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Pichalinee Kanivichaporn for the degree of Master of Science in Electrical and Computer Engineering presented on March 16, 2005.

Title: A Wireless Communications System based on Space Time Block Coding and QAM Transmission

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Dr. Mario E. Magaña

Modern applications in wireless communications systems demand the capability of high transmission rate. Engineers have to find optimum solutions to transmit the signal over fading environments. In this study, we propose the combination of space-time diversity techniques and optimum QAM constellation design, in order to further improve performance.

Specifically, we use prior knowledge of the channel behavior to design adequate mapping techniques to generate a QAM constellation from a MPSK constellation typically used in coherent space time block coding application, so that superior performance is achieved. We do this be designing QAM constellation sets that minimize the symbol error probability from the known information of the transmission medium. We study the system performance for each constellation set in both AWGN and quasi-static Rayleigh fading channels. Various multi-level QAM systems are evaluated using Monte Carlo simulation and maximum likelihood detection. System performance is assessed using bit error rate (BER) and symbol error rate (SER) versus SNR plots.

A Wireless Communications Systems based on Space Time Block Coding and QAM Transmission

by

Pichalinee Kanivichaporn

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I am deeply grateful to my parents, Dr. Pichai and Mrs. Nalinee Kanivichaporn for their generous financial support and supreme encouragement. Thank you very much for letting me know how much I am loved.

CONTRIBUTION OF OTHER RESEARCHERS

The idea of a Space Time Block coding system presented in this thesis was originally developed by Siavash M. Alamouti in 1998. Since then it has been further developed by various other researchers. Some extended the concept by introducing multiple antennas at both receiver and transmitter and some others introduced differential modulation.

This thesis presents a new space time block coding system design approach that improves SER and BER performance by employing custom designed QAM mappings. Moreover, further improvement strategies are also developed and presented in this thesis.

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1. INTRODUCTION

1.1 Space Time Block Code

Digital communications through wireless channels have been the focus of intense research in the recent past, primarily for their advantages over traditional analog communications. The latest wireless communication systems are now able to transmit information data at very high rates. Different forms of digital information such as digitized voice or images, electronic mail (E-mail), file transfer, world-wide-web services (www), etc., can be transmitted via wireless mobile channels with a high quality and speed. The need for high data rate together with the constraint on the allotted communication bandwidth demands that the communication link be more reliable than ever before. Many techniques have been used to achieve the necessary increase in reliability of the system. Among these techniques, the one that achieves diversity in space and time has been the subject on extensive research in recent years. This diversity technique can significantly improve signal-to-interference plus noise ratio (SINR) of the channel. As a consequence, the bit error rate (BER) performance of the system can be improved.

Introduced by Aloumouti [1], the space-time-block code (STBC) has attracted much interest [8-18] and has been adopted as the transmit diversity scheme in the third generation (3G) wireless communication system standard, i.e., the wideband code division multiple access (W-CDMA) [2, pp.16-17]. The STBC scheme can improve the performance (BER) of a wireless communication system significantly, especially under fading and AWGN conditions. Many papers related to the STBC have discussed its application to Code Division Multiple Access (CDMA) based system. The system described in [3] is one of them. The basic principle of the STBC is the generation of two statistically independent paths with the same average power gain by employing two antenna elements at the transmitter. Therefore, the overall diversity gain of the STBC can be twice that of a system with a single transmit antenna. The STBC decoder decouples the transmitted signal with the simple linear combination algorithm, thereby achieving spatial diversity gain.

Tarokh and Jafarkhani [4] proposed a differential STBC scheme for a slow Rayleigh fading channel with two transmit antennas. Differentially encoded STBC [4] generates a 5-point PAM signal constellation set when BPSK symbols are encoded. On the other hand, when QPSK is differentially encoded, the result will be a 9-QAM constellation set. The probability distribution of the symbols after differential encoding is non-uniform. There are also a lot of studies regarding the differential STBC [19-24].

However, BER performance of the mobile communication system degrades when the transmission medium changes from AWGN to fading or from slow fading to fast fading. In a fast fading channel, the probability density function (pdf) of the received signal will be Raleigh, i.e., most of the signal energy will be in the lower part of the PDF. Introducing a scheme that reduces the probability of the low energy bit will result in an improvement of the BER performance of the system.

1.2. Thesis Contribution

The study of optimized constellation sets for transmitting over a Rayleigh fading channel is a relatively new topic in the communications community [25-29]. In this work, we study the use of optimized constellation sets in conjunction with Space Time Block Codes (STBC). We study the system in an AWGN and in a quasi-static Rayleigh fading environment. The performance of each constellation set is evaluated via BER and SER.

1.3. Thesis Overview

Chapter 2 introduces a brief tutorial of diversity techniques including the concept of Space Time Block Code. In Chapter 3, we explain the channel model and the designed constellation sets which are used in the communication system. Simulation results and analysis will be presented in Chapter 4. Finally, in Chapter 5, we draw the conclusions and discuss the scope of further research.

2. A BRIEF TUTORIAL OF DIVERSITY TECHNIQUES, AND SPACE TIME BLOCK CODE

In a mobile wireless communication system, the channel is subject to time-varying deterioration produced by noise, interference from the delayed signal itself, and interference from other users. These channels make the performance and reliability of the system degrade especially when we transmit high speed data. Many techniques have been developed to improve the reliability of the system and overcome the adverse effect of the channel.

Among such techniques, diversity-based techniques have been the focus of much of the research to improve performance of mobile wireless communication systems in recent years. In section 2.1, we discuss the basic principle behind diversity techniques, including time, frequency and space diversity. The Space Time Block Code will be discussed in section 2.2.

2.1 Basic Principle of Diversity Techniques

Diversity techniques have been developed in order to improve the performance and reliability of mobile wireless communication systems. Basic diversity techniques can be divided into 3 schemes.

Time Diversity: When the same data are sent over the channel at different time instants, the received signals can be uncorrelated if the time separations are large enough. The required time separation is at least as great as the reciprocal of the fading bandwidth, which is two times the speed of the mobile station divided by the wavelength. Hence, the time separation is inversely proportional to the speed of the mobile station. When the mobile station is stationary, time diversity (temporal diversity) in terms of multiple transmissions of the same symbol is not as useful. This is in contrast to all of the other diversity types listed above, because they are independent of the speed of the mobile station.

Frequency Diversity: Signals with different carrier frequencies far apart with each other are possibly independent. The carrier frequencies must be separated enough so that the fading associated with the different frequencies are uncorrelated. For frequency separations of more than several times the coherence bandwidth, the signal fading would be essentially uncorrelated.

Space Diversity: If the receiver or the transmitter has multiple antennas, the distance between the antennas is made large enough to ensure independent fading. This arrangement is called space diversity. Space separation of half of the wavelength is sufficient to obtain two uncorrelated signals.

In a DS-CDMA system the diversity that has been used is temporal diversity through a RAKE receiver and space diversity by using multiple antennas at the base station. Before we proceed with STBC, we will discuss the multi-path structure of the communication channel, the ability of the RAKE receiver and the space diversity, in order to understand the diversity technique used in current CDMA system.

Multi-path Channel and RAKE Receiver

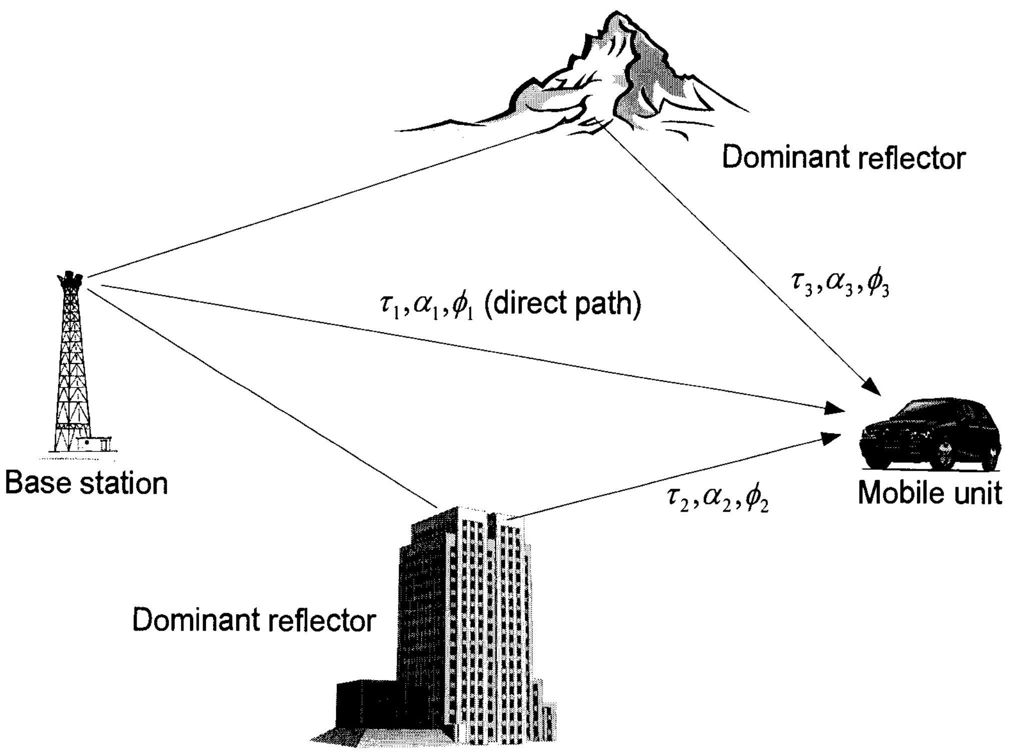

Fig 2.1: Multi-path fading channel environment

Figure 2.1 illustrates the channel model of a wireless communication system between base station (transmitter) and mobile set (receiver). The signal that has been transmitted from the base station does not arrive at the mobile unit via a single path. Obstacles such as buildings or mountains cause reflections of the transmitted signal. At the receiver, the received signal will be the summation of the signals from possibly a direct path (the signal travel directly from transmitted antenna to received antenna) and the signals from the non-line-of-sight paths. Usually the signals from additional (reflected) paths are weaker compared to the direct one. The signals propagated through the additional paths will arrive at the receiver with additional delay because of the longer distance. Besides the direct and additional paths, there are local scatterers near the mobile set. Many objects such as houses, cars, trees, etc., can generate variations in both phase and amplitude of the received signal. This phenomenon is called fading. "Multi-path fading channel" is the name that we use for the transmission model that concerns the effect of additional paths and the effect of fading. The mathematical model of the channel is often represented by a complex-valued impulse response. Specifically, this channel can be represented mathematically by attenuation coefficients α_n , phase shifts ϕ_n , and time delays τ_n , $n = 1,2,\cdots,L$. The index n denotes the path index, .i.e., it identifies a specific path.

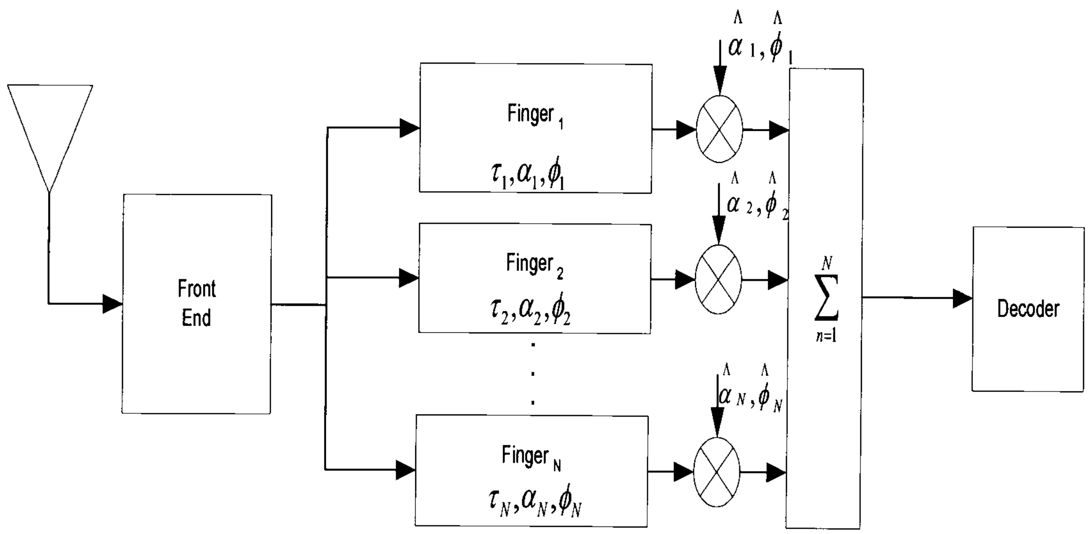

Fig 2.2: RAKE receiver

Received signals from more than one path will cause inter-symbol interference (or inter-chip interference in DS-CDMA) in a digital communication system. The RAKE receiver has been designed to overcome the problem of multi-path propagation and has been used in the second generation (2G) system of IS-95 standard. The concept of a RAKE RECEIVER is illustrated in Fig 2.2. The basic idea is to collect enough signal energy from every path in order to increase the demodulation gain. Each path of the signal is processed by the different receiver (different finger). The estimated version of channel information ($\hat{\alpha}_n$, $\hat{\phi}_n$, and $\hat{\tau}_n$) will be provided to the receiver for each finger. In a practical system only a few dominant paths are used to increase the overall signal gain.

Space Diversity

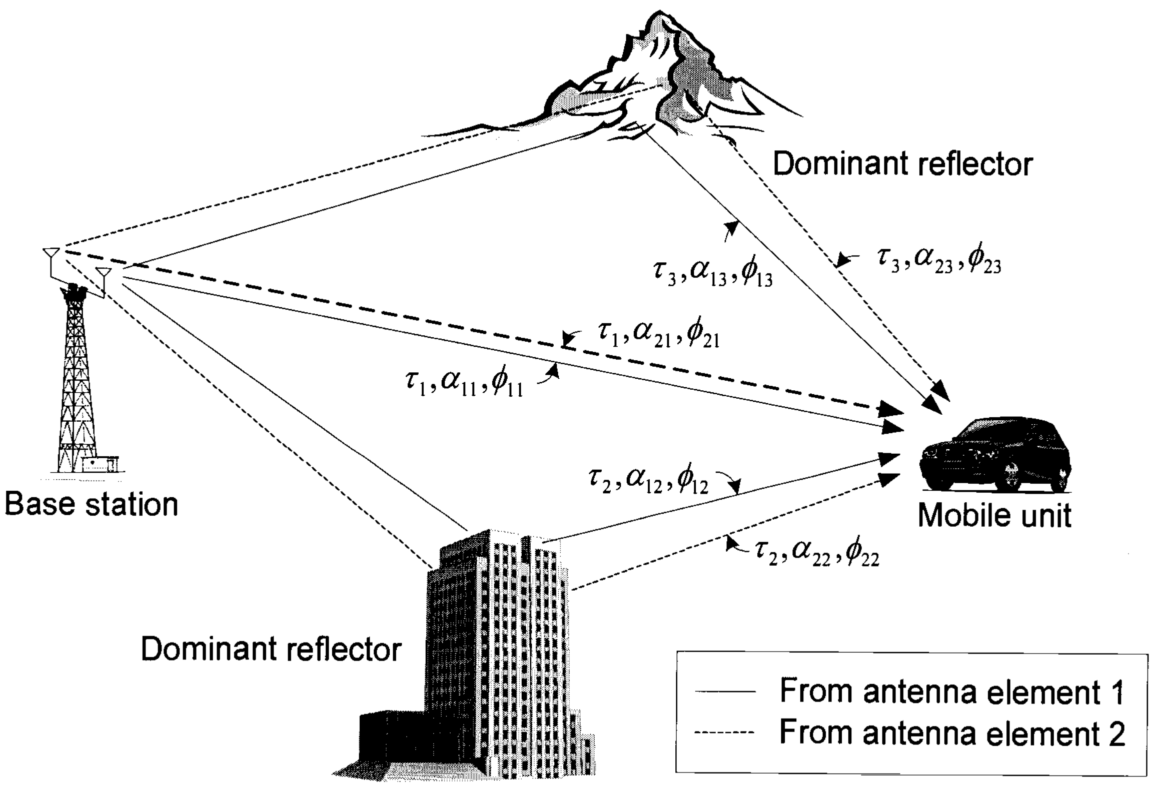

Fig 2.3: Multi-path fading channel with multiple antennas

Space diversity uses multiple transmit antennas that are separated sufficiently so that the fading paths for the received signals from each antenna are independent. The space diversity can be used in both transmitter and receiver. In receive diversity, the receiver may use the selection diversity or linear combining. The best signal, based on a certain quality will be selected by the select diversity. Linear combining diversity uses a linear combination of the weighted replica of all received signals. Introducing the receive diversity will make the receiver end more complex.

