

Modification to the Generic QOSTBCs with Rotated Constellations

Dissertation submitted in the partial fulfilment of requirements for the award of degree of

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Submitted by

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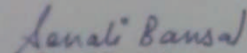
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DECLARATION

I hereby declare that the work, which is being presented in the dissertation, entitled "Modification to the Generic Quasi Orthogonal Space Time Block Codes with Rotated Constellations" in partial fulfilment of the requirements for the award of degree of Master in Engineering in Electronics and Communication Engineering submitted at Electronics and Communication Engineering department of Thapar University, Patiala, is an authentic record of my own work carried out under the guidance of Dr. Sanjay Sharma (Professor and Head), Electronics and Communication Department and refers other research's work which are duly listed in reference section. The matter presented in this dissertation has not been submitted in any other University/Institute for the award of degree.

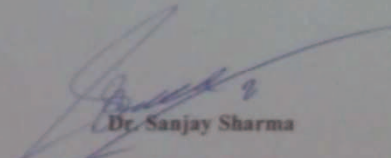
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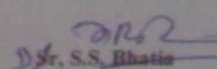
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ABSTRACT

In the recent time, the need of high data rates intended for mobile communication has been increasing considerably. To persuade these huge demands of communications exploration of new methods are going which exploits the restricted resources like bandwidth and power in the best efficient way possible. MIMO (Multiple input and Multiple Output) scheme is one of the techniques having multiple antenna elements on both the link ends which mark an efficient solution for upcoming field of wireless communications systems because it provides higher data rates by utilizing the space domain beneath the various constraints of limited bandwidth as well as transmit power. Space-Time Block Coding (STBC) is one of the MIMO transmit approach that will exploit the transmit diversity. STBCs can be classified into two major categories, named as, Orthogonal Space-Time Block Codes (OSTBCs) and Non-Orthogonal Space-Time Block Codes (NO-STBCs). The Quasi-Orthogonal Space-Time Block Codes (QO-STBCs) comes under the category of non-orthogonal-STBCs and is currently an intensive area of research. The Orthogonal-STBCs are able to achieve full diversity along with lower decoding complexity, although at the price of loss in some data rate. Achieving full data rate along with full diversity in case of two antennas is possible. Taking more than two transmit antennas and achieving full data rate in case of QSTBCs is not possible; there would be some loss in the diversity gain.

The foremost aim of this work is to grant an integrated theory of QSTBCs considering four transmit antennas and one receive antenna. The dissertation majorly consists of two parts: In the former part detailed analysis is done for various QOSTBC's with a technique employed of removing interference and also assuming that there is no knowledge of channel on the transmitter front. And in the later part rotation mechanism is applied on a specific code to achieve full rate, diversity as well as pairwise decoding.

In the first part of this dissertation a description of QSTBC considering four transmit antennas is shown. It is also revealed that by applying linear transformations on various QOSTBC's and other conventional codes transformation into one another can be done. The major transformation of converting (4×1) MIMO channel into its equivalent virtual (4×4) MIMO highly structured channel is shown for the same quasi orthogonal space time block codes. The arrangement and the composition of the equivalent channel are of key significance for measuring the performance of these QSTBCs. The off-diagonal elements of this virtual channel matrix are accountable for self-interference at the receiver from other signals.

The closer these off-diagonal elements of the virtual channel matrix are to zero, the closer is the code to an orthogonal one. After that, outcome of decoding technique is illustrated upon the performance of the code. And finally, a new QOSTBC is anticipated whose decoding complexity is much less in comparison to the conventional QOSTBC without any loss in its performance.

In the second part of this dissertation rotation mechanism is employed upon the existing QOSTBC. . In order to achieve full transmit diversity a novel criteria of rotated constellations for different transmitted symbols is explored. This scheme outperforms transmit antenna shuffling (TAS) and also removes its disadvantage of dedicated feedback channel for CSI. The final code thus achieved is very influential as it grants code rate one, full diversity as well as pairwise decoding. Finally, simulations show that bit error rate performance is significantly improved by gain of 3dB in comparison to TAS scheme

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LIST OF ABBREVIATIONS

MIMO	Multiple Input Multiple Output
STC	Space-Time Coding
STBC	Space-Time Block Code
O-STBC	Orthogonal Space-Time Block Code
QO-STBC	Quasi- Orthogonal Space-Time Block Code
LOS	Line of Sight
QPSK	Quadrature Phase Shift Keying
MRRC	Maximum Ratio Receiver Combining
EVC	Equivalent Virtual MIMO Channel Matrix
CR	Constellation Rotation
ML	Maximum Likelihood
CGD	Coding Gain Distance
SVD	Singular Value Decomposition
TAS	Transmit Antenna Shuffling
CSI	Channel State Information
ML	Maximum Likelihood
ZF	Zero Forcing

1.1 Advancement in Wireless Communication

Wireless communications is, by any appraise, the highest growing fragment of the communications industry. Cellular systems also experience an exponential enlargement over the past decades and now currently there are approximately two billions of users all around the world. Design of this wireless networks differs essentially from other wired network designs because of the character of the channels through which transmission takes place. This channel is an erratic and a difficult medium of communication. Firstly, the available radio spectrum is a limited resource which must be allocated to different applications as well as systems. And due to this reason our spectrum is proscribed by the various regulatory bodies present regionally as well as globally.

Needs for increase in capacity in wireless communications, determined by cellular mobile phones, Internet and various multimedia forces are increasing rapidly. Whereas, the accessible spectrum is restricted in nature and the capacity requirements cannot be fulfilled without any noteworthy enhancement in spectral efficiency. Advancements in coding make it possible to move toward the actual Shannon limit of capacity in the systems having single link of antenna. Momentous auxiliary advancements in spectral efficiency are also obtainable when the number of antennas is increased on both transmitter and the receiver side. These capacity limits highlights the probable spectral efficiency of multiple input multiple output channels, which are growing approximately on a linear scale with the number of antennas, if we assume the ideal propagation. The capacity is further expressed by the utmost attainable data rate for a randomly low likelihood of error, provided that the signal may be programmed by an arbitrary long space-time code.

In the last decade, the stipulate for higher data rates in cellular communication, wireless local area networks (WLANs) and high-definition (HD) audio and video broadcasting services is increasing [4]. To provide an access to the Internet and the multimedia services we require an increase in data rate which is of orders of magnitude beyond the capacity of today's technology. One of the most noteworthy and talented advancements in digital infrastructure which could meet this growing demand of higher rates is by

using numerous antennas on the transmitter side in addition to on the receiver side. Further deploying multiple antennas on the transmitter along with receiver will create a MIMO channel that not only offers higher transmission rates, but also improves our system's consistency and sturdiness to noise in comparison to single antenna systems.

1.1.1 Motivation of Thesis

Besides achieving good code rate and high diversity levels, the decoding and implementation complexity for STBC codes is also an important concern. In this thesis, we focus upon creating popular conventional codes with minimum interference and secondly introduced the concept of rotated constellation further to improve the results. Specifically, we will attempt to address the following research issues concerning to STBC.

- QOSTBC experiences interference from neighbouring signals and channels, thus causing higher decoding complexity.
- With QOSTBC, we are able to achieve full rate but not full diversity, concept of rotated constellation can be employed to achieve both.
- We found that CR technique, which enables QOSTBC to achieve full diversity, actually increases the decoding complexity. Hence this trade-off to achieve both is an interesting research issue.

1.2 Technical Challenges

Despite the fact that assorted applications have diverse specifications and there use of different wireless technologies, where most of them will face similar challenges [5]. The preference of the different challenges through wireless communications might not be identical for different applications; however, the list is applied to approximately all applications. Some of these challenges in wireless communications are:

- A need for high data rates
- Portability
- Quality of service
- Mobility
- Privacy/security
- Interference from other users

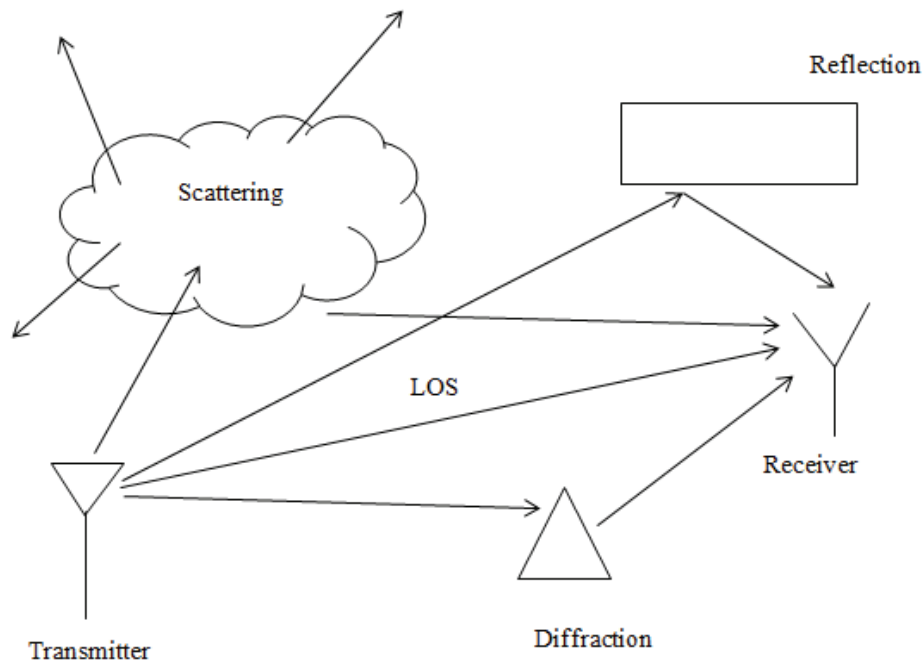


Fig1.1: An example showing different paths travelled in a wireless channel

Several demands, let's say the requirement of higher data rates and quality of service, are not unique to wireless communications. But, some of these challenges are particular to wireless communication systems. The signal transmitted from transmitter follows different paths before reaching the receiver as shown in figure 1.1. Taking example, the portability necessity will result in the need of batteries and the limitation in the battery life created a challenge to find algorithms having lower power consumptions. This required unique awareness in the design of transmitters and receivers. Since the base station is not operating upon batteries and does not require the same power restrictions, it may be principally advantageous to have asymmetric complexities in special ends.

An additional example of such challenges within wireless communications is the connectivity in wireless networks. The power of this received signal is proportional upon the distance present involving the transmitter as well as the receiver. Consequently, it is significant to make sure that if, for the reason that the mobility of the nodes, as their distance increases, there nodes remain connected. Furthermore, owing to this rapid changing temperament of the wireless channel, mobility also brings number of new challenges into the depiction. An additional important confront in a wireless channel is the intervention from various other users or other sources of electromagnetic waves. In a wired system, the communication atmosphere is an additional under control and the

interference would be less damaging. Whilst the claim of having high data rates as the performance of the signal processors increases exponentially, the spectrum as well as the bandwidth is limited [4,5]. This limited amount of bandwidth for the wireless channels will add an increased impairment. The growth is slow for the increase in battery power and there is a rising demand for less important terminals and handset strategies. Whereas on the other hand, the users desire the superiority of wire-line communication and the wire-line information rates increases rapidly. Researchers have faced much challenge to satisfy these elevated expectations through the slender duct of wireless channels.

1.3 Large-Scale Fading

Large-scale fading or attenuation occurs by numerous factors together with propagation losses, antenna losses, as well as filter losses. The average received signal, or the large-scale fading factor, generally decreases logarithmically as per the distance. This logarithm factor, or else the path gain exponent, is dependent upon the medium through which propagation takes place and also the surroundings present between the transmitter and the receiver. Taking an example, considering free space environment, as that of satellite communications, the exponent is two. Here d symbolizes the distance linking the transmitter as well as the receiver. For various other propagation environments, let say urban area, the path loss exponent is usually considered greater than 2. Or in other words, if the average power transmitted is P_T , the average received power is shown as:

$$P_R = \beta d^{-\nu} P_T \quad (1.1)$$

Here ν represents the path loss exponent and β is the parameter which depends upon the frequency and various other factors. This sometimes is also referred to as the log-distance path loss model because the path loss and the distance are having logarithmic relationship between them.

1.4 Small-Scale Fading

Small-scale fading or equivalently fading is caused due to interference present involving two or more translations of the transmitted signal which appear on the receiver front at marginally diverse times. These signals, called as multipath waves, combines on the receiver front. This consequential signal varies widely with amplitude and phase for a very short period of time, equally a short journey distance, is such that this large-scale

path loss effects might get ignored. This unpredictability causes multipath effects and the fading also results in the application of different algebraic arguments which model these wireless channels. For that reason, these statistical models are required to examine the behaviour that is amplitude and power of the signal received.

1.5 Statistical Models meant for Fading Channels

Because of the range of features which are involved in the propagation in cellular portable surroundings, it is appropriate to relate arithmetical procedures for describing signal deviations. In case of narrowband arrangement, the transmitted signal occupies a bandwidth which is lesser in comparison to the channel's coherence bandwidth, which is further referred to as the frequency range upon which the channel's mechanism in fading is correlated to each other [7]. It means that, all the spectral components present in the transmitted signals suffers from same fading and same attenuation. This type of fading which occurs is commonly called as frequency non selective or frequency flat fading. Whereas on the other side of the coin, if the bandwidth of the signal transmitted is little greater in comparison to the coherence bandwidth of our channel, the spectral components of this signal transmitted which are having frequency separation a lot more than that of coherence bandwidth will fade without any kind of help. This spectrum of the signal received becomes fuzzy, because the associations among the diverse spectral components are not exactly equivalent to that present in case of transmitted signal itself. This observable fact is referred to frequency selective fading. In case of wideband systems, frequency selective fading occurs mainly to the transmitted signals.

1.5.1 Rayleigh Fading

We consider that the signals with the single tone are transmitted with invariable amplitude. In a typical land mobile radio channel, we are assuming that the straight wave is hindered and the cellular phone unit will receive only the waves reaching after reflection. Whilst the number of waves which are reflected is large, accordingly to the central limit theorem, it states that when any two quadrature components of the signals received are uncorrelated in nature with Gaussian random processes and having zero mean with variance equal to σ_s^2 . And as a consequence, the envelope of this received signal calculated at any direct of time interval will go through a Rayleigh probability distribution and its phase will abide by unvarying distribution present among the periods

from $-\pi$ to π . The probability density function (pdf) of the Rayleigh distribution is given as:

$$p(a) = \begin{cases} \frac{a}{\sigma_s^2} \cdot e^{-a^2/2\sigma_s^2} & a \geq 0 \\ 0 & a < 0 \end{cases} \quad (1.2)$$

If we normalize the probability density function such that the standard signal power ($E[a^2]$) is put to unity, then the Rayleigh distribution when normalized will become:

$$p(a) = \begin{cases} 2ae^{-a^2} & a \geq 0 \\ 0 & a < 0 \end{cases} \quad (1.3)$$

The pdf for any normalized Rayleigh distribution is shown in the fig1.2 below:

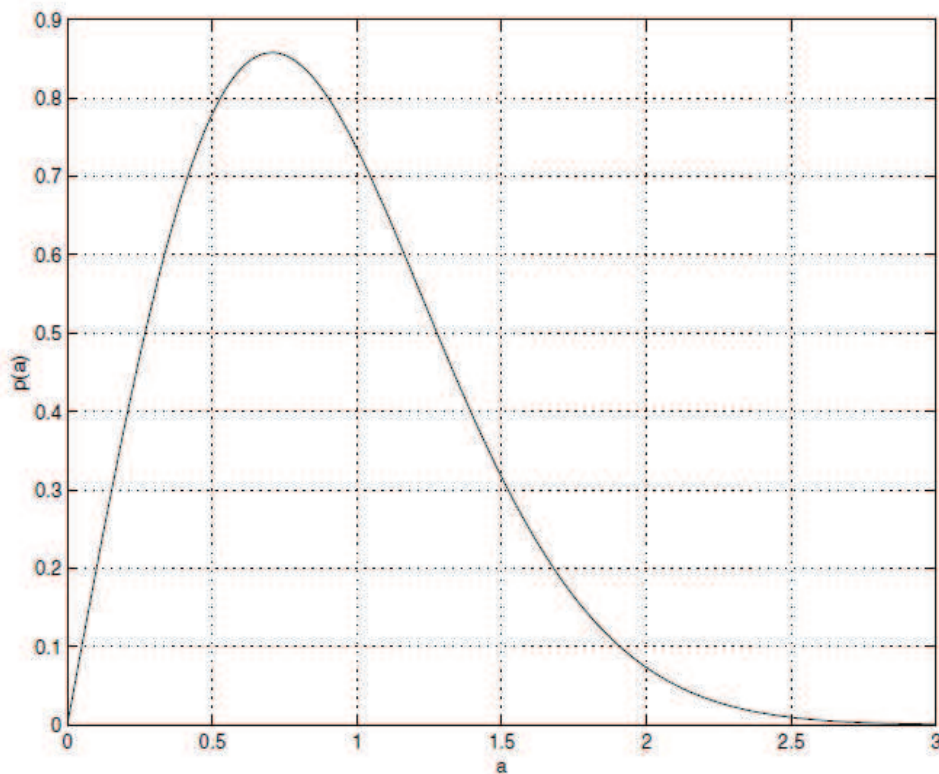


Fig1.2 Distribution showing Rayleigh Fading

1.5.2 Ricean Fading

In some of the propagation circumstances, like that of satellite or any micro cellular radio channel, there has been an effectively no interruption on the line-of-sight (LOS) or direct path. This received signal is composed of direct wave as well as reflected waves. The direct wave received is a motionless non faded signal with constant amplitude, whereas

the waves which are reflected employ self-sufficient random signals. Their accumulation is referred to as the scattered constituent of the received signal. Whilst the quantity of reflected waves is outsized, the quadrature components of this scattered signal could be illustrated as a Gaussian random development having its mean as zero with variance equal to σ_s^2 . When a direct signal having constant amplitude as well as Rayleigh disseminated scattered signal are summated it would result in a signal having Ricean envelope distribution. The pdf of any Ricean distribution is given by:

$$p(a) = \begin{cases} \frac{a}{\sigma_s^2} \cdot e^{-\frac{(a^2+D^2)}{2\sigma_s^2}} I_0\left(\frac{aD}{\sigma_s^2}\right) & a \geq 0 \\ 0 & a < 0 \end{cases} \quad (1.4)$$

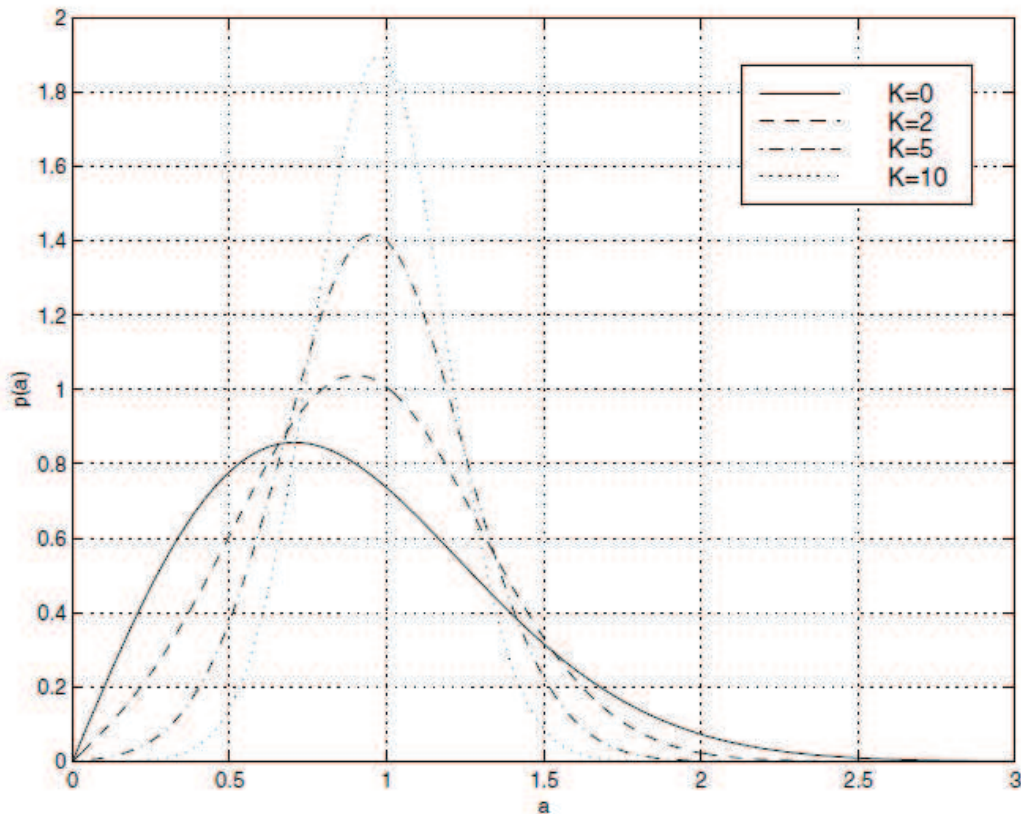


Fig1.3 Distribution showing Ricean Fading

1.6 Diversity Techniques

In wireless cellular phone communications, there are many diversity techniques which are widely used for reducing the possessions of multipath fading and advance the dependability of transmission without any increase in the transmitted power or any kind

of sacrifice in the bandwidth. The diversity technique is having numerous replicas of the signals to be transmitted on the receiver front, which are holding the similar information except that of small altitude of correlation present in their fading statistics. The basic initiative following the diversity technique is that, whenever two or more than autonomous samples of a particular signal are taking, the samples would lighten in any kind of uncorrelated manner, for example there are little samples which are rigorously faded but others are least attenuated in comparison to that. This meant that the likelihood of all these samples simultaneously underneath some given level are much lower when compared to the probability of any other individual sample beneath that level. Thus, when proper combination of the various samples is done, it results in deeply reduced harshness of fading, and in the same way, improves consistency of transmission. In the majority of wireless communication systems there are many methods of diversity which are used in arrangement to achieve essential performance. Accordingly the province where diversity was introduced, diversity techniques can be classified into time, frequency and space diversity.

