SPACE TIME BLOCK CODE FOR MIMO-OFDM SYSTEM

A DISSERTATION

Submitted in partial fulfillment of the requirements for the award of the degree
of
MASTER OF TECHNOLOGY
in
ELECTRONICS AND COMMUNICATION ENGINEERING
(With Specialization in Communication Systems)

By

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CANDIDATE'S DECLARATION

I hereby declare that the work, which is presented in this dissertation report, entitled **"PERFORMANCE EVALUATION OF SPACE TIME BLOCK CODE FOR MIMO OFDM SYSTEMS WITH FREQUENCY OFFSET"**, being submitted in partial fulfillment of the requirements for the award of the degree of **MASTER OF TECHNOLOGY** with specialization in **COMMUNICATION SYSTEMS**, in the Department of Electronics and Computer Engineering, Indian Institute of Technology, Roorkee is an authentic record of my own work carried out from July 2008 to May 2009, under guidance and supervision of **Dr. ANSHUL TYAGI**, assistant professor, Department of electronics and Computer Engineering, Indian Institute of Technology, Roorkee.

The results embodied in this dissertation have not submitted for the award of any other Degree or Diploma.

Date 27/05/09.

Place: Roorkee

ASHOK KUMAR

CERTIFICATE

This is to certify that the statement made by the candidate is correct to the best of my knowledge and belief.

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ABSTRACT

Multiple-input-multiple-output (MIMO) antenna architecture has the ability to increase capacity and reliability of a wireless communication system. Orthogonal frequency division multiplexing (OFDM) is another popular technique in wireless communication which is famous for the efficient high speed transmission and robustness to frequency selective channels. Therefore, the integration of the two technologies probably has the potential to meet the ever growing demands of future communication systems. Because of aforementioned merits of these two techniques, this thesis was about MIMO and MIMO-OFDM system, and it addressed three different issues. Firstly it has investigated the performance of MIMO-STBC systems, Secondly, it has focused on performance of MIMO-OFDM system. In last part, BER comparison of space-time block coded OFDM system (STBC) was calculated for different frequency offset using simulation software. The results proved that the reliability of the wireless link increases as the number of transmits and received antenna increase.

OFDM is a special form of multicarrier modulation (MCM), where a single data stream is transmitted over a number of lower rate subcarriers. The basic principle of OFDM is to split a high-data-rate sequence into a number of low-rate sequences that are transmitted simultaneously over a number of subcarriers

OFDM is relatively higher sensitive to carrier frequency offset (CFO) errors, compared to a single carrier system. The frequency offset error is caused by the misalignment in carrier frequencies at the receiver due to fluctuations in receiver RF oscillators or a channel's Doppler frequency. This frequency offset can destroy the subcarrier orthogonality of the OFDM signal, introducing inter-carrier interference (ICI). The ICI results in severe degradation of the bit-error-rate (BER) performance of OFDM systems.

Since OFDM subcarriers are closely packed compared to the system bandwidth the amount of tolerable frequency offset is a small fraction of the OFDM bandwidth. CFO introduces both attenuation/rotation of the useful signal and ICI which destroys the orthogonality of demodulated subcarriers. ICI and attenuation/phase rotation effects if not compensated, increase the system error rate and reduce the overall throughput.

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Table of Abbreviations

MIMO	Multiple Input Multiple Output
OFDM	Orthogonal Frequency Division Multiplexing
WLAN	Wireless Local Area Network
OSTBC	Orthogonal Space Time Block Code
FFT	Fast Fourier Transform
STBC	Space Time Block Code
BER	Bit Error Rate
CP	Cyclic Prefix
SISO	Single Input Single Output
STC	Space Time Code
SNR	Signal to Noise Ratio
EGC	Equal Gain Combining
MRC	Maximum Ratio Combining
STTC	Space Time Trellis Code
CSI	Channel State Information
PSK	Phase Shift Keying
ISI	Inter Symbol Interference
MCM	Multi Carrier Modulation
ICI	Inter Carrier Interference
IFFT	Inverse Fast Fourier Transform
DFT	Discrete Fourier Transform
IDFT	Inverse Discrete Fourier Transform
EQ	Equalizer
FDM	Frequency Division Multiplexing
SCM	Single Carrier Multiplexing
SFC	Space Frequency Code
STFC	Space Time Frequency Code

Contents

CHAPTER 1

INTRODUCTION

1.1 Introduction

Physical limitations of the wireless medium create a technical challenge for reliable wireless communication. Techniques that improve spectral efficiency and overcome various channel impairments such as signal fading and interference have made an enormous contribution to the growth of wireless communications. Moreover, the need for high-speed wireless Internet has led to the demand for technologies delivering higher capacities and link reliability than achieved by current systems. Multiple -input multiple-output (MIMO) based communication systems are capable accomplishing these objectives.