Transmit diversity, on the other hand, will put the processing burden at the transmitter (base station) and leave the receiver (mobile set) the simple design. Fig 2.3 illustrates the multi-path fading channel with multiple antennas at the transmitter. Compared to Fig 2.2, the number of transmission path in Fig 2.3 is twice due to the additional transmit antenna. Both of the transmission paths from two antennas have the same delay since the space between the antennas is negligible compared to the length of transmission paths. Attenuation coefficients and phase shifts of each fading channel from different antennas are statistically independent. Index n in Fig 2.3 represent transmit antenna index and path index will be denote by index m.

2.2 Space Time Block Code (Base System)

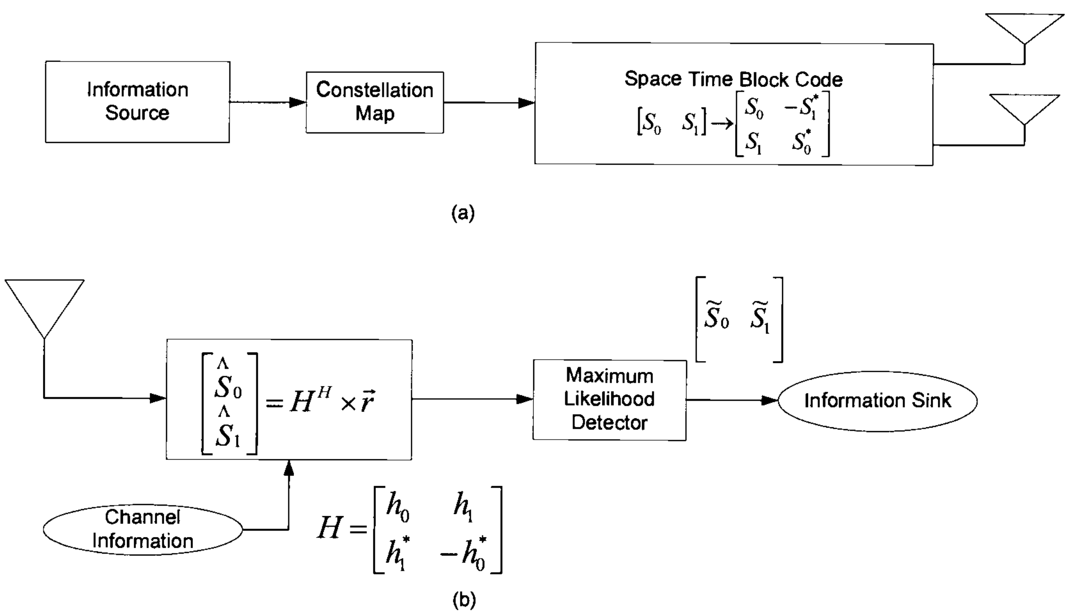

Fig 2.4: Space-Time block coding Model

Fig 2.4 illustrates the system model of Space Time Block Coding (STBC). The sequence of symbol $S_0\ S_1\ S_2\ S_3 \ldots S_m$ to be transmitted will be mapped by the STBC encoder to get the transmit matrix

$$CS_1 = \begin{bmatrix} s_0 & s_1 \\ -s_1^* & s_0^* \end{bmatrix}, \tag{2.1}$$

Where

$$CS_k = \begin{bmatrix} s_{(2*k)-2} & s_{(2*k)-1} \\ -s^*_{(2*k)-1} & s^*_{(2*k)-2} \end{bmatrix}.$$

The vectors $[s_0 \quad s_1]$ and $[-s_1^* \quad s_0^*]$ have unit length and are orthogonal to each other. It follows that

$$CS_k CS_k^H = I_2, \tag{2.2}$$

where the subscript “H” denotes Hermitian operation (transpose conjugate) and I_2 is the 2*2 identity matrix.

For basic one-tap channel in a STBC communication system with an impulse response coefficient and noise for each path (assume 2 transmit antennas and 1 receive antenna), the system works as follows: Let h_0 be the channel coefficient from T_{x0} to receive antenna and h_1 be the channel coefficient from T_{x1} to receive antenna, then at the received antenna the receive signal vector can be written as

$$\begin{bmatrix} R_0 \\ R_1 \end{bmatrix} = \begin{bmatrix} s_0 h_0 + s_1 h_1 \\ s_0^* h_1 - s_1^* h_0 \end{bmatrix} + \text{noise}, \tag{2.3}$$

R_0 and R_1 are the received signals which are the result of multiplying the channel coefficient matrix and the transmitted signal vector.

Taking the complex conjugate of signal R_1, and neglecting the noise component, we can re-write equation (2.3) as follows

$$\vec{r} = \begin{bmatrix} R_0 \\ R_1^* \end{bmatrix} = H(h_0, h_1)\begin{bmatrix} s_0 \\ s_1 \end{bmatrix}, \tag{2.4}$$

Where $H(h_0, h_1)$ is a channel coefficient matrix described by

$$H(h_0, h_1) = \begin{bmatrix} h_0 & h_1 \\ h_1^* & -h_0^* \end{bmatrix}. \tag{2.5}$$

Assuming perfect estimation at the receiver, we premultiply the received signal vector $\vec{r}$ by the matrix $H^H(h_0, h_1)$, where H^H is the complex conjugate transpose of matrix $H(h_0, h_1)$. By the orthogonal of matrix we will get

$$\begin{bmatrix} \hat{s}_0 \\ \hat{s}_1 \end{bmatrix} = H^H \times \vec{r} = \begin{bmatrix} (|h_0|^2 + |h_1|^2)\, s_0 \\ (|h_0|^2 + |h_1|^2)\, s_1 \end{bmatrix}, \tag{2.6}$$

$\hat{s}_0$ and $\hat{s}_1$ in equation (2.6) will be used to estimate the values of s_0 and s_1 which have been transmitted. The maximum likelihood (ML) decoding rule is given by

$$\tilde{s} = \arg\min_{s \in S} |\hat{s} - s|^2. \tag{2.7}$$

The STBC provides a higher diversity gain without expanding the channel bandwidth than the RAKE temporal diversity scheme. The advantage of STBC over a RAKE receiver can be explained as follows: In the RAKE receiver, the contribution of the second path signal (or finger) from the reflection is much smaller compared to the signal from the direct path. Compared to STBC, the paths from the different antennas will be transmitted with equal power and make the diversity gain (from space diversity) increase. The STBC can be combined with the RAKE receiver to take full advantage of space and temporal diversity.

3. CHANNEL MODEL AND CONSTELLATION DESIGN

In this chapter, we discuss the basic ideas used throughout the thesis as well as modifications to suit our space-time block coding system design and implementation. First, the channel model is presented. The additive white Guassian noise (AWGN) plus Raylaigh fading are assumed. The AWGN plus Rayleigh fading channel model is the common channel scheme used for mobile wireless communications. Transmission through such channel reduces overall system performance, e.g., rate of transmission and bit error rate.

3.1 Channel Model

We assume that the channel used by the system is a Rayleigh fading plus additive white Gaussian noise channel. In our system, the users are mobile stations. At any instant of time, a user suffers from signal fading and interference caused by the propagation environment and by the reception of unwanted signals from other users as well as replicas of the signal of interest. The fading channel degrades the signal strength dramatically and may result in communication failure. Thus, we need to study the nature of the channel and to build a channel model in mathematical form, in order to construct an appropriate design. Only the fading channel model is discussed in this section, since the AWGN channel model is well known. However, both fading and AWGN channels are used in our design process.

Consider the simplified wireless communication system block diagram

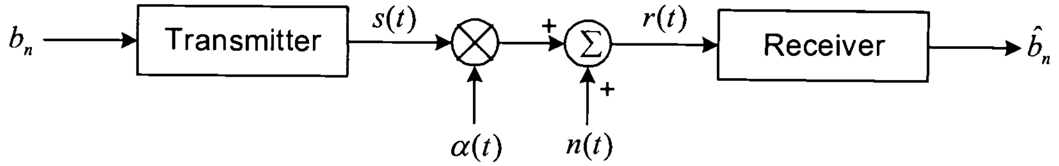

Figure 3.1: Simplified communication system block diagram

The received signal is described by

$$r(t) = \alpha(t)s(t) + n(t)\,, \tag{3.1}$$

where

$$s(t) = \mathrm{Re}\{g(t)e^{j2\pi f_c t}\} \tag{3.2}$$

and *g(t), r(t), α(t), s(t)* and *n(t)* are the baseband pulse, the received signal, the fading coefficient, the transmitted signal and noise, respectively. The noise *n(t)* is assumed to be zero-mean white and Gaussian; therefore, it will not be further discussed.

Let us describe the fading coefficient tern *α(t)*. Signal strength at the receiver antenna varies with respect to movement of the mobile user itself or other objects nearby. In a city where buildings, cars and other sizable objects are present, it is impossible for mobile clients to communicate with the base station using line of sight. Consequently, multipath reception of multiple replicas of the original signal at various propagation delays $\tau_i(t)$ occur, where i represents the i^{th} path. Therefore, the effect of the fading channel enormously degrades overall system performance. The received signal *r(t)* can be written, neglecting the noise component as

$$r(t) = \sum_n \alpha_n(t)s(t - \tau_n(t))\,, \tag{3.3}$$

where *n* denotes the number of reflecting signals

The equivalent baseband of the received signal is,

$$z(t) = \sum_n \sigma_n(t)e^{-j2\pi f_c \tau_n(t)} g(t - \tau_n(t)) \tag{3.4}$$

If we assume that an unmodulated carrier is being transmitted, *g(t)* = 1, the net received envelope, the summation of all reflected signals, is expressed as

$$z(t) = \sigma(t)e^{-j\theta(t)} \quad (3.5)$$

where σ(t) and θ(t) are the resultant amplitude and phase, respectively.

Figure 3.2 [5] shows the effect of multipath reception of the original transmitted signal. The received signal is not what was originally transmitted but the summation of the transmitted and the reflected signals. The result is the distortion of both amplitude and phase. The reflected signal, as shown in Figure 3.2, consists of the 2 orthogonal components $x_n(t)$ and $y_n(t)$, i.e.,

$$x_n(t) + jy_n(t) = \sigma_n(t)e^{-j\theta_n(t)} \quad (3.6)$$

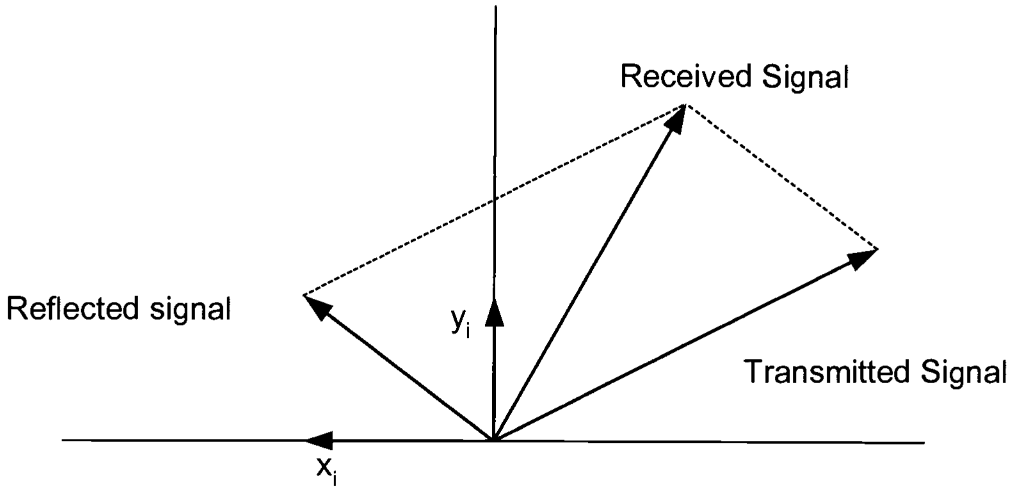

Figure 3.2: The effect of multipath reception

At any given time where the number of reflected signals is large, the random variables $x(t)$ and $y(t)$ resulting from an addition of all reflected signals will have a Gaussian pdf (Central Limit Theorem). Therefore, the amplitude of the fading component $a(t)$ is,

$$\alpha(t) = \sqrt{x^2(t) + y^2(t)} \tag{3.7}$$

The *α(t)* is a Rayleigh random variable when there is not a line-of-sight component.

3.2 Quadrature Amplitude Modulation (QAM) Performance Evaluation

3.2.1 AWGN Channel:

Suppose that the QAM symbols do not occur with equal probability, M is the cardinality of the signal set which is an arbitrary integer, and coherent detection is performed at the receiver. Let the AWGN be zero mean and have power spectral density $N_0/2$, for all frequencies.

Let the passband transmitted signal be described by

$$s_{pi}(t) = a_i \cos(\omega_c t + \theta_c) - b_i \sin(\omega_c t + \theta_c), \tag{3.8}$$

where a_i and b_i are the in-phase and quadrature components, respectively, of the i^{th} symbol, i = 1, ..., M, ω_c is the carrier frequency and θ_c is a random phase angle.

Let the equivalent complex baseband signal be described by

$$s_{bi}(t) = a_i + jb_i, \quad nT_s \le t \le (n+1)T_s \ . \tag{3.9}$$

Then, the equivalent complex baseband signal seen at the input of the receiver detector (after filtering and downconversion) is described by

$$r(t) = ai + jbi + n(t), \quad nT_s \le t \le (n+1)T_s , \tag{3.10}$$

Where

$$n(t) = n_I(t) + jn_Q(t) . \tag{3.11}$$

Performance will depend on the type (shape) of constellation that is used at the transmitter and on the detection scheme used at the receiver. Regardless of the constellation shape, if the detector uses the maximum likelihood criterion, it selects symbol s_j when the observed measurement r(t) lies on region R_j of the observation space.