1.6.1 Time Diversity

Time diversity is attained if we transmit alike messages but in dissimilar time slots and this would consequence in an uncorrelated faded signals present on the receiver front. The essential time severance should be at least equivalent to the coherence time of the channel, or else the reciprocal taken of the fading rate. Error control coding is habitually used in digital communication systems if we have to provide a coding gain which is relative to various uncoded systems. In cellular communications, error control coding is united with interleaving which finally leads to achievement of time diversity. In such cases, the duplicated copies of the signals transmitted are commonly provided towards the receiver in the outline of redundancy in the time sphere which was introduced in error control coding.

1.6.2 Frequency Diversity

In case of frequency diversity, there exists a quantity of diverse frequencies which would transmit an identical message. Such kind of frequencies needs to be estranged enough to make this fact that certain independent fading takes place with every frequency. The frequency disconnection of the order of numerous times it guarantees the coherence

bandwidth in case of any channel that the statistics as observed in for dissimilar frequencies are fundamentally uncorrelated. The coherence bandwidth is different for different propagation environments. In mobile communications, the replicas are taken in case of transmitted signals which are usually given to the receiver front in the form of redundancy during the frequency domain commenced by spread spectrum for instance direct sequence spread spectrum (DSSS), frequency hopping and multicarrier modulation.

1.6.3 Space Diversity

Space diversity is an admired skill in the field of wireless microwave communications. Besides calling it space diversity, it is also referred to as antenna diversity. It is characteristically implanted with the concern of several amounts of antennas or antenna arrays which have already been approved collectively in space for the purpose of transmission and/or reception. Such kind of multiple antennas are physically separated by applying a proper aloofness with the intention of the individual signals remaining uncorrelated in nature.

1.7 Organization of Dissertation

This dissertation comprises of six chapters including introduction which is as follows:

- *Chapter 2* deals with literature review where study is done upon existing space time block codes and their various decoding techniques are also discussed. All the work done in this field is summarized in this chapter.
- *Chapter 3* provides a methodical discussion in the field of STCs and a more explicit conversation on space-time block coding (STBC). Firstly, Alamouti STBC has been explained which grants a transmit diversity of two. Various Orthogonal designs are shown and presented with their performance evaluations shown by simulations. Later, QOSTBCs are defined in detail. They provide the comprehensive structure of QO-STBC and presented pairwise decoding. Taking conventional codes as Jafarkhani, ABBA and Tirkkonen illustrations are provided.
- *Chapter 4* is dedicated to the study of quasi-orthogonal space time block codes in open-loop transmission systems. For QO-STBCs only examples have been accounted in the narrative devoid of structured scrutiny and specific definition.

The matrices of various known QO-STBCs have been analysed and their new adaptations of these codes have been obtained. Also, the designing of the receiver configuration for the QO-STBC projected has been studied extensively. After that, the discussion on the performance of various receivers under QO-STBC transmission is shown. And finally, this chapter presents a novel QO-STBC scheme.

- *Chapter 5* provides very simple method to improve the QO-STBC transmission strategy. In order to achieve full transmit diversity a novel criteria of rotated constellations for different transmitted symbols is explored. This scheme outperforms transmit antenna shuffling (TAS) and also removes its disadvantage of dedicated feedback channel for CSI. The final code thus achieved is very influential as it grants code rate one, full diversity as well as pairwise decoding. Finally, simulations show that bit error rate performance is significantly improved by gain of 3dB in comparison to TAS scheme
- *Chapter 6* highlights the conclusion of the dissertation and tells about future prospects.

1.8 Problem Formulation

Since several designs of STBS are known in literature, by comparison and summarizing various properties various observations have been drawn:

- Low rate OSTBC and QOSTBC: The current designs available of OSTBC and QOSTBC though possesses low decoding complexity (maximum likelihood) but they have the major limitation of rates less than one; the rate equals to one exists only for the OSTBC with 2 transmit antennas and for QOSTBC with 4 transmit antennas only.
- High Complexity full rate STBC: Codes with full data rate can achieve full diversity but their decoding complexity is high since all the symbols transmitted have to be decoded jointly in order to achieve full diversity. In practical cellular communication systems, the number of antennas employed on the cell phone should be less than the number of antennas employed on the base station; in such case the maximum symbol rate achievable is equal to the number of antennas present on the receiver front.

Accordingly, achieving full diversity and low decoding complexity is the major issue pertaining with code rate equal to one. Thus designing of such STBC is one of the main challenges in this thesis.

1.9 Objectives

The major objectives of this thesis are to design quasi orthogonal space time block codes with lower decoding complexity, to achieve full transmit diversity for both conventional proposed QOSTBCs and for CR techniques in quasi static flat fading channel, Specifically we aim to:

- To propose a generalized scheme to eliminate interference terms achieved in detection matrix with lower decoding complexity.
- To decode a signal with lowest possible computational complexity utilizing algebraically the structure of the signal points of the constellation.
- Design a new constellation rotation technique to optimize the decoding performance of QOSTBC for higher rate and full diversity.
- Comparison of CR technique with TAS mechanism and showing that CR outperforms the later.

Chapter 2

Literature Review

The literature review reviews, deduces, and evaluates various existing "literature" (or available material) in command to set up current understanding of a subject. The reason for doing so relates to in progress research to build up that knowledge. The literature review may determine a disagreement, establish the need for further research, and define a topic of inquisition.

Tarokh *et al.* [2] in this paper, a new paradigm for communication called space time coding was introduced using multiple transmit antennas. Data is encoded using space time coding and then encoded data is split in such a way that n streams are transmitted using n Transmit antennas. Orthogonal structure of space time block code is used which leads to linear processing at the receiver by using maximum likelihood decoding. Then generalization of these orthogonal designs is made for both real as well complex constellations for any number of antennas. With this, he was able to achieve maximum possible transmission rate.

S.M. Alamouti. [8] In this paper, he presents a transmit diversity scheme for two branches, with two transmit and one receiver antenna. This scheme provides same diversity order as that of maximal ratio receiver combining (MRRC) with transmit and two receiver antennas. It has also been proved that this scheme can be generalised to two transmit and M receiver antennas to achieve diversity order of $2M$. Also, it does not require any feedback from receiver to transmitter, no bandwidth expansion and achieve computation complexity similar to MRCC. However, if total radiated power is to be kept same, transmit diversity scheme has disadvantage of 3dB because of transmission of two distinct symbols from two distinct antennas.

Yuen *et al.* [10] in this paper, they have considered Quasi-Orthogonal space time block code with its minimum decoding complexity (MDC-QC-STBC), by formulating its algebraic structure and proposing a systematic method for its construction. The decoding used here is ML decoding which require joint detection of any two real symbols. Using any modulation scheme like square or rectangular quadratic amplitude modulation

(QAM), Multiple Phase Shift keying (MPSK), an optimal constellation rotation angle is obtained which helps us to achieve full diversity and optimum coding gain. It has been shown that proposed MD-QC-STBC is able to achieve desirable properties, such as better scalability in adjusting of transmit antennas, more even distribution of power among antennas in comparison to coordinate interleaved orthogonal design (CIOD) and Asymmetric CIOD i.e., ACIOD

Yier Yan *et al.* [12] in this paper, a new method to transmit more information by Orthogonal Space time block code (QSTBC) has been proposed for four antennas. Here two different STBC matrices are used for transmitting one additional bit to achieve code rate as high as 9/8. Also coding gaining matrix maintains its full rank and full diversity. A property of Forbenius norms for received signals is considered with fast Maximum Likelihood (ML) decoding, it helps to reduce complexity by checking non-null values of Forbenius norm on the Transmitter side.

Sandipan Kundu *et al.* [13] In this process, the lowest computational-complexity Maximum Likelihood (ML) decoder is presented for 4x4 full-diversity Quasi Orthogonal Space Time Block codes (QO-STBC), symbols taken from rectangular or square Quadrature Amplitude Modulation (QAM) constellations. Simplified quadratic maximum likelihood decoding is being presented and signal points of QAM Constellations are being utilized which results in saving of complexity. Simulation study also demonstrates the same idea. If size of signal constellation increases, complexity gap between the previous and the developed ML algorithm grows. Explaining this with an example, for 256 QAM the presented algorithm requires less than a third of multiplication and addition, whereas for 64 QAM, it is less than half.

Ezio-Biglieri *et al.* [14] In this paper, focus is on full rate and fast decodable space time block code (STBC's) for 4x2 and 2x2 multiple-input multiple-output (MIMO) antennas. Firstly conditions are derived and then a criterion is designed for reduced complexity ML decodable 2x2 STBC's, hence applying them to two families of codes recently discovered. Finally, a novel 4x2 STBC is derived, which outperforms all previously known codes having certain constellations. Comparison is shown for minimum determinant, shaping property CER performance and examining how families behave.

Zhonding *et al.* [15] in this paper, a new class of QOSTBC is proposed with two transmit antennas and three time slots, where Alamouti is not applicable due to odd number of time-slots. The proposed Quasi-Orthogonal Code is able to achieve full code rate and full diversity with low decoding complexity. The proposed design shows excellent properties in other prospects also, like compatibility with single transmission mode, low transmitter encoding complexities and lower power fluctuation. The decoder used is maximum likelihood whose complexity is one order lower. Simulation results are taken for simple and flat Rayleigh fading channels. Its performance is compared with hybrid transmission, and it performs it with much higher value at high SNRs. The performance gap is mainly due to difference in spectral efficiency.

Samer J Alabed *et al.* [16] in this paper, a sub-optimal decoder of lower complexity is proposed for coherent as well as non-coherent quasi-orthogonal STBCs with three or four Transmit Antennas. This decoder has approximately linear complexity with same performance as optimum ML decoder. Simulation and numerical results shows that the proposed scheme provides sustainable improved trade-off between performance and complexity in comparison to known decoding schemes. Also, the numbers of operations are substantially reduced, which does not depend on the constellation size. Advantages of the proposed decoder in comparison to others, is that it approaches coherent and non-coherent decoding.

Hardip K Shah *et al.* [17] Space time block codes are widely used in MIMO for improvement of link performance whereas for more than two transmit antennas Quasi-Orthogonal (QO-STBC) and constellation rotation (CR-QOSTBC) are appropriate. Various schemes are evaluated for Rayleigh fading channels, whereas correlation among them is not considered. In this paper, performance is evaluated over correlated channels, which is created by inadequate distance among antennas. The effect of correlation over QO-STBC and CR-QOSTBC is shown by varying the antenna spacing parameter in Kronecker Channel Mode, to verify their applicability in realistic environment. The performance of QO-STBC and CR-QO-STBC systems over correlated wideband channels is analyzed. Also the scheme is combined with OFDM to mitigate the effect of multipath fading. Comparison is also done for correlated and uncorrelated channel on the basis of simulations.

Jafarkhani et al. [18] in this paper complex orthogonal design is shown which does not provide full and full transmission rate for more than two antennas. Concentration is upon generalizing these designs for full diversity and high rate. Here full code rate codes are designed with partial diversity. Also decoder does not work with single symbol, instead works with pair of transmitted symbols. An appropriate modulation scheme is used providing code rate one for 4-PSK and $\frac{1}{2}$ for 16 QAM. Simulation results that full code rate are important for high BERs and low SNRs, whereas full diversity is required for high SNRs and low BERs.

Zafar Q Taha et al. [19] a class of Quasi-Orthogonal Space time block code can achieve full code rate, but the conventional decoder of such types experience interference terms due to neighbouring signals due to their detection. These interference terms cause increase in the complexity of the decoder with a decrease in its performance. In this paper a modification is proposed to conventional decoding scheme, which improves performance, improves robustness against channel errors and improve performance. Also computational load at the receiver is reduced. Properties of Grammian matrix are used in a way, such that there is linear decoding on the receiver side without sacrificing the bandwidth. There is a gain in computation time, which translates to more battery time, lower complexity, less latency.

Amar Ismail et al. [20] in this paper, a new family of codes has been proposed combining the multi-group decodable and the fast decodable (FD) codes and resulting in fast group decodable (FGD) codes. A new construction scheme for 2^a transmits antennas for rate-1 FGD codes are also proposed. The coding gain of these codes is optimized with constellation stretching. Next these rate-1 FGDs are multiplexed to obtain rate-2-FGDs with the help of unitary matrices. The comparison is done with the existing codes to show that our rate $\frac{3}{2}$ code has lower average decoding complexity in low SNR region. If decoding is done with sphere decoder, our proposed code outperforms the existing ones.

Weifing et al. [21] in this paper, they are able to achieve full diversity with orthogonal STBC at high SNRs. Here, half of the symbols which are to be transmitted are chosen from one signal constellation set A and other half are chosen from rotated constellation

Set $Ae^{j\theta}$. The resulting code achieves full diversity and fast maximum likelihood decoding. Also the angles of rotation are chosen optimally to achieve desired results, without extra cost. A general modulation scheme is also proposed with full diversity. The optimality for the rotation angle is done specifically for QPSK constellation. Here main focus is to maximize the diversity product or the minimum of metric, which is an optimal approach. Comparison is done with Jafarkhani and TBH scheme, which are outperformed at both low and high SNRs.

Naotoshi Yoda *et al.* [22] In this paper, a new scheme is proposed which makes possible the decoding the symbol one by one, and the scheme gains full transmission diversity by upsetting rotated constellation and balance of transmit power. The channel chosen is quasi-static fading channel with modulation scheme used as QPSK and 16 QAM. The proposed schemes are compared with Jafarkhani code and a rotated quasi orthogonal STBC, and we found that its decoding complexity is lower in comparison to them.

When compared to R-QOSTBC which achieves full transmission diversity with pairwise decoding, constellation rotation and power imbalance respectively demonstrates 0.2dB and 1dB penalty for QPSK, 0.1dB and 0.7 dB penalty respectively for 16 QAM with 10^{-4} BER.

Gokee Harioglu *et al.* [23] here in this paper they propose a rate-1 space time transmit diversity scheme. A second order diversity is achieved by transmitting real as well as imaginary parts of symbols from two different antennas, and symbols are not detected jointly in these applications. Eighth order transmit diversity can also be obtained by using both Alamouti coding and Hadamard spreading diversity with the proposed method. This proposed scheme is suitable for OFDM systems also. On comparison we found Alamouti performs better than the proposed one for the 2nd order of transmits diversity, but order is limited to this much only. However when all the methods use same detection complexity, the proposed scheme performs better than Hadamard spreading diversity as well as QOSTBC.

Tiang Pen ren *et al.* [24] Space Time Block Code (STBC) in multiple input-multiple output systems for high rate provides both diversity gain and spatial multiplexing gain, with higher decoding with ML decoder. In this paper, they present a systematic method

to construct group decodable high code rate STBC with arbitrary number of transmit antennas. The proposed STBC achieves high rate which increases almost linearly with the number of transmit antennas and slope of this graph increases with the code-length. Comparisons are done with high rates such as algebraic STBCs and low rates, to show reduction in decoding complexity and better performance. Study shows that by optimizing the constellation rotation, the proposed STBC achieves same diversity as that of algebraic STBC.

Don Torrieri *et al.* [25] in this paper, by using genetic algorithm, STBCs of rate one which accommodates simple processing at the receiver are achieved but at the cost of reduced diversity. On simulation they show that when these evolved codes are combined with outer codes, better performance over fading channel is achieved. But when fading is made more severe than Rayleigh fading, for specified efficiency, evolved codes outperforms existing orthogonal Space time block code. Also the codes which are evolved have transmission matrices which are not orthogonal and its decoding is suboptimal and simple at practical values of signal to noise ratio in a fading environment.

Haiquan Wang *et al.* [26] in this paper, study of general linear transformation of symbols to be transmitted for QOSTBC is done to achieve full diversity as well as real pairwise ML decoding. Optimal transmission matrices are presented such that we want optimal diversity product for general rectangle QAM signal as well as general square QAM signal constellations. Also they present the optimal transformations for CIOO that is co-ordinate interleaved orthogonal designs proposed by Rajan Khan for rectangular constellations. For non-square RQAM constellations, optimal diversity products are better than the ones presented by Yuen-Guan with their optimal rotations. In this article, we have assumed that MIMO channels are uncorrelated with no feedback.

Jeongchang Kim *et.al.* [27] A mixed bag of full-rate, full-diversity qualities quasi orthogonal space-time block codes (QO-STBCs) have been outlined. In spite of the fact that the quasi orthogonality lessens the decoding complexity many-sided quality from request of MQN to $MQN/2$ for Q -ary QAM balance by utilizing N transmit and M reception antennas, there is still the requirement for development. A few optimum calculations were proposed however they experience the ill effects of execution over

maximum likelihood (ML) decoding. In this paper they have proposed a calculation for QO-STBCs in view of interference cancellation (IC) with decoding complexity of the request of $MQN/2-1$ with execution practically indistinguishable to that of maximum Likelihood. From this base calculation, it is created an alteration equivalent to $1/\sqrt{Q}$ times that of the base calculation with little measure of loss in performance.

T. Jung *et al.* [28] Quasi-orthogonal space-time block codes (QO-STBCs) are greatly preferred because of the presence of independent and joint maximum-likelihood (ML) decoding of four real or two complex symbols is obtainable on the receiver side. On the other hand if we want to reduce decoding complexity, it is more significant to curtail the length of the code-word. This implies that under some circumstance of full diversity, the time duration of a code-word matrix must be indistinguishable to the amount of transmit antennas, which is referred to as delay-optimal. Nevertheless, the already existing QO-STBCs are delay-optimal under certain conditions that is when we have $N = 2n$ number of transmit antennas. For previous antennas, the codes were constructed by removing certain number of columns from these so called delay-optimal codes. Therefore, taking into account the decoding complexity of maximum likelihood and data latency at the receiver, it is momentous to design a code-word matrix of square in shape and size.

Cunliang Jiang *et al.* [29] In this paper, an approach called array processing is proposed for the decoding of quasi-orthogonal space-time block codes (QO-STBC). The decoding complexity has been reduced and the decoding method proves to be extremely effective, and thus the system latency is reduced by a great amount which is created by the decoding process. This approach works by means of a single symbol and can figure out the delay which is generated by the decoder in addition to the decoding computational load. In this scheme, we are transmitting the signals using different antennas and detached by void space, which are decoded linearly. Numerical results show that the decoding complexity of quasi orthogonal space-time block codes can be significantly reduced with the proposed decoding method.

U. park *et al.* [30] proposed a new form of quasi-orthogonal space time block coding (QO-STBC) format which makes use of simple and linear decoding process. A conventional Quasi Orthogonal-STBC scheme is capable of achieving full rate, but at the

charge of decoding complexity as well as diversity gain. These shortcomings of the conventional QO-STBC proposal are mainly result of various interference terms present in the detection matrix. This new proposed QO-STBC scheme will eliminate all the interference terms present in the detection matrix. The proposed method also achieves improvement in the diversity gain with reverence to the original conventional QOSTBC scheme, over and above a great reduction is achieved in their decoding complexity.

Z. Chen *et al.* [31] proposed a novel technique of extending any type of QO-STBC which is constructed for 4 antennas on the transmitter side to some closed-loop scheme which is based upon circulant matrix. It is also shown that with the support of multiplying various entries present in QO-STBC code words by diversified and suitable phase factors which depends upon their channel information, the projected scheme also improves its transmit diversity by giving feedback of one bit. The observation is that the performance of the newly proposed method is extended from Jafarkhani's quasi orthogonal code and also work is done for the rotation scheme which is taken from an optimum constellation. And on simulation the results suggests that there is significant improvement in SNR which is an additional advantage in our newly proposed scheme and also its reproduction is simpler.

Y. Li *et al.* [29] examined the performance of some quasi orthogonal codes by considering the multiple input and multiple output channels as dual-polarized. It has been shown that when we use QO-STBC with antennas which are dual-polarized, there is a requirement of connecting the power amplifiers (PAs) to the various differently polarized antennas carefully, with the aim of avoiding more than 1dB performance loss at higher cross-polar discrimination (XPD). Moreover, it is shown that, if the Quasi Orthogonal-STBC with rotated constellations are used, which achieves full diversity in case of uni-polarized multiple input multiple output channels, the performance loss because of the bad power amplifiers to the antenna connections is alleviated. On the other hand, when the cross polar discrimination is very high, the QO-STBC devoid of full diversity, when connected properly, knows how to achieve enhanced performance in comparison to the QO-STBC having full diversity while connected badly. This shows how much important is the proper connection required at cross polar discrimination. Consequently to assure vigorous performance of QO-STBC at any XPD, in addition to designing of codes to

enclose full diversity and good diversity product, linking various power amplifiers and antennas in a proper situation is also very important.

D. Wang *et al.* [33] proposed optimal angles for rotation of quasi-orthogonal space-time block codes (STBC) by using multiple phase shift keying (MPSK) symbols, where the term optimality is defined such that the optimal diversity product is achieved. The most favourable diversity products of various quasi-orthogonal space time block codes are compared with that of rotated MPSK constellations which are already given. And on later stage the sensitivity is studied. The diversity product and the corresponding angle of rotation and also angle offset are studied.