Multiple input multiple output (MIMO) systems take advantage of spatial diversity obtained through the spatially separated antennas in a dense multipath scattering environment [1]. Spatial diversity can increase the gain diversity consequently increase the reliability of the wireless link. Theoretical studies indicate that the capacity of MIMO systems grows linearly with the number of transmit antennas used. Many recent works have focused on exploiting the added spatial dimension to increase capacity. In particular, the revolutionary vertical Bell Laboratory Layered Space Time architecture proposed by Foschini achieved the theoretical capacity limits of the MIMO architecture [4].

The multiple antennas configuration exploits the multipath effect to accomplish the additional spatial diversity. However, the multipath effect also causes the negative effect of frequency selectivity of the channel. Orthogonal frequency division multiplexing (OFDM) is a promising multi-carrier modulation scheme that shows high spectral efficiency and robustness to frequency selective channels. In OFDM, a frequency-selective channel is divided into a number of parallel frequency-flat sub channels, thereby reducing the receiver signal processing of the system. The combination of OFDM and MIMO is a promising technique to achieve high bandwidth efficiencies and system performance [2].

1.2 Motivation

MIMO-OFDM has the potential to meet the increasing high speed and reliability demands of the future. However, this technology to truly succeed in commercial deployment there are still several technical obstacles that must be tackled. A major impediment in MIMO-OFDM is the complicated receiver signal processing. The simultaneous emission of the signals from the multiple transmit antennas increases the mutual interference imposed on the signals, therefore, much more complex detection schemes are required to extract the transmitted signals. For example, the complexity of a maximum likelihood detector increases exponentially with the number of transmit antennas. Spatial equalizers and space time coding has been proposed to simplify the detection for MIMO-OFDM systems. Note, coherent detection requires knowledge of the channel; therefore, accurate channel estimation is crucial in realizing the full potential of MIMO-OFDM. Channel estimation for OFDM has been well researched in literature. The extension of the results to MIMO-OFDM channel estimation is substantially more complicated. In a MIMO system, multiple channels have to be estimated simultaneously. The increased number of channel unknowns significantly increases the computational complexity of the channel estimation algorithm.

1.3 Objectives

The dissertation has three objectives. The first one was performance analysis of space-time orthogonal block coded MIMO system. For doing so, Orthogonal Space Time Block Coded Transmission was considered in the system. The block code was 2 by 2 complex Alamouti code. The number of transmit and receive antenna was assumed N_T and N_R respectively. Second objective of the project was to assess the performance of MIMO system and MIMO-OFDM with perfect channel state information. In this part comparison analysis in the system with different number of receiver antenna and with different FFT length has been made. Third objective of the project was analysis of performance of STBC MIMO-OFDM with a known frequency offset. The first two objectives were supporting this part technically. For the last objective, structure of the system was assumed MIMO-OFDM with STBC transmission. OFDM system makes the MIMO structure robust in frequency selective channel.

1.4 Scope of works

The dissertation was based on theoretical results and software modeling. MIMO system had STBC structure with $2 \times N_R$ Antenna constellation. Channel was assumed Rayleigh flat fading with additive white Gaussian noise with zero mean and variance one. Simulation has been done at baseband. The performance of the system investigated using the BER parameter. In first phase of dissertation the literature survey of the Basic space time code, STBC for OFDM and the Effect of carrier frequency offset to the performance of STBC OFDM. Second phase of the dissertation simulation has been done for three different systems with different conditions. The first one was MIMO system with flat fading channel condition. The performance for STBC and basic Alamouti scheme was simulated. Second part of phase two, the system was assumed MIMO-OFDM system and simulation for the performance of STBC OFDM system. For this part antenna constellation was 2 by N_R, Channel was frequency selective with additive white Gaussian noise. And last part of this phase was to simulate the performance for the STBC OFDM systems with carrier frequency offset. Also Channel was assumed quasistatic within two transmission block. Number of sub-carrier and CP (cyclic prefix) has been assumed to 64, 128, 256 and 16, 32, 64 respectively.