System performance can be evaluated as follows: Suppose that at the n^{th} observation interval r(t) falls on the decision region R_i, then the probability of a correct decision, given that symbol s_i was transmitted, is described by

$$P\{correct \;\; decision|s_i\} = \iint_R f(r_1, r_2|s_i)dr_1 dr_2 , \qquad (3.12)$$

where

$$r_1 \equiv \Re\{r(t)\} = a_i + n_I(nT_s)$$

$$r_1 \equiv \mathrm{Im}\{r(t)\} = b_i + n_Q(nT_s)$$

Let $r \equiv \begin{bmatrix} r_1 \\ r_2 \end{bmatrix}$ and $n \equiv \begin{bmatrix} n_I \\ n_Q \end{bmatrix} = \begin{bmatrix} n_I(nT_s) \\ n_Q(nT_s) \end{bmatrix}$, with $E\{n_I\} = E\{n_Q\} = 0$.

$E\{n_I^2\} = E\{n_Q^2\} = N_0$, and $E\{n_I n_Q\} = 0$.

Equivalently,

$$P\{\text{correct decision} \mid s_i\} = P\{\|r - s_i\|^2 \le \|r - s_j\|^2 \mid s_i\} \quad j \ne i, i, j = 1,...,M \qquad (3.13)$$

is the probability that the obeserved vector r is closer to the signal constellation point s_i than to s_j, given that symbol s_i was transmitted.

Conversely, the probability of an incorrect decision (select symbol s_j), given that symbol s_i was transmitted is given by

$$P\{\text{incorrect decision} \mid s_i\} = P\{\|r - s_i\|^2 > \|r - s_j\|^2 \mid s_i\} \quad j \ne i, i, j = 1,...,M . \qquad (3.14)$$

But, when symbol s_i is transmitted,

$$\|\boldsymbol{r}-\boldsymbol{s}_i\|^2 = \|\boldsymbol{n}\|^2 = n_I^2 + n_Q^2 \tag{3.15}$$

And

$$\|\boldsymbol{r}-\boldsymbol{s}_j\|^2 = (a_i - a_j)^2 + (b_i - b_j)^2 + 2(a_i - a_j)n_I + 2(b_i - b_j)n_Q + n_I^2 + n_Q^2. \tag{3.16}$$

Hence, the event $\|\boldsymbol{r}-\boldsymbol{s}_i\|^2 > \|\boldsymbol{r}-\boldsymbol{s}_j\|^2$ given that s_i was transmitted implies that

$$n_I^2 + n_Q^2 > (a_i - a_j)^2 + (b_i - b_j)^2 + 2(a_i - a_j)n_I + 2(b_i - b_j)n_Q + n_I^2 + n_Q^2 \tag{3.17}$$

or

$$(a_j - a_i)n_I + (b_j - b_i)n_Q > \frac{1}{2}\left[(a_i - a_j)^2 + (b_i - b_j)^2\right]. \tag{3.18}$$

Define v_{ji} and c_{ji} as follows:

$v_{ji} \equiv (a_j - a_i)n_I + (b_j - b_i)n_Q$ and $c_{ji} \equiv \frac{1}{2}\left[(a_i - a_j)^2 + (b_i - b_j)^2\right]$.Then v_{ji} is Gaussian with zero mean and variance $\sigma_{v_{ji}}^2 = \left[(a_i - a_j)^2 + (b_i - b_j)^2\right]N_0 = 2c_{ji}N_0$.

Hence,

$$\begin{aligned} P\{\text{incorrect decision} \mid s_i\} &= P\{v_{ji} > c_{ji} \mid s_i\} = \frac{1}{\sqrt{2\pi}\sigma_{v_{ji}}} \int_{c_{ji}}^{\infty} e^{-\frac{v_{ji}^2}{2\sigma_{v_{ji}}^2}} dv_{ji} \\ &= Q\left(\frac{c_{ji}}{\sigma_{v_{ji}}}\right) = Q\left(\sqrt{\frac{c_{ji}}{2N_0}}\right) \\ &= Q\left(\sqrt{\frac{(a_i - a_j)^2 + (b_i - b_j)^2}{4N_0}}\right), \quad j \neq i,\ i, j = 1, \ldots, M \end{aligned} \tag{3.19}$$

Let $P\{\text{incorrect decision} \mid \boldsymbol{s}_i\} = P\{\text{error} \mid \boldsymbol{s}_i\}_j$, $j \neq i, i, j = 1,\ldots,M$, i.e., the probability of selecting symbol s_j when symbol s_i was transmitted. Then, the average probability of selecting the wrong symbol when symbol s_i was transmitted $P\{\text{error} \mid \boldsymbol{s}_i\}$ is upper bounded by

$$P\{\text{error} \mid \boldsymbol{s}_i\} \leq \sum_{\substack{j=1 \\ j \neq i}}^{M} P\{\text{error} \mid \boldsymbol{s}_i\}_j, \quad i = 1,\ldots,M \ . \tag{3.20}$$

Finally, an upper bound of the average symbol error probability is given by

$$P_{se} = \sum_{i=1}^{M} P\{\text{error} \mid \boldsymbol{s}_i\} P\{\boldsymbol{s}_i\} \leq \sum_{i=1}^{M} \sum_{\substack{j=1 \\ j \neq i}}^{M} Q\left(\sqrt{\frac{c_{ji}}{2N_0}}\right) P\{\boldsymbol{s}_i\}. \tag{3.21}$$

Note that eq. (3.21) is an upper bound of the symbol error probability when the decision metric is the minimum Euclidean distance and the probability of symbol transmission is arbitrary. Also, the total number of symbols M is arbitrary.

3.2.2 Rayleigh Fading plus AWGN Channel:

Consider now the transmission of QAM over a single path Rayleigh fading plus AWGN channel. Using the same minimum distance detection scheme as in the previous AWGN case, the observed vector at the output of the detector is described by

$$\boldsymbol{r} = A\boldsymbol{s}_i + \boldsymbol{n} = \begin{bmatrix} r_1 \\ r_2 \end{bmatrix}, \tag{3.22}$$

where $A = \begin{bmatrix} \alpha_1 & 0 \\ 0 & \alpha_2 \end{bmatrix}$, $\boldsymbol{s}_i = \begin{bmatrix} a_i \\ b_i \end{bmatrix}$, and $f(\alpha_k) = \frac{\alpha_k}{\sigma_k^2} e^{-\frac{\alpha_k^2}{2\sigma_k^2}}, \quad \alpha_k \geq 0, \quad k = 1, 2$.

Using the minimum distance as the decision metric at the output of the detector, the conditional probability of a decision error is given by

$$P\{\text{incorrect decision} \mid \boldsymbol{s}_i, \boldsymbol{\alpha}\} = P\left\{\|\boldsymbol{r}-\boldsymbol{s}_i\|^2 > \|\boldsymbol{r}-\boldsymbol{s}_j\|^2 \mid \boldsymbol{s}_i, \boldsymbol{\alpha}\right\}$$

$$= P\left\{v_{ji|\alpha} > c_{ji|\alpha}\right\} = Q\left(\frac{c_{ji|\alpha}}{\sigma_{v_{ji|\alpha}}}\right), \quad j \neq i,\, i, j = 1, \ldots, M \qquad (3.23)$$

where $v_{ji|\alpha} \equiv \alpha_1 (a_j - a_i) n_I + \alpha_2 (b_j - b_i) n_Q$ and

$c_{ji|\alpha} \equiv \frac{1}{2}\left[\alpha_1^2 (a_i - a_j)^2 + \alpha_2^2 (b_i - b_j)^2\right]$.

Now, for a given $\boldsymbol{\alpha} = \begin{bmatrix} \alpha_1 \\ \alpha_2 \end{bmatrix}$, $v_{ji|\alpha}$ is Gaussian distributed with zero mean and variance

$\sigma^2_{v_{ji|\alpha}} = \left[\alpha_1^2 (a_i - a_j)^2 + \alpha_2^2 (b_i - b_j)^2\right] N_0 = 2c_{ji|\alpha} N_0$.

Therefore, similar to the AWGN case, we get

$$P\{\text{incorrect decision} \mid \boldsymbol{s}_i, \boldsymbol{\alpha}\} = Q\left(\sqrt{\frac{c_{ji|\alpha}}{2N_0}}\right) = Q\left(\sqrt{\frac{\alpha_1^2 (a_i - a_j)^2 + \alpha_2^2 (b_i - b_j)^2}{4N_0}}\right),$$

$$j \neq i,\, i, j = 1, \ldots, M \qquad (3.24)$$

Let $P\{\text{incorrect decision} \mid \boldsymbol{s}_i\} = P\{\text{error} \mid \boldsymbol{s}_i\}_j, \quad j \neq i,\, i, j = 1, \ldots, M$, then

$$P\{\text{error} \mid \boldsymbol{s}_i\}_j = \int_0^\infty \int_0^\infty P\{\text{incorrect decision} \mid \boldsymbol{s}_i, \boldsymbol{\alpha}\} f(\boldsymbol{\alpha}) d\boldsymbol{\alpha}, \quad j \neq i,\, i, j = 1, \ldots, M . \qquad (3.25)$$

If the in-phase and quadrature fading components are statistically independent of each other, then

$$f(\boldsymbol{\alpha}) = \prod_{k=1}^{2} \frac{\alpha_k}{\sigma_k^2} e^{-\frac{\alpha_k^2}{2\sigma_k^2}} \qquad (3.26)$$

and

$$
\begin{aligned}
P\{\text{error} \mid \boldsymbol{s}_i\}_j &= \int_0^\infty\!\!\int_0^\infty Q\left(\sqrt{\frac{c_{ji|\alpha}}{2N_0}}\right)\prod_{k=1}^{2}\frac{\alpha_k}{\sigma_k^2}e^{-\frac{\alpha_k^2}{2\sigma_k^2}}d\boldsymbol{\alpha} \\
&= \int_0^\infty\!\!\int_0^\infty\left[\frac{1}{\pi}\int_0^{\pi/2} e^{-\frac{c_{ji|\alpha}}{4N_0\sin^2\theta}}d\theta\right]\prod_{k=1}^{2}\frac{\alpha_k}{\sigma_k^2}e^{-\frac{\alpha_k^2}{2\sigma_k^2}}d\boldsymbol{\alpha}, \quad j\neq i,\, i,j=1,\ldots,M \\
&= \frac{1}{\pi}\int_0^{\pi/2}\left[\int_0^\infty\frac{\alpha_1}{\sigma_1^2}e^{-\left[\frac{1}{2\sigma_1^2}+\frac{(a_i-a_j)^2}{8N_0\sin^2\theta}\right]\alpha_1^2}d\alpha_1\right]\left[\int_0^\infty\frac{\alpha_2}{\sigma_2^2}e^{-\left[\frac{1}{2\sigma_2^2}+\frac{(b_i-b_j)^2}{8N_0\sin^2\theta}\right]\alpha_2^2}d\alpha_2\right]d\theta
\end{aligned}
\quad . \quad (3.27)
$$

Let $\lambda_1 \equiv 1+\dfrac{(a_i-a_j)^2\sigma_1^2}{4N_0\sin^2\theta}$ and $\lambda_2 \equiv 1+\dfrac{(b_i-b_j)^2\sigma_2^2}{4N_0\sin^2\theta}$, then

$$
\begin{aligned}
P\{\text{error} \mid \boldsymbol{s}_i\}_j &= \frac{1}{\pi}\int_0^{\pi/2}\left[\int_0^\infty\frac{\alpha_1}{\sigma_1^2}e^{-\left(\frac{\lambda_1}{2\sigma_1^2}\right)\alpha_1^2}d\alpha_1\right]\left[\int_0^\infty\frac{\alpha_2}{\sigma_2^2}e^{-\left(\frac{\lambda_2}{2\sigma_2^2}\right)\alpha_2^2}d\alpha_2\right]d\theta \\
&= \frac{1}{\pi}\int_0^{\pi/2}\frac{1}{\lambda_1\lambda_2}d\theta = \frac{1}{\pi}\int_0^{\pi/2}\frac{\sin^4\theta}{(\sin^2\theta+\gamma_1)(\sin^2\theta+\gamma_2)}d\theta \\
&= \frac{1}{\pi(\gamma_1-\gamma_2)}\int_0^{\pi/2}\left[\frac{\gamma_1\sin^2\theta}{\sin^2\theta+\gamma_1}-\frac{\gamma_2\sin^2\theta}{\sin^2\theta+\gamma_2}\right]d\theta \\
&= \frac{1}{2}\left[1-\left(\frac{\gamma_1}{\gamma_1-\gamma_2}\right)\sqrt{\frac{\gamma_1}{1+\gamma_1}}+\left(\frac{\gamma_2}{\gamma_1-\gamma_2}\right)\sqrt{\frac{\gamma_2}{1+\gamma_2}}\right], j\neq i,\, i,j=1,\ldots,M
\end{aligned}
\quad , \quad (3.28)
$$

where $\gamma_1 \equiv \dfrac{(a_i-a_j)^2\sigma_1^2}{4N_0}$ and $\gamma_2 \equiv \dfrac{(b_i-b_j)^2\sigma_2^2}{4N_0}$.

Therefore,

$$
\begin{aligned}
P\{\text{error} \mid \boldsymbol{s}_i\} &\leq \sum_{\substack{j=1\\ j\neq i}}^{M} P\{\text{error} \mid \boldsymbol{s}_i\}_j \\
&= \frac{1}{2}\sum_{\substack{j=1\\ j\neq i}}^{M}\left[1-\left(\frac{\gamma_1}{\gamma_1-\gamma_2}\right)\sqrt{\frac{\gamma_1}{1+\gamma_1}}+\left(\frac{\gamma_2}{\gamma_1-\gamma_2}\right)\sqrt{\frac{\gamma_2}{1+\gamma_2}}\right]
\end{aligned}
\quad (3.29)
$$

and an upper bound of the average symbol error probability is given by

$$P_{se} = \sum_{i=1}^{M} P\{\text{error} \mid s_i\} P\{s_i\}$$
$$\leq \frac{1}{2} \sum_{i=1}^{M} \sum_{\substack{j=1 \\ j \neq i}}^{M} \left[1 - \left(\frac{\gamma_1}{\gamma_1 - \gamma_2} \right) \sqrt{\frac{\gamma_1}{1+\gamma_1}} + \left(\frac{\gamma_2}{\gamma_1 - \gamma_2} \right) \sqrt{\frac{\gamma_2}{1+\gamma_2}} \right] P\{s_i\} \quad (3.30)$$

Clearly, the bound could be made tighter if only the nearest neighbors were considered in the probability of making a decision error, since distant symbols from that which was transmitted are highly unlikely to be selected.

The idea is to select a two-dimensional lattice Λ that will yield a minimum P_{se}. Let the constellation S be obtained from a region enclosed by a circle which contains the desired M symbols, such that $P_{se}(S) \leq P_{se}(\Lambda)$, which takes into account the edge effects of the constellation boundary.

3.3 Constellation Sets for Space-Time Block Coding

Signal constellation sets play major roles in both bit error rate (BER) and symbol error rate (SER) of every communication system under certain power constraints. In fact, there are plenty of existing constellation sets which have been designed to optimize communication system performance. For our system, we consider five categories of the constellations.

3.3.1 M-ary Phase Shift Keying (M-PSK)

The symbols $d_{i,k}^{(t)} \in A$, where A is defined as

$$A = e^{j2\pi(m-1)/M} \big| m = 1,2,\ldots,M \, ,$$

While $M = 2^n$, is the number of constellation points. One symbol represents n bits.