3.1 Space Time Coding

We know that there is always a requirement of having smaller in addition to cost effective mobile receivers, thus assembling receiver diversity on the receiver front on the mobile terminal is not a practical formulation. This has motivated numerous amounts of researchers to think upon transmit diversity by employing enormous antennas on the transmitter side that is at the base station. By deploying multiple antennas on transmitter side grants transmit diversity, bit error rate can be noticeably abridged, the BER (bit error rate) curve decays at a much faster rate with SNR (signal to noise ratio). This is because when numerous of transmit antennas are used higher amount of spatial diversity is achieved. However, when diversity combining is employed on the receiver side, contrasting receive diversity can be achieved, some type of signal processing is needed when we are using transmit diversity, commonly called as space time coding, on the signals to be transmitted in charge for achieving augmentation of the signal on the receiver front. Space-Time Coding (STC) is a technique which combines coding, signal processing and modulation to achieve the transmit diversity.

The first STC anticipated is Space-Time Trellis Code (STTC), which is having good decoding performance but its decoding complexity increases exponentially with transmission code rate. In addressing this issue of decoding complexity of STTC, Space-Time Block Code (STBC) was subsequently proposed.

Although O-STBC grants full transmit diversity but at much lower computational cost, however [5] it has also been shown that it suffers from loss in its capacity when (i) there are numerous receive antennas, (ii) the transmission rate is much less than one. In this article, a generalised scheme to eliminate these interference terms is proposed by employing it on Tirkkonen and ABBA code which also decreases its decoding complexity. In order to achieve full transmit diversity a novel criteria of rotated constellations for different transmitted symbols is explored. This scheme outperforms transmit antenna shuffling (TAS) and also removes its disadvantage of dedicated feedback channel for CSI. The final code thus achieved is very powerful as it provides code rate of one, full diversity and pairwise decoding. Finally, simulations show that bit

error rate performance is significantly improved by gain of 3dB in comparison to TAS scheme.

3.1.1 Space Time Code Design Criteria

A mapping from input bits to output bits which are transmitted symbols is called a code. All the symbols are transmitted from different antennas simultaneously. To study the performance of such codes, we then derive some bounds on them with the analysis called as Asymptotic Analysis. And these different asymptotic assumptions finally results in a code criteria. The block diagram of space time coding is shown diagrammatically in fig 3.1. We consider that the transmission takes place on binary symmetric channels and

information is transmitted as blocks. And the error which prevails in that system could be calculated on the basis of Hamming distance denoted by d_{\min} . A code having minimum hamming distance d_{\min} is able to correct the error patterns with weight equal to or less than $\lfloor (d_{\min} - 1)/2 \rfloor$. The design criterion basically maximizes the minimum possible hamming distance present among all the code-word pairs, because “good code” has high minimum hamming distance. Similarly for an AWGN channel, maximizing the minimum Euclidian distance among code-word pairs is a good design criterion.

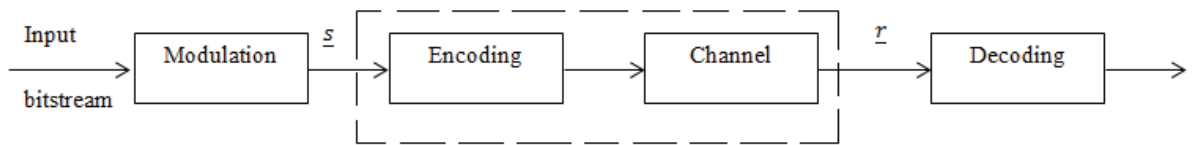


Fig 3.1 Block Diagram of space-time coding system

Let us consider we transmit a code-word C^1 .

$$C^1 = \begin{bmatrix} C_{1,1}^1 & C_{1,2}^1 & \cdots & C_{1,N}^1 \\ C_{2,1}^1 & C_{2,2}^1 & \cdots & C_{2,N}^1 \\ \vdots & \vdots & \vdots & \vdots \\ C_{T,1}^1 & C_{T,2}^1 & \cdots & C_{T,N}^1 \end{bmatrix} \quad (3.1)$$

Where, N is number of Transmitters

T is Time Slots

Let the decoder mistakenly proves that another code-word say C^2 was transmitted, where C^2 is

$$C^2 = \begin{bmatrix} C_{1,1}^2 & C_{1,2}^2 & \cdots & C_{1,N}^2 \\ C_{2,1}^2 & C_{2,2}^2 & \cdots & C_{2,N}^2 \\ \vdots & \vdots & \vdots & \vdots \\ C_{T,1}^2 & C_{T,2}^2 & \cdots & C_{T,N}^2 \end{bmatrix} \quad (3.2)$$

If we assume that all the code-words in a code-book contains only C^1 and C^2 , we say error has occurred and this pairwise error probability by transmitting C^1 and detecting C^2 is denoted as $P(C^1 \rightarrow C^2)$. In general, if there are total of I code-words transmitted,

applying union bound on probability of error we have

$$P(\text{error}|C^1 \text{ is sent}) \leq \sum_{i=2}^I P(C^1 \rightarrow C^2) \quad (3.3)$$

This overall bound is upper bound. To calculate this pairwise error probability, we assume the channel matrix H as fixed and calculate average error from expected value over H's distribution.

$$r = CH + N \quad (3.4)$$

From each antenna, the average transmission power of each symbol is $E_s = 1/N$ and variance of Noise sample is given by

$$E = [n_{t,m}^t] = N_o = 1/\gamma \quad (3.5)$$

The distribution of signals received from some known channel matrix H and code-word C is $f(r|C, H)$

$$\begin{aligned} f(r|C, H) &= \frac{1}{(\pi N_o)^{\frac{T_x M}{2}}} \exp \left\{ \frac{-\text{Tr}[(r - CH)^H (r - CH)]}{N_o} \right\} \\ &= \left(\frac{\gamma}{\pi} \right)^{\frac{T_x M}{2}} \exp \{ -\gamma \text{Tr}[(r - CH)^H (r - CH)] \} \end{aligned} \quad (3.6)$$

Applying Frobenius norms here which are defined as

$$\|A_F\| = \sqrt{\text{Tr}(A^H \cdot A)} = \sqrt{\text{Tr}(A \cdot A^H)} \quad (3.7)$$

We rewrite the equation as

$$f(r|C, H) = \left(\frac{\gamma}{\pi}\right)^{\frac{T \times M}{2}} \exp \left\{ -\gamma \sum_{t=1}^T \sum_{m=1}^M |(r - CH)_{t,m}|^2 \right\} \quad (3.8)$$

Applying maximum likelihood decoding and deciding in favour of such code-word which maximizes $f(r|C, H)$, writing the received vector as $r^1 = C^1 H + N^1$, calculating pairwise error probability as

$$P(C^1 \rightarrow C^2 | H) = P([\|r^1 - C^1 H\|_F^2 - (\|r^1 - C^2 H\|_F^2)] > 0 | H) \quad (3.9)$$

With some calculations and applying Q function we get pairwise error probability as

$$\begin{aligned} P(C^1 \rightarrow C^2 | H) &= Q \left(\frac{(\|(C^2 - C^1)H\|_F^2)}{\sqrt{2/\gamma} (\|(C^2 - C^1)H\|_F^2)} \right) \\ &= Q \left(\sqrt{\frac{\gamma}{2}} \|(C^2 - C^1) \cdot H\|_F \right) \end{aligned} \quad (3.10)$$

where, we define Q function as

$$Q(x) = \frac{1}{2\pi} \int_x^\infty e^{-\frac{y^2}{2}} dy \quad (3.11)$$

We now define an error or difference matrix as $D(C^1, C^2) = C^2 - C^1$

We then write this pairwise error probability in terms of its Eigen values of matrix as

$$\begin{aligned} A(C^1, C^2) &= D(C^1, C^2)^H \cdot D(C^1, C^2) \\ &= (C^2 - C^1)^H (C^2 - C^1) \end{aligned} \quad (3.12)$$

Because $D(C^1, C^2)$ is square root of $A(C^1, C^2)$, then Eigen values of $A(C^1, C^2)$ denoted as λ_n , $n=1, 2, \dots, N$ are real numbers which are non-negative. Applying singular value decomposition on it,

$$A(C^1, C^2) = V^H \cdot \Lambda \cdot V \quad (3.13)$$

Where Λ is = $\text{diag}(\lambda_1, \lambda_2, \dots, \lambda_N)$. Hence

$$\|(C^2 - C^1) \cdot H\|_F^2 = \text{Tr}[H^H \cdot A(C^1, C^2) \cdot H]$$

$$= \text{Tr}[\mathbf{H}^H \cdot \mathbf{V}^H \cdot \mathbf{\Lambda} \cdot \mathbf{V} \cdot \mathbf{H}] \quad (3.14)$$

Since elements of \mathbf{H} are Gaussian random variables, elements of $\mathbf{V} \cdot \mathbf{H}$ also become Gaussian. The (n,m) is element of $\mathbf{V} \cdot \mathbf{H}$ is denoted by $\beta_{n,m}$. Therefore writing this all in matrix form,

$$\begin{aligned} & (\|(\mathbf{C}^2 - \mathbf{C}^1) \cdot \mathbf{H}\|_F^2) \\ &= \text{Tr} \left\{ \begin{bmatrix} \beta_{1,1}^* & \beta_{2,1}^* & \dots & \beta_{N,1}^* \\ \beta_{1,2}^* & \beta_{2,2}^* & \dots & \beta_{N,2}^* \\ \vdots & \vdots & \vdots & \vdots \\ \beta_{1,M}^* & \beta_{2,M}^* & \dots & \beta_{N,M}^* \end{bmatrix} \begin{bmatrix} \lambda_1 & 0 & \dots & 0 \\ 0 & \lambda_2 & \dots & 0 \\ \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & 0 & \lambda_N \end{bmatrix} \begin{bmatrix} \beta_{1,1} & \beta_{1,2} & \dots & \beta_{1,M} \\ \beta_{2,1} & \beta_{2,2} & \dots & \beta_{2,M} \\ \vdots & \vdots & \vdots & \vdots \\ \beta_{N,1} & \beta_{N,2} & \dots & \beta_{N,M} \end{bmatrix} \right\} \\ &= \text{Tr} \left[\begin{pmatrix} \sum_{n=1}^N \lambda_n |\beta_{n,1}|^2 & & & \\ & \sum_{n=1}^N \lambda_n |\beta_{n,2}|^2 & & \\ & & \dots & \\ & & & \sum_{n=1}^N \lambda_n |\beta_{n,m}|^2 \end{pmatrix} \right] \end{aligned} \quad (3.15)$$

On calculating trace in last equation (3.15),

$$\|(\mathbf{C}^2 - \mathbf{C}^1) \cdot \mathbf{H}\|_F^2 = \sum_{m=1}^M \sum_{n=1}^N \lambda_n |\beta_{n,m}|^2 \quad (3.16)$$

Putting this back in equation (3.10), we get

$$P(\mathbf{C}^1 \rightarrow \mathbf{C}^2 | \mathbf{H}) = Q \left(\sqrt{\left(\frac{\gamma}{2}\right) \sum_{m=1}^M \sum_{n=1}^N \lambda_n |\beta_{n,m}|^2} \right) \quad (3.17)$$

Applying upper bound on conditional pairwise error probability

$$P(\mathbf{C}^1 \rightarrow \mathbf{C}^2 | \mathbf{H}) = \frac{1}{2} \exp \left(-(\gamma/4) \sum_{m=1}^M \sum_{n=1}^N \lambda_n |\beta_{n,m}|^2 \right) \quad (3.18)$$

Since $\beta_{n,m}$ is Gaussian in nature and its magnitude $|\beta_{n,m}|$ Rayleigh with probability density function given by

$$f(|\beta_{n,m}|) = 2|\beta_{n,m}| \exp(-|\beta_{n,m}|^2) \quad (3.19)$$

If $A(C^1, C^2)$ is of full rank, none of its Eigen values will be zero. And on the other hand, if rank is $r < N$, we have $\lambda_1 \geq \lambda_2 \geq \dots \geq \lambda_r > 0$ and $\lambda_{r+1} = \dots = \lambda_N = 0$. At high values of SNRs, we neglect denominator of inequality, and writing the upper bound based on non-zero Eigen Values:

$$P(C^1 \rightarrow C^2) = \frac{4^{rM}}{(\prod_{n=1}^r \lambda_n)^M \gamma^M} \quad (3.20)$$

Rank Criterion: It states that the minimum determinant value of the $A(C^i, C^j) = D(C^i, C^j)^H \cdot D(C^i, C^j)$ needs to be larger to obtain high value of coding gain. And coding gain is related to the product of all the Eigen values $\prod_{n=1}^N \lambda_n$.

Determinant Criterion: It states that coding gain $\prod_{n=1}^N \lambda_n$, associated with pairwise error probability is equivalent to determinant of the matrix. Thus high coding gain is achieved by maximizing the minimum of the determinant.

3.2 Alamouti Space Time Code

The Alamouti scheme is the very first space time block code for two transmit antennas which provides full transmit diversity. There are delay diversity schemes also available which can achieve full diversity, but it introduces interference between symbols, moreover complex detectors are required on the receiver end.

3.2.1 Alamouti Space Time Encoding

Let us assume we are using M-ary modulation scheme, where m-information bits are first modulated $m = \log_2 M$. The encoder will take blocks of two modulated symbols x_1 and x_2 in its encoding operation and then mapping is done to transmit antennas according to code matrix as:

$$X = \begin{bmatrix} x_1 & -x_2^* \\ x_2 & x_1^* \end{bmatrix} \quad (3.21)$$

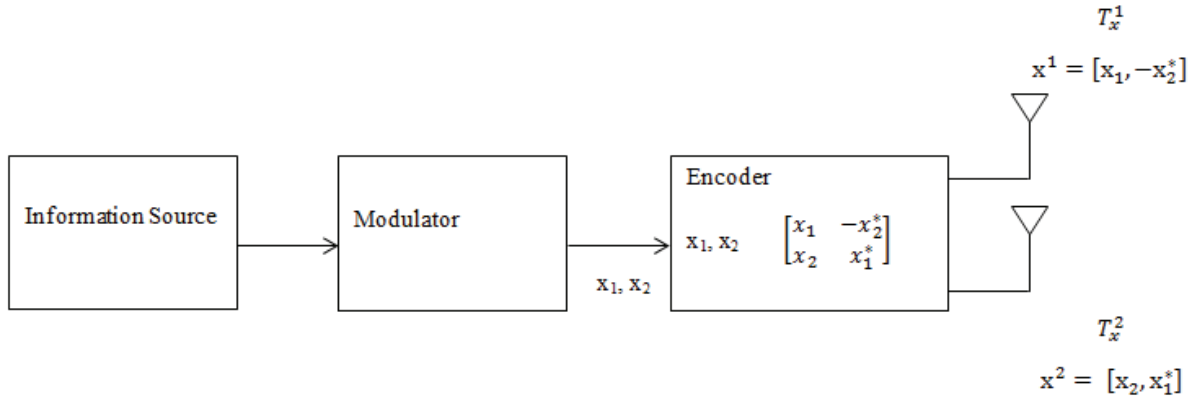


Fig3.2 Representation of Alamouti Scheme

During the first transmission, two symbols x_1 and x_2 are transmitted from two different antennas showing that encoding is done in both time and space domain.

Let us denote the sequence to be transmitted from two antennas as x^1 and x^2 ,

$$x^1 = [x_1, -x_2^*] \quad (3.22)$$

$$x^2 = [x_2, x_1^*] \quad (3.23)$$

The main property of this Alamouti scheme is orthogonality i.e. the sequence which is transmitted from different antennas is orthogonal in nature, because inner product of sequences x^1 and x^2 is zero. i.e.

$$x^1 \cdot x^2 = x_1 x_2^* - x_2^* x_1 = 0 \quad (3.24)$$

The code matrix has the property as follows:

$$\begin{aligned} X \cdot X^H &= \begin{bmatrix} |x_1|^2 + |x_2|^2 & 0 \\ 0 & |x_1|^2 + |x_2|^2 \end{bmatrix} \\ &= (|x_1|^2 + |x_2|^2) I_2 \end{aligned} \quad (3.25)$$

Where, I_2 is a 2x2 identity matrix. The channel is considered as flat-fading, and its channel coefficients at any time t are denoted as $h_1(t)$ and $h_2(t)$ from first and second transmit antennas respectively. We also assume that these fading coefficients are constant during consecutive symbol transmission.

$$h_1(t) = h_1(t + T) = |h_1| e^{j\theta_1} \quad (3.26)$$

$$h_2(t) = h_2(t + T) = |h_2| e^{j\theta_2} \quad (3.27)$$

Where, $|h_i|$ represents the amplitude gain and θ_i represents the phase shift for the path between transmit antennas i and the receiver antennas, and T represents symbol duration where $i=0,1$.

The received signal at the receiver antennas is given by r_1 and r_2 at time t and $(t+T)$ respectively,

$$r_1 = h_1x_1 + h_2x_2 + n_1 \quad (3.28)$$

$$r_2 = -h_1x_2^* + h_2x_1^* + n_2 \quad (3.29)$$

Where, n_1 and n_2 represent Gaussian noise samples at time t and $(t+T)$ respectively.

3.2.2 Combining and Maximizing Likelihood Decoding

Considering that channel fading coefficients are perfectly recovered on the receiver front, decoder uses them as Channel State Information (CSI). Assuming all the signals to be equiprobable, ML decoder will choose those pairs of signals (\hat{x}_1, \hat{x}_2) which will minimize the distance matrix.

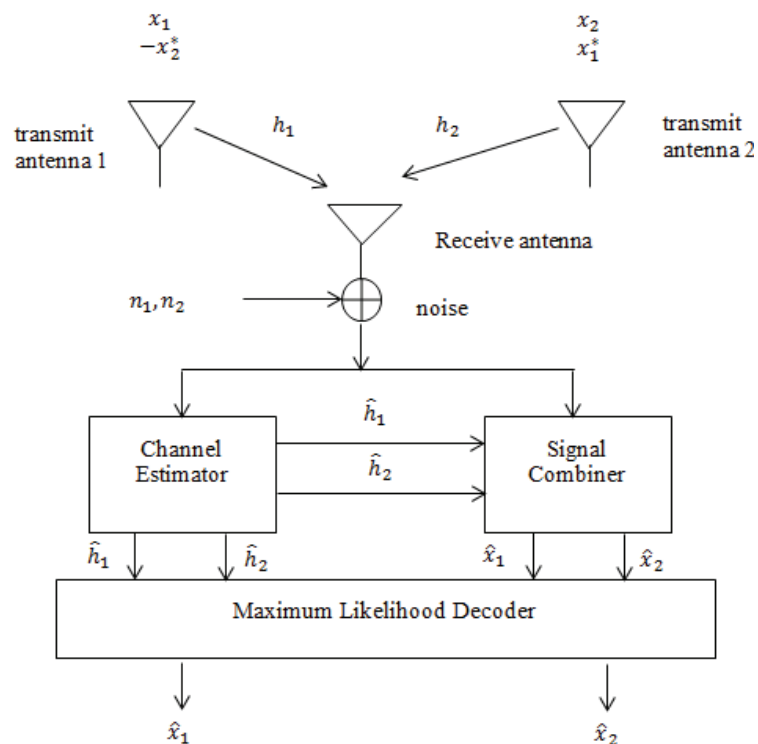


Fig 3.3 Receiver for Alamouti Scheme

$$\begin{aligned}
& d^2(r_1, h_1 \hat{x}_1 + h_2 \hat{x}_2) + d^2(r_2, -h_1 \hat{x}_2^* + h_2 \hat{x}_1^*) \\
& = |r_1 - h_1 \hat{x}_1 - h_2 \hat{x}_2|^2 + |r_2 + h_1 \hat{x}_1^* - h_2 \hat{x}_2^*|^2
\end{aligned} \tag{3.30}$$

For all possible values of \hat{x}_1 and \hat{x}_2 , on substituting the values for ML decoding and considering \tilde{x}_1 and \tilde{x}_2 as two decision statistics, with CSI on receiver.

$$\tilde{x}_1 = h_1^* r_1 + h_2 r_2^* \tag{3.31}$$

$$\tilde{x}_2 = h_2^* r_1 - h_1 r_2^* \tag{3.32}$$

On substituting the values of r_1 and r_2 , decision statistics are written as:

$$\tilde{x}_1 = (|h_1|^2 + |h_2|^2)x_1 + h_1^* n_1 + h_2 n_2^* \tag{3.33}$$

$$\tilde{x}_2 = (|h_1|^2 + |h_2|^2)x_2 - h_1 n_2^* + h_2^* n_1 \tag{3.34}$$

If the channel coefficient h_1 and h_2 are known, the ML decoding leads to two decoding syntax for x_1 and x_2 as:

$$\hat{x}_1 = \arg \min_{\hat{x}_1 \in \mathcal{S}} (|h_1|^2 + |h_2|^2 - 1)|\hat{x}_1|^2 + d^2(\tilde{x}_1, \hat{x}_1) \tag{3.35}$$

$$\hat{x}_2 = \arg \min_{\hat{x}_2 \in \mathcal{S}} (|h_1|^2 + |h_2|^2 - 1)|\hat{x}_2|^2 + d^2(\tilde{x}_2, \hat{x}_2) \tag{3.36}$$

For all the signal points, in M-PSK signal constellation neglecting the constraints $(|h_1|^2 + |h_2|^2 - 1)|\hat{x}_i|^2$ where $i=1, 2$ simplifies our decision rule as

$$\hat{x}_1 = \arg \min_{\hat{x}_1 \in \mathcal{S}} d^2(\tilde{x}_1, \hat{x}_1) \tag{3.38}$$

$$\hat{x}_2 = \arg \min_{\hat{x}_2 \in \mathcal{S}} d^2(\tilde{x}_2, \hat{x}_2) \tag{3.39}$$

3.2.3 Performance of Alamouti Scheme:

As we are transmitting the sequences from two different antennas, we show that orthogonality between it leads Alamouti to achieve full transmit diversity. To show this,

we consider two different code sequences as X and \hat{X} which are generated with inputs (x_1, x_2) and (\hat{x}_1, \hat{x}_2) respectively. We achieve the code-word distance matrix as

$$B(X, \hat{X}) = \begin{bmatrix} x_1 - \hat{x}_1 & -\hat{x}_2^* + \hat{x}_2^* \\ x_2 - \hat{x}_2 & \hat{x}_1^* - \hat{x}_1^* \end{bmatrix} \quad (3.39)$$

Rows of above code-word matrix are orthogonal in nature; same should be with code-word difference matrix as well

$$A(X, \hat{X}) = B(X, \hat{X}).B^H(X, \hat{X}) = \begin{bmatrix} |x_1 - \hat{x}_1|^2 + |x_2 - \hat{x}_2|^2 & 0 \\ 0 & |x_1 - \hat{x}_1|^2 + |x_2 - \hat{x}_2|^2 \end{bmatrix} \quad (3.40)$$

The determinant of this matrix is given by

$$\det(A(X, \hat{X})) = (|x_1 - \hat{x}_1|^2 + |x_2 - \hat{x}_2|^2)^2 \quad (3.41)$$

The rank of above matrix is two, because $(x_1, x_2) \neq (\hat{x}_1, \hat{x}_2)$. And hence according to determinant criteria, we get full transmit diversity that is two. The channel considered here is slow Rayleigh fading channel and the fading is mutual independent from transmitter to receiver and also perfect CSI is also present on receiver side.