1.5 Thesis outline

In chapter 2, the background theoretical information about transmit and receive diversity is given. The different diversity schemes are like time diversity, frequency

CHAPTER 2

DIVERSITY

2.1 Diversity Techniques

In wireless mobile communications, diversity techniques are widely used to reduce the effects of multipath fading and improve the reliability of transmission without increasing the transmitted power or sacrificing the bandwidth [3]. The diversity technique requires multiple replicas of the transmitted signals at the receiver, all carrying the same information but with small correlation in fading statistics. The basic idea of diversity is that, if two or more independent samples of a signal are taken, these samples will fade in an uncorrelated manner, e.g., some samples are severely faded while others are less attenuated. This means that the probability of all the samples being simultaneously below a given level is much lower than the probability of any individual sample being below that level. Thus, a proper combination of the various samples results in greatly reduced severity of fading, and correspondingly, improved reliability of transmission. In most wireless communication systems a number of diversity methods are used in order to get the required performance. According to the domain where diversity is introduced, diversity techniques are classified into time, frequency and space diversity.

2.1.1 Time Diversity

Time diversity can be achieved by transmitting identical messages in different time slots, which results in uncorrelated fading signals at the receiver. The required time separation is at least the coherence time of the channel, or the reciprocal of the fading rate $1/f_d = c/vf_c$. The coherence time is a statistical measure of the period of time over which the channel fading process is correlated. Error control coding is regularly used in digital communication systems to provide a coding gain relative to uncoded systems. In mobile communications, error control coding is combined with interleaving to achieve time diversity. In this case, the replicas of the transmitted signals are usually provided to the receiver in the form of redundancy in the time domain introduced by error control coding [4]. The time separation between the replicas of the transmitted signals is provided by time interleaving to obtain independent fades at the input of the decoder. Since time interleaving results in decoding delays, this technique is usually effective for fast fading environments where the coherence time of the channel is small. For slow fading channels, a large interleaver can lead to a significant delay which is intolerable for delay sensitive applications such as voice transmission. This constraint rules out time diversity for some mobile radio systems. For example, when a mobile radio station is stationary, time diversity cannot help to reduce fades. One of the drawbacks of the scheme is that due to the redundancy introduced in the time domain, there is a loss in bandwidth efficiency [3].

2.1.2 Frequency Diversity

In frequency diversity, a number of different frequencies are used to transmit the same message. The frequencies need to be separated enough to ensure independent fading associated with each frequency. The frequency separation of the order of several times the channel coherence bandwidth will guarantee that the fading statistics for different frequencies are essentially uncorrelated. The coherence bandwidth is different for different propagation environments. In mobile communications, the replicas of the transmitted signals are usually provided to the receiver in the form of redundancy in the frequency domain introduced by spread spectrum such as direct sequence spread spectrum, multicarrier modulation and frequency hopping. Spread spectrum techniques are effective when the coherence bandwidth of the channel is small. However, when the coherence bandwidth of the channel is larger than the spreading bandwidth, the multipath delay spread will be small relative to the symbol period. In this case, spread spectrum is ineffective to provide frequency diversity. Like time diversity, frequency diversity induces a loss in bandwidth efficiency due to a redundancy introduced in the frequency domain.

2.1.3 Space Diversity

Space diversity has been a popular technique in wireless microwave communications. Space diversity is also called antenna diversity. It is typically implemented using multiple antennas or antenna arrays arranged together in space for transmission and/or reception. The multiple antennas are separated physically by a proper distance so that the individual signals are uncorrelated. The separation requirements vary with antenna height, propagation environment and frequency. Typically a separation of a few wavelengths is enough to obtain uncorrelated signals. In space diversity, the replicas of the transmitted signals are usually provided to the receiver in the form of redundancy in the space domain. Unlike time and frequency diversity, space diversity does not induce any loss in bandwidth efficiency. This property is very attractive for future high data rate wireless communications. Polarization diversity and angle diversity are two examples of space diversity. In polarization diversity, horizontal and vertical polarization signals are transmitted by two different polarized antennas and received by two different polarized antennas. Different polarizations ensure that the two signals are uncorrelated without having to place the two antennas far apart [5]. Angle diversity is usually applied for transmissions with carrier frequency larger than 10 GHz. In this case, as the transmitted signals are highly scattered in space, the received signals from different directions are independent to each other. Thus, two or more directional antennas can be pointed in different directions at the receiver site to provide uncorrelated replicas of the transmitted signals [4].