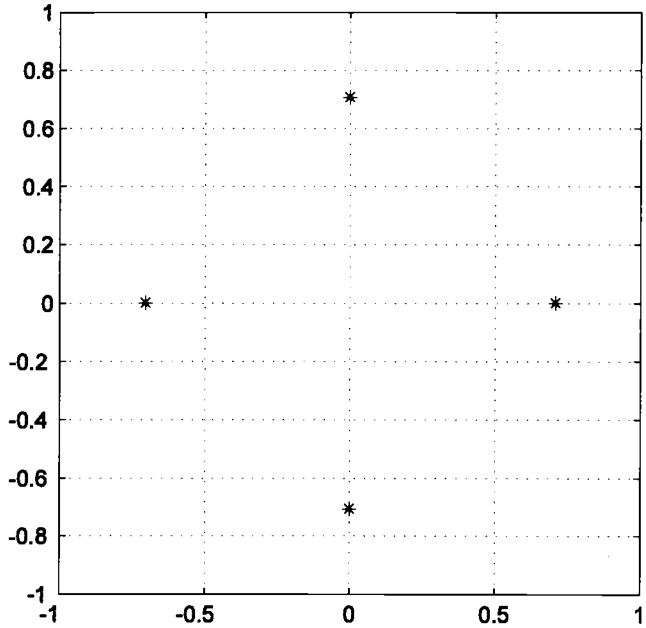

Figure 3.3: QPSK Constellation Set

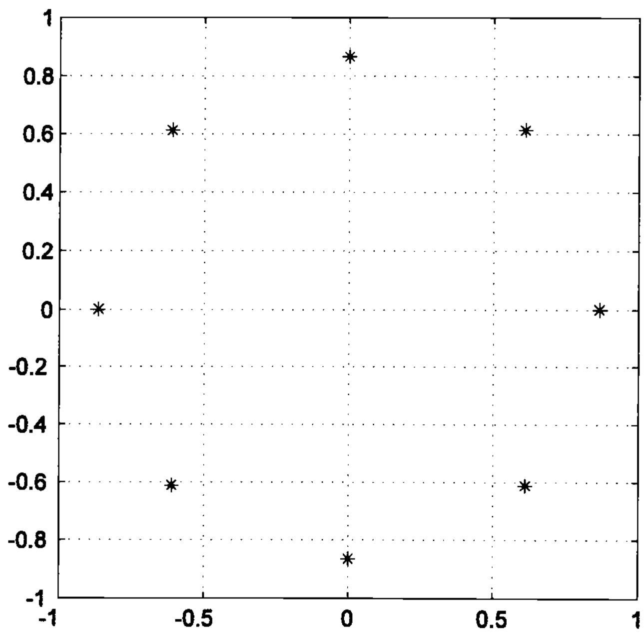

Figure 3.4: 8- PSK Constellation Set

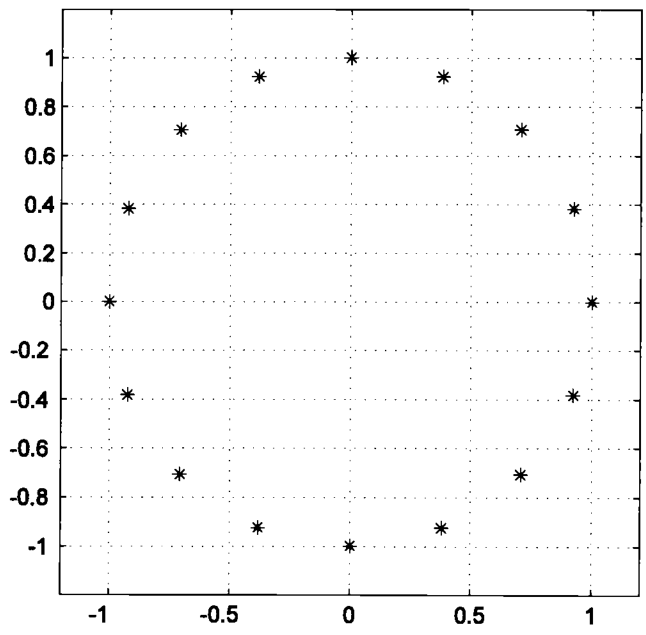

Figure 3.5: 16 - PSK Constellation Set

3.3.2. Constellations from Gradient Search Algorithm

As shown in Figure 3.3-3.5, the M-PSK constellation set is not the densest constellation set available. A hexagonal lattice constellation set yields the densest constellation set while maintaining the minimum distance to the nearest neighbors. Hence, an algorithm called Gradient Search algorithm [6] is used to design the signal constellation sets. According to the Gradient search algorithm along with our simulation results based on the algorithm, the constellation set generated from such an algorithm tends toward hexagonal lattice sphere packed constellations. The algorithm is implemented in software to generate constellation sets under the assumptions of AWGN and uniformly distributed signal sets.

Equations (3.31a) and (3.31b) are the basis of the gradient search algorithm, where δ_m is the positive step size, and $\nabla P_e(\cdot)$ is the gradient of P_e with respect to S_m of the constellation set.

$$S^*_{m+1} = S_m - \delta_m \nabla P_e(S_m) \tag{3.31a}$$

$$S_{m+1} = \frac{S^{*}{}_{m+1}}{\left\| S^{*}{}_{m+1} \right\|} \tag{3.31b}$$

The form of the gradient $\nabla P_e(\cdot)$ is the following:

$$\nabla P_e = \begin{bmatrix} g_1 \\ g_2 \\ \vdots \\ g_N \end{bmatrix}, \tag{3.32}$$

where

$$g_k \sim -\sum_{i \neq k} \exp\left[-\frac{\left\| s_k - s_i \right\|}{8N_0} \right] \left[\frac{1}{\left\| s_k - s_i \right\|^2} + \frac{1}{4N_0} \right] I_{s_k - s_i} \tag{3.33}$$

where $I_{s_k - s_i}$ is a two-dimensional unit vector in the direction of s_k-s_i, and s denotes signal point. While $M = 2^n$, is the number of constellation points. One symbol represents n bits.

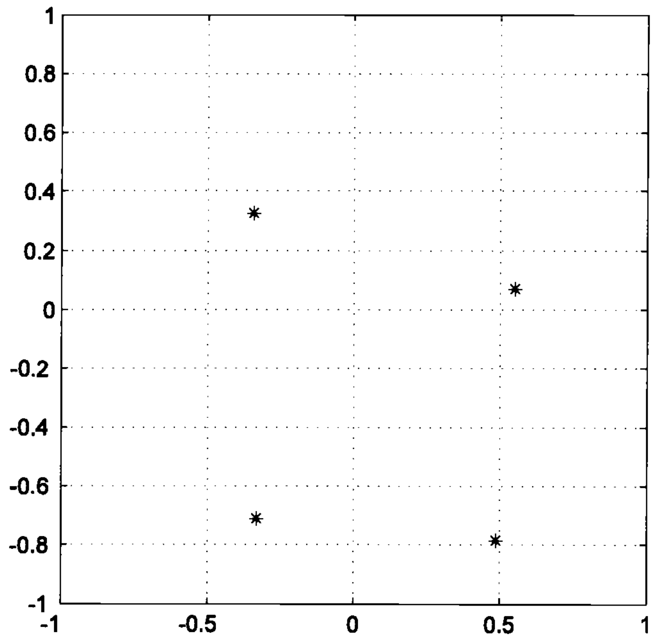

Figure 3.6: 4 points Gradient Searched Constellation Set

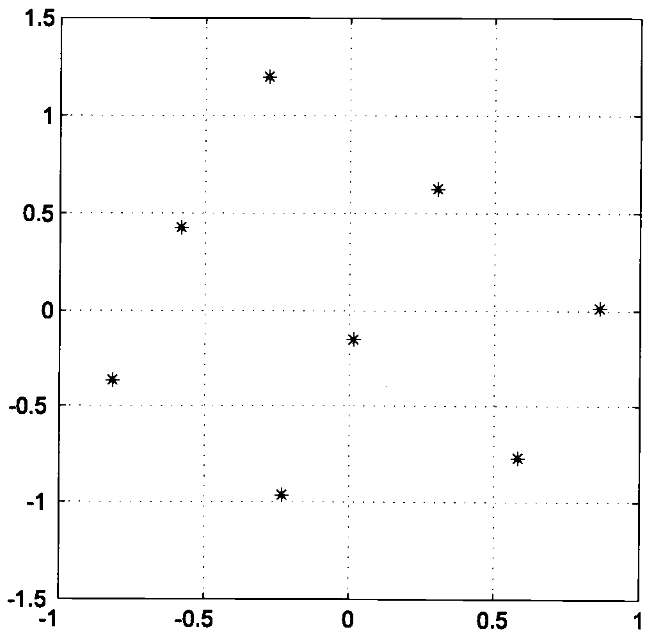

Figure 3.7: 8 points Gradient Searched Constellation Set

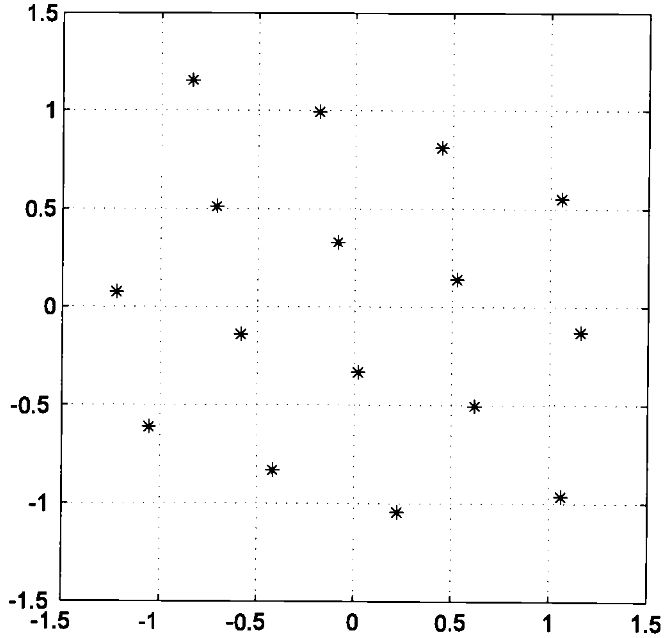

Figure 3.8: 16 points Gradient Searched Constellation Set

3.3.3 Constellation carved from lattice Λ_{21}

It generates a hexagonal lattice with a fundamental volume of $\sqrt{3}/2$ minimum squared Euclidean distance $d^2_{E\min} = 1$, and diversity $L = 1$ [7], since the vector (1,0) belongs to the lattice. And it is characterized by the lattice generator matrix

$$G = \begin{bmatrix} 1 & 0 \\ \frac{1}{2} & \frac{\sqrt{3}}{2} \end{bmatrix} \tag{3.34}$$

And $M = 2^n$, is the number of constellation points. One symbol represents n bits.

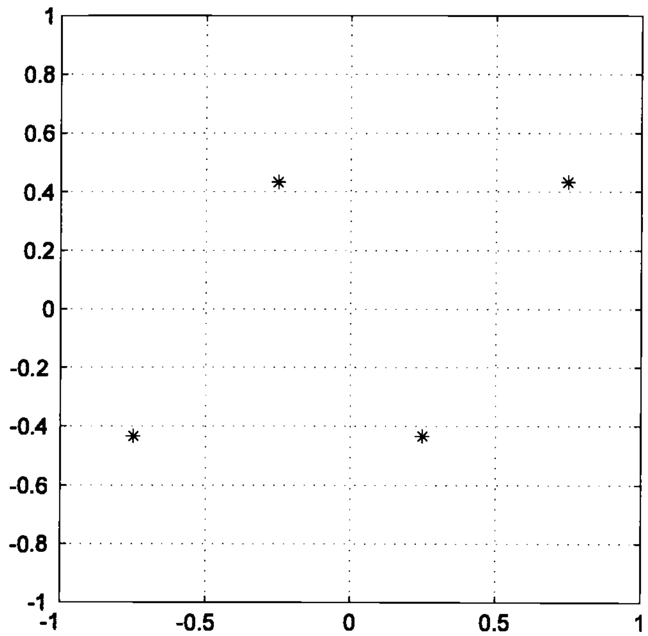

Figure 3.9: 4 points Constellation Set carved from lattice Λ_{21}

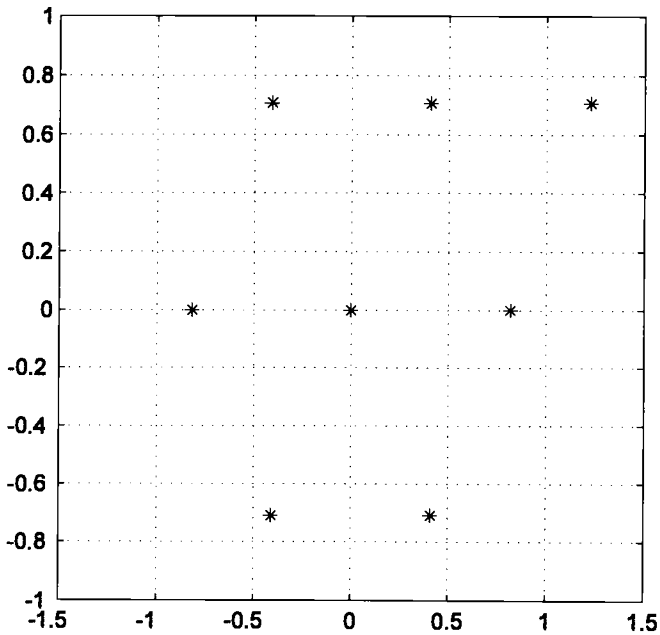

Figure 3.10: 8 points Constellation Set carved from lattice Λ_{21}

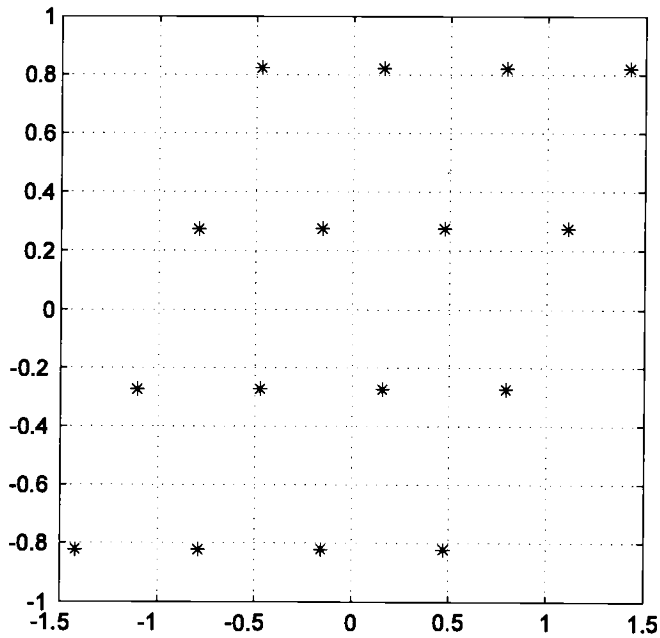

Figure 3.11: 16 points Constellation Set carved from lattice Λ_{21}

3.3.4. Constellation from lattice Λ_{22}

It generates a hexagonal lattice with a fundamental volume of $\sqrt{5}$ minimum squared Euclidean distance $d^2_{E\min} = 2$, and diversity $L = 2$ [7], since the vector (1,0) belongs to the lattice. And it is characterized by the lattice generator matrix

$$G = \begin{bmatrix} 1 & 1 \\ \dfrac{1+\sqrt{5}}{2} & \dfrac{1-\sqrt{5}}{2} \end{bmatrix} \tag{3.35}$$

For $M = 2^n$, is the number of constellation points. One symbol represents n bits.