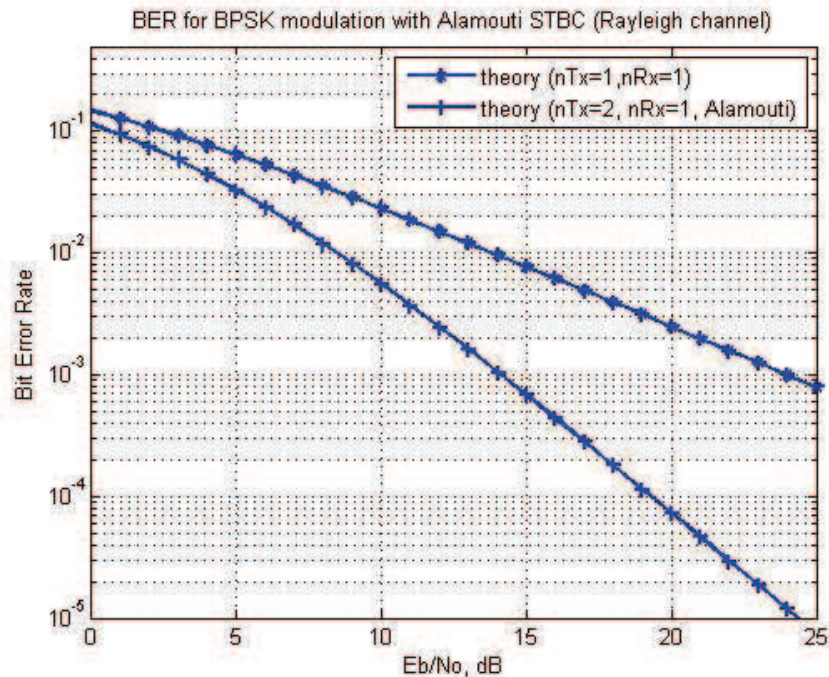


Fig3.4 BER for BPSK modulation with Alamouti STBC (Rayleigh channel)

We assume that the total power transmitted from two antennas in Alamouti scheme is exactly same as that transmitted to single antenna in MRC diversity scheme. Simulation shows that Alamouti and MRC have same slopes and thus same diversity, but there is 3dB worse performance for Alamouti scheme.

The Alamouti code provides two important properties.

- 1) Simple decoding: Each symbol is decoded separately using only linear processing.
- 2) Maximum diversity: The code satisfies the rank criterion and therefore provides the maximum possible diversity.

This penalty in performance is due to the fact that powers given to one transmit antenna is exactly half as that given in MRC.

3.3 Orthogonal Space Time Block Code

The discovery of Alamouti, which is a simple and effective transmit diversity, promoted the idea that similar schemes can be extended and generalized to more number of antennas which could provide higher transmit diversity. In OSTBC, vectors taken from any column, for transmission or coding matrix are orthogonal in nature. Orthogonality helps us to achieve linear decoding at the receiver. If the number of transmit antennas are increased more than two, our goal is to achieve higher rate.

Let us consider transmission matrix G , with code rate $\frac{1}{2}$. With regard to this, let four symbols (S_1, S_2, S_3, S_4) are taken from a complex constellation and transmitted from four antennas during eight periods as:

$$G = \begin{bmatrix} S_1 & S_2 & S_3 & S_4 \\ -S_2 & S_1 & -S_4 & S_3 \\ -S_3 & S_4 & S_1 & -S_2 \\ -S_4 & -S_3 & S_2 & S_1 \\ S_1^* & S_2^* & S_3^* & S_4^* \\ -S_2^* & S_1^* & -S_4^* & S_3^* \\ -S_3^* & S_4^* & S_1^* & -S_2^* \\ -S_4^* & -S_3^* & S_2^* & S_1^* \end{bmatrix} \quad (3.34)$$

We assume a flat fading channel over eight time slots and one receiver antenna, received signals are:

$$R = [r_1 r_2 r_3 r_4 r_5 r_6 r_7 r_8]' \quad (3.35)$$

Where r_i is the received signal at i^{th} time slot.

The effect of noise is different in all channels, that is noise effects different channels by different amount which depends upon its signal to noise ratio

$$N = [n_1 n_2 n_3 n_4 n_5 n_6 n_7 n_8]' \quad (3.36)$$

Where n_i represents complex Gaussian noise at i^{th} time.

The channel coefficients for multiple channels are given by

$$h = [h_1 h_2 h_3 h_4]' \quad (3.37)$$

The signal when received on the receiver front is given by

$$R = G \times h + N \quad (3.38)$$

We rewrite this equation as

$$W = \begin{bmatrix} r_1 \\ r_2 \\ r_3 \\ r_4 \\ r_5^* \\ r_6^* \\ r_7^* \\ r_8^* \end{bmatrix} = H \times \begin{bmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \\ n_3 \\ n_4 \\ n_5^* \\ n_6^* \\ n_7^* \\ n_8^* \end{bmatrix} \quad (3.39)$$

Where H is derived by applying complex conjugate operation on fifth, sixth, seventh, eighth time-slot, which are also orthogonal.

$$G = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2 & -h_1 & h_4 & -h_3 \\ h_3 & -h_4 & -h_1 & h_2 \\ h_4 & h_3 & -h_2 & -h_1 \\ h_1^* & h_2^* & h_3^* & h_4^* \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & -h_4^* & -h_1^* & h_2^* \\ h_4^* & h_3^* & -h_2^* & -h_1^* \end{bmatrix} \quad (3.40)$$

For orthogonal STBC's, received signals are decoded with detection matrix D as

$$D = H^H H \quad (3.41)$$

And detection matrix D here is

$$D = \begin{bmatrix} 2 \times \sum_{j=1}^4 \|h_j\|^2 & & & \\ & 2 \times \sum_{j=1}^4 \|h_j\|^2 & & \\ & & 2 \times \sum_{j=1}^4 \|h_j\|^2 & \\ & & & 2 \times \sum_{j=1}^4 \|h_j\|^2 \end{bmatrix} \quad (3.42)$$

We observe that detection matrix is a diagonal matrix, which leads to simple linear decoding. Upon reception, the signals are estimated as

$$S = [s_1 s_2 s_3 s_4]' \quad (3.43)$$

$$\tilde{S} = H^H \times W \quad (3.44)$$

The only disadvantage with this was we could not achieve full code rate.

3.4 Quasi-Orthogonal Space Time Block Code

Orthogonal space time block code is a vital class of Space time block codes as it helps us with full diversity, and in addition enables linear decoding at the receiver. However, for complex constellations, maximum achievable rate for OSTBCs with two transmit antennas was two [15]. For higher number of transmit antennas, the rate achieved was less than one which imply loss of bandwidth efficiency.

Let us write the generator matrix for Alamouti code

$$G(x_1, x_2) = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix} \quad (3.45)$$

Its main properties were full diversity and simple decoding. But if we want to design codes with full rates, we switch to other types of codes. Let us take following Q-OSTBC.

$$G = \begin{bmatrix} G(x_1, x_2) & G(x_3, x_4) \\ -G^*(x_3, x_4) & G^*(x_1, x_2) \end{bmatrix} = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \\ -x_2^* & x_1^* & -x_4^* & x_3^* \\ -x_3^* & -x_4^* & x_1^* & x_2^* \\ x_4 & -x_3 & -x_2 & x_1 \end{bmatrix} \quad (3.46)$$

Where G^* is complex conjugate of G .

$$G^*(x_1, x_2) = G(x_1^*, x_2^*) = \begin{pmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{pmatrix} \quad (3.47)$$

For variable x_1, x_2, x_3, x_4 we have

$$\langle v_1, v_2 \rangle = \langle v_1, v_3 \rangle = \langle v_2, v_4 \rangle = \langle v_3, v_4 \rangle = 0 \quad (3.48)$$

Where $\langle v_i, v_j \rangle$ represents inner product of vectors v_i and v_j . This is because the subspace created by v_1 and v_4 is orthogonal to subspace created by v_2 and v_3 . This is the justification behind the name ‘‘Quasi-Orthogonal’’..

$$\min_{s_1, s_2, s_3, s_4} \{H^H \cdot C^H \cdot C \cdot H - H^H \cdot C^H \cdot r - r^H \cdot C \cdot H\} \quad (3.49)$$

Where C is obtained by replacing x_k by s_k . Simple manipulation results in following sum:

$$f_{14}(s_1, s_4) + f_{23}(s_2, s_4) \quad (3.50)$$

Where

$$\begin{aligned} f_{14}(s_1, s_4) = \sum_{m=1}^M \left[(|s_1|^2 + |s_4|^2) \left(\sum_{n=1}^4 |\alpha_{n,m}|^2 \right) \right. \\ + 2R\{(-\alpha_{1,m}r_{1,m}^* - \alpha_{2,m}^*r_{2,m} - \alpha_{3,m}^*r_{3,m} - \alpha_{4,m}r_{4,m}^*)s_1 \\ + (-\alpha_{4,m}r_{1,m}^* + \alpha_{3,m}^*r_{2,m} + \alpha_{2,m}^*r_{3,m} - \alpha_{1,m}r_{4,m}^*)s_4\} \\ \left. + 4R\{\alpha_{1,m}\alpha_{4,m}^* - \alpha_{2,m}^*\alpha_3\}R\{s_1s_4^*\} \right] \end{aligned} \quad (3.51)$$

$$\begin{aligned} f_{23}(s_2, s_3) = \sum_{m=1}^M \left[(|s_2|^2 + |s_3|^2) \left(\sum_{n=1}^4 |\alpha_{n,m}|^2 \right) \right. \\ + 2R\{(-\alpha_{2,m}r_{1,m}^* + \alpha_{1,m}^*r_{2,m} - \alpha_{4,m}^*r_{3,m} + \alpha_{3,m}r_{4,m}^*)s_2 \\ + (-\alpha_{3,m}r_{1,m}^* - \alpha_{4,m}^*r_{2,m} + \alpha_{1,m}^*r_{3,m} + \alpha_{2,m}r_{4,m}^*)s_3\} \\ \left. + 4R\{\alpha_{2,m}\alpha_{3,m}^* - \alpha_{1,m}^*h_4\}R\{s_2s_3^*\} \right] \end{aligned} \quad (3.52)$$

Clearly $f_{14}(s_1, s_4)$ is independent of (s_2, s_3) and $f_{23}(s_2, s_3)$ are independent of (s_1, s_4) , and also the pairs (s_1, s_4) and (s_2, s_3) are decoded separately. Therefore on applying ML decoding, we have to minimize $f_{14}(s_1, s_4)$ over all values of s_1 and s_4 , and minimizing $f_{23}(s_2, s_3)$ over all values of s_2 and s_3 .

When comparison is done between quasi orthogonal STBC and uncoded system it is clearly observed that QOSTBC performs much better showing minimum error in comparison to the latter.

Disadvantage: Although QO-STBC can achieve higher code rates than O-STBC, it generally does not provide full transmit diversity directly. As a result, their performance degrades at high SNR

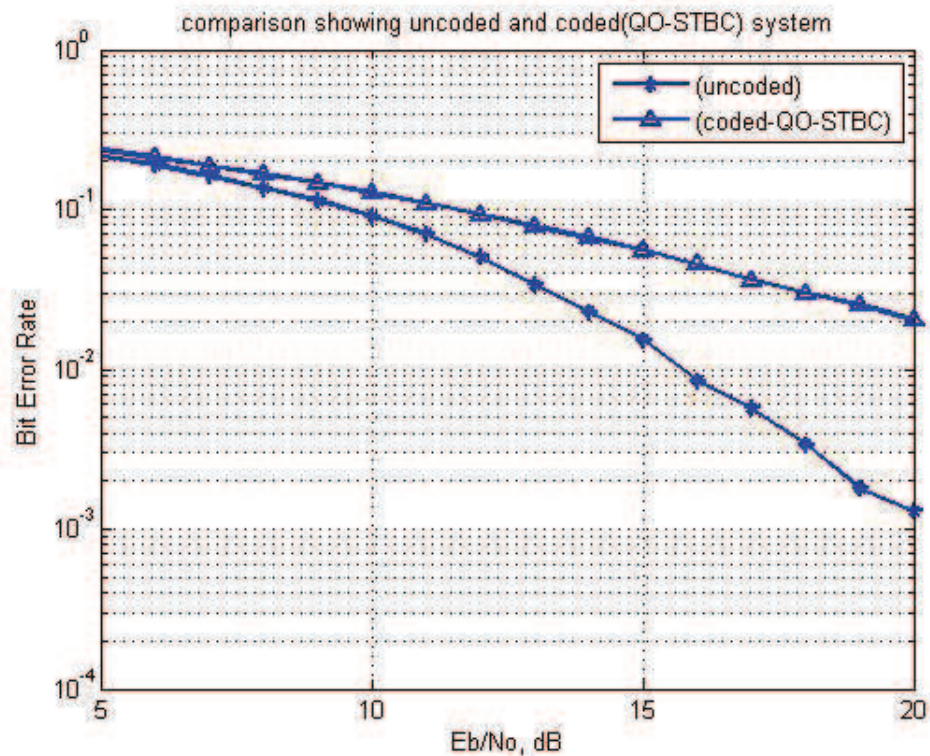


Fig 3.5 Performance of QOSTBC

Design of conventional QO-STBCs with different approach

4.1 Various QO-STBCs.

4.1.1 Jafarkhani code

Among various quasi-orthogonal designs, Jafarkhani proposed QO-STBC by himself which was an extension of already proposed Alamouti scheme. As Alamouti was for two antennas, he extended it to four antennas straight forwardly.

$$S_J = \begin{bmatrix} Al_{12} & Al_{34} \\ -Al_{34}^* & Al_{12}^* \end{bmatrix} = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & -s_3 & -s_2 & s_1 \end{bmatrix} \quad (4.1)$$

Where Al_{12} and Al_{34} represents the Alamouti scheme for two transmit antennas, such as

$$Al_{12} = \begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{bmatrix} \quad Al_{34} = \begin{bmatrix} s_3 & s_4 \\ -s_4^* & s_3^* \end{bmatrix} \quad (4.2)$$

4.1.2 ABBA Code

ABBA codes, class of Quasi orthogonal space time block codes allow us with low decoding complexity and full diversity. Its decoding is done as pairwise with special signal mapping. Again just like Jafarkhani a popular QO-STBC introduced for four Transmit antennas is:

$$S_A = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ s_3 & s_4 & s_1 & s_2 \\ -s_4^* & s_3^* & -s_2^* & s_1^* \end{bmatrix} \quad (4.3)$$

4.1.3 Tirkkonen Code

On a similar basis, proposed Tirkkonen code for the four transmits antennas as:

$$S_T = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ s_2^* & -s_1^* & s_4^* & -s_3^* \\ s_3 & s_4 & s_1 & s_2 \\ s_4^* & -s_3^* & s_2^* & -s_1^* \end{bmatrix} \quad (4.4)$$

4.2 Equivalent virtual channel Matrix

A highly structured and unique way to find an equivalent and virtual channel matrix H is done by EVCN approach [19]. Let us apply this approach on any one of the above code as Jafarkhani and then calculate H matrices for two other codes too. Let us consider a QO-STBC denoted as S and a flat Rayleigh faded MISO channel h for (4×1) antennas. The received vector is then given by:

$$r = Sh + n \quad (4.5)$$

Where, r represents the received signals at different instants of time.

Let the channel is denoted by $h = [h_1, h_2, h_3, h_4]^T$ and $n = [n_1, n_2, n_3, n_4]^T$ as noise vector.

$$\begin{bmatrix} r_1 \\ r_2 \\ r_3 \\ r_4 \end{bmatrix} = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & -s_3 & -s_2 & s_1 \end{bmatrix} \begin{bmatrix} h_1 \\ h_2 \\ h_3 \\ h_4 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \\ n_3 \\ n_4 \end{bmatrix} \quad (4.6)$$

On expanding this we get

$$\begin{aligned} r_1 &= s_1 h_1 + s_2 h_2 + s_3 h_3 + s_4 h_4 \\ r_2 &= -s_2^* h_1 + s_1^* h_2 - s_4^* h_3 + s_3^* h_4 \\ r_3 &= s_3^* h_1 + s_4 h_2 - s_1^* h_3 - s_2^* h_4 \\ r_4 &= s_4 h_1 - s_3 h_2 - s_2 h_3 + s_1 h_4 \end{aligned} \quad (4.7)$$

We modify the received signals for third and second time slot, and rewrite above equations:

$$\begin{aligned} r_1 &= s_1 h_1 + s_2 h_2 + s_3 h_3 + s_4 h_4 \\ r_2^* &= s_2 h_1^* - s_1 h_2^* - s_4 h_3^* - s_3 h_4^* \\ r_3^* &= s_3 h_1^* + s_4 h_2^* - s_1 h_3^* - s_2 h_4^* \\ r_4^* &= s_4 h_1 - s_3 h_2 - s_2 h_3 + s_1 h_4 \end{aligned} \quad (4.8)$$

Thus in matrix form we have

$$\begin{bmatrix} r_1 \\ r_2^* \\ r_3^* \\ r_4 \end{bmatrix} = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & h_4^* & -h_1^* & -h_2^* \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2^* \\ n_3^* \\ n_4 \end{bmatrix} \quad (4.9)$$

Now, our modified received signal vector y is

$$y = \begin{bmatrix} r_1 \\ r_2^* \\ r_3^* \\ r_4 \end{bmatrix} = Hs + \bar{n} \quad (4.10)$$

With S as $[s_1, s_2, s_3, s_4]^T$ and $\bar{n} = [n_1, n_2^*, n_3^*, n_4]^T$ and H represents equivalent channel matrix of size (4×4) which is

$$H_j = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & h_4^* & -h_1^* & -h_2^* \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix} \quad (4.11)$$

So we are able to achieve equivalent channel matrix H for corresponding code Jafarkhani, with four transmit antennas and four virtual receiver antennas. Thus this EVCM approach simplifies our analysis to a greater extent.

4.2.1 Properties of the EVCM

i) The equivalent virtual channel matrix obtained above as H is having block structure very much analogous to the corresponding non orthogonal matrix S .

ii) The Matrix obtained by multiplying H with its hermitian, leads to a matrix referred as Grammian matrix or detection matrix, specified by

$$D = H^H H \quad (4.12)$$

Just as in Jafarkhani we will see the detection matrix obtained is

$$D_A = \begin{bmatrix} a & 0 & b & 0 \\ 0 & a & 0 & b \\ b & 0 & a & 0 \\ 0 & b & 0 & a \end{bmatrix} \quad (4.13)$$

iii) The matrix D is a meagre matrix having gain factor a as real valued located at its main diagonal and a self-interference feature b present at off-diagonal positions.

iv) The self-interference parameter b could be of nature as real or purely imaginary that depends upon which type of quasi orthogonal codes we are using.

v) If any kind of linear transformations are applied on these codes, then the quasi-orthogonal structure of D remains same and does not change, just its value of b changes.

vi) The self-interference factor b is responsible for all the non-orthogonality present in the code.

vii) Also the factor b is the only phrase which will change its value when the linear transformations are applied upon b, which also depends on the exacting channel recognition further influencing directly on the performance of code leading to complex and computationally pricey decoding.

4.3 EVCM and Grammian matrix for these QO-STBCs

4.3.1 EVCM for Jafarkhani

We have been able to achieve an equivalent channel matrix for the corresponding code as

$$H_j = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & h_4^* & -h_1^* & -h_2^* \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix} \quad (4.14)$$

For orthogonal codes, we decode the received signal by calculating detection matrix as

$$D = H^H H \quad (4.15)$$

This detection matrix is diagonal in nature and contains only diagonal elements representing the channel gains, whereas here for QO-STBCs, the detection matrix not only contains diagonal elements but also non-diagonal elements representing the interference terms achieved from other channels.