Depending on whether multiple antennas are used for transmission or reception, we can classify space diversity into two categories: receive diversity and transmit diversity [6]. In receive diversity, multiple antennas are used at the receiver site to pick up independent copies of the transmit signals. The replicas of the transmitted signals are properly combined to increase the overall received SNR and mitigate multipath fading. In transmit diversity, multiple antennas are deployed at the

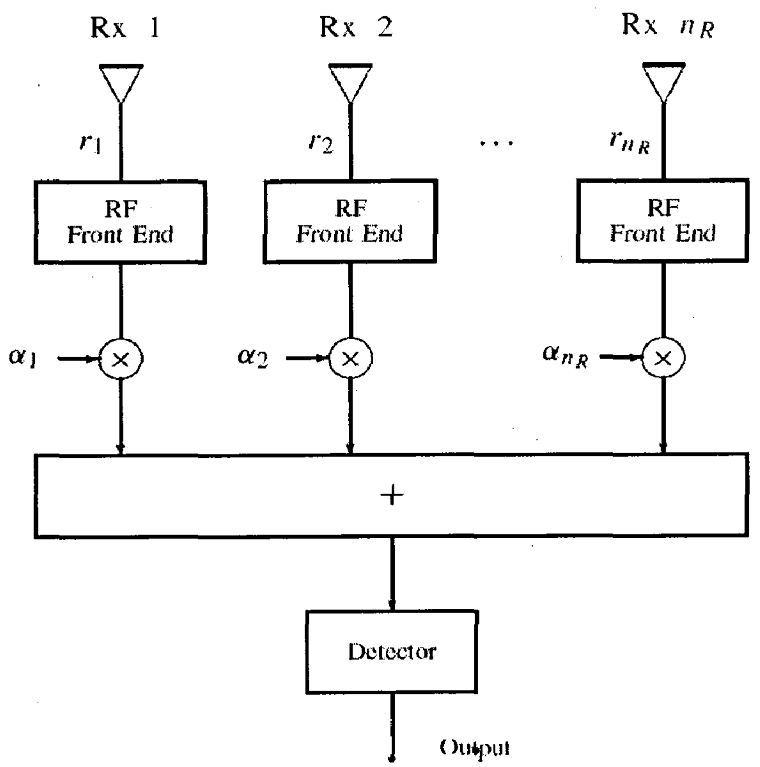


Fig.: - 2.3 Maximal Ratio Combining

where r_i is the received signal at receive antenna i, and α_i is the weighting factor for receive antenna i. In maximum ratio combining, the weighting factor of each receive antenna is chosen to be in proportion to its own signal voltage to noise power ratio. Let A_i and φ_i be the amplitude and phase of the received signal r_i, respectively. Assuming that each receive antenna has the same average noise power, the weighting factor r_i can be represented as

$$\alpha_i = A_i e^{-j\phi_i}$$

This method is called optimum combining since it can maximize the output SNR. It is shown that the maximum output SNR is equal to the sum of the instantaneous SNRs of the individual signals [3].

In this scheme, each individual signal must be co-phased, weighted with its corresponding amplitude and then summed. This scheme requires the knowledge of channel fading amplitude and signal phases. So, it can be used in conjunction with coherent detection, but it is not practical for noncoherent detection.

2.2.4 Equal Gain Combining

Equal gain combining is a suboptimal but simple linear combining method. It does not require estimation of the fading amplitude for each individual branch. Instead, the receiver sets the amplitudes of the weighting factors to be unity.

$$\alpha_i = A_i e^{-j\emptyset_i}$$

In this way all the received signals are co-phased and then added together with equal gain. The performance of equal-gain combining is only marginally inferior to maximum ratio combining. The implementation complexity for equal-gain combining is significantly less than the maximum ratio combining.