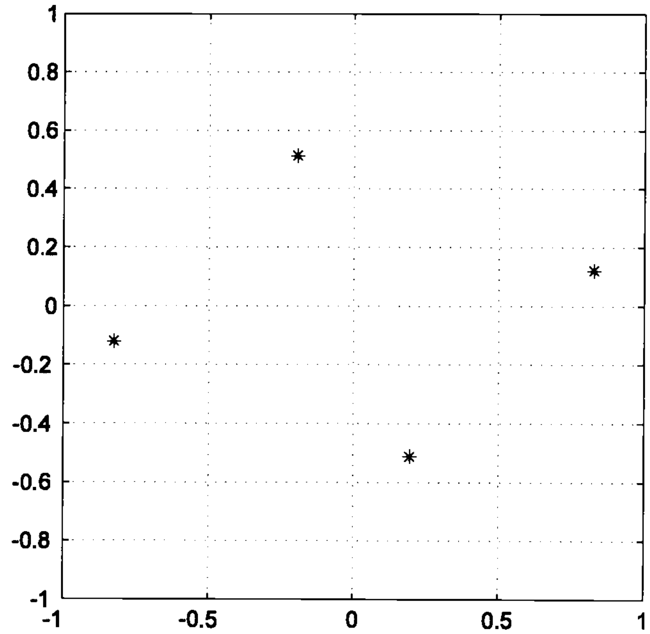

Figure 3.12: 4 points Constellation Set carved from lattice Λ_{22}

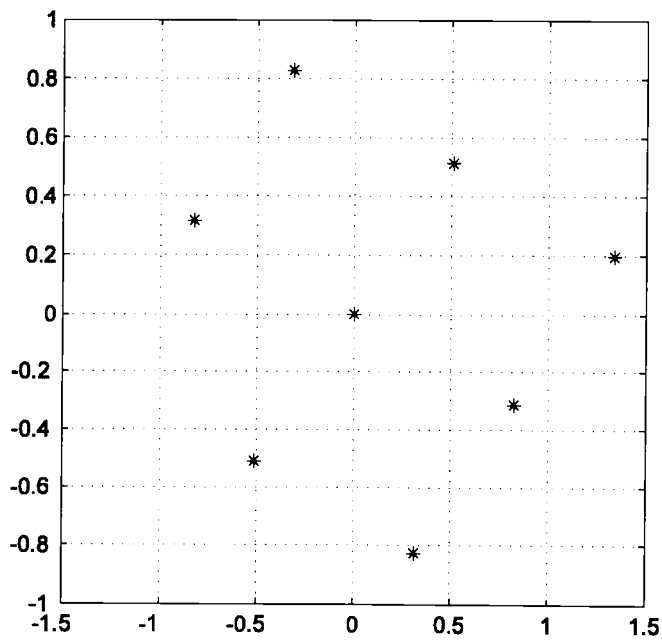

Figure 3.13: 8 points Constellation Set carved from lattice Λ_{22}

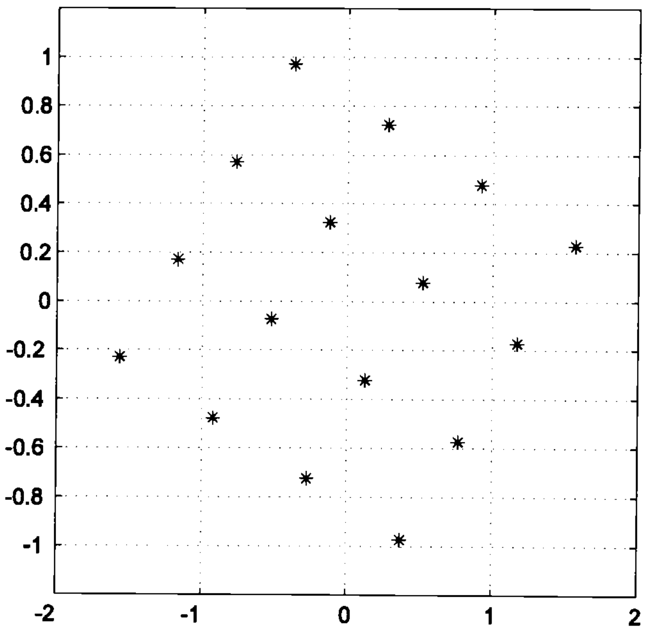

Figure 3.14: 16 points Constellation Set carved from lattice Λ_{22}

3.3.5. M-ary Quadrature Amplitude Modulation (M-QAM)

We define $M = 2^n$ as the number of constellation points. Each symbol contains n bits. There are several possible constellation sets for 8-QAM. From the simulations, the 8-QAM constellation set used in this thesis is chosen because of its superior performance over the others.

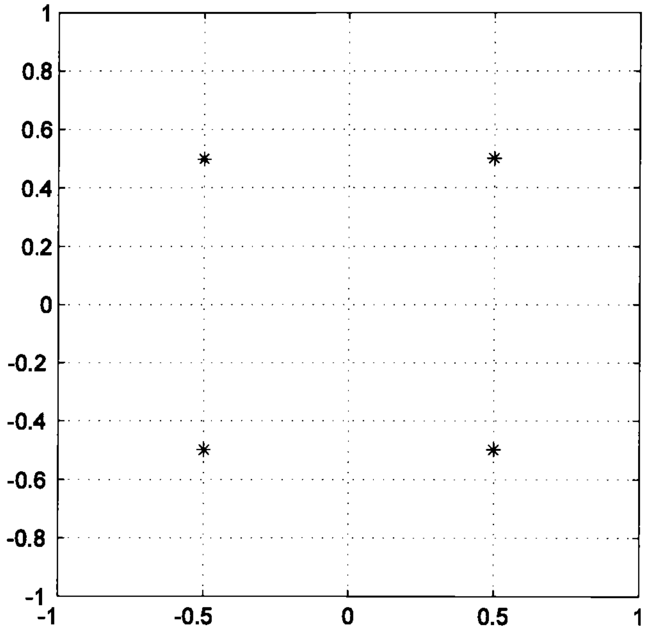

Figure 3.15: 4-QAM Constellation Set

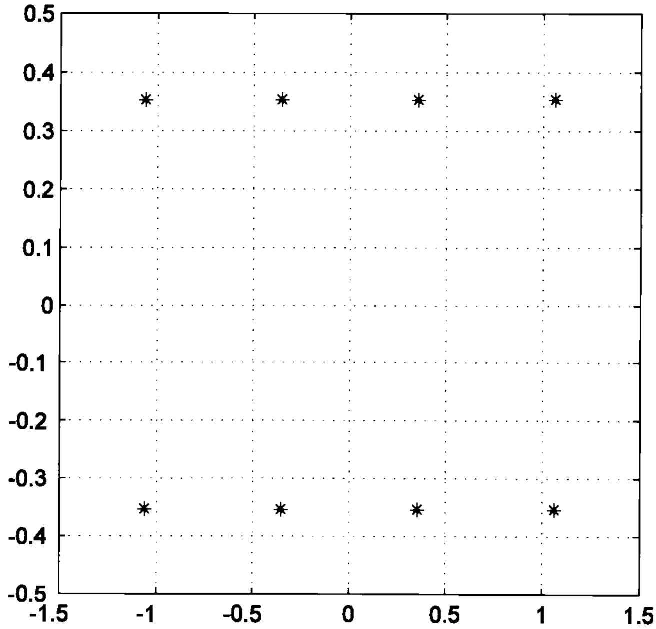

Figure 3.16: 8-QAM Constellation Set

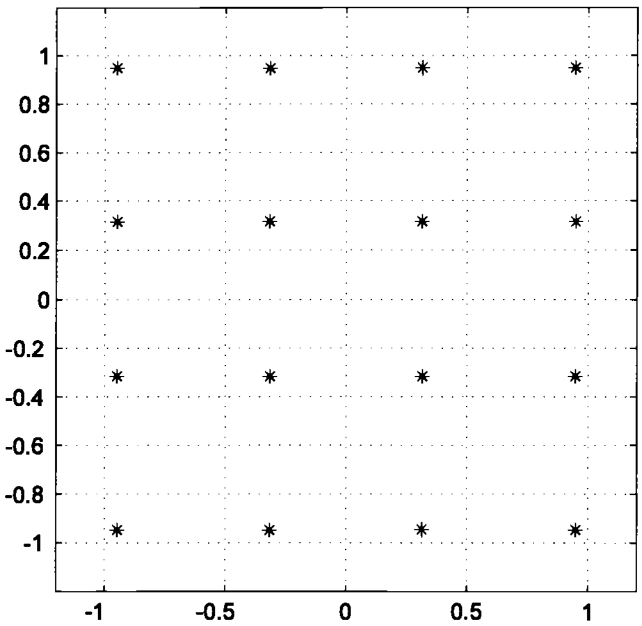

Figure 3.17: 16-QAM Constellation Set

4. SYSTEM PERFORMANCE ANALYSIS

This chapter presents various system performance analyses of STBC based on designed constellation sets for the AWGN and the AWGN plus Rayleigh fading environments. Specifically study the performance in quasi-static fading (the path gains are constant during 2 time slots). In the evaluation of system performance, downlink channel is assumed, i.e., from base station to mobile stations. Moreover, in a Rayleigh fading environment, where the channel is time-varying, perfect channel estimation is assumed.

The signal from the base station is modulated using the proposed augmented STBC scheme, mapped to the designed constellation sets by Grey coding scheme, and then transmitted through the communication channel, i.e., through an AWGN plus Rayleigh fading channel, to mobile stations under the same signal-to-noise ratio and fading constraints. The information is then demodulated at the mobile receiver using maximum-likelihood (ML) decision strategy.

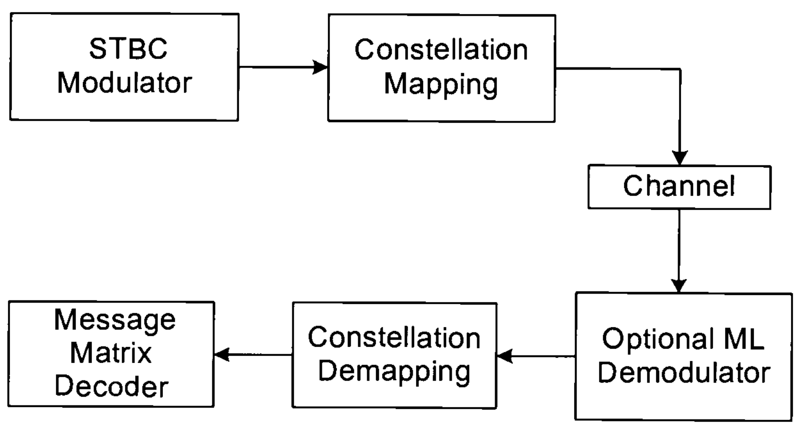

Figure 4.1: A simplified proposed system

We studied the performance of the designed constellation sets in an AWGN channel and also in a quasi-static fading channel where the path gains are constant during 2 time slots, and statistically independent in each k^{th} time slot.

4.1 Performance of STBC in 4-points constellation system.

The constellation sets we used can refer in appendices. One symbol represents 2 message bits.