Here for QO-STBC we call the detection matrix as Grammian Matrix also. So

$$D_J = H_J^H H_J = \begin{bmatrix} a & 0 & 0 & b \\ 0 & a & -b & 0 \\ 0 & -b & a & 0 \\ b & 0 & 0 & a \end{bmatrix} \quad (4.16)$$

With $a = \sum_{i=1}^4 |h_i|^2$ (Channel Gain) (4.17)

And $b = 2\text{Re}(h_1 h_4^* - h_2 h_3^*)$ (Interference Terms) (4.18)

Both a and b are real in nature, thus leading to D as Symmetric matrix.

4.3.2 EVCM/Grammian matrix for ABBA code

The virtual channel matrix is also obtained for ABBA code by applying the similar EVCM approach and we get it as

$$H_A = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3 & h_4 & h_1 & h_2 \\ h_4^* & -h_3^* & h_2^* & -h_1^* \end{bmatrix} \quad (4.19)$$

With the same approach as applied above, the detection matrix or Grammian matrix obtained is calculated as

$$D_A = H_A^H H_A \quad (4.20)$$

$$D_A = \begin{bmatrix} a & 0 & b & 0 \\ 0 & a & 0 & b \\ b & 0 & a & 0 \\ 0 & b & 0 & a \end{bmatrix} \quad (4.21)$$

Where $a = \sum_{i=1}^4 |h_i|^2$ (4.22)

And $b = 2\text{Re}(h_1 h_3^* + h_2 h_4^*)$ (4.23)

Here also both these constant are real in nature, so D_A is symmetric in nature.

4.3.3 EVCM/Grammian matrix for Tirkkonen Code

The EVCM approach lets the channel matrix as

$$H_T = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3 & h_4 & h_1 & h_2 \\ h_4^* & -h_3^* & h_2^* & -h_1^* \end{bmatrix} \quad (4.24)$$

The Grammian matrix here is given by

$$D_T = H_T^H H_T = \begin{bmatrix} a & 0 & b & 0 \\ 0 & a & 0 & b \\ b & 0 & a & 0 \\ 0 & b & 0 & a \end{bmatrix} \quad (4.25)$$

$$\text{With } a = |h_1|^2 + |h_2|^2 + |h_3|^2 + |h_4|^2 \quad (4.26)$$

$$\text{And } b = 2\text{Re}(h_1 h_3^* + h_2 h_4^*) \quad (4.27)$$

D_T is also real symmetric because constant nature of 'a' and 'b'.

4.4 Receiver Algorithm for QO-STBCs

4.4.1 Linear Receivers

There are mainly two types of linear receivers as

- (i) Zero forcing
- (ii) Minimum mean square error

These minimum receivers decrease the decoding complexity to much extent but suffer from one drawback as noise enhancement. The formula is given by

$$\hat{S} = (H^H H + \mu I)^{-1} z \quad (4.28)$$

Where, μ is zero for zero forcing

And $\mu = \sigma_n^2$ for MMSE, where σ_n^2 is noise variance.

Practically it is difficult to calculate actual value of this parameter which is σ_n^2 , so MMSE is not used much.

With zero forcing we have

$$\begin{aligned} \hat{S} &= (H^H H)^{-1} z \\ &= S + (H^H H)^{-1} H^H \bar{n} \\ &= S + \tilde{n} \end{aligned} \quad (4.29)$$

Zero forcing will decode symbols independently but undoubtedly it can degrade performance significantly.

4.4.2 Maximum Likelihood Channel

This receiver minimizes the probability of error to a greater extent, but its decoding complexity increases with the constellation size and the number of transmit antennas. The principle in which it is, is that it will select a signal vector S , and tries to minimize the distance between all the possible output vectors and the received vectors. Let us take S be a transmit vector possible, denote distance as D , then

$$D(S) = \|y - H^H\|^2 \quad (4.30)$$

When we are considering QPSK constellation, there are total of $4^4=256$ combinations possible. So it means ML decoder will calculate 256 distance matrices and try to find the one with the least distance. And finally correspondingly best possible code S is selected.

4.4.3 Maximal ratio combining

The simplest way to decode any QO-STBC is with this method. The received signal vector y is multiplied with H^H , and z-decision vector is calculated as

$$\begin{aligned} z &= H^H y \\ &= H^H (Hs + \bar{n}) \\ &= H^H Hs + H^H \bar{n} \\ &= Ds + H^H \bar{n} \end{aligned} \quad (4.31)$$

Where $D=H^H H$ is a detection matrix. Considering Jafarkhani type, the detection matrix D has the following form:

$$D = a \begin{bmatrix} 1 & 0 & 0 & X \\ 0 & 1 & -X & 0 \\ 0 & -X & 1 & 0 \\ X & 0 & 0 & 1 \end{bmatrix} \quad (4.32)$$

Here X is representing self-interference gain with $X = \beta/\alpha$

Decoding on the receiver front is performed by splitting up decision vector z into subset z_1 and z_4 , z_2 and z_3 . Mathematically,

$$z = \alpha \begin{bmatrix} S_1 + XS_4 \\ S_2 - XS_3 \\ S_3 - XS_2 \\ S_4 + XS_1 \end{bmatrix} + H^H \bar{n} \quad (4.33)$$

Making groups of two, we get

$$\begin{bmatrix} z_1 \\ z_4 \end{bmatrix} = \alpha \begin{bmatrix} 1 & X \\ X & 1 \end{bmatrix} \begin{bmatrix} S_1 \\ S_4 \end{bmatrix} + \begin{bmatrix} \tilde{n}_1 \\ \tilde{n}_4 \end{bmatrix} \quad (4.34)$$

$$\begin{bmatrix} z_2 \\ z_3 \end{bmatrix} = \alpha \begin{bmatrix} 1 & X \\ X & 1 \end{bmatrix} \begin{bmatrix} S_2 \\ S_3 \end{bmatrix} + \begin{bmatrix} \tilde{n}_2 \\ \tilde{n}_3 \end{bmatrix} \quad (4.35)$$

Where, \tilde{n}_i , $i=1$ to 4 represents received noise vector after MRC.

When these decoders are compared it is seen that zero forcing is better than the maximum likelihood decoder. But still we prefer maximum likelihood because zero forcing has the disadvantage of noise amplification.

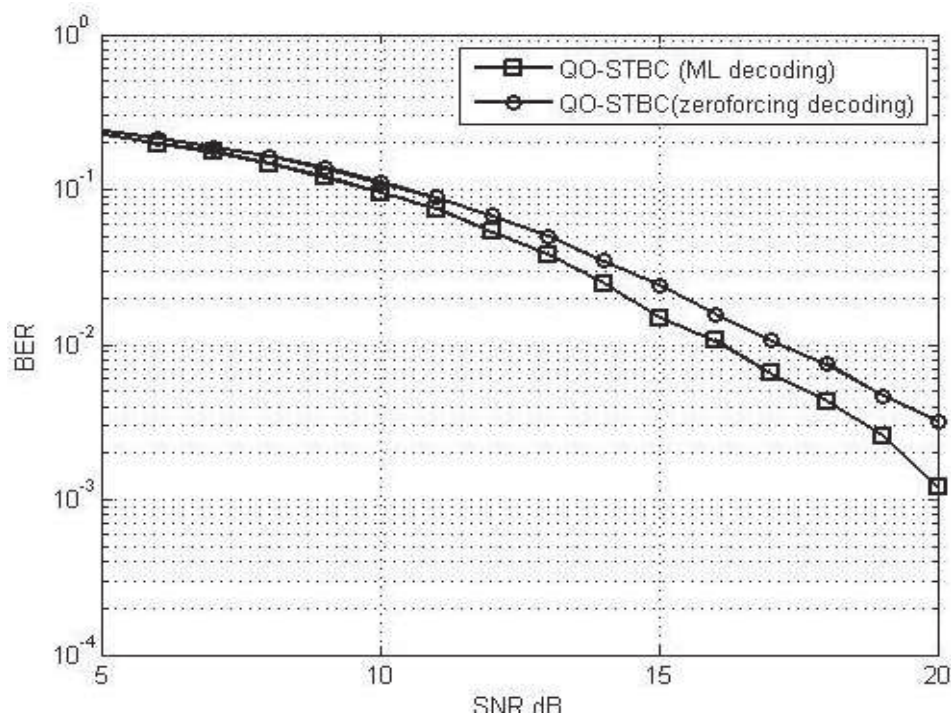


Fig 4.1 Comparison of QO-STBC with different decoding techniques.

4.5 Generalized proposed scheme for Conventional QO-STBC

The detection matrix which we have obtained contains interference terms along with channel gains which lead to complex decoding on receiver front. So we proposed a generalized scheme for various popular QO-STBC, which could eliminate these interference terms and we get linear decoding as mentioned above, our detection matrix is symmetric in nature i.e., $D^T = D$.

And any symmetric real matrix can be expressed as

$$D = QD_nQ^T \quad (4.36)$$

With Q as Orthogonal Matrix and D_n as diagonal matrix, the diagonal elements of whose are Eigen-values of D .

Now pre and post-multiplying the detection matrix with Q^T and Q , the matrix D_n obtained in the form of diagonal matrix, is free from interference terms. In Taha Scheme, similar concept is used but this value of Q taken is a general unitary matrix. Here, we apply the concept of singular value decomposition to calculate the value of Q .

4.5.1 Jafarkhani Code

As discussed previously Jafarkhani for four Transmit antennas and one receiver antenna is:

$$S_J = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & -s_3 & -s_2 & s_1 \end{bmatrix} \quad (4.37)$$

Its corresponding equivalent channel matrix obtained is

$$H = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & h_4^* & -h_1^* & -h_2^* \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix} \quad (4.38)$$

The received vector y is written as

$$y = HS + \bar{n} \quad (4.39)$$

$$s = [s_1, s_2, s_3, s_4]^T \quad (4.40)$$

$$\bar{n} = [n_1, n_2^*, n_3^*, n_4]^T \quad (4.41)$$

The Q obtained by SVD is

$$Q = \begin{bmatrix} 0.5 & 0.5 & 0.5 & -0.5 \\ 0.5 & -0.5 & 0.5 & 0.5 \\ -0.5 & 0.5 & 0.5 & 0.5 \\ 0.5 & 0.5 & -0.5 & 0.5 \end{bmatrix} \quad (4.42)$$

In matrix form, we write it as

$$D_n = Q = \begin{bmatrix} 0.5 & 0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & 0.5 & 0.5 & -0.5 \\ -0.5 & 0.5 & 0.5 & 0.5 \end{bmatrix} \begin{bmatrix} a & 0 & 0 & b \\ 0 & a & -b & 0 \\ 0 & -b & a & 0 \\ b & 0 & 0 & a \end{bmatrix} \begin{bmatrix} 0.5 & 0.5 & 0.5 & -0.5 \\ 0.5 & -0.5 & 0.5 & 0.5 \\ -0.5 & 0.5 & 0.5 & 0.5 \\ 0.5 & 0.5 & -0.5 & 0.5 \end{bmatrix} \quad (4.43)$$

The new determinant matrix D_n free from interference is:

$$D_n = \begin{bmatrix} a + b & 0 & 0 & 0 \\ 0 & a + b & 0 & 0 \\ 0 & 0 & a - b & 0 \\ 0 & 0 & 0 & a - b \end{bmatrix} \quad (4.44)$$

Corresponding to this, we derive new channel matrix and new coding matrix as

$$\begin{aligned} D_n &= Q^T \cdot D \cdot Q = Q^H \cdot D \cdot Q \\ &= Q^H \cdot H^H \cdot H \cdot Q \\ &= (HQ)^H (HQ) \end{aligned} \quad (4.45)$$

So new channel matrix obtained is

$$H_n = HQ \quad (4.46)$$

When this modified jafarkhani code (which is obtained from rotation mechanism) is sent from the transmitter better results are obtained as observed in fig 4.2.

$$\begin{aligned}
& H_n \\
& = \begin{bmatrix} \frac{(h_1 + h_2 - h_3 + h_4)}{2} & \frac{(h_1 - h_2 + h_3 + h_4)}{2} & \frac{(h_1 + h_2 + h_3 - h_4)}{2} & \frac{(-h_1 + h_2 + h_3 + h_4)}{2} \\ \frac{(h_2^* - h_1^* - h_4^* - h_3^*)}{2} & \frac{(h_2^* + h_1^* + h_4^* - h_3^*)}{2} & \frac{(h_2^* - h_1^* + h_4^* + h_3^*)}{2} & \frac{(-h_2^* - h_1^* + h_4^* - h_3^*)}{2} \\ \frac{(h_3^* + h_4^* + h_1^* - h_2^*)}{2} & \frac{(h_3^* - h_4^* - h_1^* - h_2^*)}{2} & \frac{(h_3^* + h_4^* - h_1^* + h_2^*)}{2} & \frac{(-h_3^* + h_4^* - h_1^* - h_2^*)}{2} \\ \frac{(h_4 - h_3 + h_2 + h_1)}{2} & \frac{(h_4 + h_3 - h_2 + h_1)}{2} & \frac{(h_4 - h_3 - h_2 + h_1)}{2} & \frac{(-h_4 - h_3 - h_2 + h_1)}{2} \end{bmatrix}
\end{aligned} \tag{4.47}$$

And new encoding matrix obtained is

$$\begin{aligned}
& S_n \\
& = \begin{bmatrix} \frac{(s_1 + s_2 + s_3 - s_4)}{2} & \frac{(s_1 - s_2 + s_3 + s_4)}{2} & \frac{(-s_1 + s_2 + s_3 + s_4)}{2} & \frac{(s_1 + s_2 - s_3 + s_4)}{2} \\ \frac{(-s_1^* + s_2^* - s_3^* - s_4^*)}{2} & \frac{(s_1^* + s_2^* + s_3^* - s_4^*)}{2} & \frac{(-s_1^* - s_2^* + s_3^* - s_4^*)}{2} & \frac{(-s_1^* + s_2^* + s_3^* + s_4^*)}{2} \\ \frac{(s_1^* - s_2^* - s_3^* - s_4^*)}{2} & \frac{(-s_1^* - s_2^* + s_3^* - s_4^*)}{2} & \frac{(s_1^* + s_2^* + s_3^* - s_4^*)}{2} & \frac{(s_1^* - s_2^* + s_3^* + s_4^*)}{2} \\ \frac{(s_1 + s_2 - s_3 + s_4)}{2} & \frac{(s_1 - s_2 - s_3 - s_4)}{2} & \frac{(-s_1 + s_2 - s_3 - s_4)}{2} & \frac{(s_1 + s_2 + s_3 - s_4)}{2} \end{bmatrix}
\end{aligned} \tag{4.48}$$

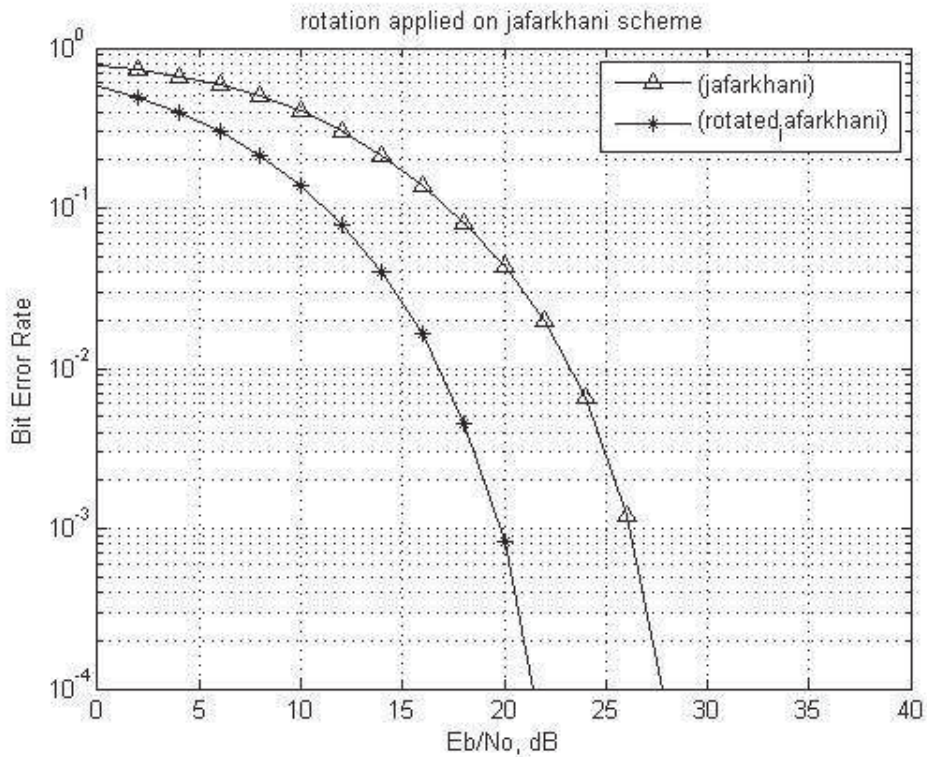


Fig 4.2 Rotation mechanism applied upon Jafarkhani Scheme

4.5.2 Tirkkonen Code

Just as described above for Jafarkhani code, the Q matrix for Tirkkonen is obtained as

$$Q = \begin{bmatrix} -0.5 & 0.5 & 0.5 & 0.5 \\ -0.5 & -0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & 0.5 & -0.5 & 0.5 \end{bmatrix} \quad (4.49)$$

Writing it in matrix form, we have

$$D_n = \begin{bmatrix} -0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & -0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & -0.5 \\ 0.5 & 0.5 & 0.5 & 0.5 \end{bmatrix} \begin{bmatrix} a & 0 & b & 0 \\ 0 & a & 0 & b \\ b & 0 & a & 0 \\ 0 & b & 0 & a \end{bmatrix} \begin{bmatrix} -0.5 & 0.5 & 0.5 & 0.5 \\ -0.5 & -0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & 0.5 & -0.5 & 0.5 \end{bmatrix} \quad (4.50)$$

The D_n obtained here is

$$D_n = \begin{bmatrix} a+b & 0 & 0 & 0 \\ 0 & a+b & 0 & 0 \\ 0 & 0 & a-b & 0 \\ 0 & 0 & 0 & a-b \end{bmatrix} \quad (4.51)$$

The new channel matrix D_n obtained here is

$$\begin{aligned} D_n &= Q^H D Q \\ &= (HQ)^H (HQ) \end{aligned} \quad (4.52)$$

So new channel matrix obtained here is $H_n = HQ$

H_n

$$= \begin{bmatrix} \frac{(-h_1 - h_2 + h_3 + h_4)}{2} & \frac{(h_1 - h_2 - h_3 + h_4)}{2} & \frac{(h_1 - h_2 + h_3 - h_4)}{2} & \frac{(h_1 + h_2 + h_3 + h_4)}{2} \\ \frac{(-h_1^* + h_2^* + h_3^* - h_4^*)}{2} & \frac{(-h_1^* - h_2^* + h_3^* + h_4^*)}{2} & \frac{(-h_1^* - h_2^* - h_3^* - h_4^*)}{2} & \frac{(h_1^* - h_2^* + h_3^* - h_4^*)}{2} \\ \frac{(h_1 + h_2 - h_3 - h_4)}{2} & \frac{(-h_1 + h_2 + h_3 - h_4)}{2} & \frac{(h_1 - h_2 + h_3 - h_4)}{2} & \frac{(h_1 + h_2 + h_3 + h_4)}{2} \\ \frac{(h_1^* - h_2^* - h_3^* + h_4^*)}{2} & \frac{(h_1^* + h_2^* - h_3^* - h_4^*)}{2} & \frac{(-h_1^* - h_2^* - h_3^* - h_4^*)}{2} & \frac{(h_1^* - h_2^* + h_3^* - h_4^*)}{2} \end{bmatrix} \quad (4.53)$$

And correspondingly the new encoding matrix is

$$\begin{aligned}
& S_n \\
& = \begin{bmatrix} \frac{(-s_1 + s_2 + s_3 + s_4)}{2} & \frac{(-s_1 - s_2 - s_3 + s_4)}{2} & \frac{(s_1 - s_2 + s_3 + s_4)}{2} & \frac{(s_1 + s_2 - s_3 + s_4)}{2} \\ \frac{(-s_1^* - s_2^* - s_3^* + s_4^*)}{2} & \frac{(s_1^* - s_2^* - s_3^* - s_4^*)}{2} & \frac{(s_1^* + s_2^* - s_3^* + s_4^*)}{2} & \frac{(-s_1^* + s_2^* - s_3^* + s_4^*)}{2} \\ \frac{(s_1 - s_2 + s_3 + s_4)}{2} & \frac{(s_1 + s_2 - s_3 + s_4)}{2} & \frac{(-s_1 + s_2 + s_3 + s_4)}{2} & \frac{(-s_1 - s_2 - s_3 + s_4)}{2} \\ \frac{(s_1^* + s_2^* - s_3^* + s_4^*)}{2} & \frac{(-s_1^* + s_2^* - s_3^* - s_4^*)}{2} & \frac{(-s_1^* - s_2^* - s_3^* + s_4^*)}{2} & \frac{(s_1^* - s_2^* - s_3^* - s_4^*)}{2} \end{bmatrix}
\end{aligned}
\tag{4.54}$$