CHAPTER 3

SPACE-TIME CODED MIMO SYSTEMS

3.1 Introduction

In This chapter the concept of MIMO system is briefly explained. STBC structure model generally has $N_T \times N_R$ antenna configuration. However, in the simulation 2 × N configuration was considered. This configuration is similar to Alamouti scheme, the different is the number of receive antenna. In Alamouti scheme maximum number of receive antenna has been investigated was two, but in this dissertation the number of receive antenna can be increased by N_R. Due to use of Alamouti schemes in the dissertation, the principle of Alamouti transmission is given from block code till detection part. One of the interesting classes of space-time coding proposed by [7] is MIMO trellis coding which is not the subject of the dissertation. Spatial multiplexing or Layered space-time coding is another subcategory of MIMO system. Brief introduction about these two transmission technique can be find at the end of this chapter.

3.2 Basic principle of MIMO systems

Multiple-input-multiple-output (MIMO) communication systems use multiple antennas at both the transmitter and the receiver fig. 3.1. Under rich multipath environments with independent multipath fading between each transmit and receive antenna pair, MIMO wireless communications systems achieve significant capacity gains over conventional single antenna systems by exploiting the plurality of modes present in the matrix channel within the same time frequency slot. Moreover MIMO systems offer significant diversity advantage over traditional wireless communication systems by exploiting both transmit and receive diversity by employing various space-time coding schemes. These have led to MIMO being regarded as one of the most promising emerging wireless technologies.

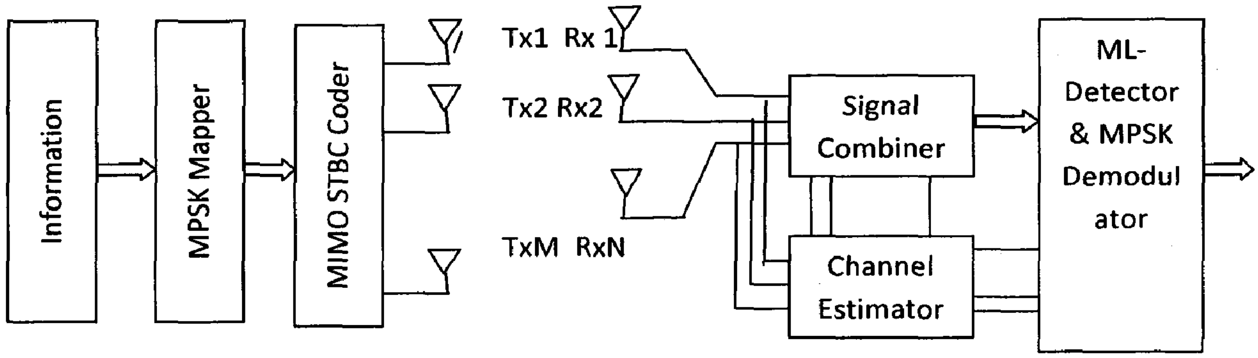


Figure – 3.1 Simple MIMO System Model

These two feature is related to two different class of MIMO namely space-time coding and spatial multiplexing. In summary first class can improve the link reliability and second class enhance the transmission rate of the wireless systems.

We consider a MIMO system with N_T transmit and N_R receive antennas. The transmitted signal is represented by an $N_T \times 1$ column matrix **s**. The total transmitted power is constrained to P, regardless of the number of transmit antennas N_T .It is assume that the signals transmitted from individual antenna elements have equal powers of P/N_T.

The transmitted signal bandwidth is narrow enough, so its frequency response can be considered as flat. The noise at the receiver is described by an $N_R \times 1$ column matrix, denoted by **n**. Its components are statistically independent complex zero-mean Gaussian variables. Receive antennas have identical noise powers of σ^2.The received signal is represented by an $N_R \times 1$ column matrix, denoted by **r**, where each complex component refers to a receive antenna.

The channel is described by an $N_T \times N_R$ complex matrix, denoted by **H**. It is assumed that the channel matrix is known to the receiver, which can be achieved, for example, by transmitting a training preamble. On the other hand, in most situations it is assumed that the channel parameters are not known at the transmitter. Matrix H has got several scenarios in below.

