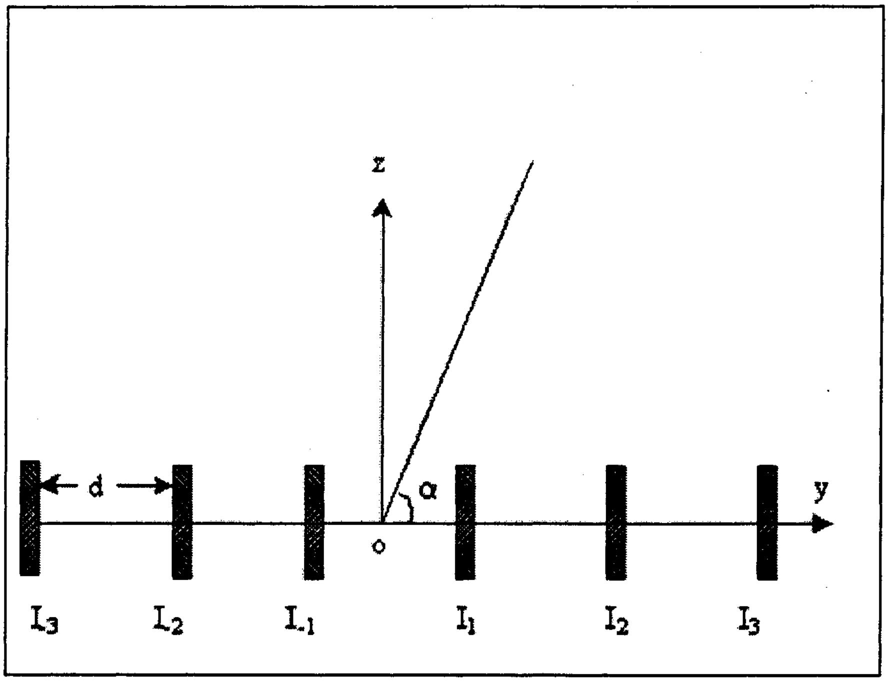


Figure 4.5: Uniform linear array

According to [68][69], each element in the ULA has an array response vector, each array response vector depends on the direction of arrival (DOA) angle, the distance between the elements array and how far its from the reference point (usually the reference point is the first antenna element). In our simulation, we will assume an ideal ULA structure, which means that the array response vectors are equal and they have a unity value. The reasons for this assumption are (1) our system is not based upon the DOA estimation, (2) simplicity in the signal processing. However, we will definitely have a loss part of the space gain that will produce from the ULA. In addition to the ULA assumption, there is no mutual coupling between the antenna elements at the base station.

In virtually all research papers that deal with multiuser detection for multiple antenna receivers, the multi-user detector is placed after the combiner, to adapt the weights of the antenna array in order to direct the antenna beam toward the desired user. By using these structures, we need a long training sequence in order to find the DOA of the desired user such as using the multi-user (PIC, SIC or adaptive multi-user) detection with antenna array. In addition, the direction of arrival of all users must be known, especially that of the desired user. A problem will arise when the MAI has the same direction of arrival as the desired user. We will find that the complexity and time required to find the DOA will increase as we try to eliminate the MAI in the system.

In figure 4.4 of our proposed receiver structure block diagram, we apply a decorrelator detector within the antenna array before it enters the combiner. Using decorrelator detectors, it will eliminate the MAI components leaving only the desired user, which will be enhanced by the MRC weights of the antenna array.

The signal received at one of the antenna elements, after passing through the channel is given by

$$r_k(t) = \sum_{i=1}^{N} \alpha_{i,k}(t)(A_p c_{ip}(t) b_{ip}(t) + A_d c_{id}(t) b_{id}(t)) + \mathrm{n(t)} \tag{4.8}$$

where $k \in \{1, \cdots, M\}$, M is the number of receiver antenna elements and all other parameters were defined in previous sections. After applying (4.8) to the despreader and the conventional matched filter of the data desired user at each branch we will have the following equation

$$h_k(t) = (A_d b_{kd} \alpha_{kk}) + \sum_{i \neq 1}^{N} \alpha_{i,k}(t)(A_p \rho_{ip}(t) b_{ip}(t) + A_d \rho_{id}(t) b_{id}(t)) + \mathrm{L_{kd}} \tag{4.9}$$

where $\rho_{ip} = \begin{cases} \frac{1}{T_b} \int_{Tb} c_{id}(t) c_{jp}(t) dt & i \neq j \end{cases}$

and

$$\rho_{id} = \begin{cases} \frac{1}{T_b} \int_{Tb} c_{id}(t) c_{jd}(t) dt & i \neq j \end{cases}$$

Where the first term of equation (4.9) is the desired part and the second term is the MAI that our proposed algorithm tries to eliminate. Also we notice that we need to find the channel gains of all users. L_{kd} is the output of the kth user's matched filter for an input of n(t). The noise components in all matched filters' outputs on the same antenna branch are correlated with one another due to the correlation between the signature waveforms of the different users [64].

Going back to equation (4.8), we have M received signals that are similar to equation (4.9) but at different antennas branches for the same desired user i, where m =[1...M], M is the number of antenna elements. We shouldn't forget that each received signal in equation (4.9) is multiplied by the weights of the antenna array (w^*) and combined using maximal ratio combining, as a result, equation (4.9) becomes:

$$h_{kq}(t) = (\sum_{m=1}^{M}((A_d b_{kd} \alpha_{kk}) + \sum_{i \neq 1}^{N} \alpha_{i,k}(t)(A_p \rho_{ip}(t) b_{ip}(t) + A_d \rho_{id}(t) b_{id}(t) + L(t))).w^* \qquad (4.10)$$

As in chapter 3, equation (4.10) is similar to the conventional multiple array except that the conventional receiver has difficulty-removing MAI from its system. Our proposed structure is a modification of the conventional receiver to eliminate the MAI components. To demonstrate our idea, before we multiple each receiver signal $h_{kq}(t)$ by their weights to gain from the space diversity; let us apply the h_{kq} to a decorrelator detector to remove completely the MAI. This is easily done by finding the inverse matrix of the cross correlation between the users data signal code signal and pilot code signal. Let us go back to equation (4.9) and try to present it in matrix view for simplification as in,

$$h_k(t) = R[\alpha_i B_i] + L \tag{4.11}$$

Where R is the cross correlation between the code users, α_i is the channel response vector, B_i is the data and pilot signals for user i and L is the correlated noise. All the above parameters are given below:

$$R = \begin{bmatrix} 1 & \rho_{21} & \cdots & \rho_{N1} \\ \rho_{12} & 1 & \cdots & \rho_{N2} \\ \vdots & \vdots & \ddots & \vdots \\ \rho_{1N} & \rho_{2N} & \cdots & 1 \end{bmatrix} \quad \text{where } \rho_{i,j} = \begin{cases} 1 & i = j \\ \dfrac{1}{T_b} \int_{Tb} ci(t)cj(t)dt & i \neq j \end{cases}$$

and

$$\alpha_i.B_i = \begin{bmatrix} \alpha_{d1}.b_{d1} \\ \alpha_{d2}.b_{d2} \\ \vdots \\ \alpha_{dN}.b_{dN} \\ \alpha_{p1}.b_{p1} \\ \alpha_{p2}.b_{p2} \\ \vdots \\ \alpha_{pN}.b_{pN} \end{bmatrix} \tag{4.12}$$

where $[\alpha_{d1} \cdots \alpha_{dN}] = \lfloor \alpha_{p1} \cdots \alpha_{pN} \rfloor$

Now if we apply the decorrelater detector array matrix which the inverse of R (R^{-1}) to equation (4.11), we will get,

$$h_k(t) = (R\alpha_i B_i).R^{-1} + (L.R^{-1}) \tag{4.13}$$

or

$$h_k(t) = (\alpha_i B_i) + (L.R^{-1}) \tag{4.14}$$

Now by using the decorrelater detector it's clear that the MAI components are removed completely, on the other hand, the penalty for this interference removal is the enhancement of the background noise.

Next, the received signal is summed using the MRC after its multiplied by the antenna weight w * that is the complex conjugate of the impulse channel vectors α_i.

$$e_k(t) = \sum_{m=1}^{M} ((\alpha_i B_i) + (L.R^{-1}).w^*) \tag{4.15}$$

We have previously assumed the channel estimation to be ideal. Now we will focus on a system, which estimates the complex channel gain. As previously mentioned, we will use the pilot signal that is spread by a signature waveform that is orthogonal to the data spreading code signal for each user to estimate the channel. Since we need a pilot signal without any MAI components to achieve best channel estimation, we easily insert the pilot signal to the decorrelator detector as in equation (4.15) as same as we do for the data signal. For the simplicity purpose, we take estimated channel vectors that are impinging to only one antenna element of the M antenna elements. Again from equation (4.15), we could easily find the pilot signal or the data signal of either the desired user or the MAI users. Since we are interested to get the pilot signal of the desired user, equation (4.15) becomes as follows,

$$p_k(t) = (\alpha_i B_{ip}) + (L.R^{-1}) \tag{4.16}$$

Since B_{ip}, a pilot signal of the desired user and it is known both for its transceiver ends with fixed value. In order to get smooth channel estimation, we apply equation (4.16) to low pass filter. Finally, for each input sample, our proposed soft decision of the received signal before it reaches the hard decision is given by:

$$g_k(t) = \sum_{m=1}^{M} ((\alpha_i B_i) + (L.R^{-1})).w_m^{*}) \tag{4.17}$$

4.2 System Parameters and Performance Evaluation Criteria

In the previous sections, we have discussed in detail the transmitter model, the channel model and the received model. Table (4.1) summarizes the transmitter, channel, and the receiver parameters that are used in our simulation.

The multiple access system used is synchronous direct sequence division multiple access (DS-CDMA) system, where each of the users has equal time delay that could detect perfectly. As stated before, we are only interested in removing the MAI within one cell in our simulations. Moreover, all the simulations were based on bit level simulations and assume perfect carrier frequency recovery.

Table 1.1: Simulation parameters

Transmitter Parameters		***Channel & Receiver Parameters***	
Modulation Scheme	BPSK	**Channel**	Slow Fading
Multiple Access Scheme	CDMA	**Multipath**	Nonselective fading
Processing Gain	[15 to 40]	**Power control**	Perfect power Control
Number of users	2,3,4 users	**Number of antenna Elements**	1,2,3,4

4.3 Simulation results and discussions

In the previous section we have explained and described our proposed system by presenting the overall system conditions and giving the mathematical expressions of the proposed system. In this section, we present our simulation results for the proposed system using MATLAB programming.

The simulation results section is organized as follows. At first, we will show the best power allocated to the pilot signal. We will compare the performances of a single-user / single antenna system for both the perfect and the proposed channel estimation scheme and compare them with the theoretical BER for validation purposes. Next we will discuss the near far problem and demonstrate how our proposed system resists the near-far problem in terms of BER curves and constellation diagrams. Then we will show the performance of the proposed system in terms of BER performance of single, two, and three users using single antenna, two antennas, three antennas and four antennas when the cross correlation between users have values in the range [0.15,0.25,0.35] and a channel cross correlations in the range [0, 0.25, 0.5] (independent channel gains are assumed when the channel cross correlation equal to zero). The comparison in performance between single antennas and three antenna elements for four users to illustrate the space diversity advantage is shown afterward. Finally, a comparison between a conventional antenna array receiver and our system in terms of BER curves and constellation diagrams will be presented in the presence of MAI.

4.3.1 Pilot Power Allocation

There is trade off in choosing the ratio of the pilot signal power to the data signal power. If we increase power to the pilot signal, we will have strong estimation of the channel gain. But at the same time allocating too much power to the pilot stream reduces the power that is allocated to the data. This will make the bit detection less reliable since the probability of bit error depends on the actual bit energy. For our channel with Doppler spread $B_d = 0.1/T_b$, we show the simulated bit error rate of PSAM using BPSK in Figure 4.6. The results show that the best percentage of power allocation to the pilot stream is 15% of the total transmitted power for the case considered.

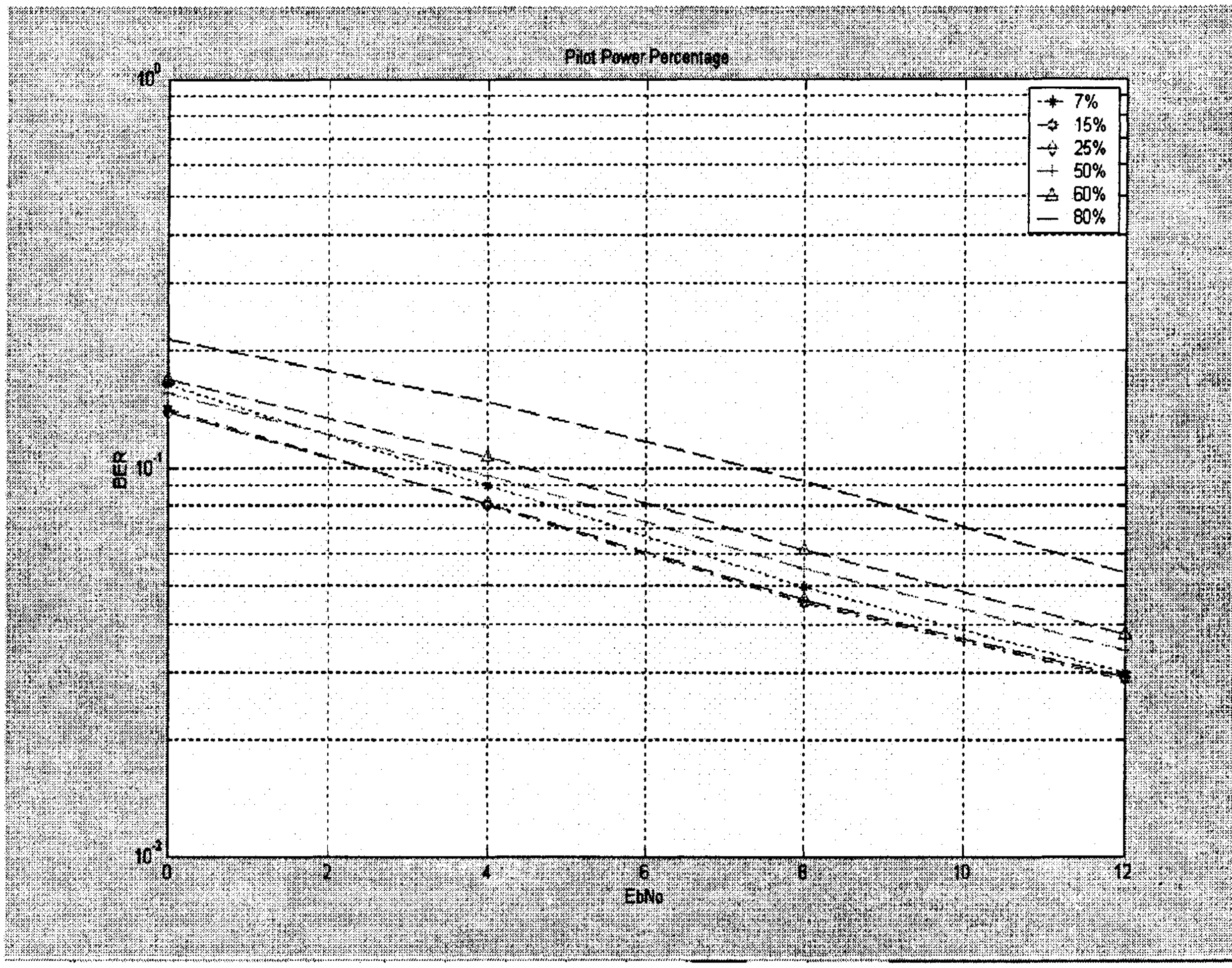


Figure 4.6: Percentage of power allocation to the pilot.

4.3.2 Channel Estimation

Figure 4.7 shows the simulation for single user (no MAI) and single antenna (no space diversity gain) to compare between the proposed channel estimation, perfect channel estimation and the theoretical value using independent channel vectors to show the validity of our simulations. We can see from figure 4.7 that at a BER of $10^{-1.5}$, simulation of the system employing perfect channel estimation is close to the theoretical case. The PSAM system performs roughly 2dB worse than the ideal case at the same BER ($10^{-1.5}$). However, by allocating 15% of the power to the pilot, we already have a 0.7dB power loss in the data signal compared to the ideal case. This means that an additional 1.3 dB is lost due to imperfect channel estimation.