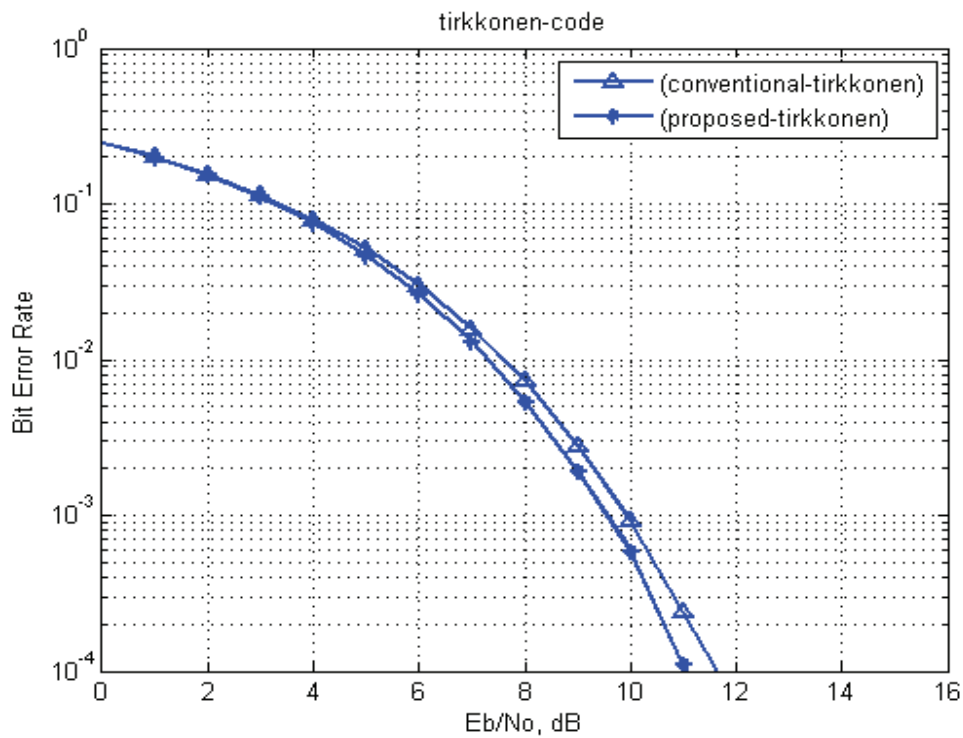


Fig 4.3 Rotation mechanism applied upon Tirkkonen Code

4.5.3 ABBA Code

A similar procedure when applied on ABBA results in unitary matrix:

$$Q = \begin{bmatrix} -0.5 & 0.5 & 0.5 & 0.5 \\ -0.5 & -0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & 0.5 & -0.5 & 0.5 \end{bmatrix}
\tag{4.55}$$

The new detection matrix is written as

$$D_n = \begin{bmatrix} -0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & -0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & -0.5 \\ 0.5 & 0.5 & 0.5 & 0.5 \end{bmatrix} \begin{bmatrix} a & 0 & b & 0 \\ 0 & a & 0 & b \\ b & 0 & a & 0 \\ 0 & b & 0 & a \end{bmatrix} \begin{bmatrix} -0.5 & 0.5 & 0.5 & 0.5 \\ -0.5 & -0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & 0.5 & -0.5 & 0.5 \end{bmatrix}$$

$$\Rightarrow D_n = \begin{bmatrix} a+b & 0 & 0 & 0 \\ 0 & a+b & 0 & 0 \\ 0 & 0 & a-b & 0 \\ 0 & 0 & 0 & a-b \end{bmatrix} \quad (4.56)$$

The new channel matrix is:

$$H_n = \begin{bmatrix} \frac{(-h_1 - h_2 + h_3 + h_4)}{2} & \frac{(h_1 - h_2 - h_3 + h_4)}{2} & \frac{(h_1 - h_2 + h_3 - h_4)}{2} & \frac{(h_1 + h_2 + h_3 + h_4)}{2} \\ \frac{(h_1^* - h_2^* - h_3^* + h_4^*)}{2} & \frac{(h_1^* + h_2^* - h_3^* - h_4^*)}{2} & \frac{(h_1^* + h_2^* + h_3^* + h_4^*)}{2} & \frac{(-h_1^* + h_2^* - h_3^* + h_4^*)}{2} \\ \frac{(h_1 + h_2 - h_3 - h_4)}{2} & \frac{(-h_1 + h_2 + h_3 - h_4)}{2} & \frac{(h_1 - h_2 + h_3 - h_4)}{2} & \frac{(h_1 + h_2 + h_3 + h_4)}{2} \\ \frac{(-h_1^* + h_2^* + h_3^* - h_4^*)}{2} & \frac{(-h_1^* - h_2^* + h_3^* + h_4^*)}{2} & \frac{(h_1^* + h_2^* + h_3^* + h_4^*)}{2} & \frac{(-h_1^* + h_2^* - h_3^* + h_4^*)}{2} \end{bmatrix} \quad (4.57)$$

And the corresponding encoding matrix is

$$S_n = \begin{bmatrix} \frac{(-s_1 + s_2 + s_3 + s_4)}{2} & \frac{(-s_1 - s_2 - s_3 + s_4)}{2} & \frac{(s_1 - s_2 + s_3 + s_4)}{2} & \frac{(s_1 + s_2 - s_3 + s_4)}{2} \\ \frac{(s_1^* + s_2^* + s_3^* - s_4^*)}{2} & \frac{(-s_1^* + s_2^* + s_3^* + s_4^*)}{2} & \frac{(-s_1^* - s_2^* + s_3^* - s_4^*)}{2} & \frac{(s_1^* - s_2^* + s_3^* + s_4^*)}{2} \\ \frac{(s_1 - s_2 + s_3 + s_4)}{2} & \frac{(s_1 + s_2 - s_3 + s_4)}{2} & \frac{(-s_1 + s_2 + s_3 + s_4)}{2} & \frac{(-s_1 - s_2 - s_3 + s_4)}{2} \\ \frac{(-s_1^* - s_2^* + s_3^* - s_4^*)}{2} & \frac{(s_1^* - s_2^* + s_3^* - s_4^*)}{2} & \frac{(s_1^* + s_2^* + s_3^* - s_4^*)}{2} & \frac{(-s_1^* + s_2^* + s_3^* + s_4^*)}{2} \end{bmatrix} \quad (4.58)$$

From here we observe that to remove interference terms from detection matrix we use the symmetry concept and reduce the desired matrix to diagonal form as done in Taha scheme. But the unitary matrix obtained is calculated as SVD. And this whole concept is generalized with a conclusion that the scheme can be applied to other codes which enhances their diversity with minimum decoding complexity.

The new encoding matrix S_n is Quasi-Orthogonal whereas the channel matrix H_n is orthogonal, so decoding is achieved via simple linear-decoding.

$$\hat{S} = (H^H H)S_n + H^H N \quad (4.59)$$

$$S_n = [s_1, s_2, s_3, s_4]^T \quad (4.60)$$

$$\bar{N} = [n_1, n_2^*, n_3^*, n_4]^T \quad (4.61)$$

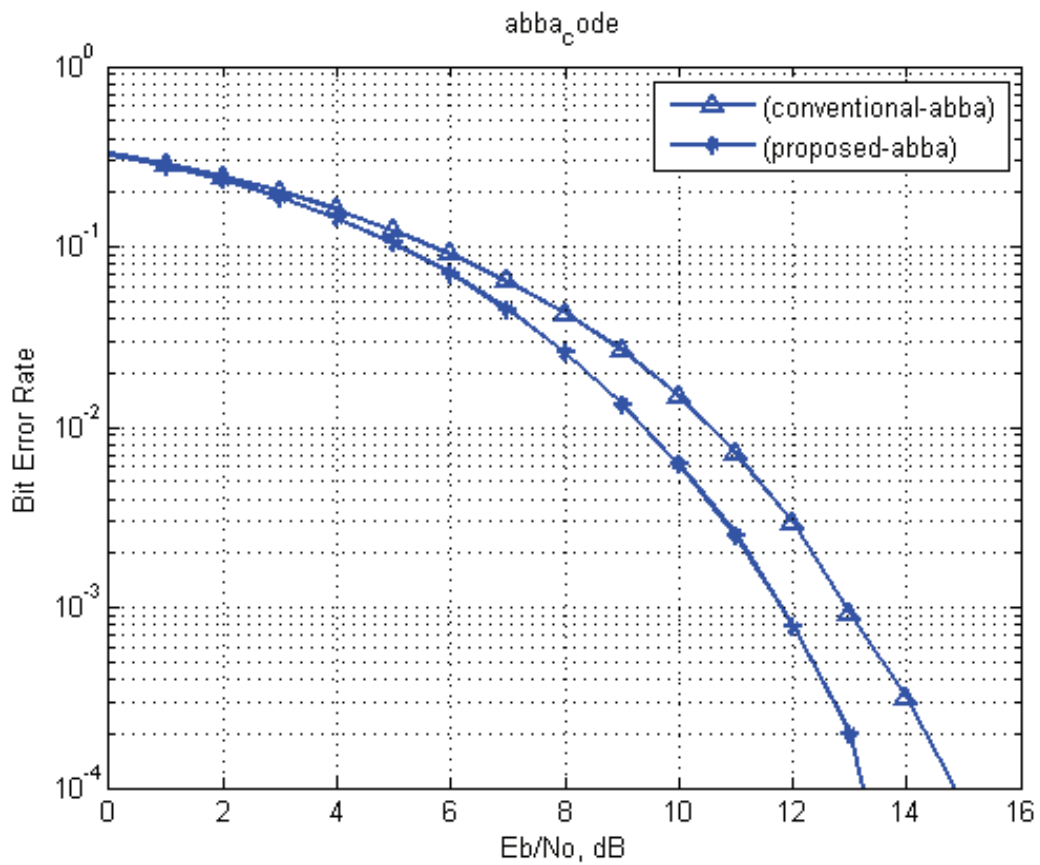


Fig 4.4 Rotation mechanism applied upon ABBA code

The comparison of all three codes Jafarkhani, Tirkkonen and ABBA is shown in fig 4.5; all of them behave differently under different channel conditions. Any of the codes can outperform the other one and hence any one code could give better results because channel conditions are random. Here in particular Jafarkhani performs better in comparison to all other two codes.

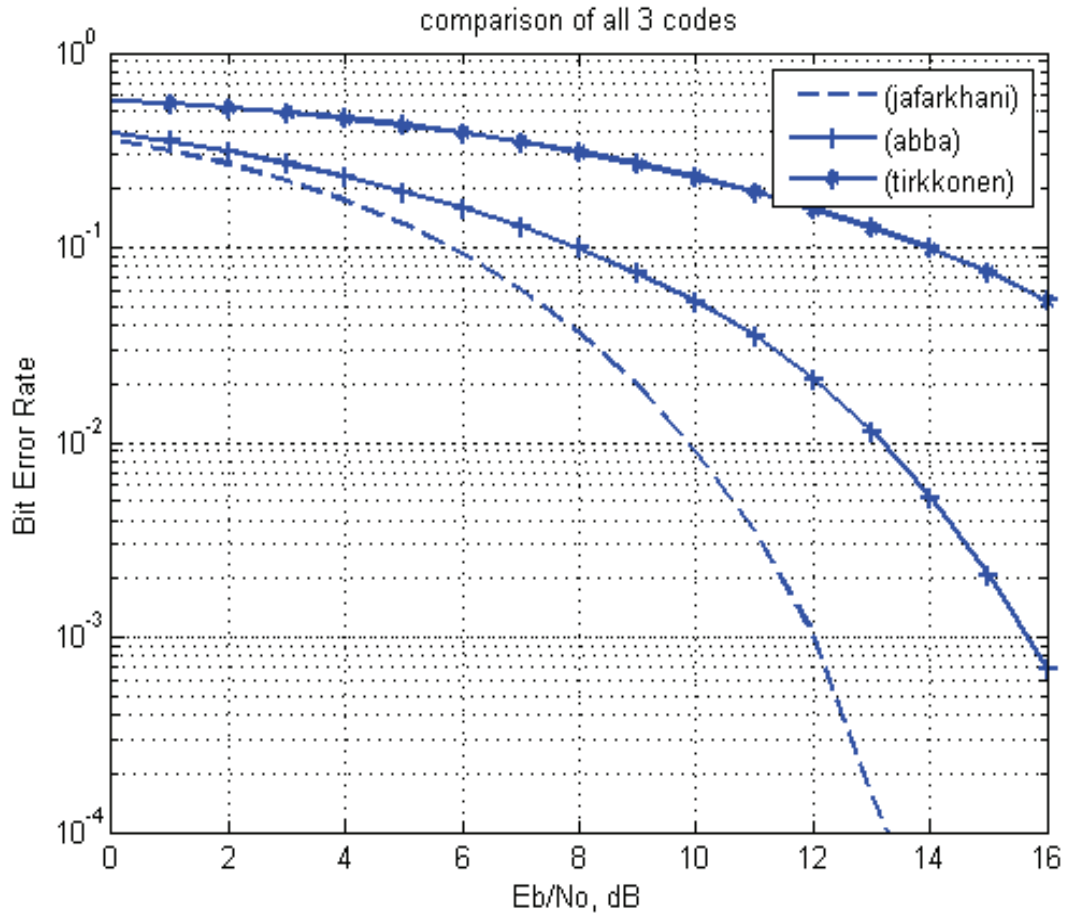


Fig 4.5 Performance showing comparison of all three codes.

ROTATED QUASI ORTHOGONAL–SPACE TIME BLOCK CODES

5.1 Introduction

Talking about regular and symmetric constellations as QAM or PSK, the rank of the difference matrix which is $D(C^i, C^j)$ came out to be two in case of QO-STBCs. Here if we have M receive antennas, we are able to achieve $2M$ diversity order, while maintaining code rate of 1. This was the case of real and complex orthogonal code. Moving to complex orthogonal codes while keeping antennas till four, we could not achieve diversity order of $4M$, with the code rate of one.

Thus to provide full diversity, different symbols are transmitted from different constellations. Let us say we rotate symbols x_3 and x_4 before transmitting them. So their rotated versions are denoted by \tilde{x}_3 and \tilde{x}_4 respectively [21]. And we will prove that on replacing (x_3, x_4) by $(\tilde{x}_3, \tilde{x}_4)$, it is possible to achieve full diversity Quasi Orthogonal STBCs. The finally achieved code will give us full diversity order, code rate 1, also pairwise decoding. We now calculate the conditions to achieve this full diversity, by deriving CGD i.e., code-word gain distance for different cases and prove that they behave in a similar manner.

We explain the general rotation mechanism by taking Jafarkhani code.

$$S_J = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & -s_3 & -s_2 & s_1 \end{bmatrix} \quad (5.1)$$

With its equivalent channel matrix for the corresponding code is

$$H_J = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & h_4^* & -h_1^* & -h_2^* \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix} \quad (5.2)$$

Here for QO-STBC we call the detection matrix as Grammian Matrix also. So

$$D_J = H_J^H H_J \quad (5.3)$$

We calculated G as generator matrix which was used to decode quasi orthogonal space time block codes, we now decode it for checking diversity and code rate. Let the pair of code-words to be transmitted in generator matrix are denoted as:

$$C = G(s_1, s_2, \tilde{s}_3, \tilde{s}_4) \quad (5.4)$$

$$C' = G(s'_1, s'_2, \tilde{s}'_3, \tilde{s}'_4) \quad (5.5)$$

CGD between these two code-words is given by the formulae:

$$CGD(C, C') = \det(D(C, C')^H \cdot D(C, C')) \quad (5.6)$$

Where, $D(C, C')$ represents the difference matrix and $D(C, C')^H$ represents the hermitian.

So now we calculate code-word distance by putting the values,

$$\begin{aligned} CGD(C, C') &= (|(s_1 - s'_1) - (\tilde{s}_4 - \tilde{s}'_4)|^2 + |(s_2 - s'_2) + (\tilde{s}_3 - \tilde{s}'_3)|^2) \\ &\times (|(s_1 - s'_1) + (\tilde{s}_4 - \tilde{s}'_4)|^2 + |(s_2 - s'_2) - (\tilde{s}_3 - \tilde{s}'_3)|^2) \end{aligned} \quad (5.7)$$

The detection matrix obtained let us say for Jafarkhani type was

$$D = \begin{bmatrix} a & 0 & 0 & b \\ 0 & a & -b & 0 \\ 0 & -b & a & 0 \\ b & 0 & 0 & a \end{bmatrix} \quad (5.8)$$

$$a = \sum_{j=1}^4 |x_j|^2 \quad (5.9)$$

$$b = 2 \operatorname{Re}(x_1 x_4^* - x_2 x_3^*) \quad (5.10)$$

Applying one of the determinant equality, we have

$$\det \begin{pmatrix} A & B \\ C & D \end{pmatrix} = \det(A) \det(D - CA^{-1}B) \quad (5.11)$$

$$\text{whose determinant is given by } \det(D) = (a^2 - b^2)^2 \quad (5.12)$$

Putting all the values and with simple algebraic manipulation, we get

$$\det(D) = (|x_1 - x_4|^2 + |x_2 + x_3|^2)^2 \times (|x_1 + x_4|^2 + |x_2 - x_3|^2)^2 \quad (5.13)$$

Now replacing x_1 with $(s_1 - s'_1)$

$$x_2 \text{ with } (s_2 - s'_2)$$

$$x_3 \text{ with } (s_3 - s'_3)$$

$$x_4 \text{ with } (s_4 - s'_4)$$

(5.14)

On putting these values of x_1, x_2, x_3, x_4 in above equation and doing simple algebraic manipulations we achieve,

$$\begin{aligned} \text{CGD} = & (|(s_1 - s'_1) - (\tilde{s}_4 - \tilde{s}'_4)|^2 + |(s_2 - s'_2) + (\tilde{s}_3 - \tilde{s}'_3)|^2)^2 \\ & \times (|(s_1 - s'_1) + (\tilde{s}_4 - \tilde{s}'_4)|^2 + |(s_2 - s'_2) - (\tilde{s}_3 - \tilde{s}'_3)|^2)^2 \end{aligned}$$

(5.15)

We could generalize this condition as

$$\text{CGD}(C, C') = \left(\sum_{k=1}^2 |(s_k - s'_k) - (\tilde{s}_{k+2} - \tilde{s}'_{k+2})|^2 \right)^2$$

(5.16)

5.1.1 Condition for full diversity QO-STBC

We can achieve full diversity only if our CGD is zero, then at least one of the two factors should become zero. We find the minimum value of CGD, by considering all possible values of $s_k, s'_k, \tilde{s}_{k+2}, \tilde{s}'_{k+2}$ where k is either 1 or 2. With all the set of values, $\sum_{k=1}^2 |(s_k - s'_k) - (\tilde{s}_{k+2} - \tilde{s}'_{k+2})|^2 = 0$, and switching \tilde{s}_{k+2} with \tilde{s}'_{k+2} , we get $\sum_{k=1}^2 |(s_k - s'_k) - (\tilde{s}_{k+2} - \tilde{s}'_{k+2})|^2 = 0$ too. We consider a case when second sum become zero, and this will contain only non-negative terms. Thus sum will be zero, when both the terms are zero. If no symbols, results in

$$s_k - s'_k = \tilde{s}_{k+2} - \tilde{s}'_{k+2} \quad (5.17)$$

None of the terms will make sum zero. Henceforth it is proven that if we choose two symbols let's say third and four from a rotated constellation full diversity order of four is

achievable.

5.2 Optimal rotation and performance of QO-STBC.

Here we use determinant criteria to calculate the optimum rotation angle for different constellations, and also this angle will maximize the minimum value of CGD among all the constellation points which are possible [22].

On transmitter side, encoder will choose or pick the four symbols (s_1, s_2, s_3, s_4) from a block of 4b input bits. In the generator matrix of QO-STBC, the two symbols i.e., s_3 and s_4 are chosen as rotated symbols i.e. $\tilde{s}_3 = e^{j\phi}s_3$ and $\tilde{s}_4 = e^{j\phi}s_4$. Our work is to find optimum value of this ϕ which will maximize the minimum value of CGD. Let us denote this minimum CGD for rotation mechanism as $\text{CGD}_{\min}(\phi)$. And this CGD is a function of angle ϕ , and our goal is to maximize this CGD among all the rotations which are possible.

$$\text{CGD}_{\min}(\phi) = \min_{(s_1, \tilde{s}_3) \neq (s_1', \tilde{s}_3')} |(s_1 - s_1')^2 - (\tilde{s}_3 - \tilde{s}_3')^2|^4 \quad (5.18)$$

Here, s_1 and s_1' are taken from original constellation and \tilde{s}_3 and \tilde{s}_3' are taken from rotated constellations.

This term $\text{CGD}_{\min}(\phi)$ represents the distance between original and rotated constellation. $d_{\min} = \min(s_1 - s_1')$ is representing the minimum Euclidean distance among all the possible points. If we choose $\tilde{s}_3 = \tilde{s}_3'$, right hand side becomes equal to power of eight of minimum Euclidean distance in constellation.

$$\text{CGD}_{\min}(\phi) \leq d_{\min}^8 \quad (5.19)$$

Here d_{\min}^8 is an upper bound. If it exists a rotation angle ϕ^* , for which CGD_{\min} becomes equal to d_{\min}^8 , then its maximum value will come for the optimum angle of rotation as ϕ^* . Also, this optimum angle is not unique; moreover this bound is not always achievable. And now if we calculate the minimum Euclidean distance d_{\min} , the minimum possible CGD is upper bound by

$$\det(A) = \left| (c_1^1 - c_1^2)^2 - (c_4^1 e^{j\phi} - c_4^2 e^{j\phi})^2 \right|^4$$

$$\begin{aligned}
&= |(c_1^1 - c_1^2)^2 - e^{2j\phi}(c_4^1 - c_4^2)^2|^4 \\
&= |d_{\min}^2 - e^{2j\phi}d_{\min}^2|^4 \\
&= |1 - e^{2j\phi}|^4 d_{\min}^8 \\
&= |2\sin\phi|^4 d_{\min}^8
\end{aligned}$$

$$CGD_{\min}(\phi) \leq |2\sin\phi|^4 d_{\min}^8$$

(5.20)

From the above two equations, the second one is smaller than the upper bound for $\phi < \pi/6$. This code-word gain distance is a non-decreasing function for the rotation of angle $0 \leq \phi \leq \pi/2$. Next, we plot the graph between minimum CGD and different values of rotation. The graphs in fig 5.1 and 5.2 are plotted considering minimum code gain distance for QPSK and 8-PSK.