1. Matrix **H** is deterministic. 2. Matrix **H** is random, fast fading. Channel matrix entries change randomly at the beginning of each symbol interval T and are kept constant during one symbol interval. Such channel is called a fast fading channel. 3. Matrix **H** is random, block fading. Its entries change randomly and are kept constant during a fixed number of symbol intervals, much shorter than a transmission block. Such channel is referred to as a block fading channel. 4. Matrix **H** is random, quasi-static. The matrix entries are random but are fixed at the start of a transmission block and kept constant during a transmission block, which means that the symbol duration is small compared to the channel coherence time. This channel is known as a slow or quasi-static fading channel.

By using the linear model the received signal can be represented as Equation (3.1). In the matrix form receive vector can be represented as Equation (3.2).

$$r_t^j = \sum_{i=1}^{N_R} h^{ji}(t).s^t(t) + v^j(t) \qquad j = 1,2,\dots.N_R \tag{3.1}$$

$$\begin{pmatrix} r_t^1 \\ r_t^2 \\ \vdots \\ r_t^{N_R} \end{pmatrix} = \begin{pmatrix} h^{11}(t) & h^{11}(t) & \dots & h^{11}(t) \\ h^{11}(t) & h^{11}(t) & \dots & h^{11}(t) \\ \vdots & \vdots & \vdots & \vdots \\ h^{11}(t) & h^{11}(t) & \dots & h^{11}(t) \end{pmatrix} \times \begin{pmatrix} s_t^1 \\ s_t^2 \\ \vdots \\ s_t^{N_T} \end{pmatrix} + \begin{pmatrix} v_t^1 \\ v_t^2 \\ \vdots \\ v_t^{N_R} \end{pmatrix} \tag{3.2}$$

$s_t = [s_t^1, \ s_t^2, \ \cdots \ s_t^{N_T}]^T$denotes the transmit symbol vector with equally distributed transmit power across the transmit power. Here, the superscript T is transposition [8].

3.3 Space-time coding

Space-time codes refer to the set of coding schemes that allow for the adjusting and optimization of joint encoding across multiple transmit antennas in order to maximize the reliability of a wireless link. Space-time coding schemes exploit spatial diversity in order to provide coding and diversity gains over an uncoded wireless link

[7]. The basics of space-time coding are given using Alamouti schemes. However, technically there are various approaches in coding algorithm and structure, including Space Time Block Codes (STBC), Space Time Trellis Codes (STTC).

STBC for the first time was proposed by Alamouti at 1997. This technique has found a lot of popularity, and a lot of researcher start working at this area of wireless communication. The second outstanding researcher in this area is Vahid Tharokh who proposed the Theory of space-time coding at 1999. STTC scheme, based on trellis coding has been proposed by Tharokh and his group. These coding schemes provide any desirable configuration for complex and real transmission block with optimum code rate [9].

3.4 Alamouti scheme

The encoder for Alamouti schemes can be seen in fig. 3.2. This scheme with two transmit antennas and two receive antenna is interpreted here. In general case, we may use N_R receive antennas. In Alamouti encoding scheme [9], during any given transmission period two signals are transmitted simultaneously from two transmit antennas.

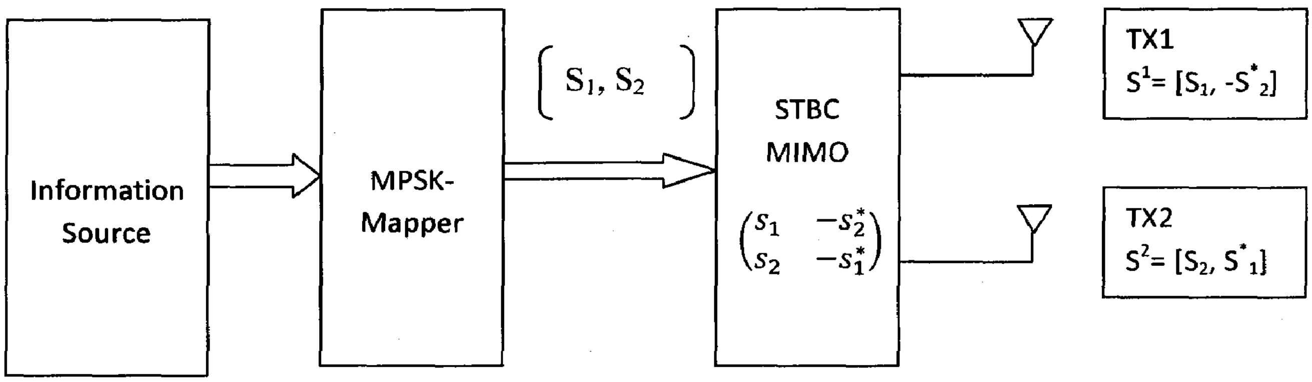


Fig.: - 3.2 Alamouti Transmitter

The encoder takes two modulated symbols s_1 and s_2 at a time. The transmit matrix S is given by Equation (3.3)