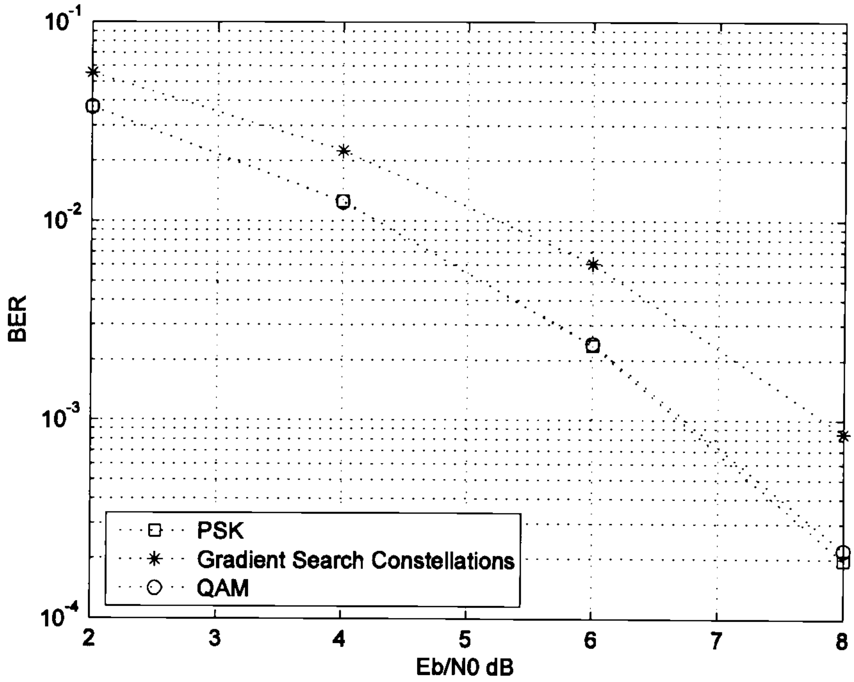

Figure 4.2: Performance of STBC in AWGN using constellation sets 1, 4, 13

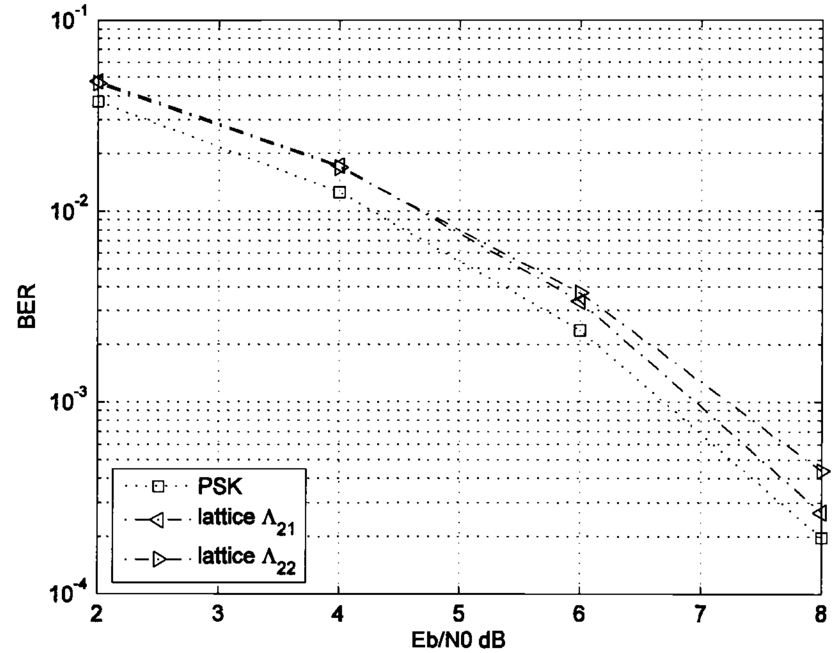

Figure 4.3: Performance of STBC in AWGN using constellation sets 1, 7, 10

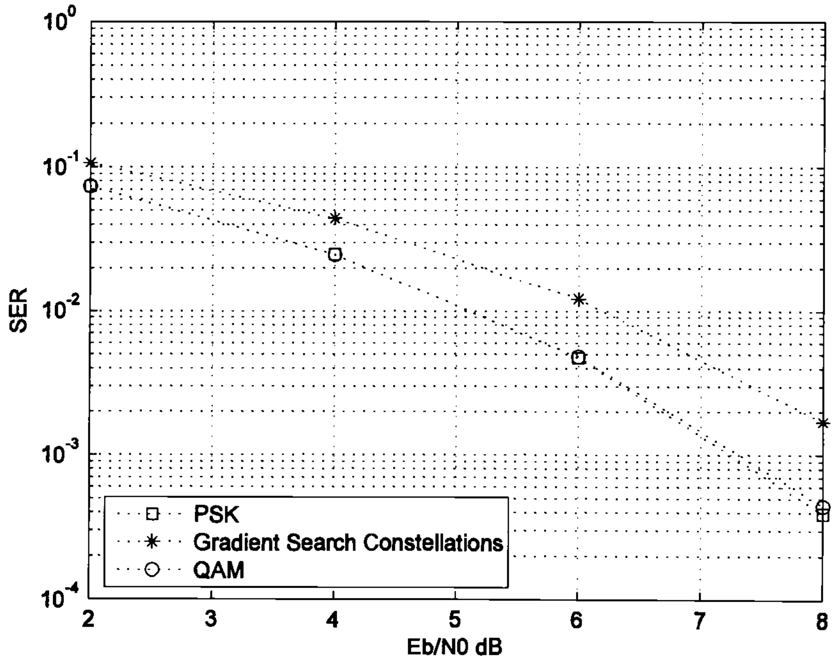

Figure 4.4: Performance of STBC in AWGN using constellation sets 1, 4, 13

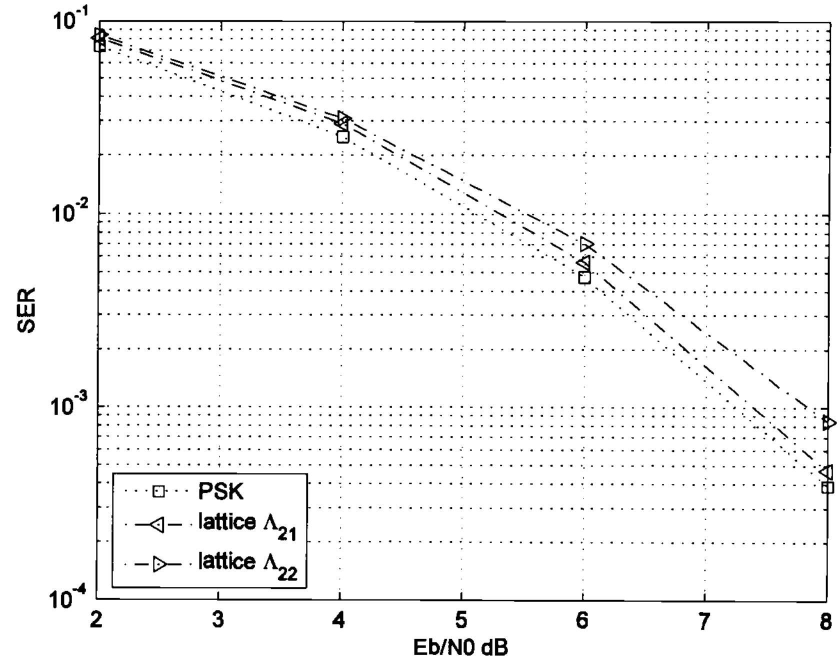

Figure 4.5: Performance of STBC in AWGN using constellation sets 1, 7, 10

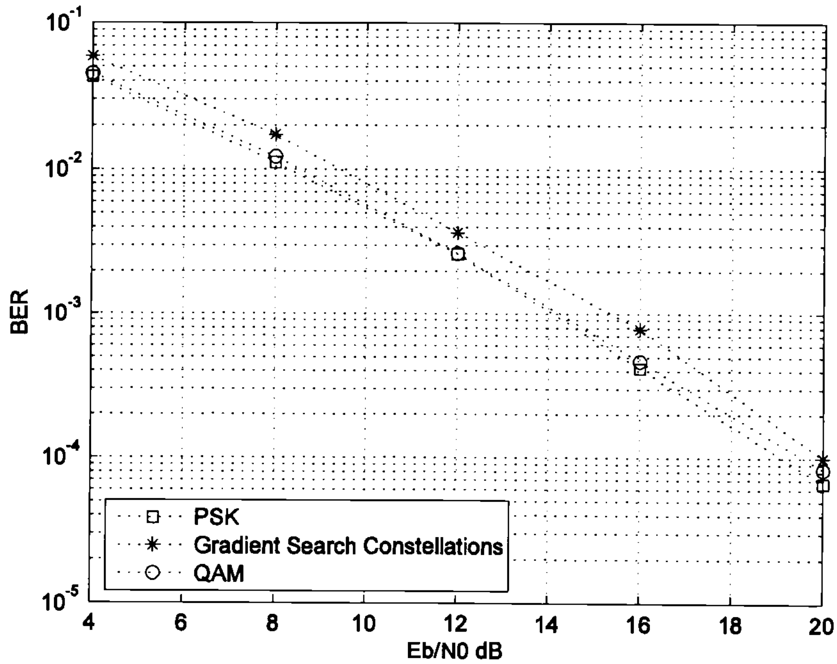

Figure 4.6: Performance of STBC in quasi-static fading using constellation sets 1, 4, 13

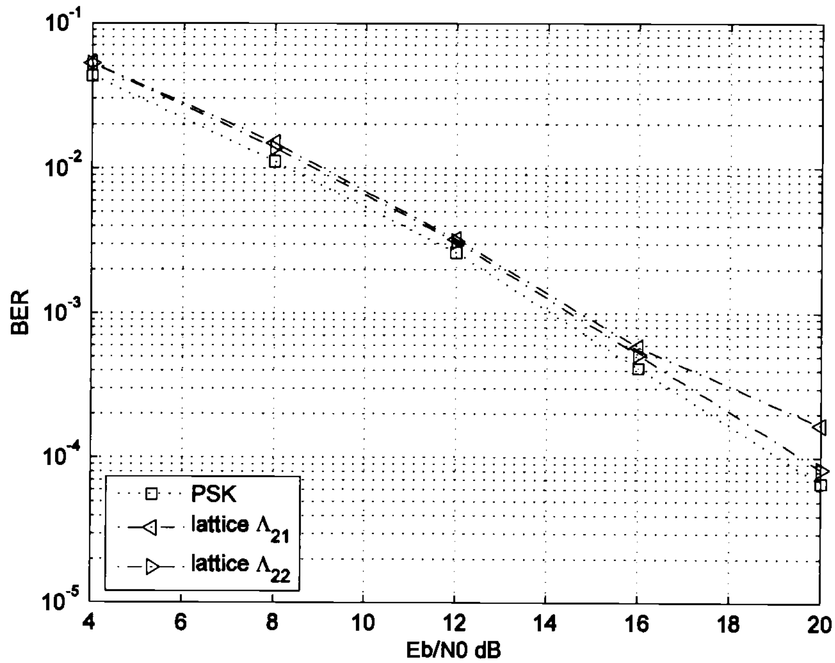

Figure 4.7: Performance of STBC with in quasi-static fading using constellation sets 1, 7, 10

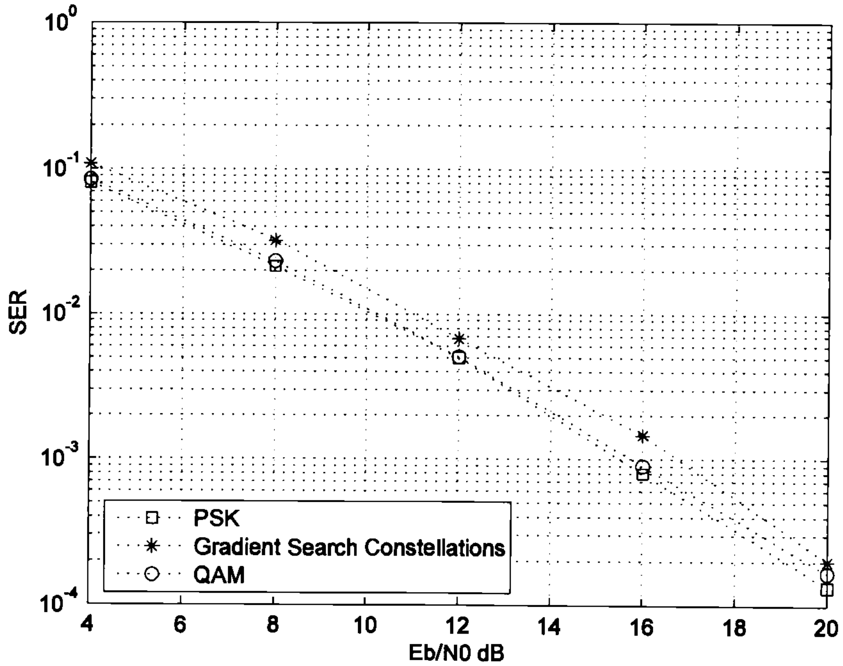

Figure 4.8: Performance of STBC in quasi-static fading using constellation sets 1, 4, 13

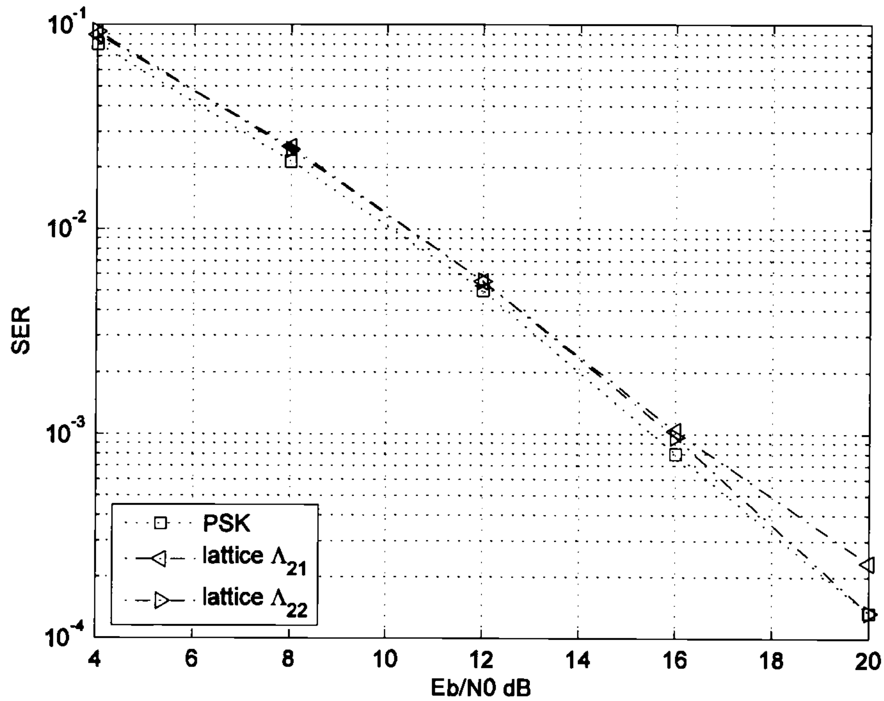

Figure 4.9: Performance of STBC in quasi-static fading using constellation sets 1, 7, 10

From figures 4.2 - 4.5, we can see that in AWGN channel, QPSK transmission and QAM transmission which have same result are the best system. As shown in figures 4.6 - 4.9, we can see that the best performance is achieved by the system with the QAM transmission, in the AWGN plus Rayleigh fading channel. However, it is not much different than those of the QPSK constellation set (the original one). According to figures 4.3 and 4.7, we can see that the constellation sets carved from a lattice, both have almost the same system performance.

4.2 Performance of STBC in 8-points constellation system.

The constellation sets we used can refer in appendix. One symbol represents 3 message bits.