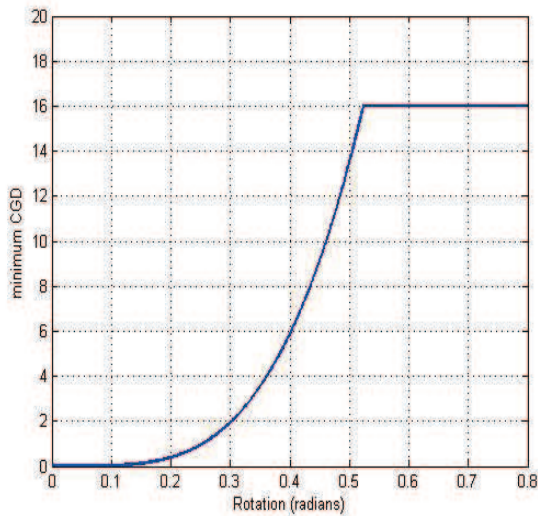


Fig 5.1 Minimum CGD using QPSK

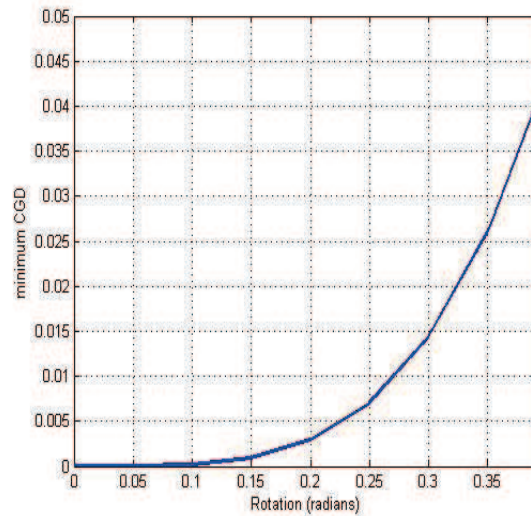


Fig 5.2 Minimum CGD using 8-PSK.

In case of QPSK constellation, theta equals to $\pi/4$ is an optimum rotation angle which will achieve its required upper bound with utmost probable $CGD_{\min} = 16$. Figure shows the minimum value of CGD for diverse rotation angles using QPSK constellation and the next Figure shows performance of Quasi Orthogonal space time block codes using QPSK with different rotations.

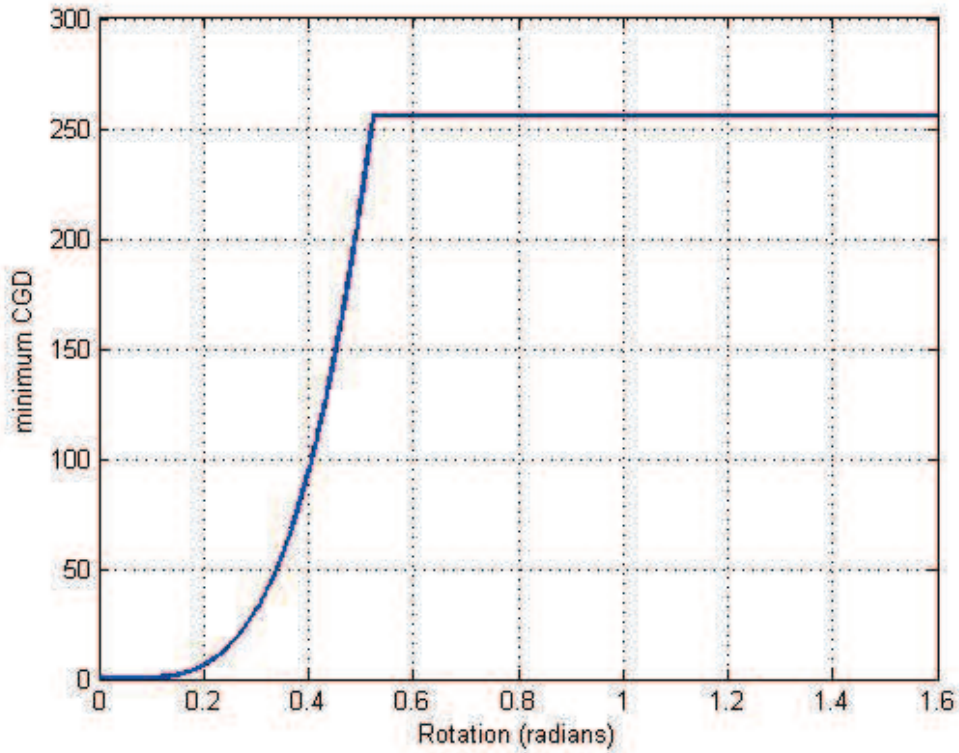


Fig 5.3: Minimum CGD for different rotation; QO-STBC using BPSK.

Whereas fig 5.3 illustrates minimum code gain distance taken for binary phase shift keying. And in case of 8-PSK, theta equals to $\pi/4$ is an optimum rotation which achieves the upper bound [23].

Thus finally the Jafarkhani code which is used as there rotated version, with third and fourth symbols taken from rotated constellations is described. Instead of denoting as C we mention it as C_CR that is code C with rotated constellations.

$$C_{CR} = \begin{bmatrix} c_1 & c_2 & c_3 e^{j\phi} & c_4 e^{j\phi} \\ -c_2^* & c_1^* & -(c_4 e^{j\phi})^* & (c_3 e^{j\phi})^* \\ -(c_3 e^{j\phi})^* & -(c_4 e^{j\phi})^* & c_1^* & c_2^* \\ c_4 e^{j\phi} & -c_3 e^{j\phi} & -c_2 & c_1 \end{bmatrix} \quad (5.21)$$

We also show here that when different values of theta are taken, different values of gain are obtained with different error rates. As can be seen in fig 5.4 at an angle of 45 degrees we get the best results. Thus our optimum angle of rotation is 45 degrees. And taking this optimum angle we get less BER for rotated QOSTBC as shown in fig 5.5.

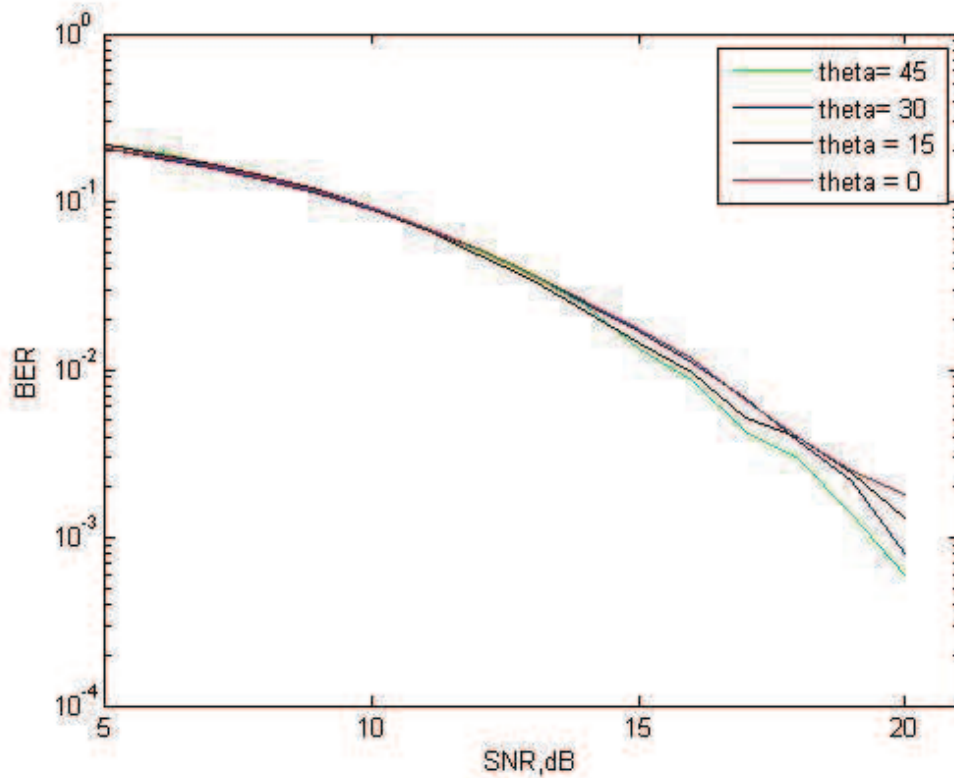


Fig. 5.4 BER curve for QO-STBC for different rotations using QPSK modulation

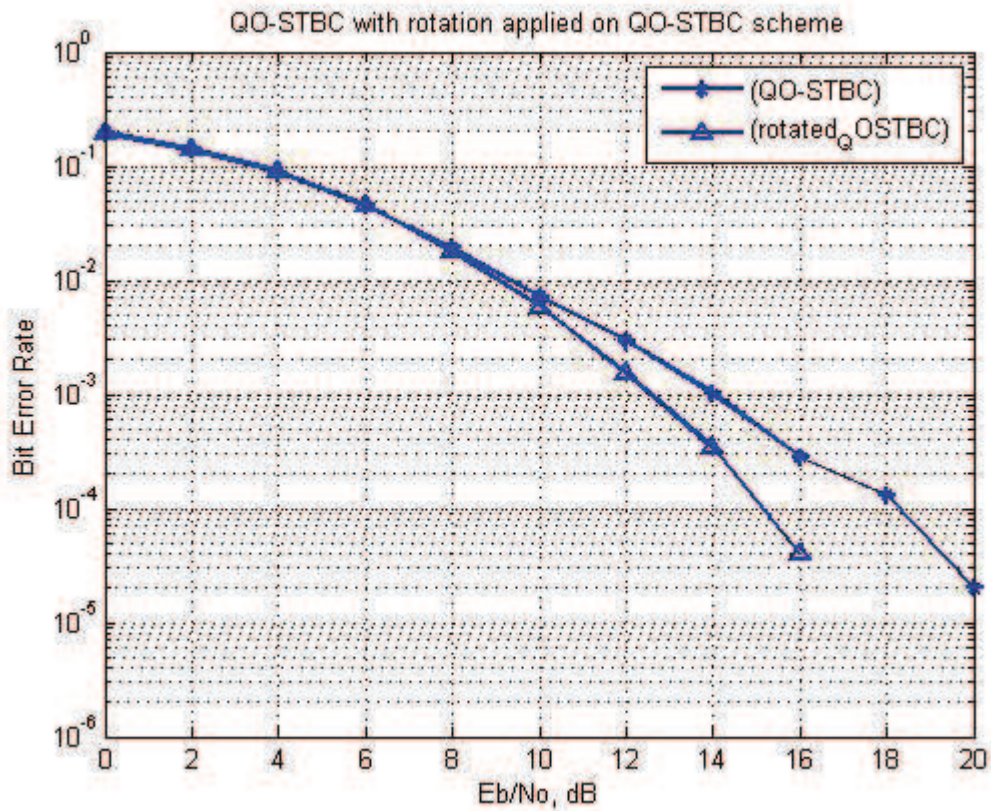


Fig 5.5: Comparison of bit error rate for rotated and non-rotated QO-STBC.

5.3 Rotation applied on Jafarkhani Scheme

As we know QOSTBC can achieve higher code rates than OSTBC, but it could not provide full-transmit diversity directly. Through rank criterion, it is easy to prove that the minimum rank of difference matrix $D(S^i, S^j)$ is two for T^X antennas with QOSTBC code, while using Symmetric Constellations with QPSK, QAM. But it is impossible to obtain maximum diversity of four if all the symbols are taken from the same constellation.

In this proposed scheme, the two symbols transmitted are, for example, \tilde{s}_3 and \tilde{s}_4 as rotated versions of s_3 and s_4 , whereas s_1 and s_2 are transmitted from the original constellation itself. The resulting code is more robust and powerful since it provides full rate, full diversity and linear decoding with good performance. Also we do not need any feedback as was vital in TAS.

Let our desired QOSTBC code is modified Jafarkhani type obtained

$$S_n = \begin{bmatrix} \frac{(s_1 + s_2 + s_3 - s_4)}{2} & \frac{(s_1 - s_2 + s_3 + s_4)}{2} & \frac{(-s_1 + s_2 + s_3 + s_4)}{2} & \frac{(s_1 + s_2 - s_3 + s_4)}{2} \\ \frac{(-s_1^* + s_2^* - s_3^* - s_4^*)}{2} & \frac{(s_1^* + s_2^* + s_3^* - s_4^*)}{2} & \frac{(-s_1^* - s_2^* + s_3^* - s_4^*)}{2} & \frac{(-s_1^* + s_2^* + s_3^* + s_4^*)}{2} \\ \frac{(s_1^* - s_2^* - s_3^* - s_4^*)}{2} & \frac{(-s_1^* - s_2^* + s_3^* - s_4^*)}{2} & \frac{(s_1^* + s_2^* + s_3^* - s_4^*)}{2} & \frac{(s_1^* - s_2^* + s_3^* + s_4^*)}{2} \\ \frac{(s_1 + s_2 - s_3 + s_4)}{2} & \frac{(s_1 - s_2 - s_3 - s_4)}{2} & \frac{(-s_1 + s_2 - s_3 - s_4)}{2} & \frac{(s_1 + s_2 + s_3 - s_4)}{2} \end{bmatrix} \quad (5.22)$$

An error will occur if the decoder mistakenly considers that we have transmitted other code S_n^2 instead of S_n^1 , where S_n^1 and S_n^2 are given by

$$S_n^1 = \begin{bmatrix} \frac{(s_1^1 + s_2^1 + s_3^1 - s_4^1)}{2} & \frac{(s_1^1 - s_2^1 + s_3^1 + s_4^1)}{2} & \frac{(-s_1^1 + s_2^1 + s_3^1 + s_4^1)}{2} & \frac{(s_1^1 + s_2^1 - s_3^1 + s_4^1)}{2} \\ \frac{(-s_1^{*1} + s_2^{*1} - s_3^{*1} - s_4^{*1})}{2} & \frac{(s_1^{*1} + s_2^{*1} + s_3^{*1} - s_4^{*1})}{2} & \frac{(-s_1^{*1} - s_2^{*1} + s_3^{*1} - s_4^{*1})}{2} & \frac{(-s_1^{*1} + s_2^{*1} + s_3^{*1} + s_4^{*1})}{2} \\ \frac{(s_1^{*1} - s_2^{*1} - s_3^{*1} - s_4^{*1})}{2} & \frac{(-s_1^{*1} - s_2^{*1} + s_3^{*1} - s_4^{*1})}{2} & \frac{(s_1^{*1} + s_2^{*1} + s_3^{*1} - s_4^{*1})}{2} & \frac{(s_1^{*1} - s_2^{*1} + s_3^{*1} + s_4^{*1})}{2} \\ \frac{(s_1^1 + s_2^1 - s_3^1 + s_4^1)}{2} & \frac{(s_1^1 - s_2^1 - s_3^1 - s_4^1)}{2} & \frac{(-s_1^1 + s_2^1 - s_3^1 - s_4^1)}{2} & \frac{(s_1^1 + s_2^1 + s_3^1 - s_4^1)}{2} \end{bmatrix} \quad (5.23)$$

And

$$S_n^2 = \begin{bmatrix} \frac{(s_1^2 + s_2^2 + s_3^2 - s_4^2)}{2} & \frac{(s_1^2 - s_2^2 + s_3^2 + s_4^2)}{2} & \frac{(-s_1^2 + s_2^2 + s_3^2 + s_4^2)}{2} & \frac{(s_1^2 + s_2^2 - s_3^2 + s_4^2)}{2} \\ \frac{(-s_1^{*2} + s_2^{*2} - s_3^{*2} - s_4^{*2})}{2} & \frac{(s_1^{*2} + s_2^{*2} + s_3^{*2} - s_4^{*2})}{2} & \frac{(-s_1^{*2} - s_2^{*2} + s_3^{*2} - s_4^{*2})}{2} & \frac{(-s_1^{*2} + s_2^{*2} + s_3^{*2} + s_4^{*2})}{2} \\ \frac{(s_1^{*2} - s_2^{*2} - s_3^{*2} - s_4^{*2})}{2} & \frac{(-s_1^{*2} - s_2^{*2} + s_3^{*2} - s_4^{*2})}{2} & \frac{(s_1^{*2} + s_2^{*2} + s_3^{*2} - s_4^{*2})}{2} & \frac{(s_1^{*2} - s_2^{*2} + s_3^{*2} + s_4^{*2})}{2} \\ \frac{(s_1^2 + s_2^2 - s_3^2 + s_4^2)}{2} & \frac{(s_1^2 - s_2^2 - s_3^2 - s_4^2)}{2} & \frac{(-s_1^2 + s_2^2 - s_3^2 - s_4^2)}{2} & \frac{(s_1^2 + s_2^2 + s_3^2 - s_4^2)}{2} \end{bmatrix} \quad (5.24)$$

To examine the decoding error probability of S_n , we calculate its code-word distance matrix.

$$A(S_n^1, S_n^2) = (S_n^2 - S_n^1)^H \cdot (S_n^2 - S_n^1) \quad (5.25)$$

The expression for A is given by

$$A = 1/4 \begin{bmatrix} \alpha & 0 & 0 & \beta \\ 0 & \alpha & -\beta & 0 \\ 0 & -\beta & \alpha & 0 \\ \beta & 0 & 0 & \alpha \end{bmatrix} \quad (5.25)$$

Where

$$\alpha = 4[\Delta_1^2 + \Delta_2^2 + \Delta_3^2 + \Delta_4^2] \quad (5.26)$$

$$\beta = 4[\Delta_1^2 + \Delta_2^2 - \Delta_3^2 - \Delta_4^2] \quad (5.27)$$

Where $\Delta_i = (s_i^1 - s_i^2)$ represents possible symbol errors & s^1 and s^2 are the symbols of code words S^1 and S^2 respectively. The determinant of the code-word distance matrix A is given as:

$$\det(A) = (\alpha + \beta)^2 (\alpha - \beta)^2 \quad (5.28)$$

With simple algebraic manipulation we get

$$\begin{aligned}
&=16[(\Delta_1^2 + \Delta_2^2 + \Delta_3^2 + \Delta_4^2) + (\Delta_1^2 + \Delta_2^2 - \Delta_3^2 - \Delta_4^2)]^2 [(\Delta_1^2 + \Delta_2^2 + \Delta_3^2 + \Delta_4^2) - (\Delta_1^2 + \Delta_2^2 - \Delta_3^2 - \Delta_4^2)]^2 \\
&=16[(\Delta_1^2 + \Delta_4^2) + (\Delta_2^2 + \Delta_3^2) + (\Delta_1^2 - \Delta_4^2) + (\Delta_2^2 - \Delta_3^2)]^2 \\
&\quad [(\Delta_1^2 + \Delta_4^2) + (\Delta_2^2 + \Delta_3^2) - (\Delta_1^2 - \Delta_4^2) + (\Delta_2^2 - \Delta_3^2)]^2
\end{aligned}
\tag{5.29}$$

The minimum value of determinant occurs when $\Delta_2 = \pm\Delta_3$ (i.e. only half the symbols have error)

$$=16[(\Delta_1^2 + \Delta_4^2)^2 - (\Delta_1^2 - \Delta_4^2)^2] \tag{5.30}$$

As per determinant criteria the minimum determinant of the code-word distance matrix has to be large to obtain high coding gains. And as per rank criteria rank of the difference matrix has to be full in order to obtain full diversity. So to achieve full rank, the determinant obtained in equation above should not be zero and only then full diversity is achievable. Hence to achieve this some particular symbols are transmitted from rotated constellations in fig 5.6 with a particular angle of rotation and $\Delta_1 \neq \Delta_4$.

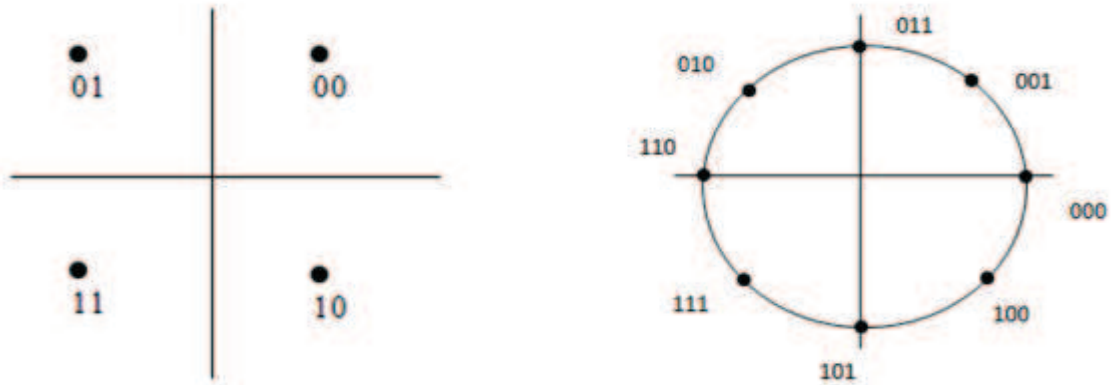


Fig.5.6: Constellation diagram for QPSK and 8-PSK

With s_4 and s_2 taken from the rotated constellation and putting the value of Δ we have,

$$\begin{aligned}
\det(A) &= 16[(s_1^1 - s_1^2)^2 + (s_4^1 - s_4^2)^2]^2 - [(\tilde{s}_1^1 - \tilde{s}_1^2)^2 - (\tilde{s}_4^1 - \tilde{s}_4^2)^2]^2 \\
&= 16 \left\{ [d_{\min}^2 + e^{2j\theta} d_{\min}^2]^2 - [d_{\min}^2 - e^{2j\theta} d_{\min}^2]^2 \right\}
\end{aligned}$$

$$= 16\{e^{4j\theta} d_{\min}^4\} \quad (5.31)$$

Now to calculate the optimum rotation angle that will maximize the code-word gain distance, we try to find the determinant for all possible values of theta and put it in the above equation. From all the values calculated, it is deduced that for theta equals to fifteen degrees ($\theta = 22.5^\circ \left(\frac{\pi}{8}\right)$), maximum CGD is obtained.