$$S = \begin{pmatrix} s_1 & -s_2^* \\ s_2 & s_1^* \end{pmatrix} \quad (3.3)$$

Where s* is complex conjugate of s. During the first transmission period, two signals s_1 and s_2 are transmitted simultaneously from antenna one and antenna two, respectively. In the second transmission period, signal $-s_2^*$ is transmitted from transmit antenna one and signal s_1^* from transmit antenna two. It is clear that the encoding is done in both space and time domain. The transmit sequence from antennas one and two is showed by $\mathbf{S}^1$ and $\mathbf{S}^2$, respectively.

$$S^1 = [s_1, \quad -s_2^*] \quad (3.4)$$

$$S^2 = [s_2, \quad s_1^*] \quad (3.5)$$

The key feature of the Alamouti scheme is given in Equation (3.6).

$$S.S^H = (|S_1|^2 + |S_2|^2)I_2 \quad (3.6)$$

I_2 is the 2×2 identity matrix.

The block diagram of the receiver for the Alamouti scheme is shown in Fig. (3.3). The fading channel coefficients from the first and second transmit antennas to the receive antenna one is denoted by $h_0(t)$ and $h_1(t)$ while to the receive antenna two is denoted by $h_2(t)$ and $h_3(t)$ respectively. Assuming that the fading coefficients are constant across two consecutive symbol transmission periods, means that channel is quasi static within transmission block.

Hence following equation (3.7), (3.8), (3.9) and (3.10) for every single channel coefficients at time interval T can be written.