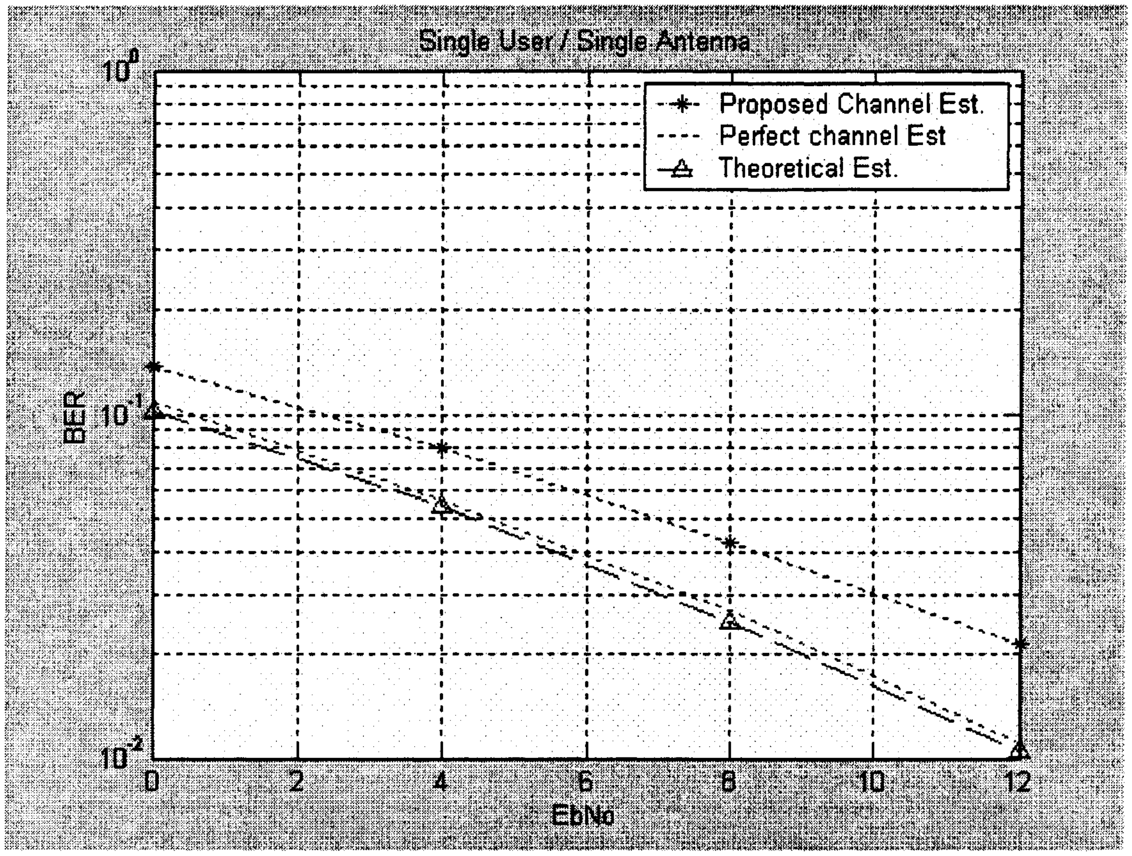


Figure 4.7: Comparison between perfect channel estimation and the proposed channel estimation

4.3.3 Near-Far Resistance

Figures 4.8 to 4.10 and Figure 4.11 to 4.13 demonstrate the degree of near-far resistance our proposed receiver has compared with the conventional receiver. The simulation was based in two transmitting users and with a three antennas receiver. In figure 4.8, 4.9 and 4.10, we assume that the interfering user's signal is received with twice as much power as the desired user. While in figure 4.11, 4.12 and 4.13 we assume that the interfering user's signal is received with five times more power than the desired user. The results in terms of BER performance and the constellation diagrams show that our proposed system is resistant to the near-far problem. This can be seen from the improved BER performance of our proposed system as well as in the sharper constellation diagrams.

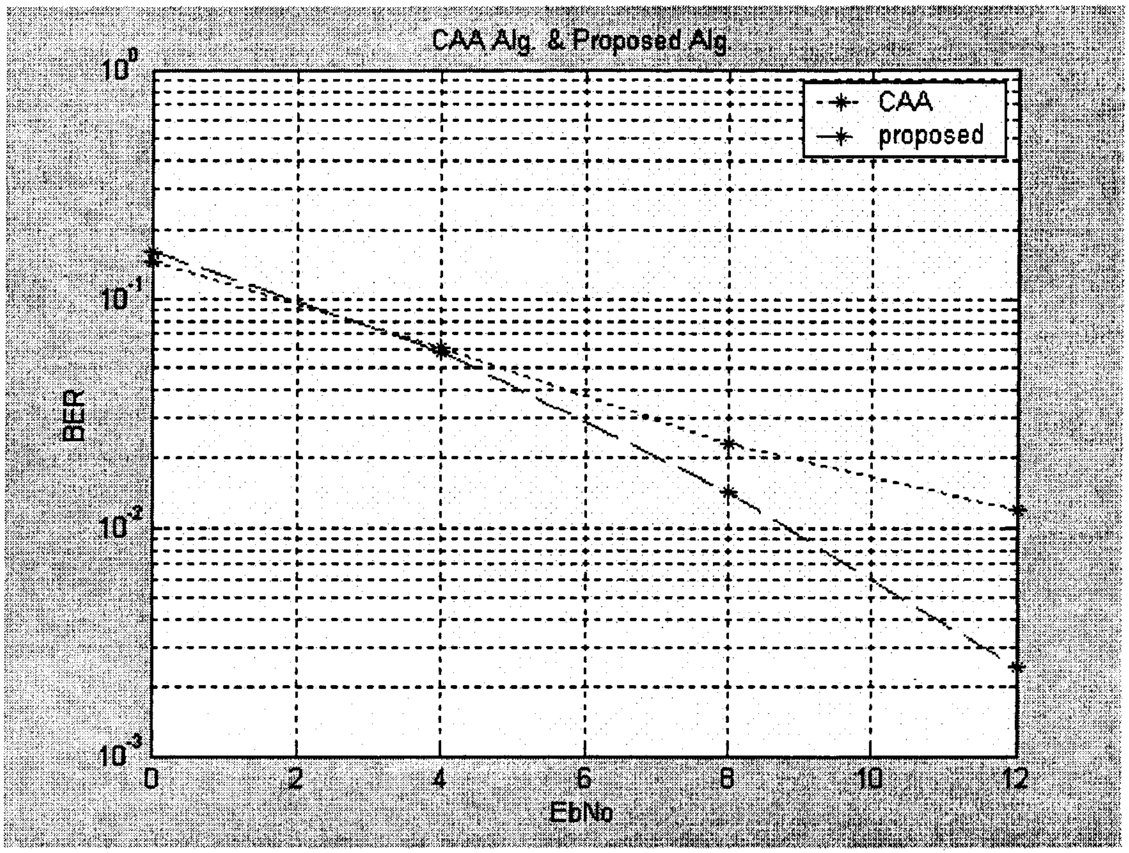


Figure 4.8: BER performance of desired user when interfering user's signal is received with twice as much power

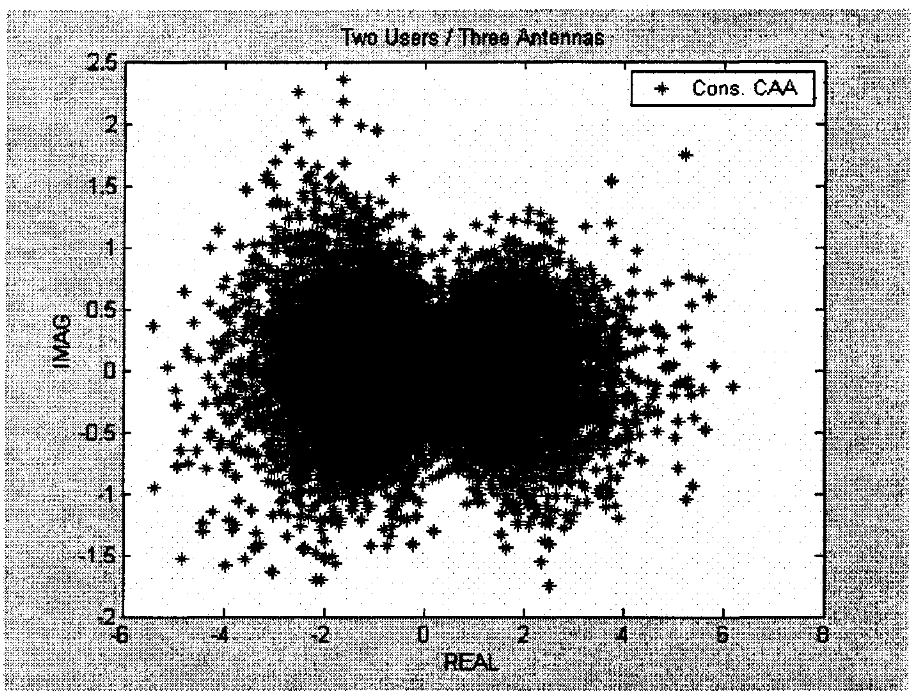


Figure 4.9: Constellation diagram for desired user employing CAA when interfering user's signal is received with twice as much power

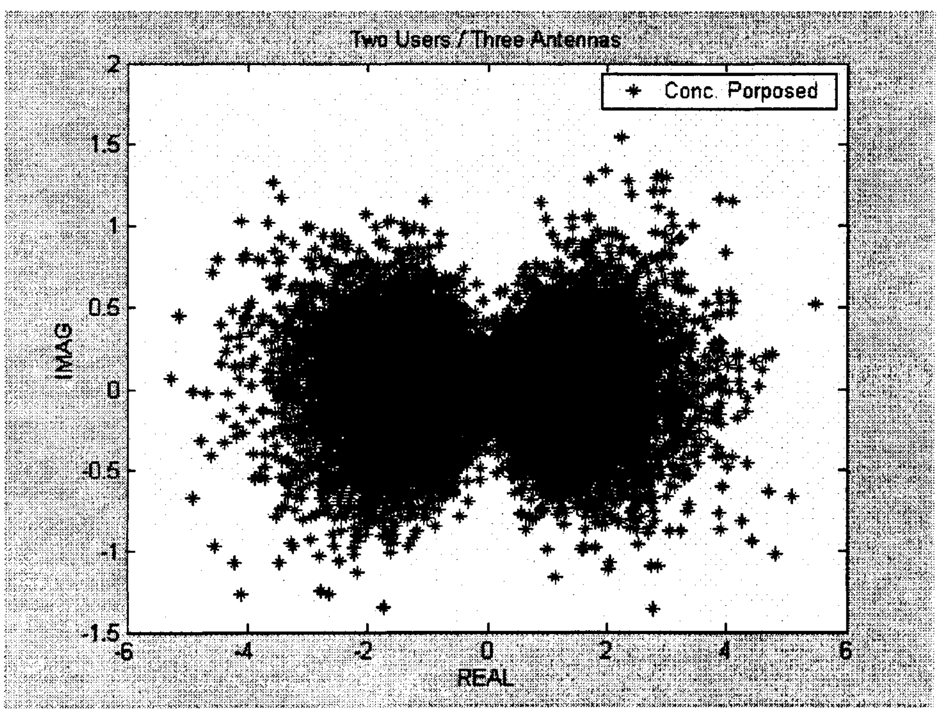


Figure 4.10: Constellation diagram of desired user employing proposed system when interfering user's signal is received with twice as much power

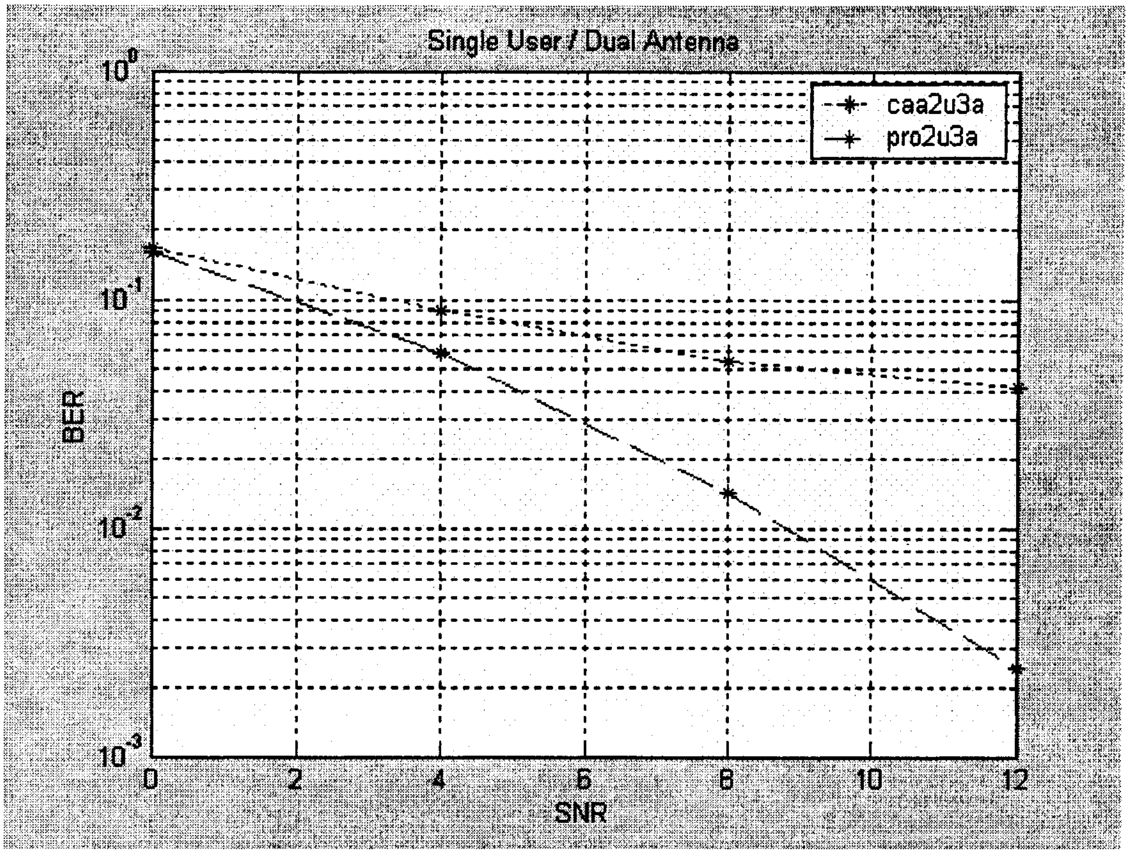


Figure 4.11: BER performance of desired user when interfering user's signal is received with five times as much power

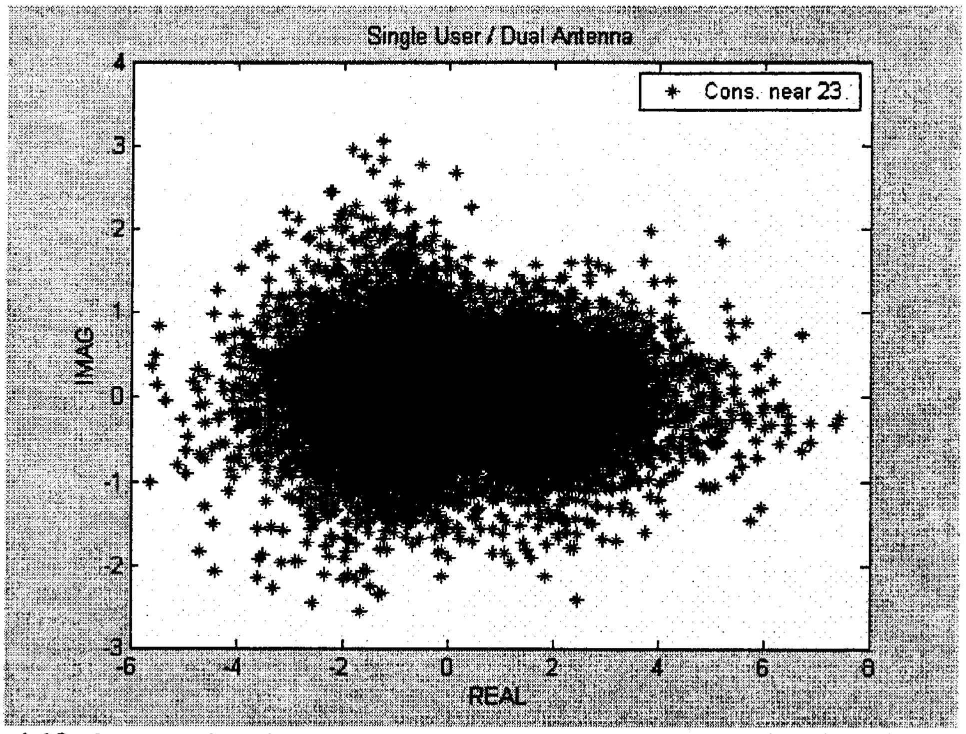


Figure 4.12: Constellation diagram for desired user employing CAA when interfering user's signal is received with five times as much power

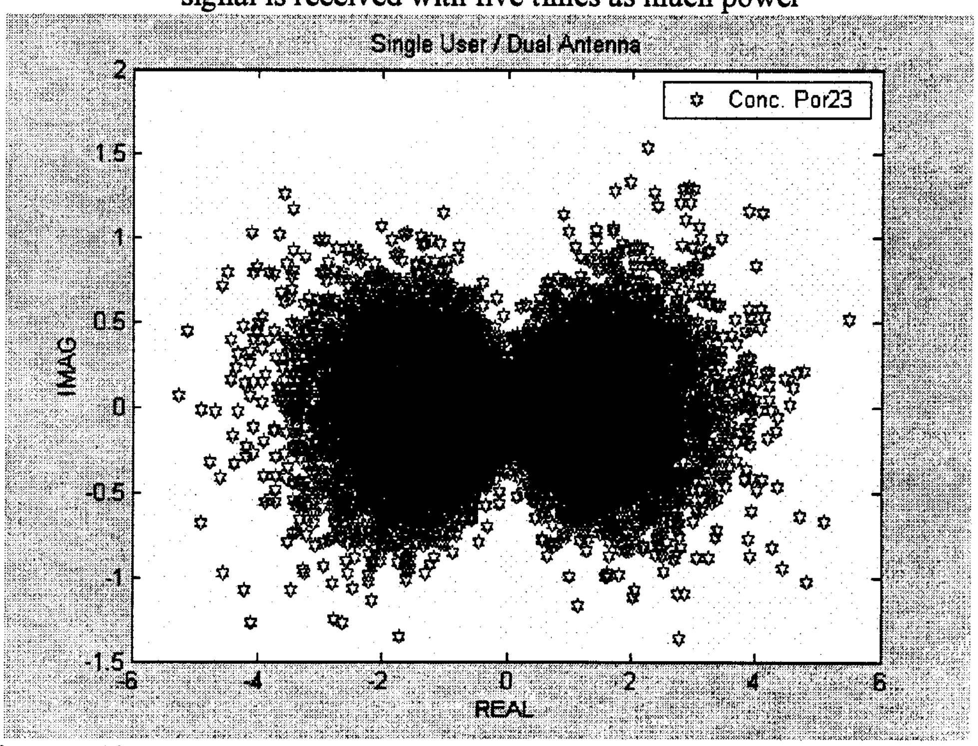


Figure 4.13: Constellation diagram of desired user employing proposed system when interfering user's signal is received with five times as much power

4.3.4 User and channel cross correlations

In this section, we used single, two, three and four receiver antennas in presence of four users sharing the same channel bandwidth. We used different cross correlation between users and in the same time we use different cross correlation between the channel paths in our BER performance curves.