Now, trying to calculate the value of determinant by taking possible values of theta and then selecting the respective angle which will provide maximum gain. With simple calculations it was observed that for theta equals to 22.5 degrees the highest code word gain distance is obtained, which finally concludes that the constellation should be rotated optimally by an angle of 22.5 degree.

5.4 Simulation Results

The rotation mechanism is applied to the modified Jafarkhani proposed in [18] and better results are achieved as shown in Fig 5.7 it is apparent that Jafarkhani code performs better when rotation is applied to it. It gives better result by if we are using symbols from rotated constellations and, hence better performance is achieved with full diversity.

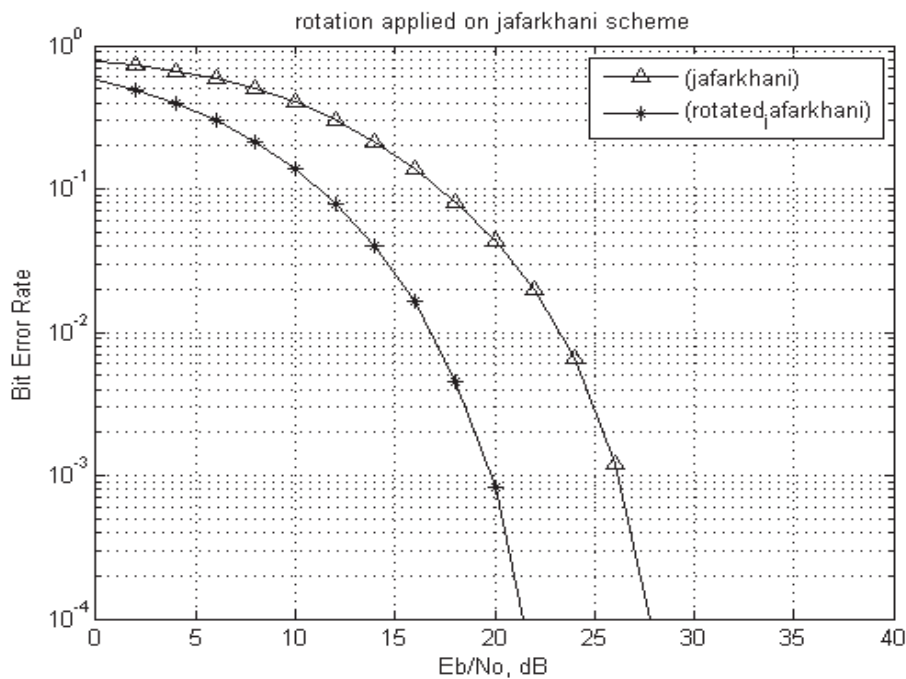


Fig 5.7: Performance of Jafarkhani scheme with rotated constellation. Angle of rotation applied is 22.5 degrees.

5.4.1 Comparison of Conventional Jafarkhani Scheme with and without TAS and Proposed rotated Jafarkhani Scheme

Transmit antenna shuffling is one of the feedback mechanism which is used for the improvement as well as the performance of any existing code. In this method, mapping of space time progressions of the code is done adaptively for the suitable number of transmit antennas which depends upon the various channel conditions taking place, such that the transmit diversity is enhanced with restricted amount of feedback. Here, transmit antenna shuffling is applied upon Jafarkhani scheme for four transmit antennas such that optimum pattern of antenna shuffling is obtained and finally selected to perk up the transmit diversity.

In order to achieve full transmit diversity a novel criteria of rotated constellations for different transmitted symbols is explored. This scheme outperforms transmit antenna shuffling (TAS) and also removes its disadvantage of dedicated feedback channel for CSI.

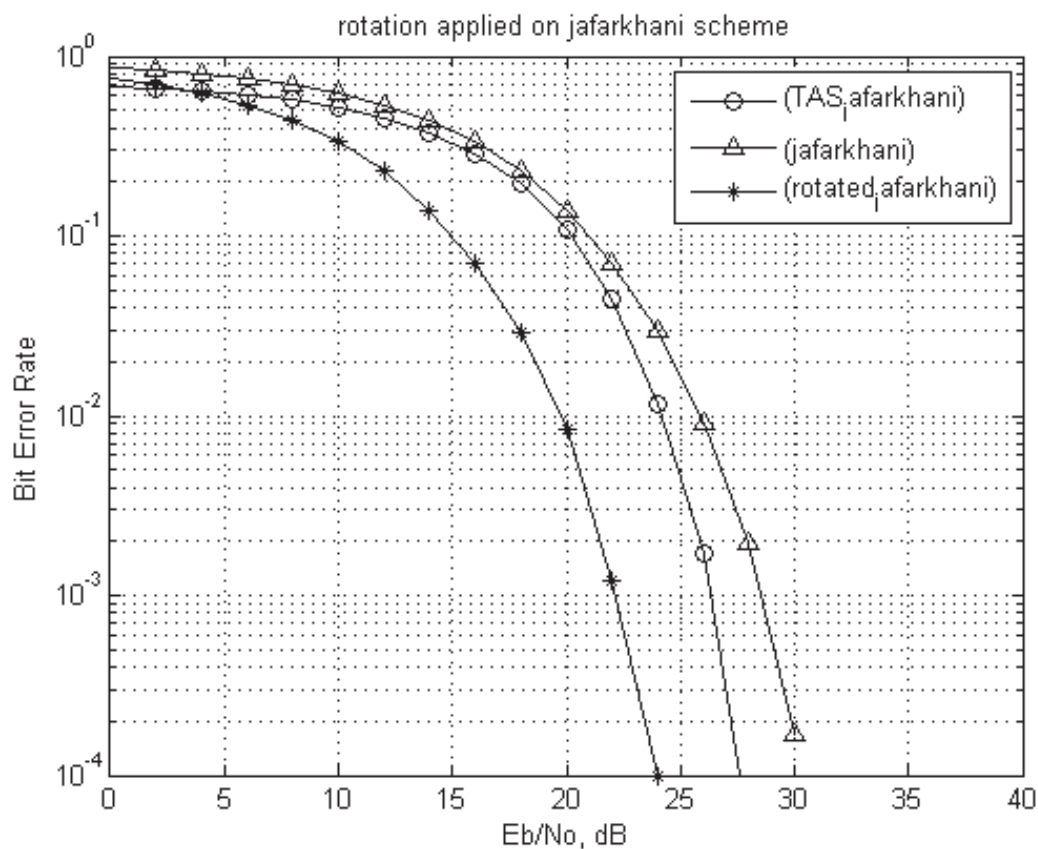


Fig 5.8: Shows comparison of non-rotated QO-STBC, with TAS scheme applied upon it, rotated QO-STBC.

The final code thus achieved is very powerful as it provides code rate of one, full diversity and pairwise decoding [25]. Finally, simulations show that bit error rate performance is significantly improved by gain of 3dB in comparison to TAS scheme.

In this work, we have generalized the scheme to Tirkkonen and ABBA code of eliminating the interference terms from detection matrix. The scheme proposed has considerably lower decoding complexity in comparison to maximum likelihood and zero-forcing-decoder. Here the new encoding matrix S_n is Quasi-Orthogonal in nature where as its equivalent channel matrix is Orthogonal in nature; causing simple linear decoding which in turn lowers the transmission delay.

Further to improve the BER performance and to obtain full diversity, the strategy of rotated QOSTBC is applied, where two particular symbols are taken from a rotated constellation of QPSK. It is clear that this proposed scheme offers lower bit error rate over the conventional one. So finally, with rotated constellation, proposed QO-STBC scheme offers minimum decoding complexity and also provide full diversity with code rate one. Simulation results illustrate that the employed scheme have enhanced the diversity gain with an amount equal to 3dB.

6.1 Conclusion

It is recognized that QO-STBCs have been derived from quasi-orthogonal designs, where the orthogonality is relaxed for providing higher code rate. Quasi Orthogonal-Space Time Block Codes permits a trade-off between both higher rate as well as decoding complexity contrasting in STBC where uncomplicated linear decoding is used.

On the whole two kinds of decoder are used to decode any symbol present in the transmission matrix of QO-STBC, one is called as maximum likelihood decoding and other one is zero forcing decoding. In maximum likelihood decoding, the decoder used, processes transmitted symbols as pairs instead of decoding a single symbol. To decode these pairs of transmitted symbols in QOSTBC, two decoders are used in parallel referred to as maximum likelihood detectors. The decision metric thus obtained in QO-STBC is the summation of two terms; and hence minimizing this decision metric is correspondent to minimize the two terms independently. Two maximum likelihood detectors used are either connected in sequence or in parallel. Consequently, decoding the pairs in case of QO-STBC is additionally complex than decoding single symbols in case of space–time block codes. This now results in higher decoding complexity on the receiver front. In particular, the decoding complexity always increases with the modulation intensity. Hence, in order to diminish this decision metric with the help of maximum likelihood technique, the receiver will compute the decision metric which is possible over all the symbols of a particular constellation or else modulation level and finally decides in favour of that particular constellation symbols which minimizes the decision metric. It is observed that as the size of the constellation enlarges, the receiver has to minimize the decision metric possible over large number of symbols. This would, consequently, increase the transmission delay when higher modulation schemes or else more number of antennas are employed. For lower modulation level, out of two decoders described maximum likelihood decoding is a superior option but practically we need to use higher modulation if we want to achieve higher data rate. On the other side, if we use zero forcing decoding there is no such problem of transmission delay as the modulation level is increased; hence it is used more frequently. But due to the practice of taking inverse of the matrix which is

used in this type of decoding, the procedure is having higher decoding complexity. Thus, there is no such need of searching new code for which this complexity in decoding is minimized with slightest transmission delay.

In this effort, a new QO-STBC has been proposed which is having much lower decoding complexity in comparison to the decoding complexity of above illustrated decoding schemes. A novel QO-STBC is also proposed with the help of symmetry property observed in the detection matrix of Jafarkhani code. This new encoding matrix S_n is Quasi-Orthogonal whereas its corresponding and equivalent virtual channel matrix H_n is Orthogonal in nature, such that simple linear decoding is achievable which also helps in reducing the transmission delay. Henceforth, decoding complexity of our proposed code is a lot lower than that of conventional QO-STBC without any loss in its BER performance.

In order to achieve full transmit diversity a novel criteria of rotated constellations for different transmitted symbols is explored. This scheme outperforms transmit antenna shuffling (TAS) and also removes its disadvantage of dedicated feedback channel for CSI. The final code thus achieved is very influential as it grants code rate one, full diversity as well as pairwise decoding. Finally, simulations show that bit error rate performance is significantly improved by gain of 3dB in comparison to TAS scheme.

6.2 Future Scope

Space time coding finds its applications in cellular communication as well as in wireless local area networks. Some of the work in space time coding focuses on explicitly by improving the performance of existing systems in terms of probability of error, signal to noise ratio, decoding complexity by employing extra transmit antennas and other research capitalizes on the promises of information theory for increasing the throughput. In particular, the construction of space-time coding schemes is to a large extent trade-off between the three conflicting goals of maintaining a simple decoding, maximizing the information rate, minimizing the error rate. This has motivated a lot of researchers including us to look for different schemes which could achieve all these desired results.

The newly proposed Quasi Orthogonal-STBC acts as a good candidate for future wireless communication systems. The proposed QO-STBCs provide high transmission rate with simple decoding algorithms and minimum transmission delay. Further improvements in code rate of QO-STBC, is another parameter for research by optimizing structure of QO-STBC. There is also chance to explore structure of QO-STBC using Clifford algebra, which is also an area of further research.

Singular value decomposition from linear algebra states that a rectangular matrix (let us say D) can be decomposed into product of three different matrices – orthogonal matrix U , a diagonal matrix S , and the transpose of orthogonal matrix V . Thus we write it as

$$D = USV^T \quad (a1)$$

Where $UU^T = I$ and $VV^T = I$, the columns of U are orthogonal Eigen vectors of DD^T and the columns of V are orthogonal Eigen vectors of $D^T D$, and S is a diagonal matrix which contains the square roots of Eigen values from V or U in descending order. From equation (6) and (10) we have the detection matrix as:

$$D = \begin{bmatrix} a & 0 & b & 0 \\ 0 & a & 0 & b \\ b & 0 & a & 0 \\ 0 & b & 0 & a \end{bmatrix} \quad (a2)$$

To find U , let's start with DD^T

$$\begin{aligned} DD^T &= \begin{bmatrix} a & 0 & b & 0 \\ 0 & a & 0 & b \\ b & 0 & a & 0 \\ 0 & b & 0 & a \end{bmatrix} \begin{bmatrix} a & 0 & b & 0 \\ 0 & a & 0 & b \\ b & 0 & a & 0 \\ 0 & b & 0 & a \end{bmatrix} \\ &= \begin{bmatrix} a^2 + b^2 & 0 & 2ab & 0 \\ 0 & a^2 + b^2 & 0 & 2ab \\ 2ab & 0 & a^2 + b^2 & 0 \\ 0 & 2ab & 0 & a^2 + b^2 \end{bmatrix} \end{aligned} \quad (a3)$$

Now to find the Eigen values and Eigen vectors of DD^T , Eigen vectors are defined by equation $A\vec{v} = \lambda\vec{v}$, applying it on DD^T

$$\begin{bmatrix} a^2 + b^2 & 0 & 2ab & 0 \\ 0 & a^2 + b^2 & 0 & 2ab \\ 2ab & 0 & a^2 + b^2 & 0 \\ 0 & 2ab & 0 & a^2 + b^2 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix} = \lambda \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix} \quad (a4)$$

To solve them, we put determinant of coefficient matrix to zero

$$\begin{bmatrix} a^2 + b^2 - \lambda & 0 & 2ab & 0 \\ 0 & a^2 + b^2 - \lambda & 0 & 2ab \\ 2ab & 0 & a^2 + b^2 - \lambda & 0 \\ 0 & 2ab & 0 & a^2 + b^2 - \lambda \end{bmatrix} = 0$$

Solving it we get,

$$\begin{aligned} (a^2 + b^2 - \lambda)x_1 + (2ab)x_3 &= 0 \\ (a^2 + b^2 - \lambda)x_2 + (2ab)x_4 &= 0 \\ (a^2 + b^2 - \lambda)x_3 + (2ab)x_1 &= 0 \\ (a^2 + b^2 - \lambda)x_4 + (2ab)x_2 &= 0 \end{aligned} \tag{a5}$$

$$\begin{aligned} (a^2 + b^2 - \lambda)[(a^2 + b^2 - \lambda)(a^2 + b^2 - \lambda)^2 + 2ab(-2ab(a^2 + b^2 - \lambda))] + \\ 2ab[-(a^2 + b^2 - \lambda)(2ab) + 2ab(2ab)^2] = 0 \end{aligned} \tag{a6}$$

This gives us the Eigen values as $\lambda = (a+b)^2$; $\lambda = (a+b)^2$; $\lambda = (a-b)^2$ and $\lambda = (a-b)^2$

Putting λ back to its original equations gives us the Eigen vectors as

For $\lambda = (a-b)^2$,

$$\begin{aligned} (a^2 + b^2 - (a-b)^2)x_1 + (2ab)x_3 &= 0 \\ (a^2 + b^2 - (a-b)^2)x_2 + (2ab)x_4 &= 0 \\ (a^2 + b^2 - (a-b)^2)x_3 + (2ab)x_1 &= 0 \\ (a^2 + b^2 - (a-b)^2)x_4 + (2ab)x_2 &= 0 \end{aligned} \tag{a7}$$

On solving these equations we get, $x_1 = -x_3$ and $x_2 = -x_4$. Thus the corresponding Eigen vectors for Eigen value $\lambda = (a-b)^2$ are $[1 \ -1 \ 1 \ -1]$ and $[1 \ 1 \ 1 \ 1]$.

For $\lambda = (a+b)^2$

$$\begin{aligned} (a^2 + b^2 - (a+b)^2)x_1 + (2ab)x_3 &= 0 \\ (a^2 + b^2 - (a+b)^2)x_2 + (2ab)x_4 &= 0 \\ (a^2 + b^2 - (a+b)^2)x_3 + (2ab)x_1 &= 0 \\ (a^2 + b^2 - (a+b)^2)x_4 + (2ab)x_2 &= 0 \end{aligned} \tag{a8}$$

On solving these equations we get, $x_1 = x_3$ and $x_2 = x_4$. And thus corresponding to Eigen value $\lambda = (a+b)^2$, Eigen vectors are $[-1 -1 1 1]$ and $[1 -1 -1 1]$.

These eigenvectors become column vectors in a matrix ordered by the size of its corresponding Eigen value. The eigenvectors for $\lambda = (a+b)^2$ are in column one and two, whereas for $\lambda = (a-b)^2$ are in column three and four.

$$\begin{bmatrix} -1 & 1 & 1 & 1 \\ -1 & -1 & -1 & 1 \\ 1 & -1 & 1 & 1 \\ 1 & 1 & -1 & 1 \end{bmatrix} \quad (\text{a9})$$

Finally on applying Gram Schmidt orthogonalization procedure to the column vector, the above matrix is being converted into an orthogonal matrix. Let's begin by applying

$$\text{For } \vec{v}_1 = [-1 -1 1 1]$$

$$\vec{u}_1 = \vec{v}_1 / |\vec{v}_1| = [-1 -1 1 1] / \sqrt{-1^2 + -1^2 + 1^2 + 1^2} = [-1 -1 1 1] / 2 = [-0.5 -0.5 0.5 0.5] \quad (\text{a10})$$

For

$$\vec{v}_2 = [1 -1 -1 1]$$

$$\vec{w}_2 = \vec{v}_2 - \vec{u}_1 \cdot \vec{v}_2 * \vec{u}_1$$

$$\vec{u}_2 = \vec{w}_2 / |\vec{w}_2| = [1 -1 -1 1] / \sqrt{1^2 + -1^2 + -1^2 + 1^2} = [1 -1 -1 1] / 2 = [0.5 -0.5 -0.5 0.5] \quad (\text{a11})$$

For

$$\vec{v}_3 = [1 -1 1 -1]$$

$$\vec{w}_3 = \vec{v}_3 - \vec{u}_1 \cdot \vec{v}_3 * \vec{u}_1 - \vec{u}_2 \cdot \vec{v}_3 * \vec{u}_2 \quad (\text{a12})$$

$$\vec{u}_3 = \vec{w}_3 / |\vec{w}_3| = [1 -1 1 -1] / \sqrt{1^2 + -1^2 + 1^2 + -1^2} = [1 -1 1 -1] / 2 = [0.5 -0.5 0.5 -0.5]$$

For

$$\vec{v}_4 = [1 1 1 1]$$

$$\vec{w}_4 = \vec{v}_4 - \vec{u}_1 \cdot \vec{v}_4 * \vec{u}_1 - \vec{u}_2 \cdot \vec{v}_4 * \vec{u}_2 - \vec{u}_3 \cdot \vec{v}_4 * \vec{u}_3 \quad (\text{a13})$$

$$\vec{u}_4 = \vec{w}_4 / |\vec{w}_4| = [1 1 1 1] / \sqrt{1^2 + 1^2 + 1^2 + 1^2} = [1 1 1 1] / 2 = [0.5 0.5 0.5 0.5]$$

All these \vec{u} 's result in an orthogonal matrix U given by:

$$\begin{bmatrix} -0.5 & 0.5 & 0.5 & 0.5 \\ -0.5 & -0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & 0.5 & -0.5 & 0.5 \end{bmatrix} \quad (\text{a14})$$

In order to find V, start with $D^T D$. As D is the symmetric matrix, so $D^T D = DD^T$. This means the value of V is calculated the same way as that of U. So V^T is given by

$$\begin{bmatrix} -0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & -0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & -0.5 \\ 0.5 & 0.5 & 0.5 & 0.5 \end{bmatrix} \quad (\text{a15})$$

For finding S square roots of the non-zero Eigen values are taken and then placed along the diagonal, putting largest one at first. So S is given by

$$\begin{bmatrix} a+b & 0 & 0 & 0 \\ 0 & a+b & 0 & 0 \\ 0 & 0 & a-b & 0 \\ 0 & 0 & 0 & a-b \end{bmatrix} \quad (\text{a16})$$

So $D = USV^T$ is given by

$$= \begin{bmatrix} -0.5 & 0.5 & 0.5 & 0.5 \\ -0.5 & -0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & 0.5 & -0.5 & 0.5 \end{bmatrix} \begin{bmatrix} a+b & 0 & 0 & 0 \\ 0 & a+b & 0 & 0 \\ 0 & 0 & a-b & 0 \\ 0 & 0 & 0 & a-b \end{bmatrix}$$

$$X \begin{bmatrix} -0.5 & -0.5 & 0.5 & 0.5 \\ 0.5 & -0.5 & -0.5 & 0.5 \\ 0.5 & -0.5 & 0.5 & -0.5 \\ 0.5 & 0.5 & 0.5 & 0.5 \end{bmatrix} \quad (\text{a17})$$

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List of Publications

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