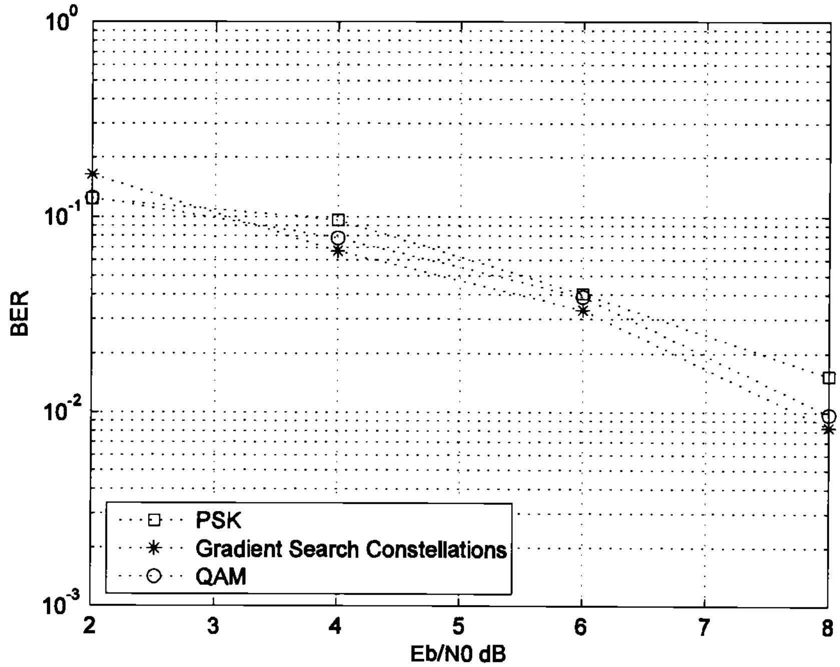

Figure 4.10: Performance of STBC in AWGN using constellation sets 2, 5, 14

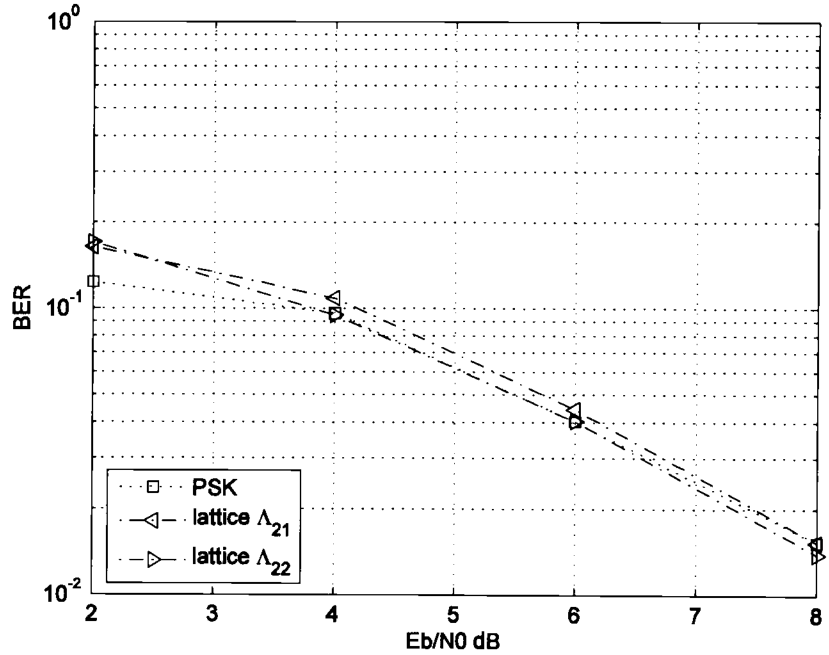

Figure 4.11: Performance of STBC in AWGN
using constellation sets 2, 8, 11

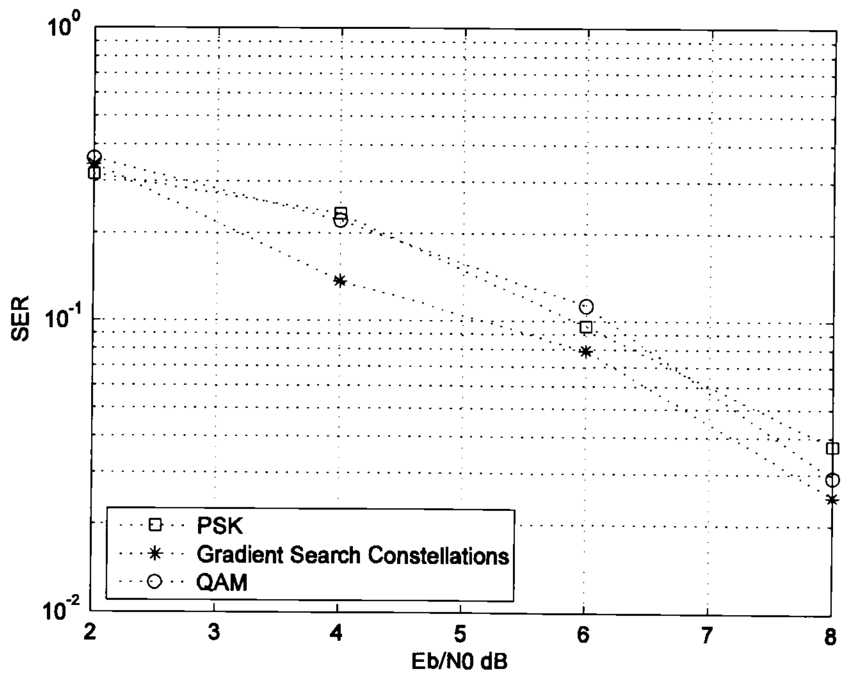

Figure 4.12: Performance of STBC in AWGN
using constellation sets 2, 5, 14

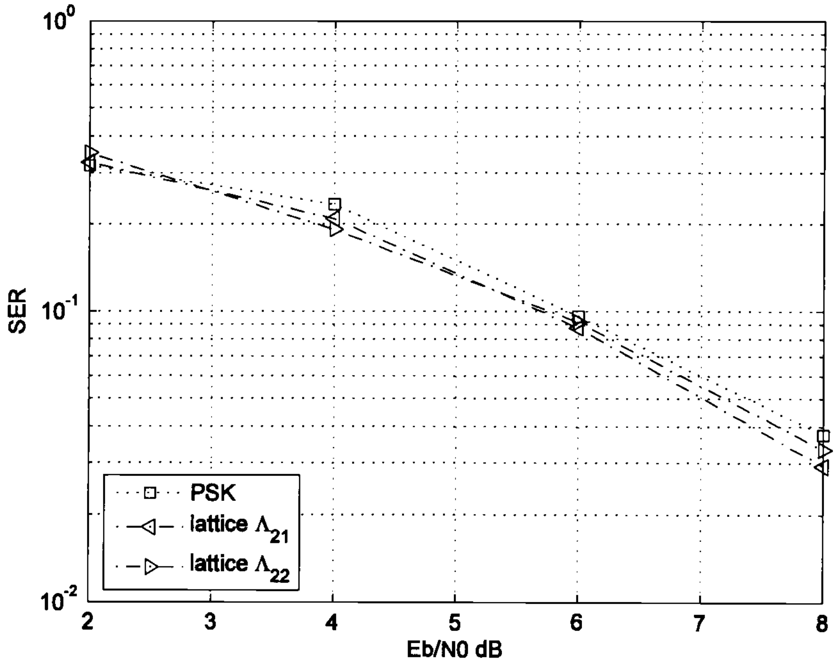

Figure 4.13: Performance of STBC in AWGN using constellation sets 2, 8, 11

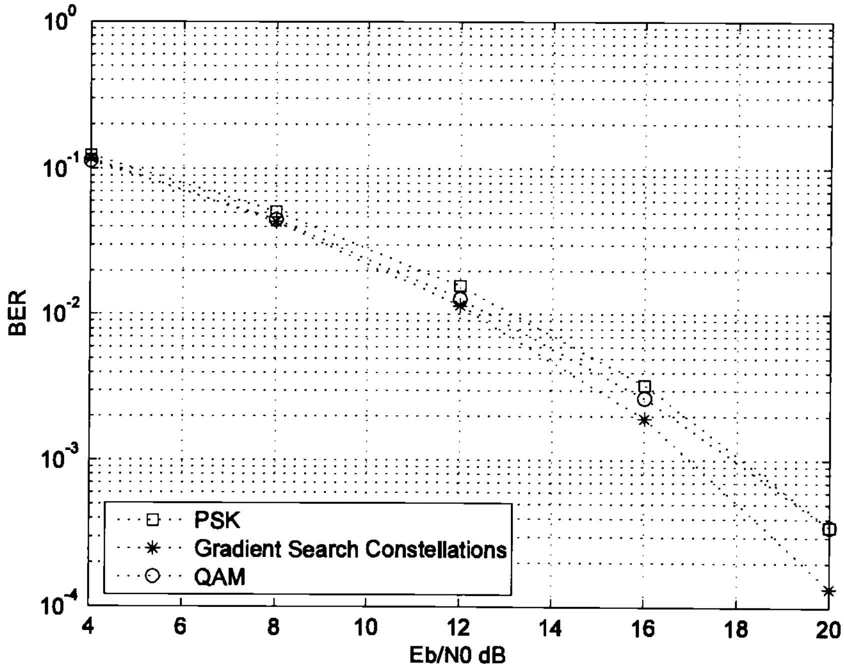

Figure 4.14: Performance of STBC in quasi-static fading using constellation sets 2, 5, 14

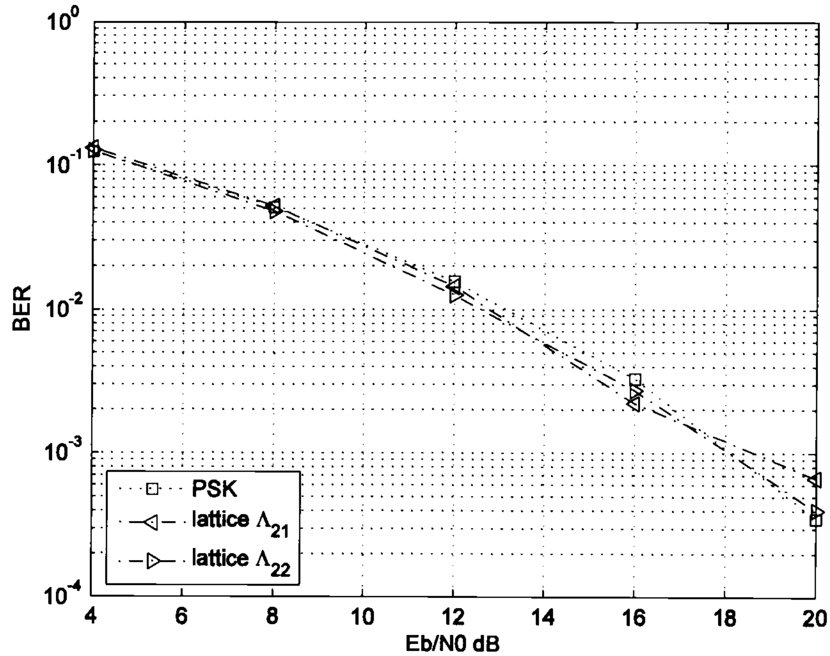

Figure 4.15: Performance of STBC in quasi-static fading using constellation sets 2, 8, 11

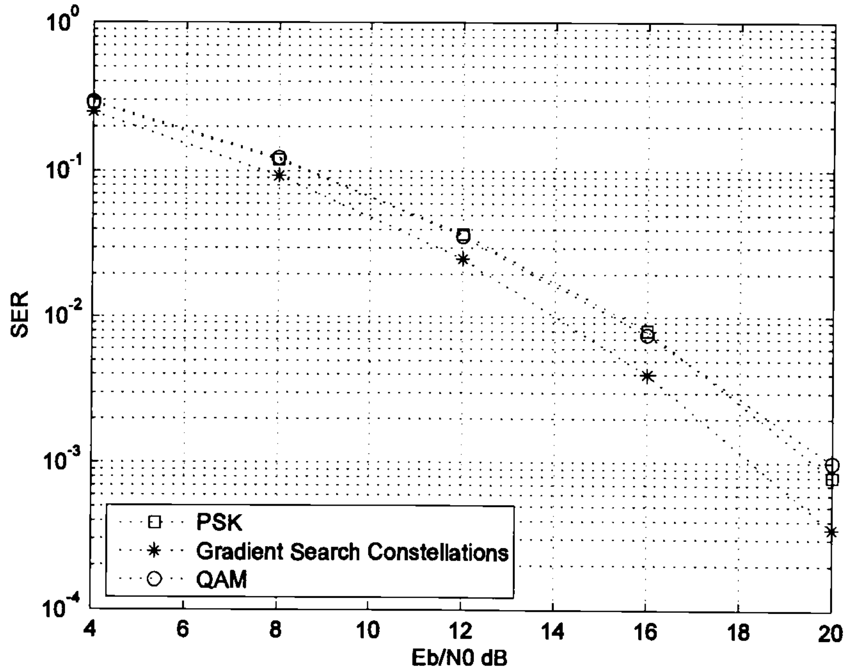

Figure 4.16: Performance of STBC in quasi-static fading using constellation sets 2, 5, 14

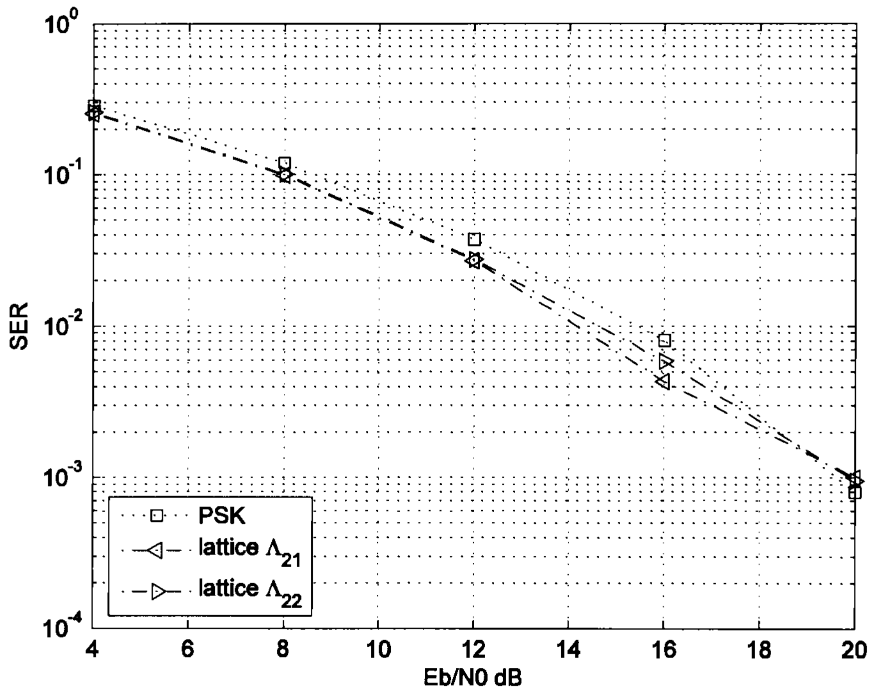

Figure 4.17: Performance of STBC in quasi-static fading using constellation sets 2, 8, 11

From figures 4.10, 4.12, 4.14, and 4.16, the constellation set which gives the best performance is the gradient search constellation set. The BER and SER of the 8-PSK constellation set are worse than the BER and SER of the 8-QAM and gradient search constellation set, respectively. The performance of the gradient search constellation set is 0.5 dB better than those of the 8-QAM constellation set, and 1 dB better than the performance of 8-PSK (the original one). From figures 4.11, 4.13, 4.15, and 4.17, the constellation sets obtained from lattice Λ_{21} and lattice Λ_{22} slightly improve the performance of the system compared to the 8-PSK constellation set. The simulation results of these two systems are not much different from each other. The performance of the systems with four designed constellation sets are much better than the original one (8-PSK), compared with the performance in the quasi-static fading channel.

4.3 Performance of STBC in 16-points constellation system.

The constellation sets we used were are described in the appendix. One symbol represents 4 message bits.