Figure 4.14 to figure 4.17 demonstrate the BER performance when the cross correlation between the spreading codes of the different users is 0.15 and the cross correlation of the fading process at different antennas is equal to 0. Figure 4.18 to figure 4.21 illustrate the BER performance for the same spreading code cross correlation, but with channel cross correlation equal to 0.5. In the same manner, figure 4.22 to 4.25 demonstrate the BER performance when the cross correlation between the spreading codes of the different users is 0.25 and cross correlation between paths is equal to 0. Figure 4.26 to Figure 4.29 show the BER performance for cross correlations between the spreading codes of the different users of 0.25, with path cross correlations equal to 0.5. Figure 4.30 to Figure 4.33, present the BER performance when the cross correlation between the spreading codes of the different users is 0.35 and path cross correlation equal to 0. Lastly, Figure 4.34 to Figure 4.37 illustrate the BER performance for the same cross correlation between the spreading codes of the users, but with path cross correlation equal to 0.5.

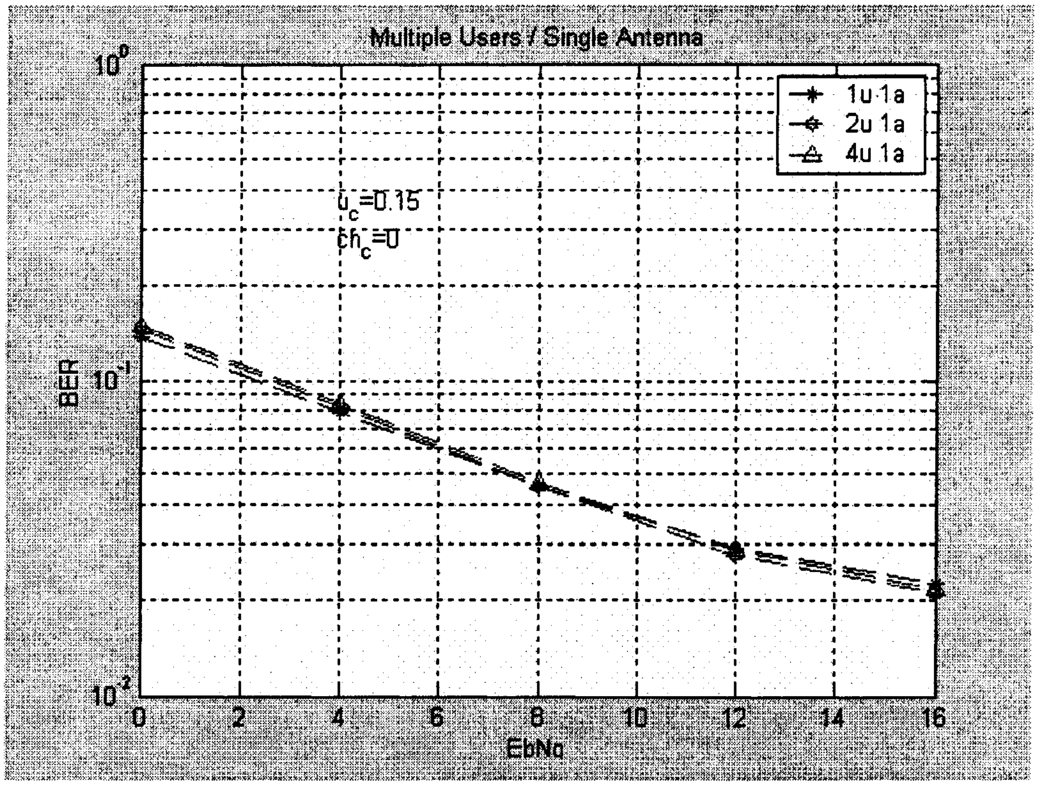


Figure 4.14: Performance of multiple users of single receive antenna with ch-corr=0 and user-corr=0.15

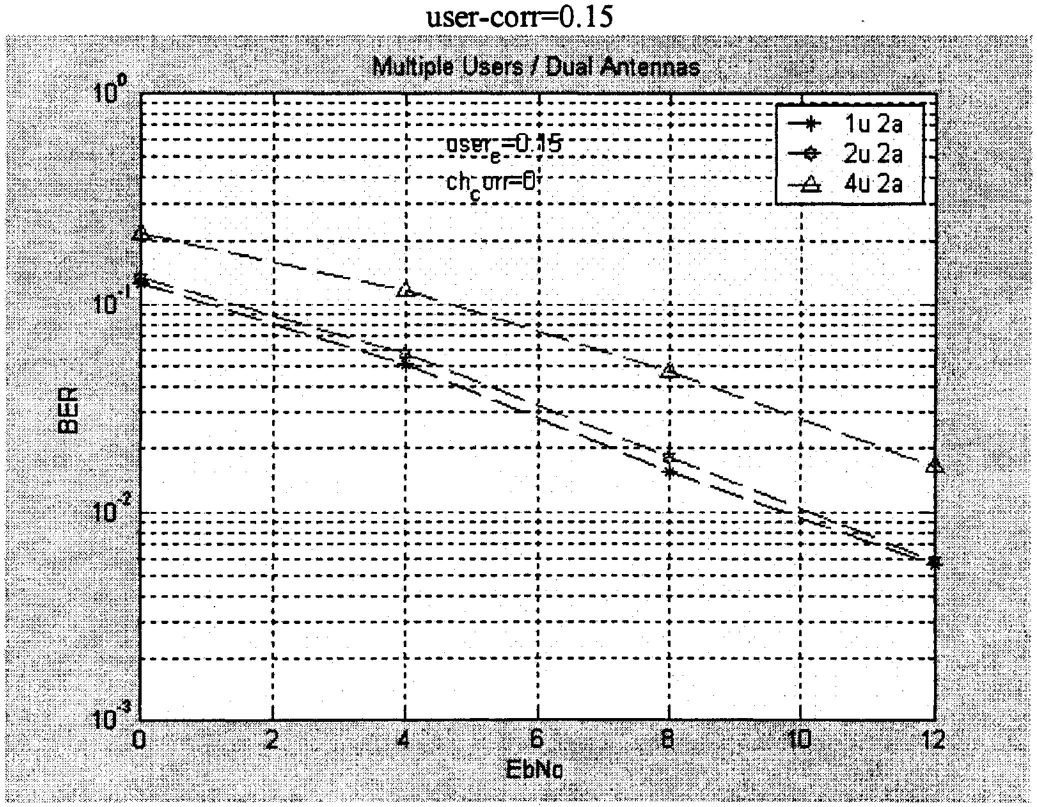


Figure 4.15: Performance of multiple users of two antennas with ch-corr=0 and user-corr=0.15

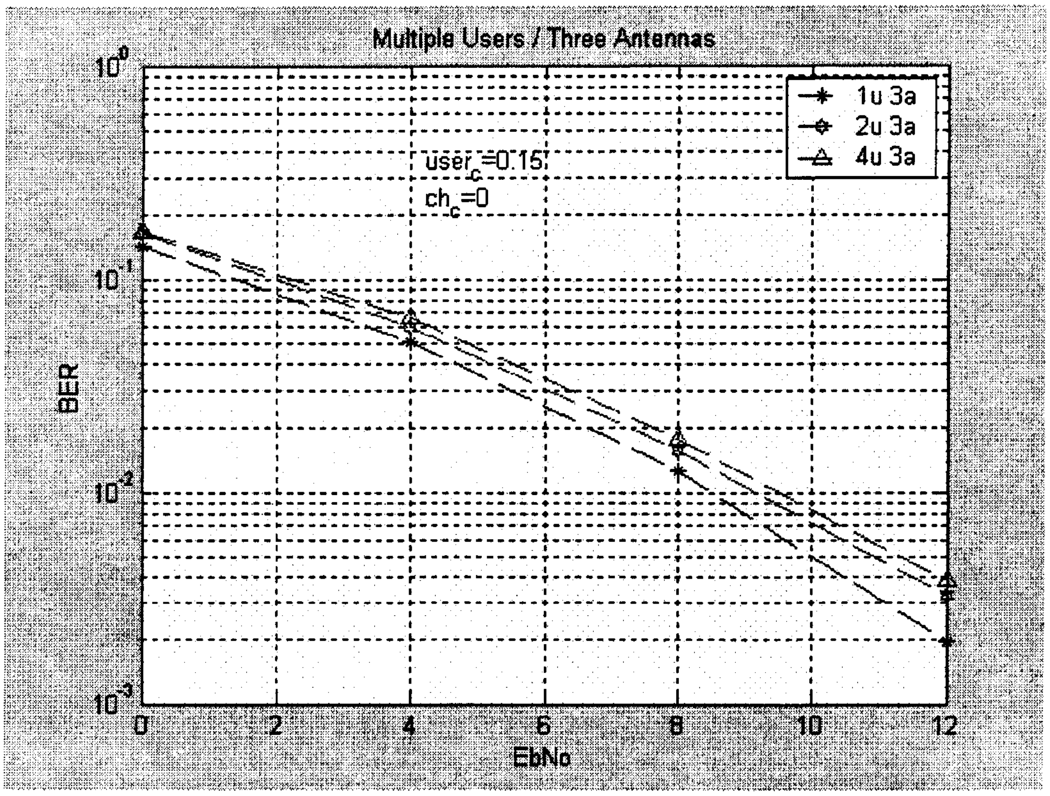


Figure 4.16: Performance of multiple users of three antenna with ch-corr=0and user-corr=0.15

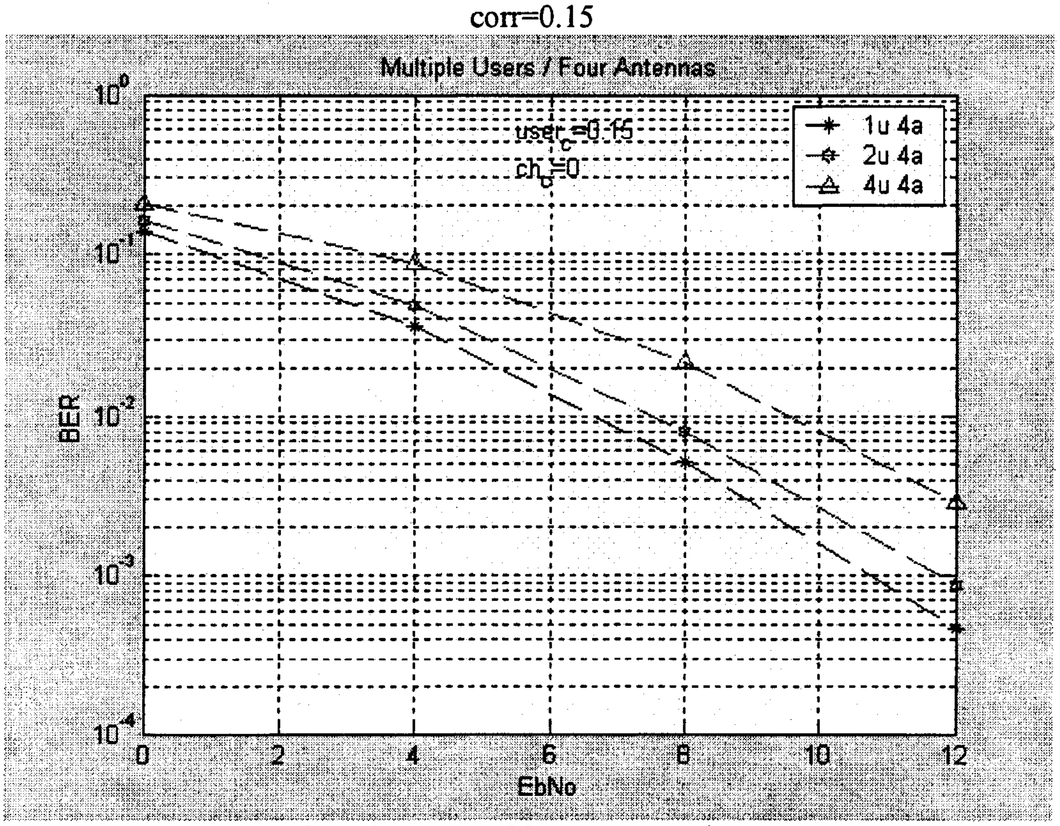


Figure 4.17: Performance of multiple users of four antennas with ch-corr=0 and user-corr=0.15

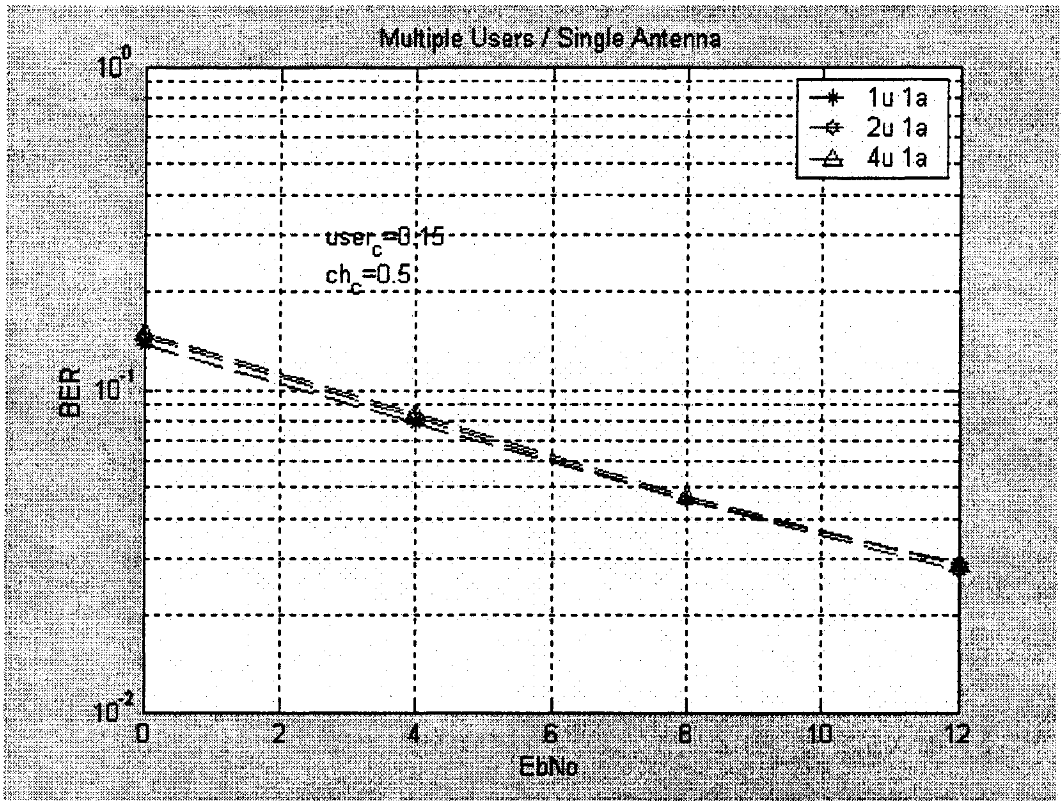


Figure 4.18: Performance of multiple users of single antenna with ch-corr=0. 5 and user-corr=0.15

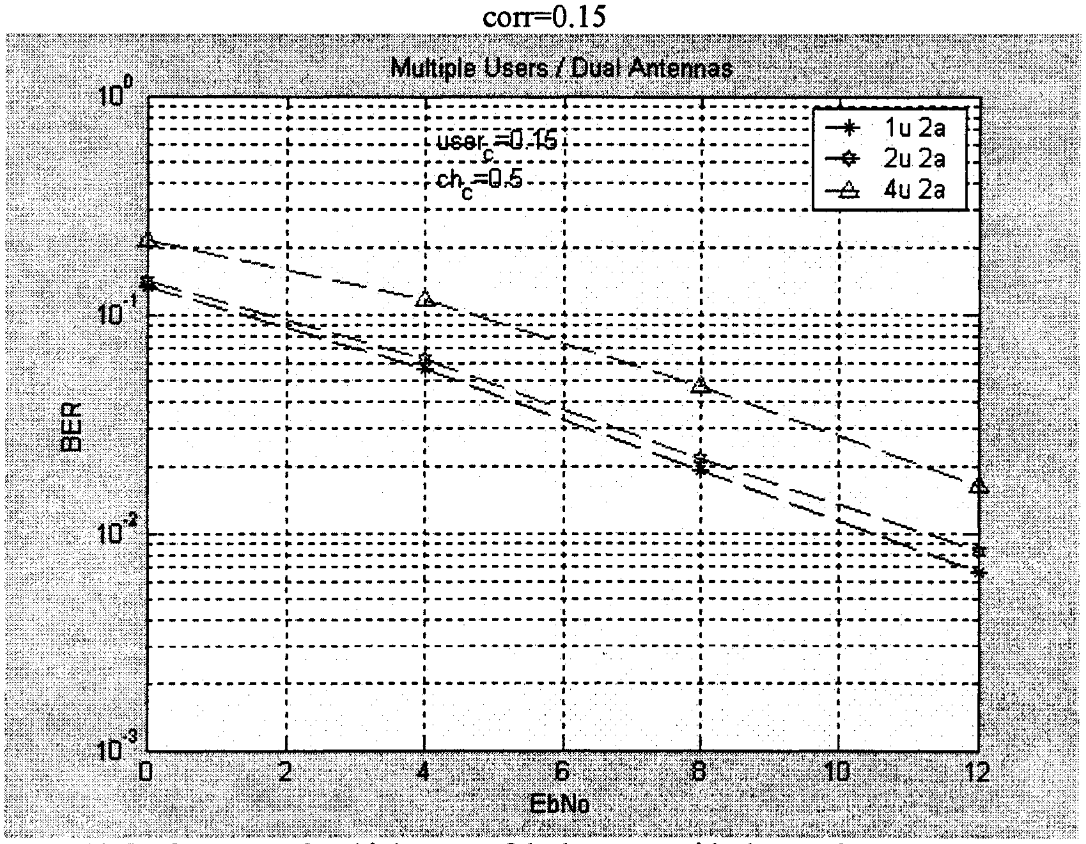


Figure 4.19: Performance of multiple users of dual antenna with ch-corr=0. 5 and user-corr=0.15

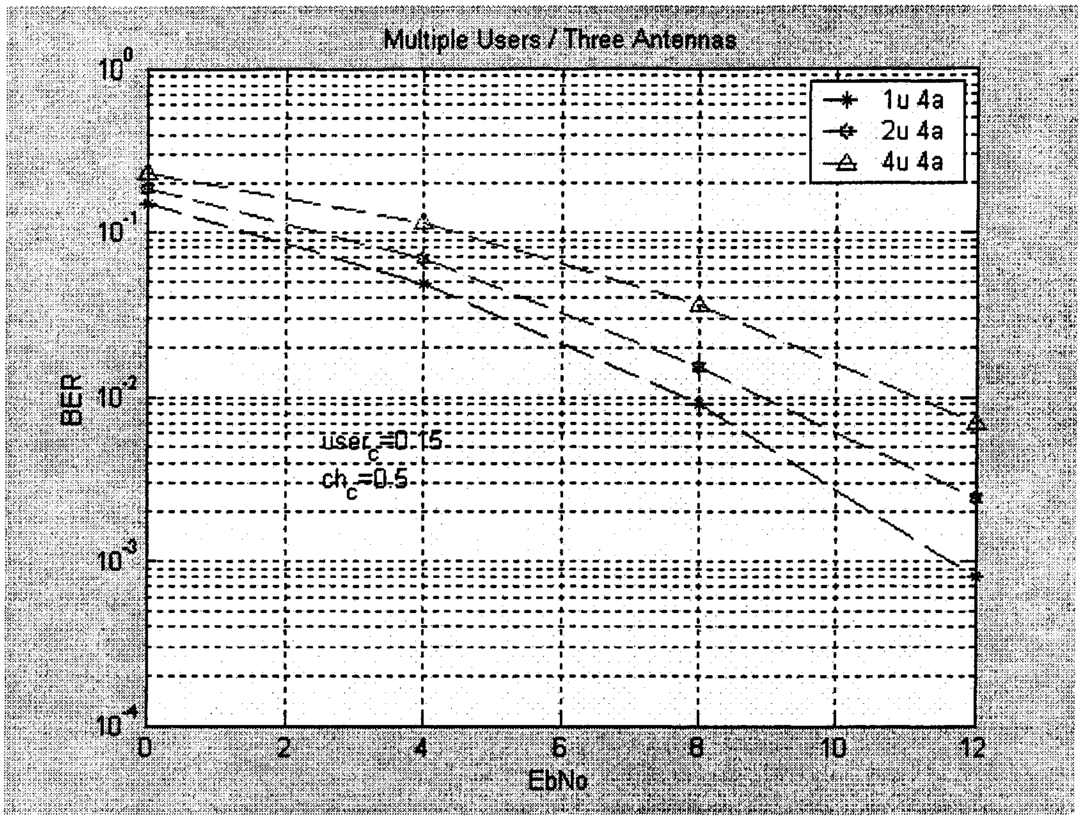


Figure 4.20: Performance of multiple users of three antenna with ch-corr=0. 5 and user-corr=0.15

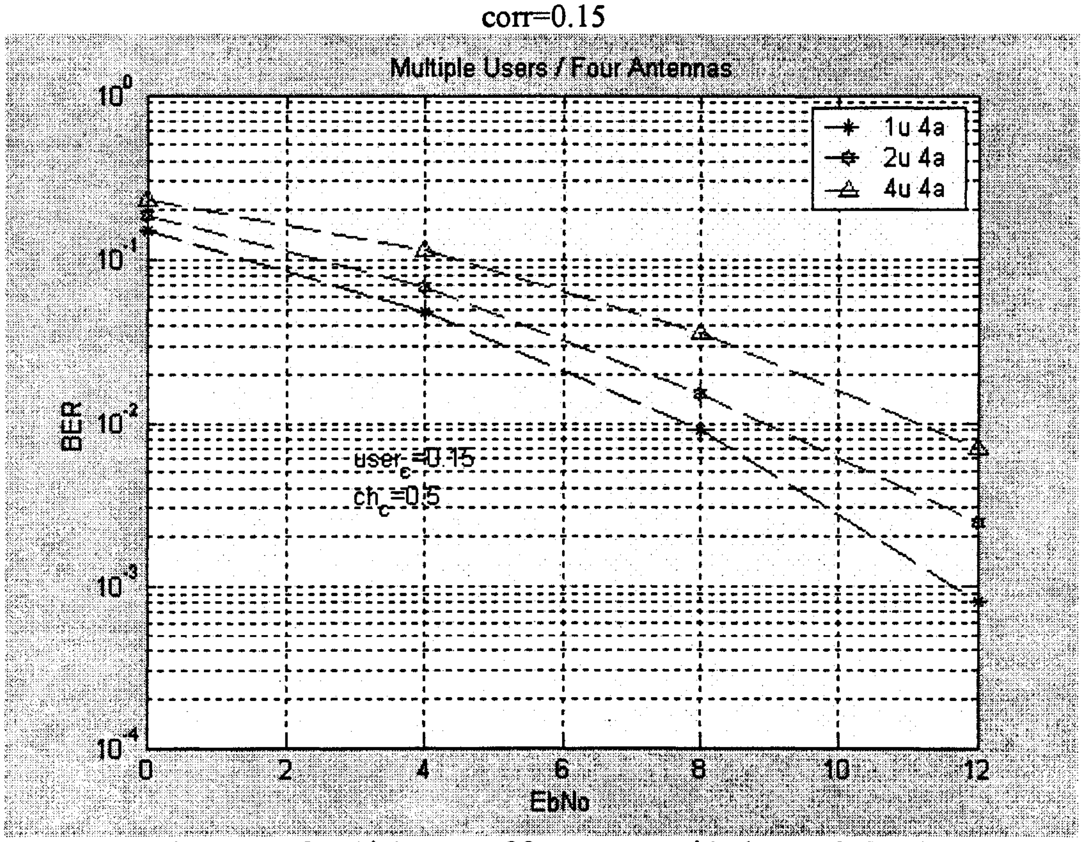


Figure 4.21: Performance of multiple users of four antenna with ch-corr=0. 5 and user-corr=0.15

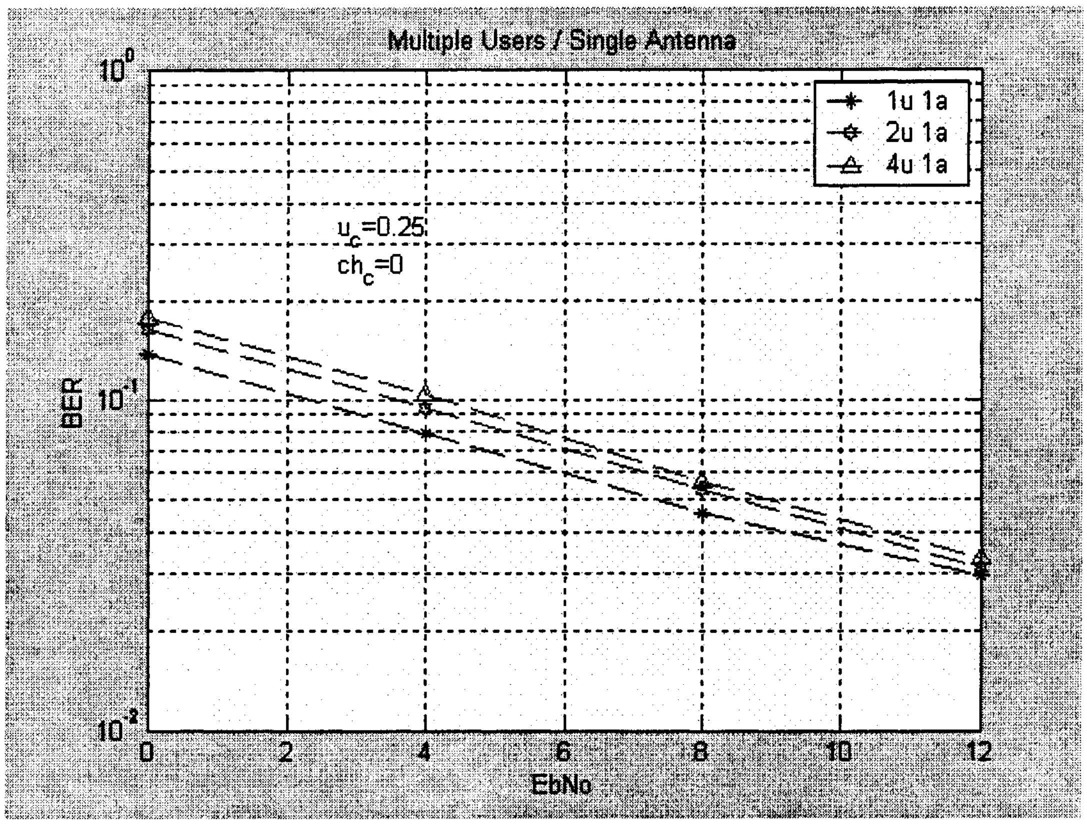


Figure 4.22: Performance of multiple users of single antenna with ch-corr=0 and user-corr=0.25

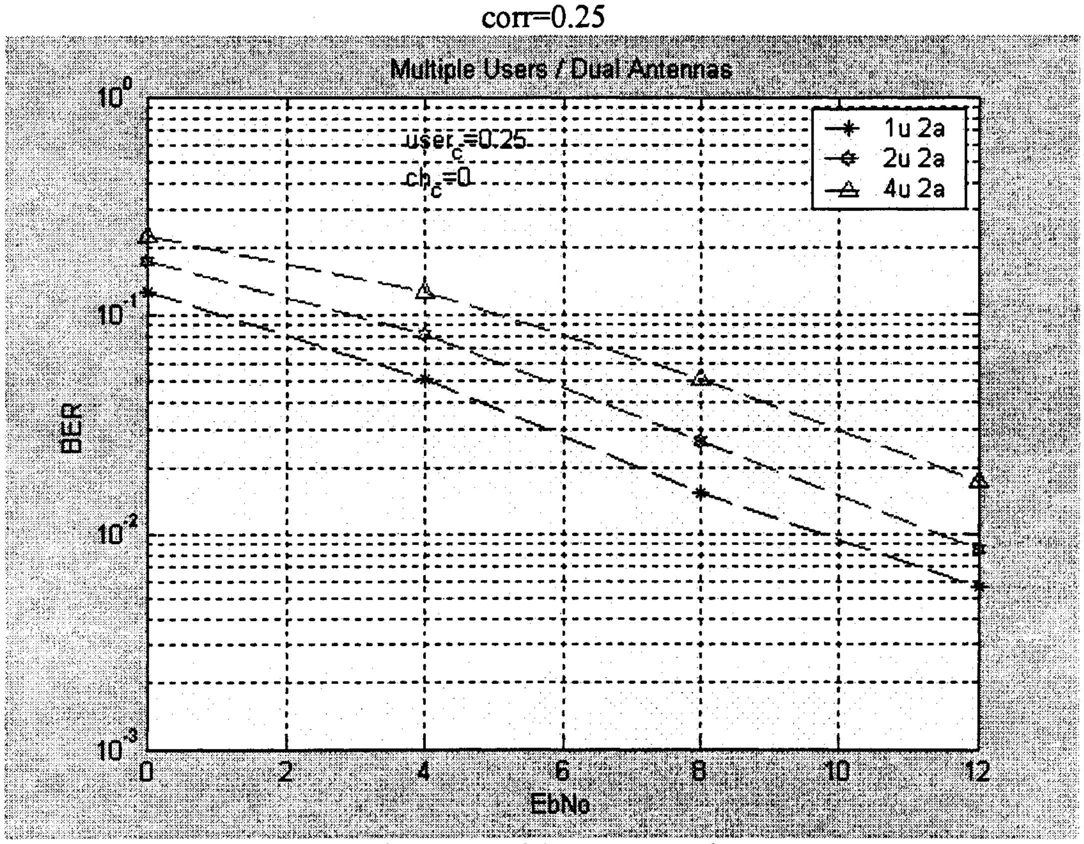


Figure 4.23: Performance of multiple users of dual antenna with ch-corr=0 and user-corr=0.25