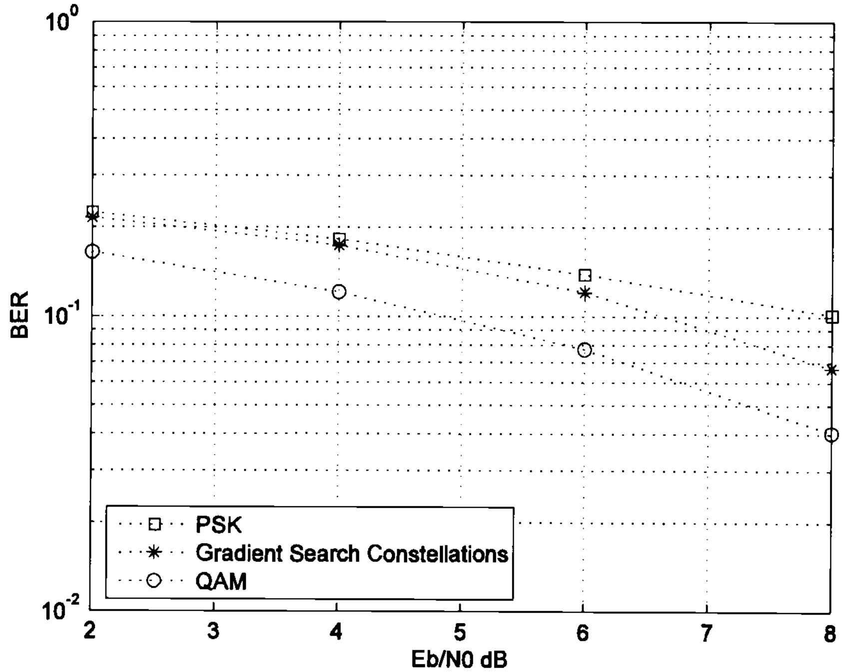

Figure 4.18: Performance of STBC in AWGN using constellation sets 3, 6, 15

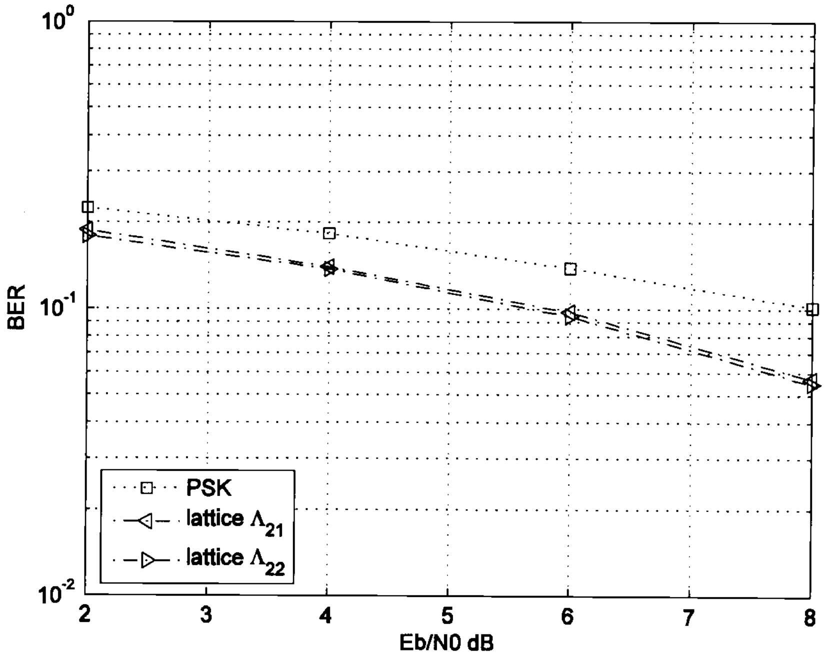

Figure 4.19: Performance of STBC in AWGN using constellation sets 3, 9, 12

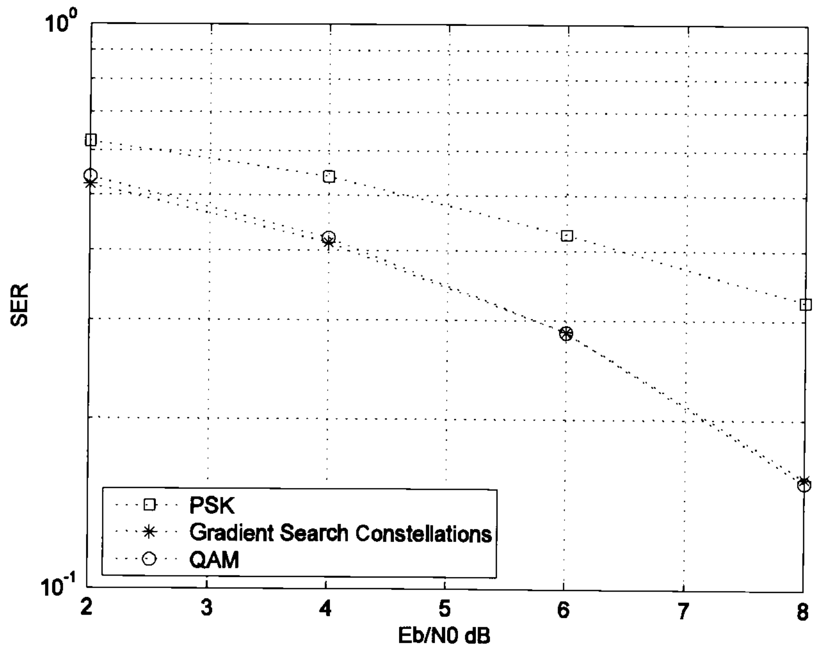

Figure 4.20: Performance of STBC in AWGN using constellation sets 3, 6, 15

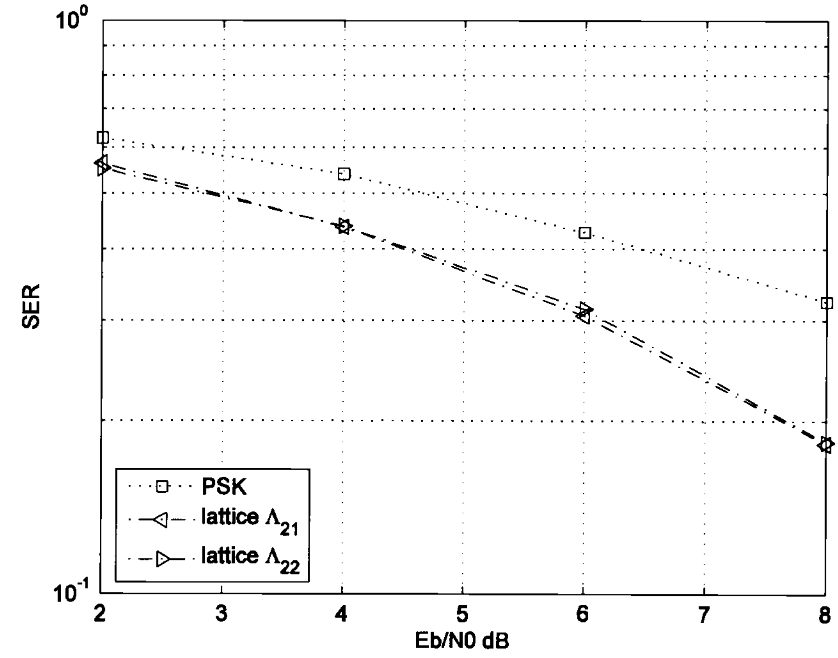

Figure 4.21: Performance of STBC in AWGN using constellation sets 3, 9, 12

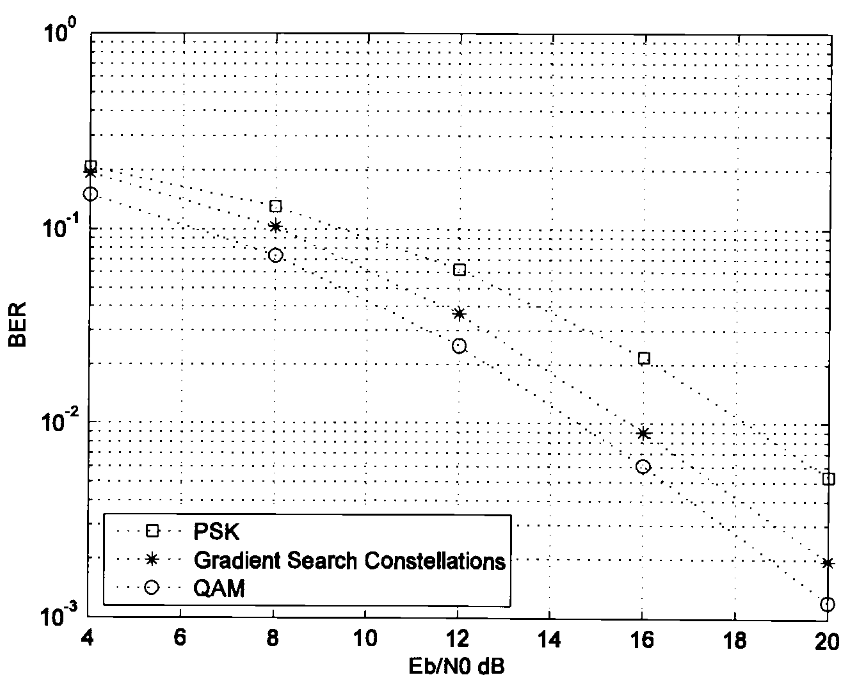

Figure 4.22: Performance of STBC in quasi-static fading using constellation sets 3, 6, 15

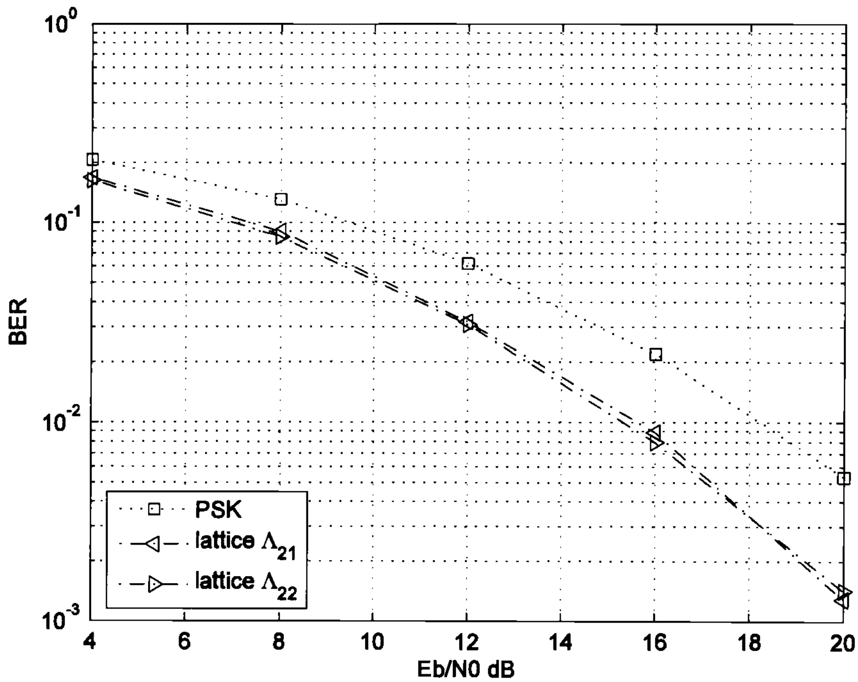

Figure 4.23: Performance of STBC in quasi-static fading using constellation sets 3, 9, 12

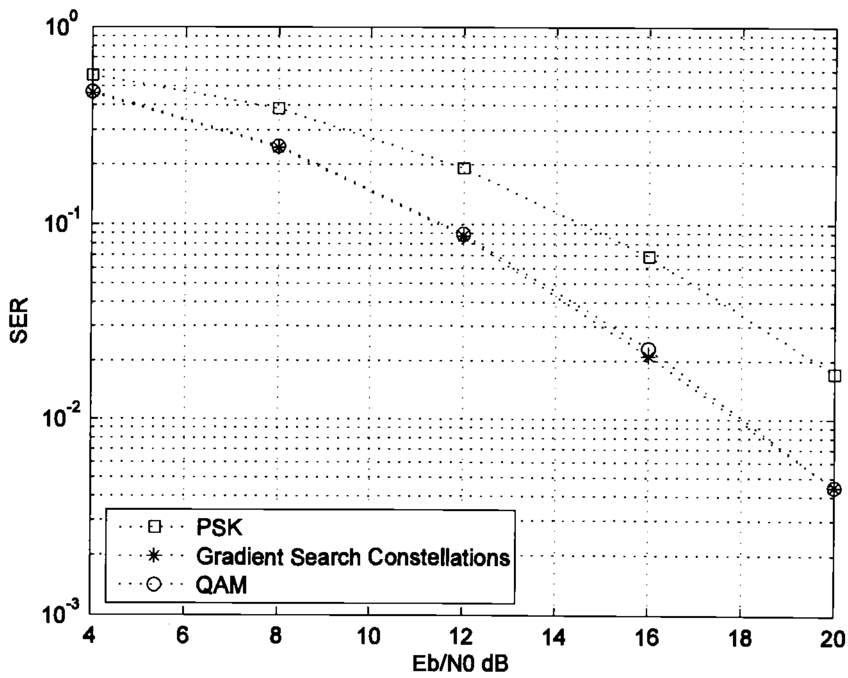

Figure 4.24: Performance of STBC in quasi-static fading using constellation sets 3, 6, 15

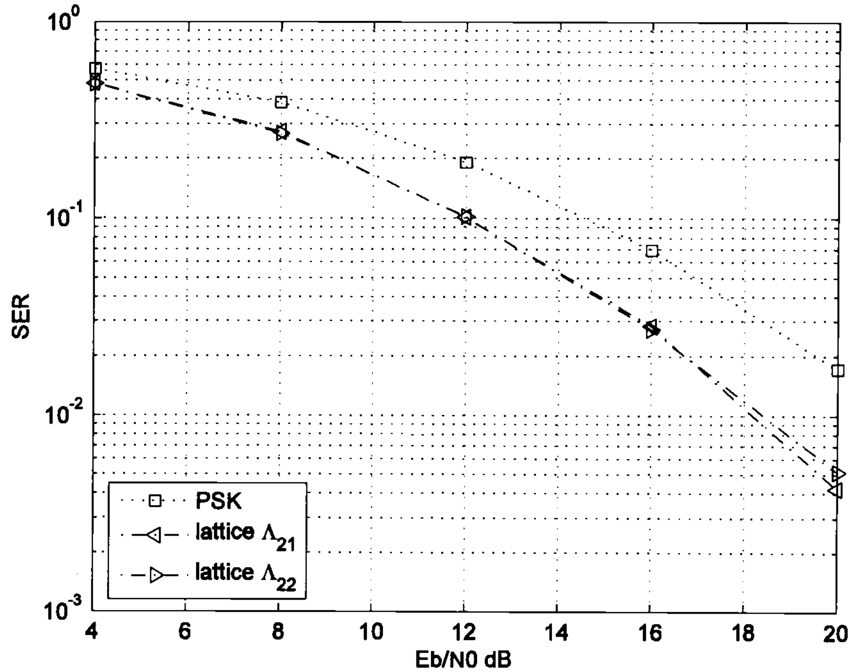

Figure 4.25: Performance of STBC in quasi-static fading using constellation sets 3, 9, 12

From figures 4.18 - 4.26, the lowest BER is from the system with the 16-QAM symbols. In an AWGN channel (figure 4.18), at BER = 0.01, the 16-QAM is 3 dB better than the 16-PSK and 1.5 dB better than the gradient search constellation set. In quasi-static fading channel (figure 4.22), at BER = 0.01, the 16-QAM is 3.6 dB better than the 16-PSK and 1.2 dB better than the gradient search constellations set. From figures 4.19, 4.21, 4.23, and 4.25, the constellation sets carved from lattice Λ_{21} and lattice Λ_{22} are better than the 16-PSK system. In an AWGN channel, at BER = 0.1, both constellation sets from those two lattice are 2 dB better than the 16-PSK constellations set. In quasi-static fading channel, at BER = 0.01, the dB gain of the systems with the constellation sets from the lattice are 2.6 dB lower than those of the 16-PSK symbol. The four systems with designed QAM constellation sets have BER lower than the system with 16-PSK. For the constellation sets carved from a lattice, both have almost the same system performance.

5. CONCLUSIONS AND SCOPE OF FURTHER RESEARCH

This research examined the performance of each designed constellation set in the system with space time block coding. We considered a communication system in a quasi-static fading environment. For the systems with 4-point and 16-point constellation sets, the performances of the designed constellation sets are better than the original one. The best system with 4-point constellation sets and 16-point constellation sets is the one with M-ary QAM constellation. For the 8-point constellation set system, the QAM constellation set from the gradient search algorithm is the best one. The systems with constellation sets carved from lattice Λ_{21} and the system with the constellation sets carved from lattice Λ_{22}, have similar results. The performance improvement of the designed constellation sets increased as the number of bits represented by one symbol increased.

As the part of future research, we recommend the study of encoding the PN code in the system. Also, the chip interleaving, which can be employed to reduce the effect of deep fading, should be investigated.

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APPENDICES

A:

Constellation 1: Figure 3.3
Constellation 2: Figure 3.4
Constellation 3: Figure 3.5
Constellation 4: Figure 3.6
Constellation 5: Figure 3.7
Constellation 6: Figure 3.8
Constellation 7 Figure 3.9
Constellation 8: Figure 3.10
Constellation 9: Figure 3.11
Constellation 10: Figure 3.12
Constellation 11: Figure 3.13
Constellation 12: Figure 3.14
Constellation 13: Figure 3.15
Constellation 14: Figure 3.16
Constellation 15: Figure 3.17