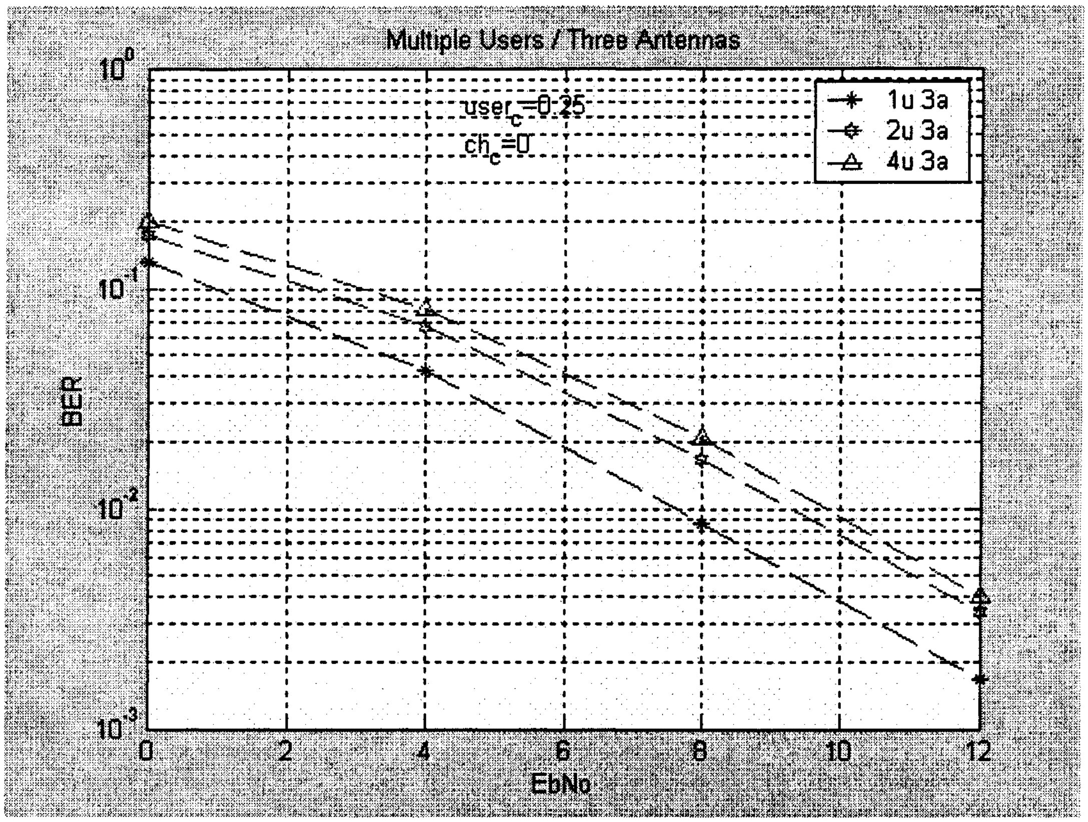


Figure 4.24: Performance of multiple users of three antenna with ch-corr=0 and user-corr=0.25

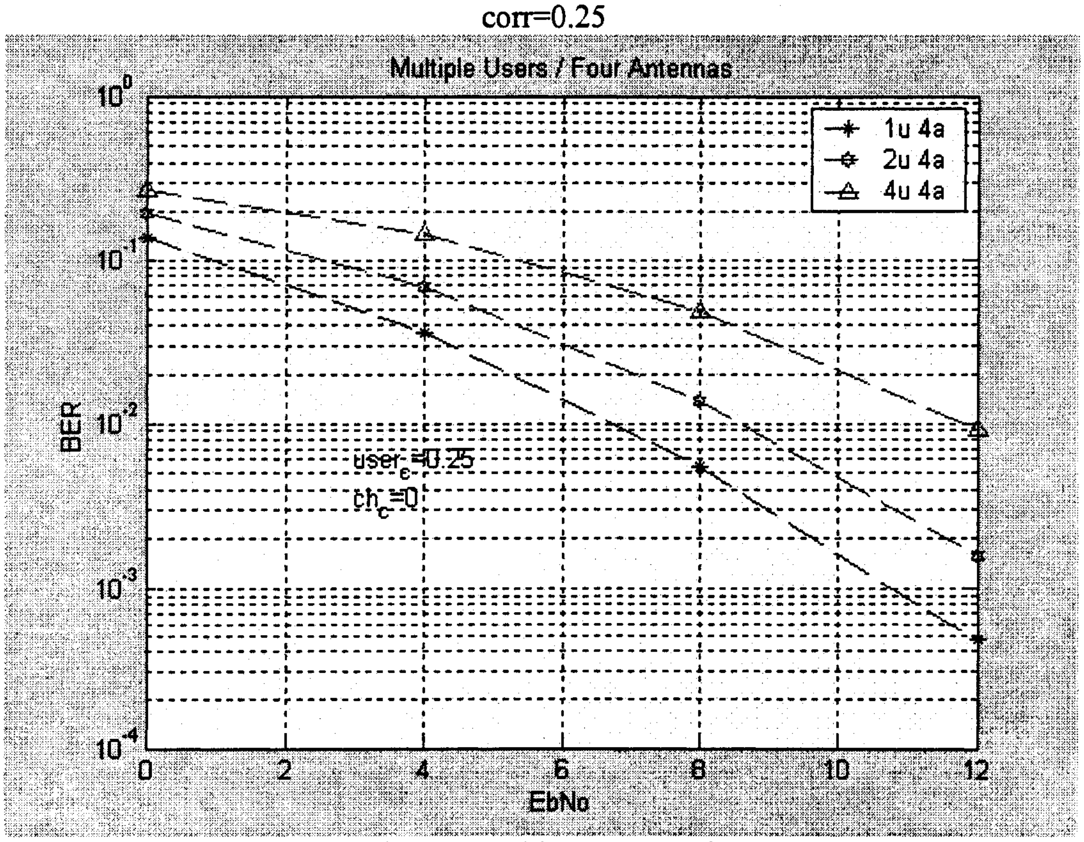


Figure 4.25: Performance of multiple users of four antenna with ch-corr=0 and user-corr=0.25

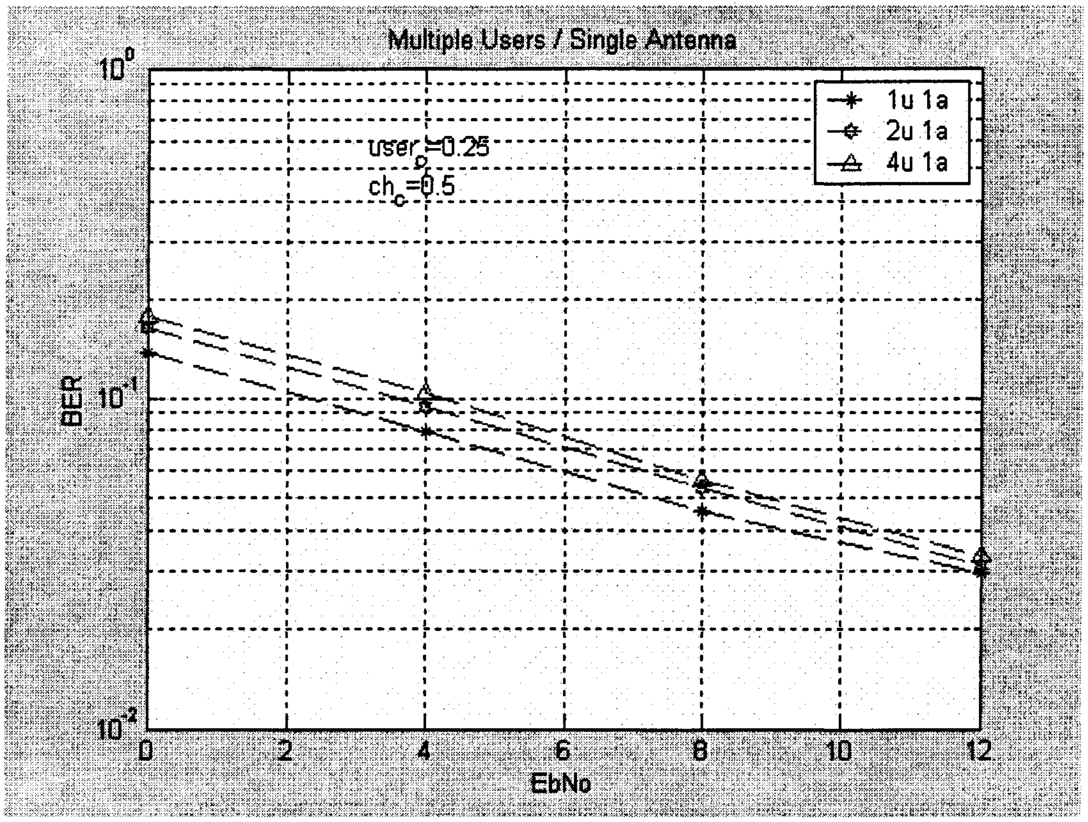


Figure 4.26: Performance of multiple users of single antenna with ch-corr=0.5 and user-corr=0.25

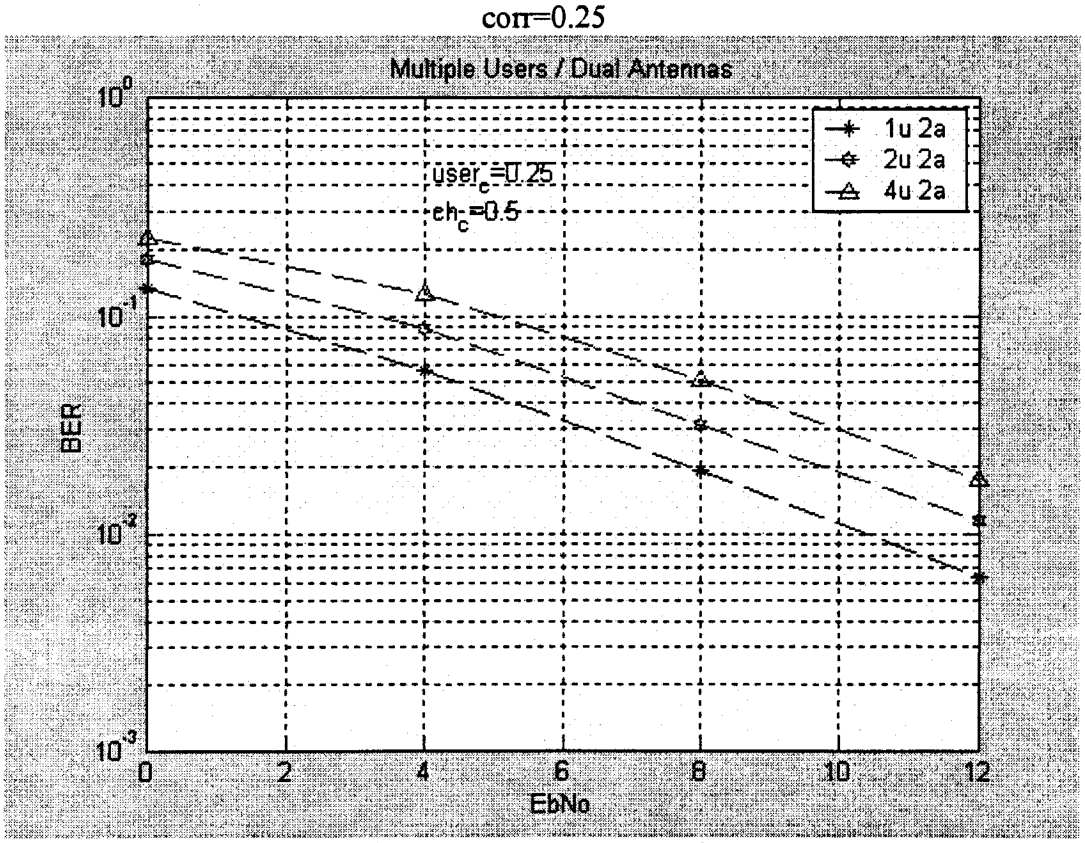


Figure 4.27: Performance of multiple users of dual antenna with ch-corr=0.5 and user-corr=0.25

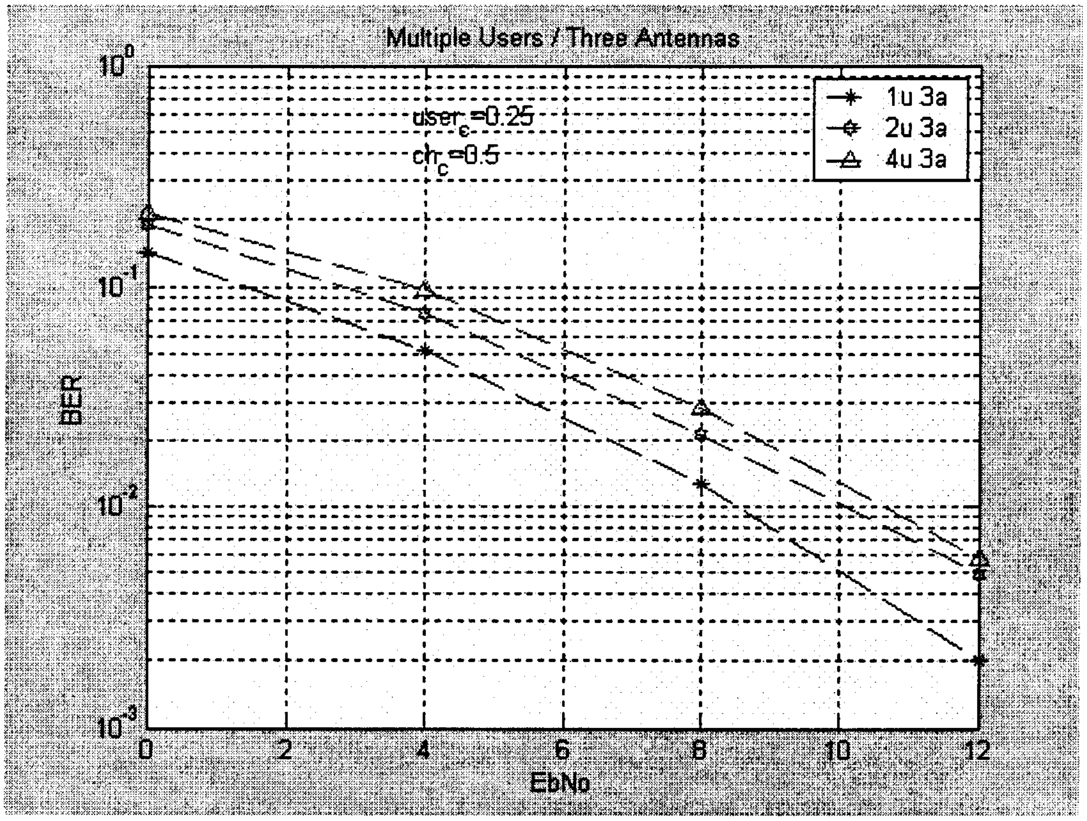


Figure 4.28: Performance of multiple users of three antennas with ch-corr=0. 5 and user-corr=0.25

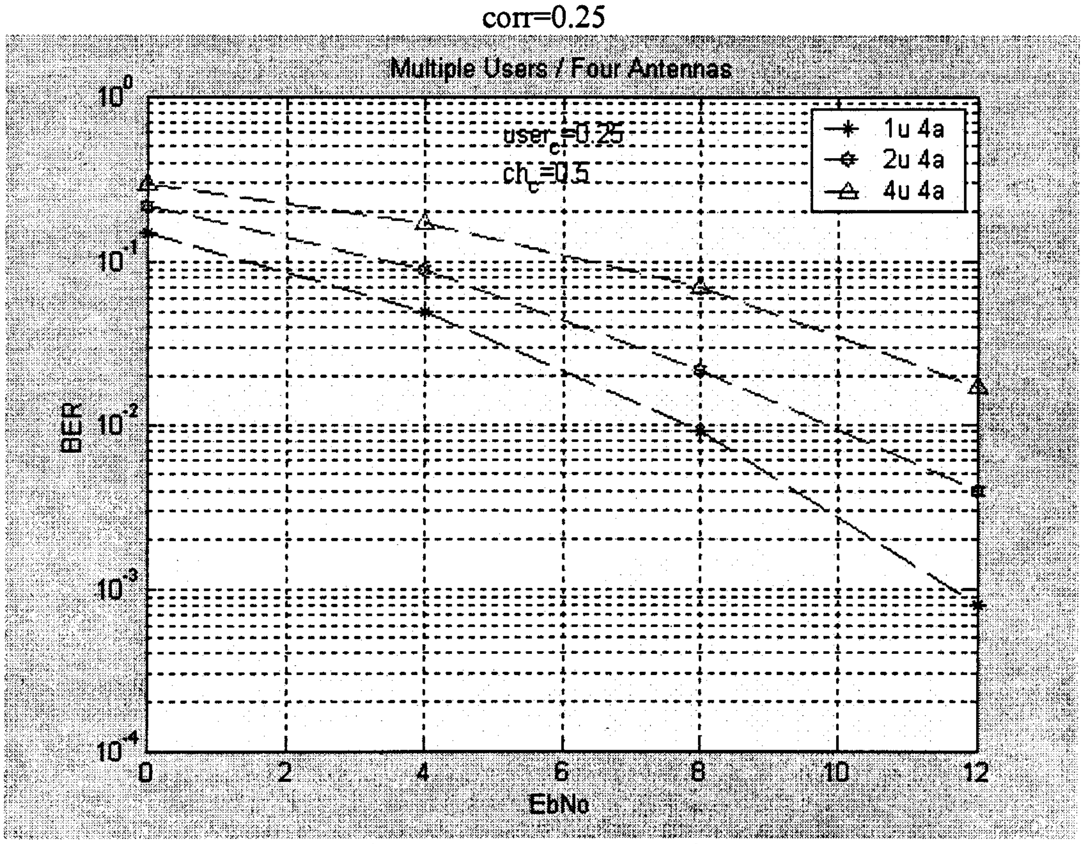


Figure 4.29: Performance of multiple users of four antenna with ch-corr=0. 5 and user-corr=0.25

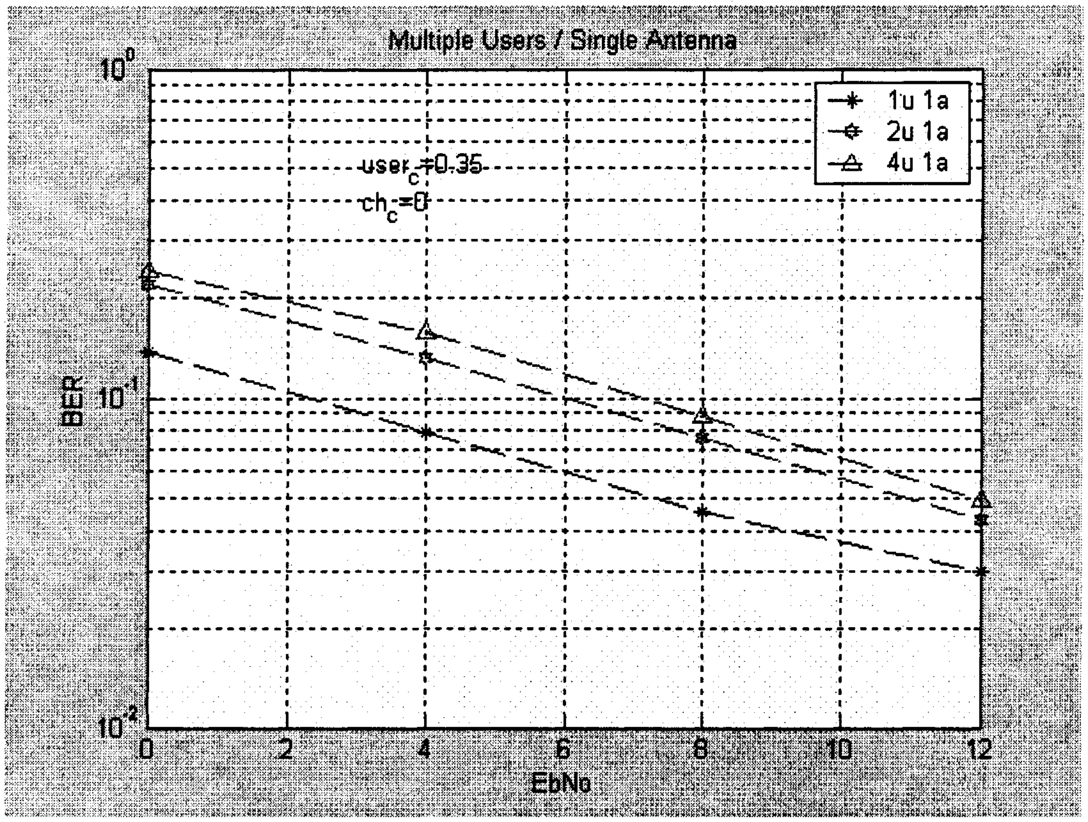


Figure 4.30: Performance of multiple users of single antenna with ch-corr=0. 5 and user-corr=0.35

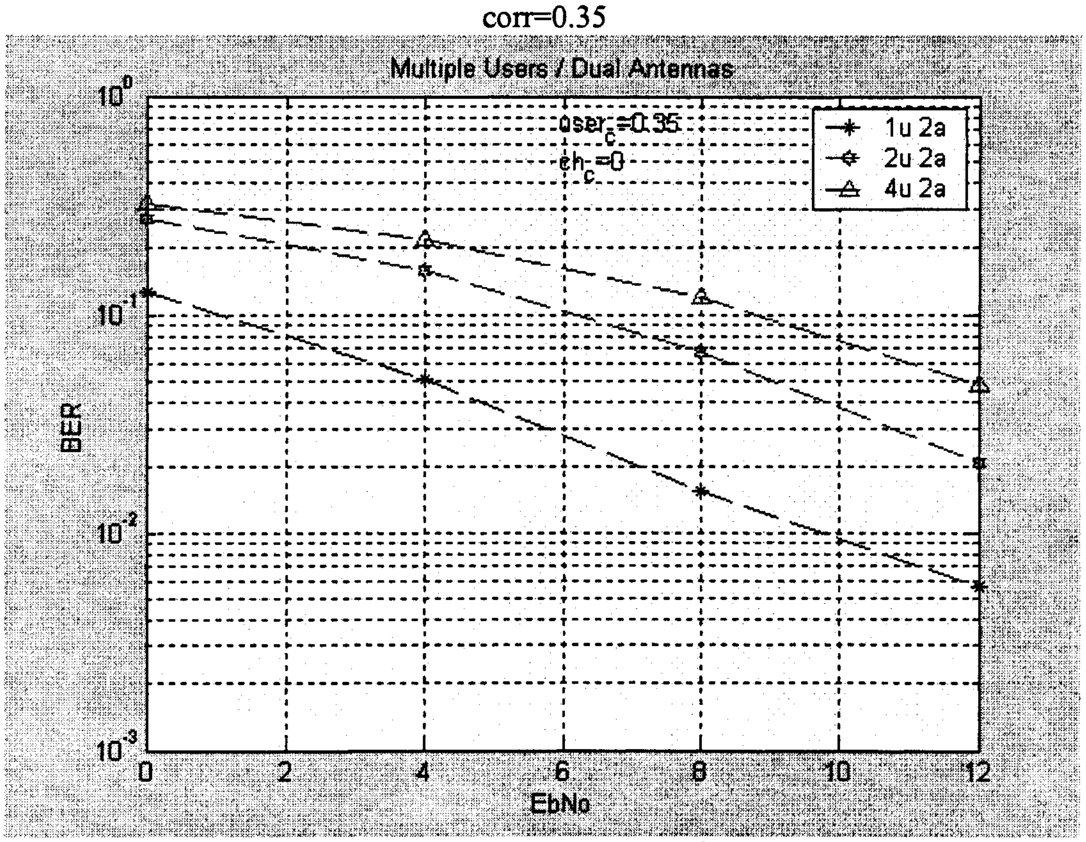


Figure 4.31: Performance of multiple users of Dual antenna with ch-corr=0. 5 and user-corr=0.35

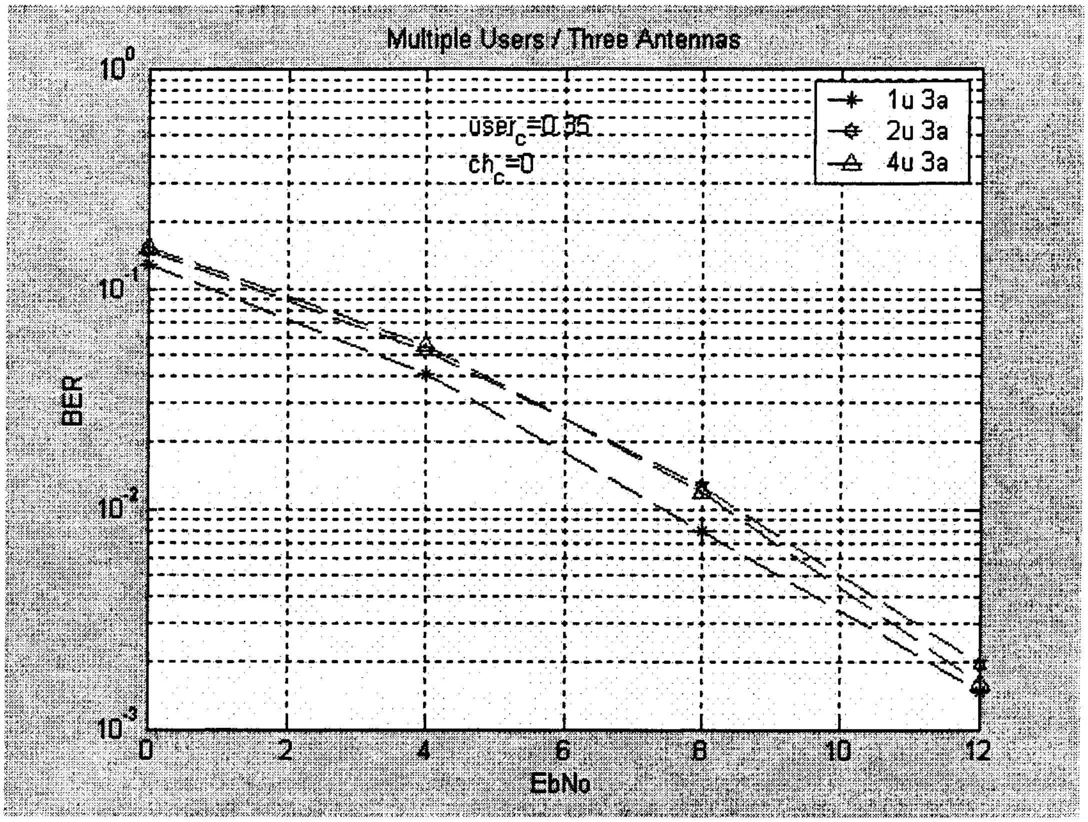


Figure 4.32: Performance of multiple users of three antenna with ch-corr=0. 5 and user-corr=0.35

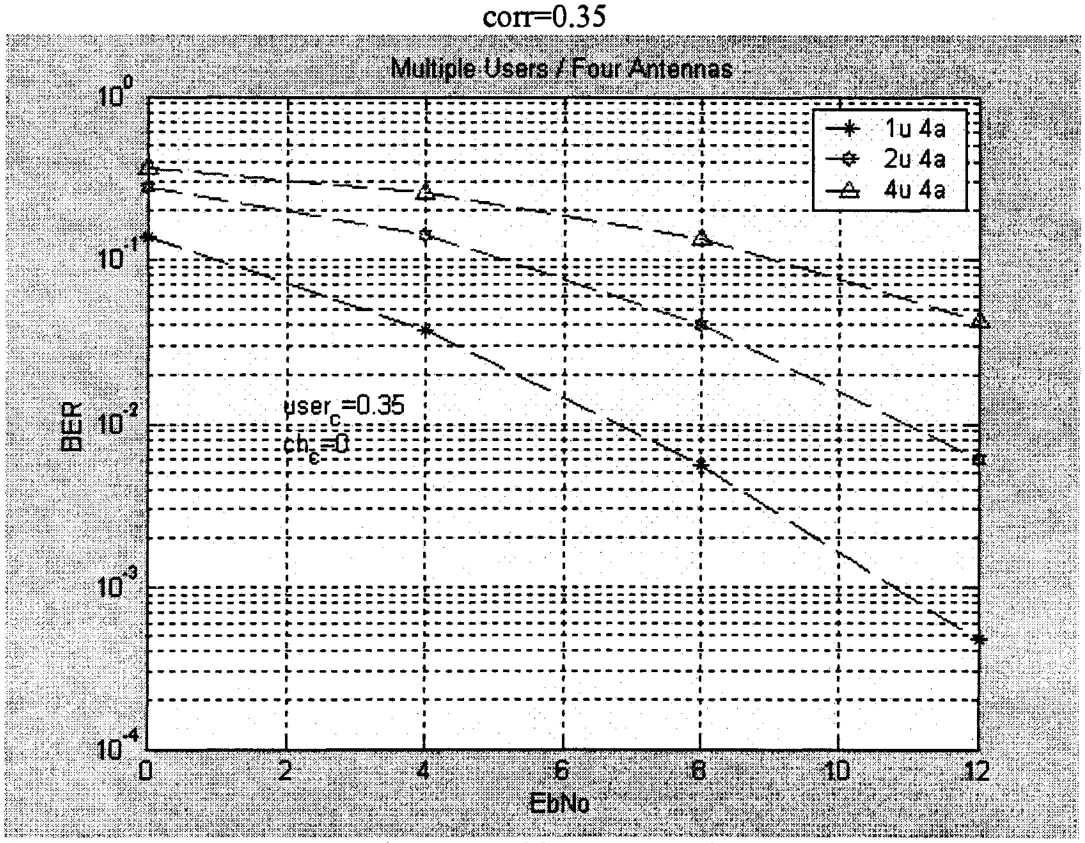


Figure 4.33: Performance of multiple users of four antenna with ch-corr=0 and user-corr=0.35

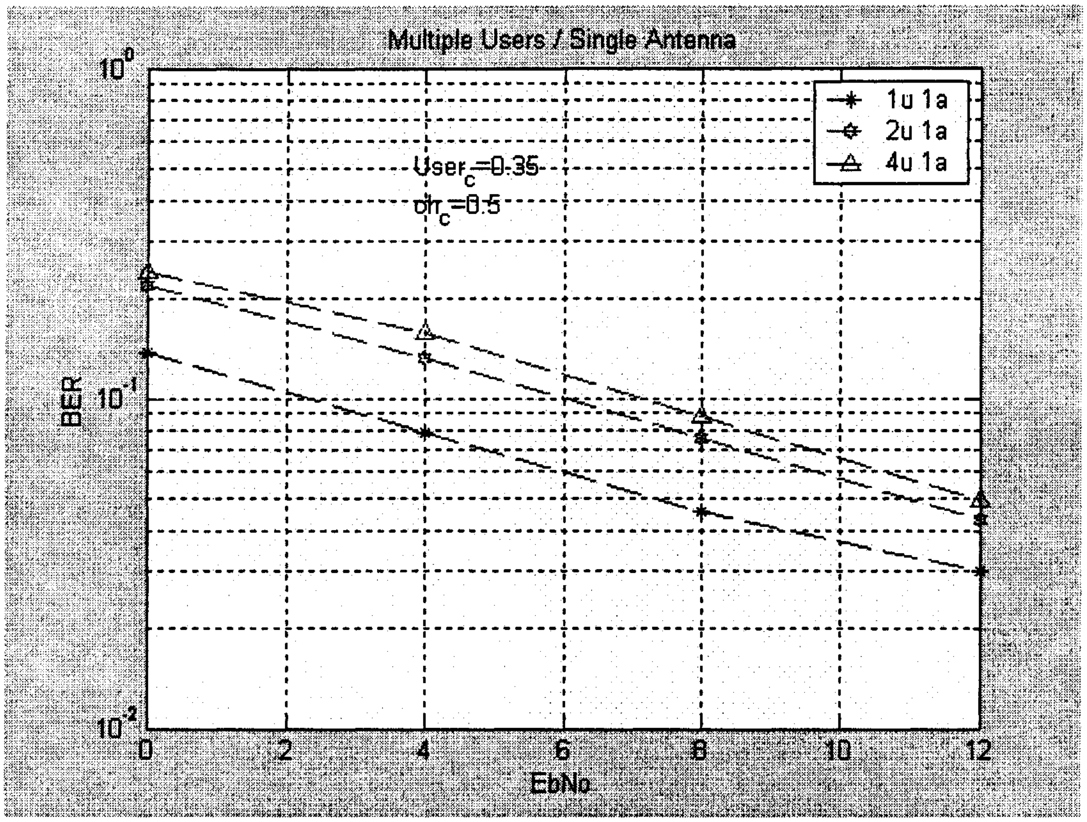


Figure 4.34: Performance of multiple users of single antenna with ch-corr=0. 5 and user-corr=0.35

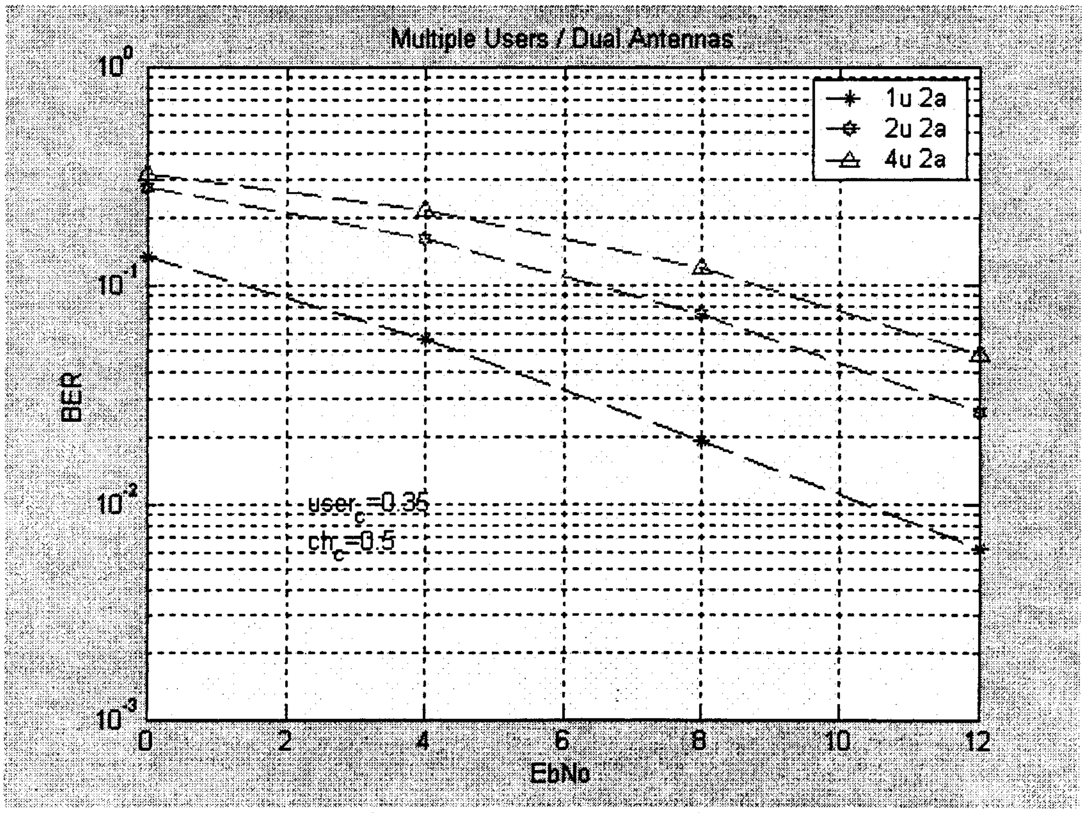


Figure 4.35: Performance of multiple users of dual antenna with ch-corr=0. 5 and user-corr=0.35

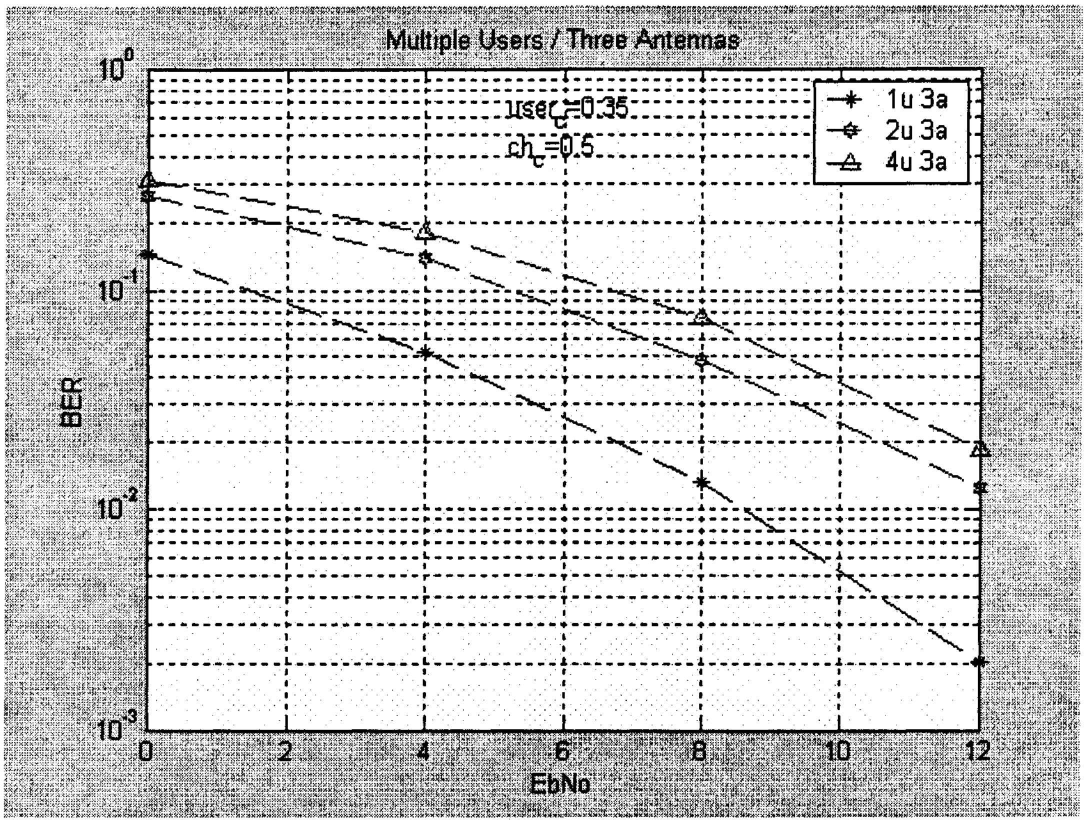


Figure 4.36: Performance of multiple users of three antenna with ch-corr=0. 5 and user-corr=0.35

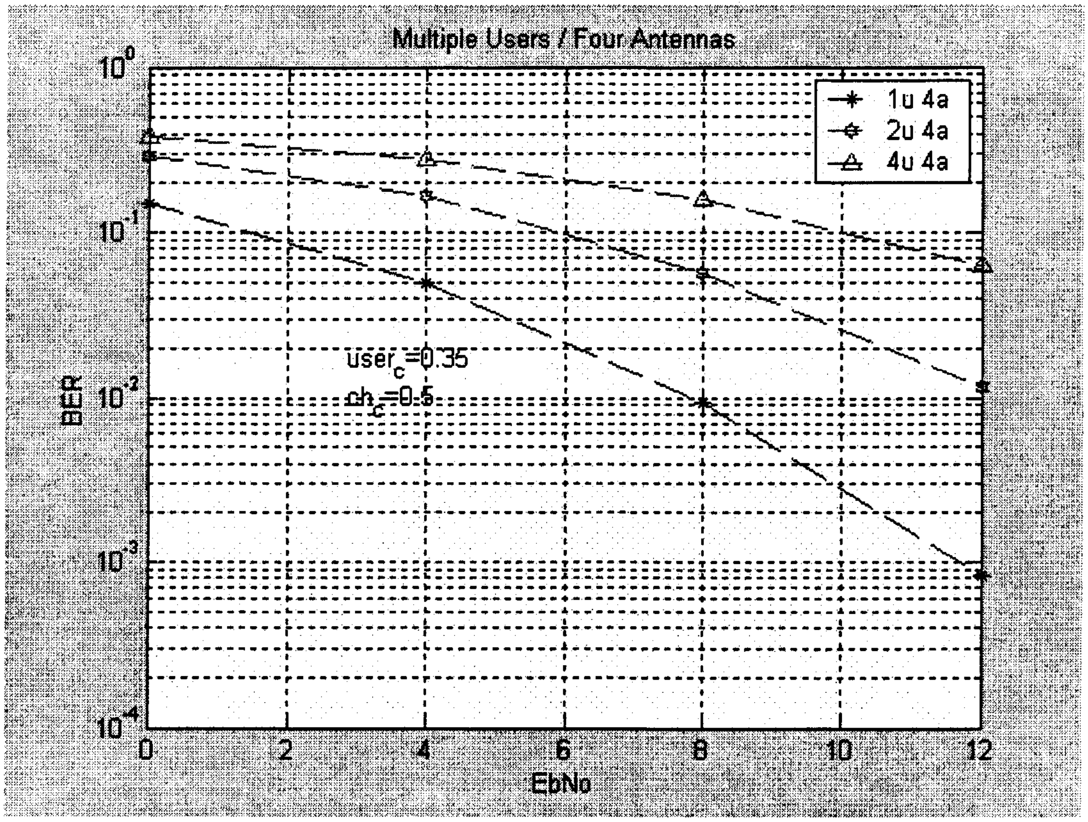


Figure 4.37: Performance of multiple users of four antenna with ch-corr=0. 5 and user-corr=0.35

When the cross correlation between the users is kept low, the ability of the system to remove the MAI is high as seen in figure 4.14 compared with figure 4.31. In figure 4.14, when the user cross correlation is about 0.15 (approximately the processing factor is 40), the performance of higher numbers of MAI users is almost similar to the performance if we have only one single user. But if we increase the user correlation to 0.35, we will see from figure 4.31 that the performance of the system with more than two interfering users shows significant degradation compared to the single user performance. The reason for this result is that as the cross correlation between signature waveforms of the different users increases, the noise enhancement caused by the decorrelator increases. This is because the noise enhancement is proportional to the cross-correlation values in the inverse correlation matrix.

The channel paths between the transmitters and the different receive antennas may be correlated. In our simulations, we use different values for the cross correlation between the channel gains, that represents how much correlation exists between paths that are departing from the same transmit antenna and arriving at different receive antennas. The simulations show a slow degradation in the performance as we increase the cross correlation from 0 (uncorrelated) to 0.5. For example, in the case of two users with four receiver antennas and for a fixed correlation between users of 0.15, to obtain a BER of 10^{-2} we need an E_b/N_o of 7 dB when the paths are uncorrelated as seen in figure 4.17. When the paths have a 0.5 correlation, for the same BER, we now need an E_b/N_o of 9 dB as seen in figure 4.21.

Another observation was that an increase in the cross correlation of channel gains has a smaller effect on the performance of our proposed system, compared to an increase in cross correlation between the spreading codes of the different users. For instance, in figure 4.37, when the channels are heavily correlated (i.e. channel correlation = 0.5), the E_b/N_o required to obtain a BER of 10^{-2} is 12dB, while for the same user correlation but with uncorrelated fading paths we need E_b/N_o = 11dB at the same BER as seen in figure 4.33. In other words, the increased correlation between paths only causes 1dB degradation. In contrast, if we fix the path cross correlation to 0.5 and change the user cross correlation from 0.25 as in figure 4.27 to 0.35, the simulation results show that at cross correlation of 0.25 we need Eb/No of 9.5dB at BER of 10^{-2}, while for cross-correlation of 0.35, E_b/N_o needs to be 12dB for a BER of 10^{-2} as seen in figure 4.37. The explanation of the results is that as we increase the amount of cross correlation between users, the background noise will be enhanced since there is strong relationship between the user cross correlation with the decorrelator cross correlation matrix, as previously mentioned. On the other hand, theoretically, since our propose system tries to have a good estimation of the channel paths, a slow degradation in the performance will appear even when the paths are correlated.

4.3.5 Space receiver diversity

Figure 4.38 shows the advantage in increasing the number of receiver antenna elements. The simulation is done for two users with cross correlation equal to 0.25 transmitting through a channel with cross correlation equal to 0 between its paths. The results shows that the performance of having two users with three antenna elements is better than having two users with one antenna as. In other word, we will achieve more space receiver diversity gain as we increase the receiver antenna array.

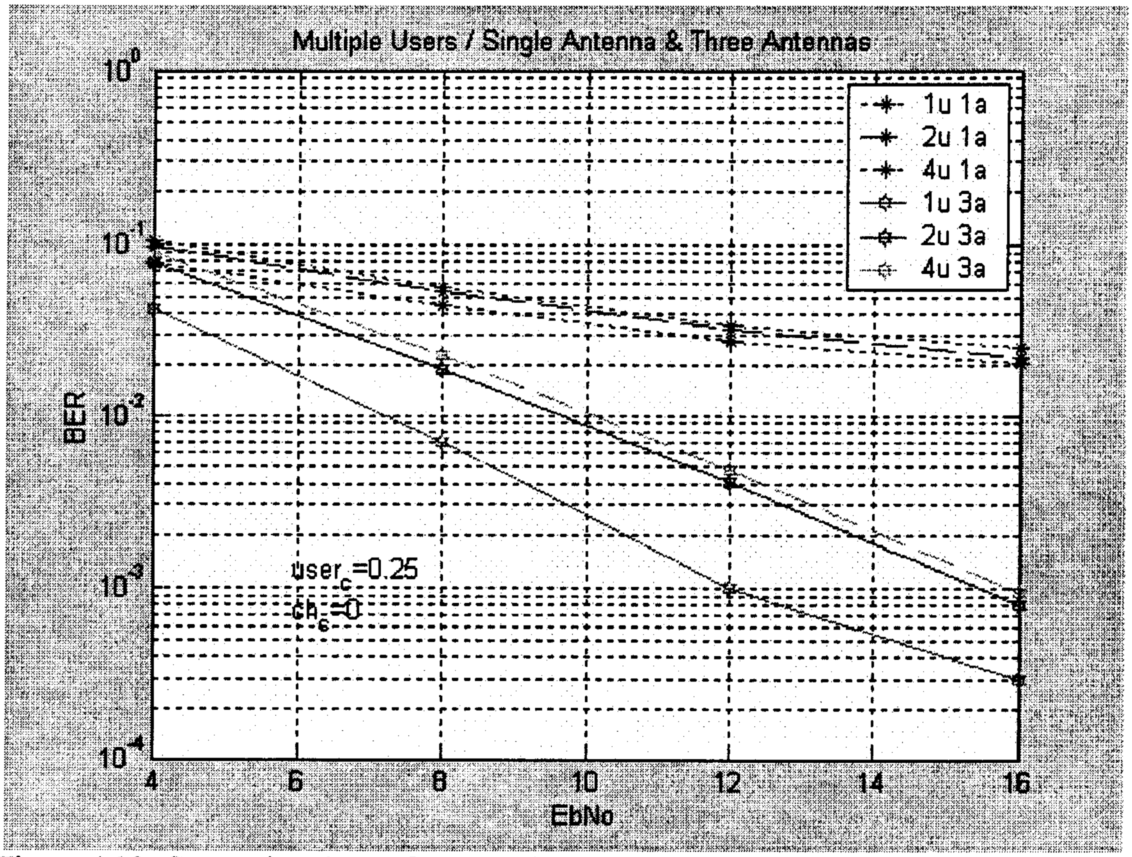


Figure 4.38: Comparing the performance between a single antenna and three antennas of our proposed receiver

4.3.6 Proposed system and conventional system

After presenting our proposed algorithm with several users combined with several antenna elements, we would like to highlight the advantages and disadvantages of using the linear decorrelator detector in our proposed system, by comparing it with the a conventional receiver represented as a conventional multiple array receiver (CAA). This simulation was done for four users and three receiver antennas elements. The constellation diagram results for the conventional multiple array receivers at 4 dB and 16dB are presented in figure 4.39 and in figure 4.41 respectively. At the same time the constellation diagram results of the proposed receiver at 4 dB and 16dB are shown in figure 4.40 and in figure 4.42 respectively. Then the BER performance for both systems is shown in figure 4.43.

The conventional receiver shows that it has better or same BER performance than our proposed system for E_b/N_o below 8dB, which can be seen in figure 4.43. For example, at 4dB, the constellation of our proposed system as in figure 4.39 is almost the same or worse than the constellation of CAA as in figure 4.40. The reason of no improvement in the performance of the proposed receiver is drawback of using the decorrelater detector in our received structure. As we mentioned earlier in chapter 3, the decorrelater structure disadvantages comes from the enhancement of background noise that it generates at low E_b/N_o. However, beyond 8 dB, figure 4.43 will show significant improvement of BER of our proposed signal processing, sharper and cleaner constellation diagram than constellation diagram of the conventional receiver at 16 dB as in figure 4.42 and figure 4.41 respectively. In the same manner, in figure 4.43, the CAA will create an error floor when the Eb/No increase beyond approximately 12 dB, while our proposed receiver will continue to benefit from those increases in E_b/N_o.

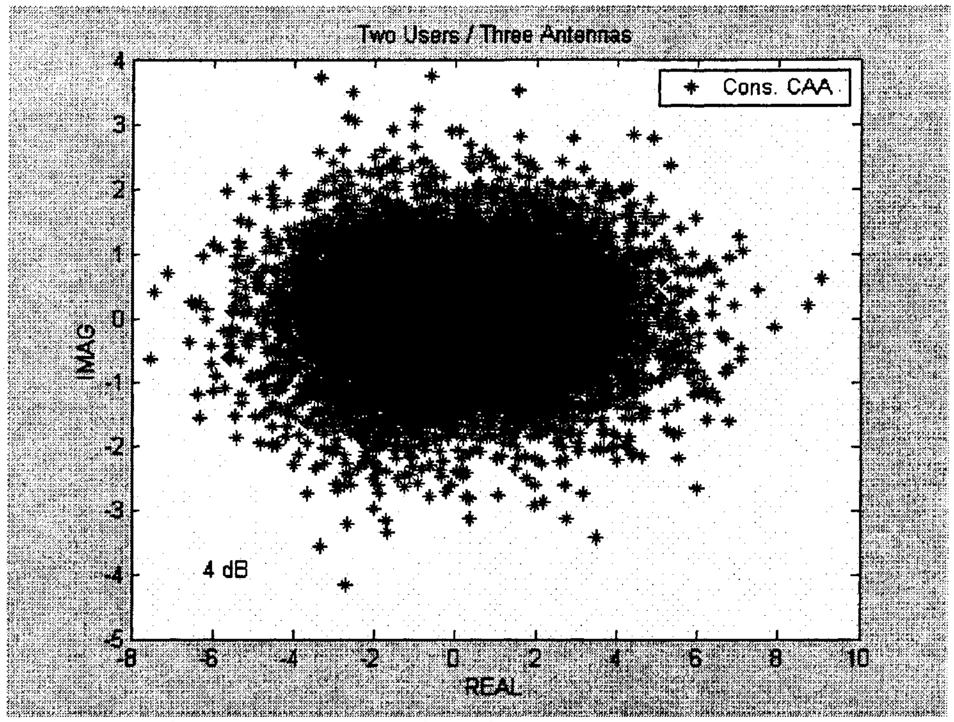


Figure 4.39: Constellation diagram of the CAA at 4 dB

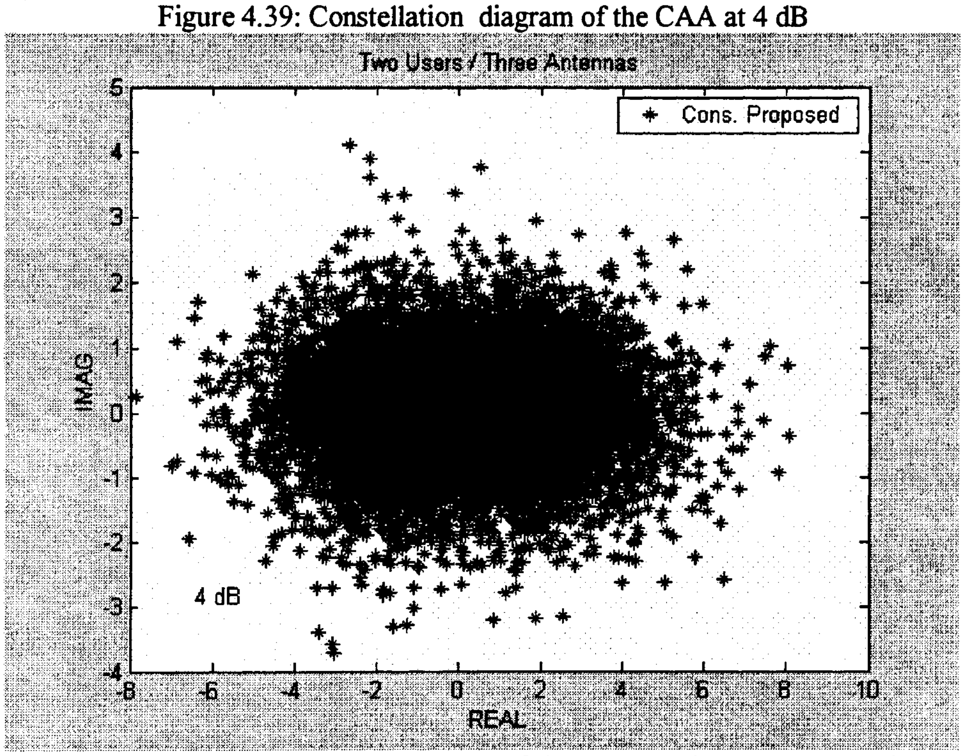


Figure 4.40: Constellation diagram of the proposed structure at 4 dB

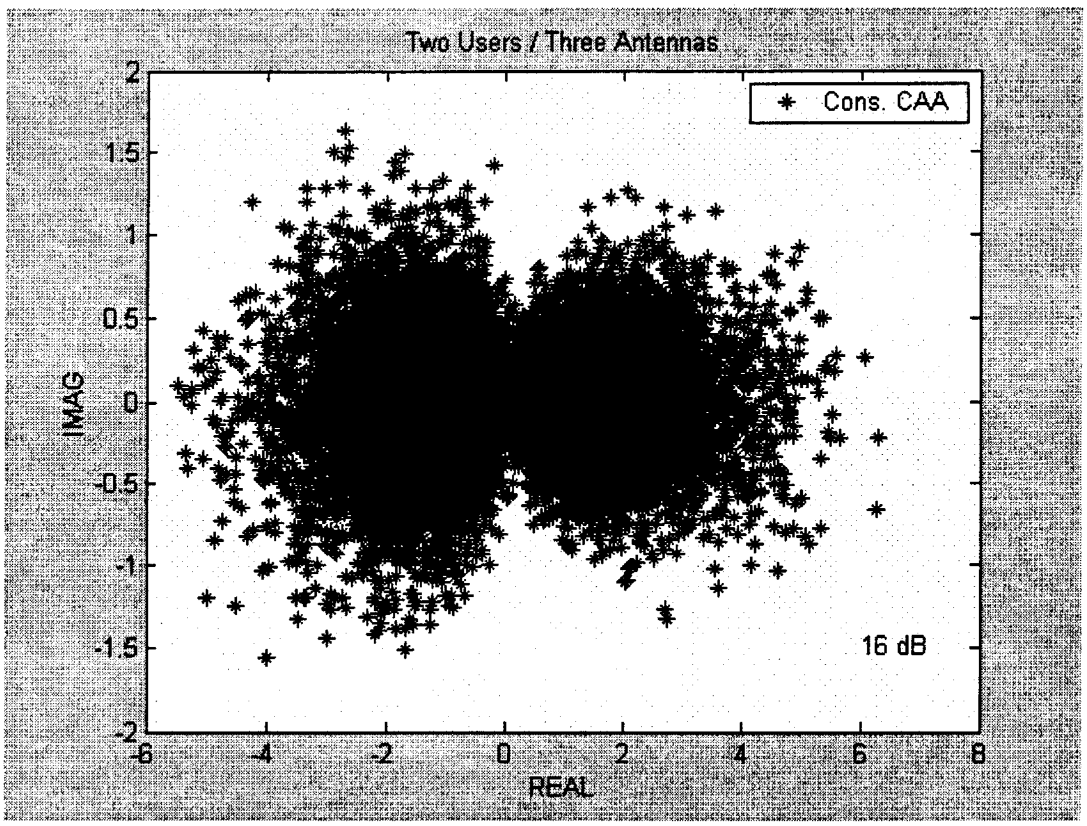


Figure 4.41: Constellation diagram of CAA at 16 dB

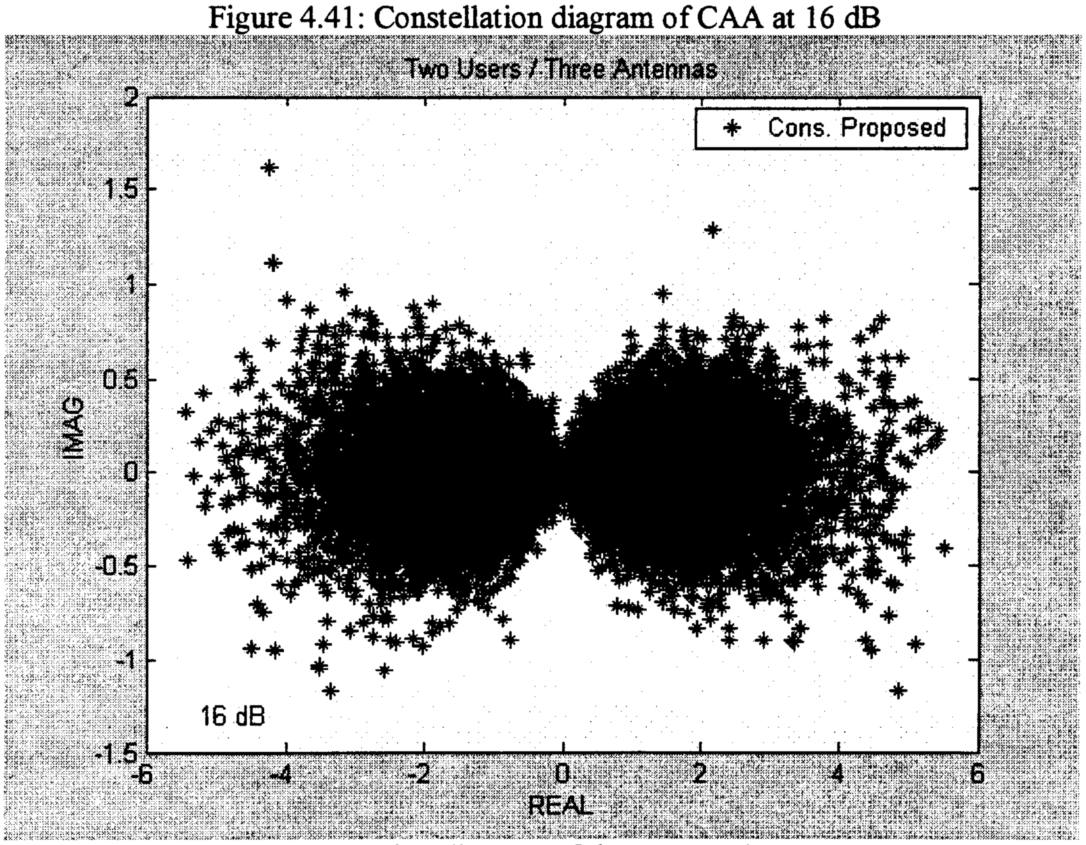


Figure 4.42: Constellation diagram of the proposed structure at 16 dB

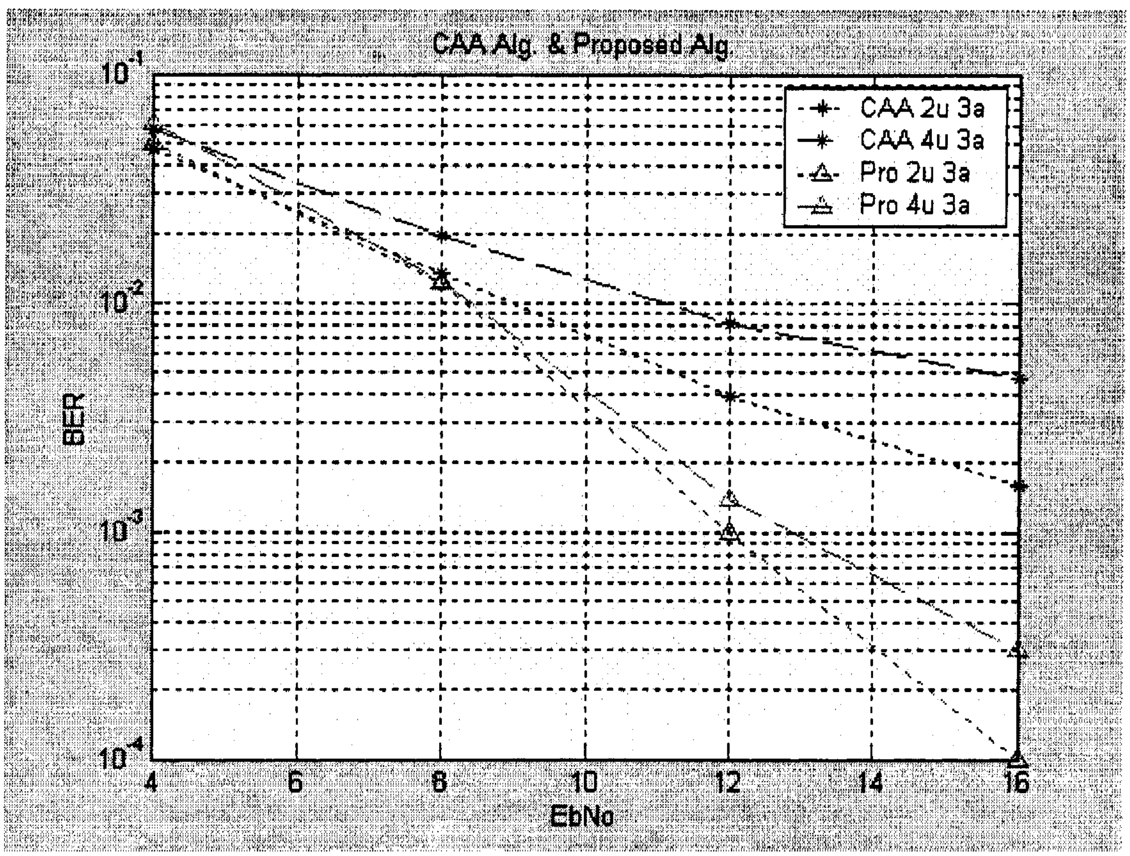


Figure 4.43: Comparing the performance between a conventional and proposed receiver with three antennas

Chapter 5

Conclusion and Future Work

5.1 Conclusion

The use of space-time processing is emerging as a powerful tool for improving the performance of wireless networks. Field deployment of this technology is likely to emerge soon since the use for more advanced multiple access techniques are needed in mobile communication. Space-time processing can improve cell coverage, enhance link quality, and increase system capacity for a rapidly fluctuating mobile channel with multipath delay and angle spreads that offers a significant challenge.

We have developed a new space-time processing system that can enhance and improve the performance of CDMA systems. The first phase of the algorithm was to smoothly join a linear multi-user detector (decorrelator detector) with multiple antenna array. The other proposed step in the algorithm was to estimate the time varying channel response by using the pilot symbol stream that is transmitted using a spreading code that is orthogonal and to the data signal's spreading code.

We have provided a detailed analysis for the proposed algorithm. The performance of the proposed algorithm has been compared with algorithms presented in literature. This algorithm showed the ability to substantially improve the BER performance compared to a conventional space-time processing detector. Our proposed system, although more complex than the conventional space-time processor discussed in Chapter 4, is low in computational complexity compared to other adaptive space-time processors, making it attractive to DSP developers. However, our proposed system suffers from background noise enhancement due to the decorrelator detector and the proposed system's performance is not optimized since we are not using beamforming weighting in spatial processing. In addition, we have some power loss in our performance since we need to allocate some of the signal's power to channel estimation.

In this thesis we demonstrated the use of a CDMA system with parallel pilot streams using multiple receive antennas with decorrelators at the output of each antenna in the array. The purpose of the decorrelator is to eliminate the MAI from the data as well as the pilot streams. The pilot streams are then used to estimate the channel gains of all users on each antenna. These estimated channel gains are used for phase correction and maximal ratio combination weighting. The important conclusions drawn from the results of this thesis are:

1) Low percentage of power needs to be assigned the the pilot stream. The proposed system needs only 15% the total transmitted power to be allocated to the pilot stream in order to estimate the channel gains for a fading rate $B_d = 0.1/T_b$.
2) A simple and efficient way to estimate the channel gains was obtained by transmitting a pilot stream using a spreading code that is orthogonal to that of the data stream and then applying the received pilots to a decorrelator, along with the

data. This yields reliable channel estimates, which in turn translates into good BER performance.

3) Space diversity was achieved. The simulation shows that as we increase the number of receiver antennas in the proposed structure, the more diversity gain we will get that in turn will improve our bit error rate.
4) The system provides excellent resistance to the near far problem. By using the decorrelator detectors at the output of each antenna array element, the simulations show that our algorithm is quite resistant to the near far problem.
5) Comparison was done between our proposed system and a conventional system. The study showed that at higher E_b/N_o our proposed system will continue to improve its performance, while an error floor occurs in the conventional antenna arrays.
6) In terms of simulations that were done for the single, dual, three, and four antenna receiver using different cross correlation values for the channel gains and for the users, the study shows that as we increase the cross correlation between users' spreading codes, the BER performance of the multi-user system degrades compared to the single user system. While increasing the cross correlation of channel vector (more dependent channel vectors), the overall performance for any users and any antenna array will slowly degrade.

5.2 Future Work

The areas of multi-user detection and multiple antennas have gained a lot of attention in the last few years. These two areas may provide the capacity needed for future wireless systems, which must support different services such as voice, data and video transmission. In this thesis we jointly combined the spatial and temporal domains and proposed several space-time structures. These two areas are not fully developed and it required a lot of research. During this research work, we found many additional questions that need more study in the future. In this section we highlight the more important ones.

Combining space-time processing with power control

As we have seen in this thesis multi-user detection techniques can alleviate or eliminate the near-far problem, which in turn will reduce the complexity associated with power control design. However, power control provides feedback to the mobile users to adjust their transmitted power to maintain equal received power at the base station, which in turn will allow mobile users to transmit only the power needed to maintain an acceptable quality of service (QOS), thus the mobile users will not consume more power than necessary from its battery.

Using Advanced ST detection with multi rate reception

As the next generation of mobile communication systems will be required to support different communication services, i.e. multimedia, voice transmission, data transmission and video transmission. They should handle multi rate transmission with different QOS.

Some of these services need higher data rates than others, such as video transmission, to keep the QOS fixed at certain level. To do that, multi power level will be assigned for each data rate. In other words, the higher the data rates the service needs, the more power will be given to that service. Therefore again, a near-far problem will rise up. The suggested solution is to use advanced space-time detection that has variable algorithms in terms of simplicity and efficiency to handle different rate reception. Again a multi rate combined with a proper space-time receiver will be essential requirement as the mobile communication jumped to the next generates.

Switching between the MMSE detector and Decorrelator Detector

As it was mentioned earlier in chapter 3 & 4, the decorrelater detector either in time domain or in space-time signal processing, suffers from the boosted background noise generated from the inverse decorrelator matrix. A suggestion to this problem is to use MMSE detector that will over come the enhancement noise power. As it was discussed in previous chapters, the problem with the MMSE is that it is more complex than the decorrelater, since it needs to know the amplitude response of all users in advance. However, we can use the switching technique between them, depending on the noise power level. In other words, if the noise level is under certain threshold level, it is better to use the decorrelater detector for it simplicity property, but if the noise level jumps above the threshold level we will smoothly switch to the MMSE detector.

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