$$h_0(t) = h_0(t+T) = h0 = |h0|e^{j\theta 0} \quad (3.7)$$

$$h_1(t) = h_1(t+T) = h1 = |h1|e^{j\theta 1} \quad (3.8)$$

$$h_2(t) = h_2(t+T) = h2 = |h2|e^{j\theta 2} \quad (3.9)$$

$$h_3(t) = h_3(t+T) = h3 = |h3|e^{j\theta 3} \quad (3.10)$$

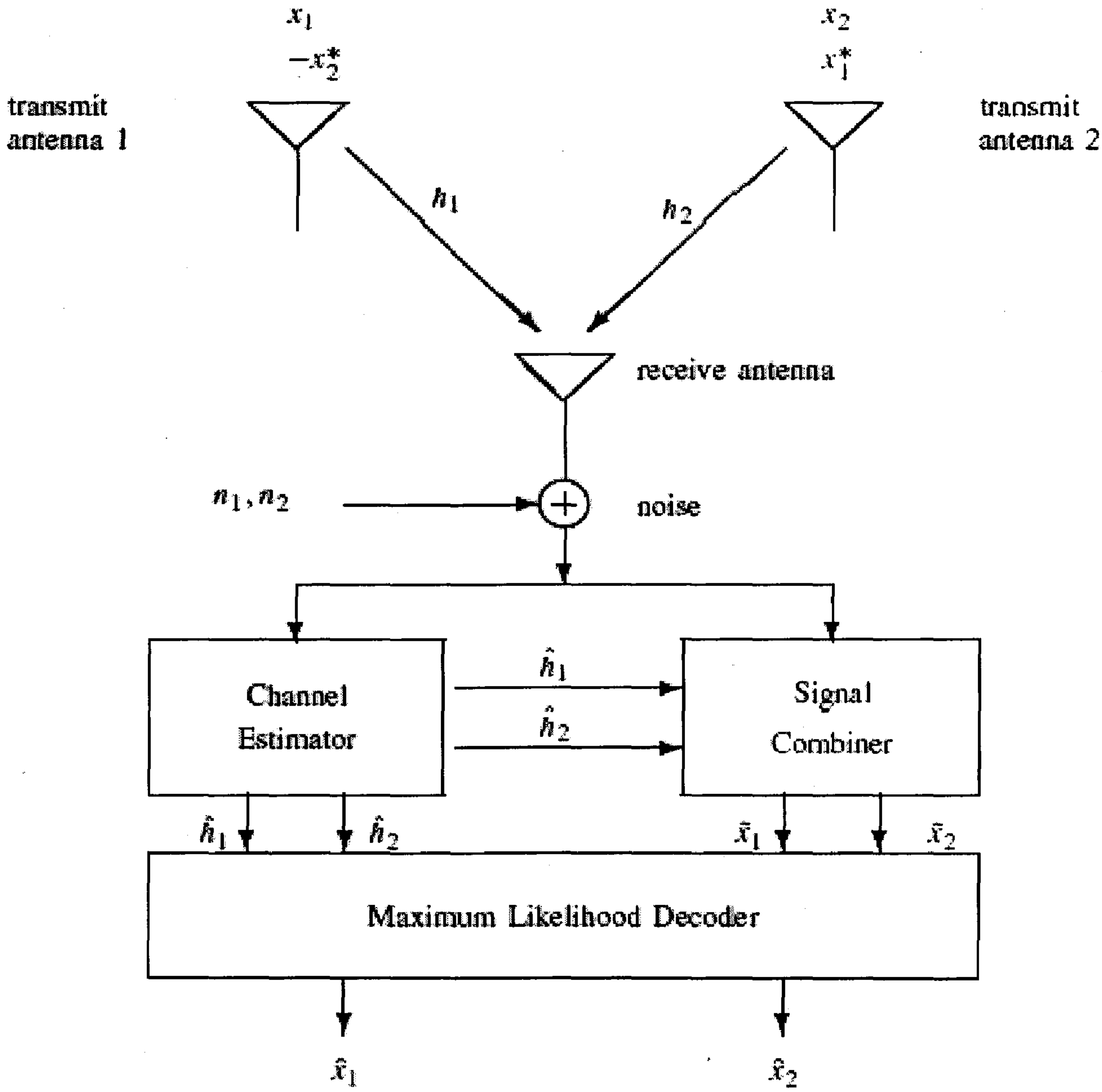


Fig.: -3.3 Receiver for Alamouti scheme

Where $|h_i|$ and θ_i, are the amplitude gain and phase shift for the path, and T is the symbol duration. At the receive antenna one, the received signals over two consecutive symbol periods, denoted by r_0 and r_1, at time t and t + T, respectively, while received signal for received antenna two, denoted by r_2 and r_3. Received signal by antenna one and two can be expressed by equation (3.11), (3.12), (3.13) and (3.14).

$$r_0 = h_0 s_0 + h_1 s_1 + n_0 \tag{3.11}$$

$$r_1 = -h_0 s_1^* + h_1 s_0^* + n_1 \tag{3.12}$$

$$r_2 = h_2 s_0 + h_3 s_1 + n_2 \tag{3.13}$$

$$r_3 = -h_2 s_1^* + h_3 s_0^* + n_3 \tag{3.14}$$

n_0, n_1, n_2, and n_3 are independent complex variables representing additive white Gaussian noise with zero mean and one-sided power spectral density N_0.

3.4.1 Maximum Ratio Combining and Decoding

If the channel fading attenuations hi can be perfectly recovered at the receiver, the receiver will use them as the channel state information (CSI) in the decoder. A combiner forms the following combined signals represented in equation (3.15) and (3.16).

$$\tilde{s}_1 = h_1^* r_0 - h_0 r_1^* + h_3^* r_2 - h_2 r_3^* \tag{3.15}$$

$$\tilde{s}_0 = h_0^* r_0 + h_1 r_1^* + h_2^* r_2 + h_3 r_3^* \tag{3.16}$$

Substituting r_i from (3.11)-(3.14), in equation (3.15) and (3.16) the combined signals can be written as equation (3.16) and (3.17)

$$\tilde{s}_0 = (|h1|^2 + |h2|^2 + |\ h3|^2 + |h4|^2)s_0 + h_0^* n_0 + h_1 n_1^* + h_2^* n_2 + h_3 n_3^* \tag{3.16}$$

$$\tilde{s}_1 = (|h0|^2 + |h1|^2 + |\ h2|^2 + |h3|^2)s_0 + h_1^* n_0 - h_0 n_1^* + h_3^* h_2 - h_2 n_2^* \tag{3.17}$$

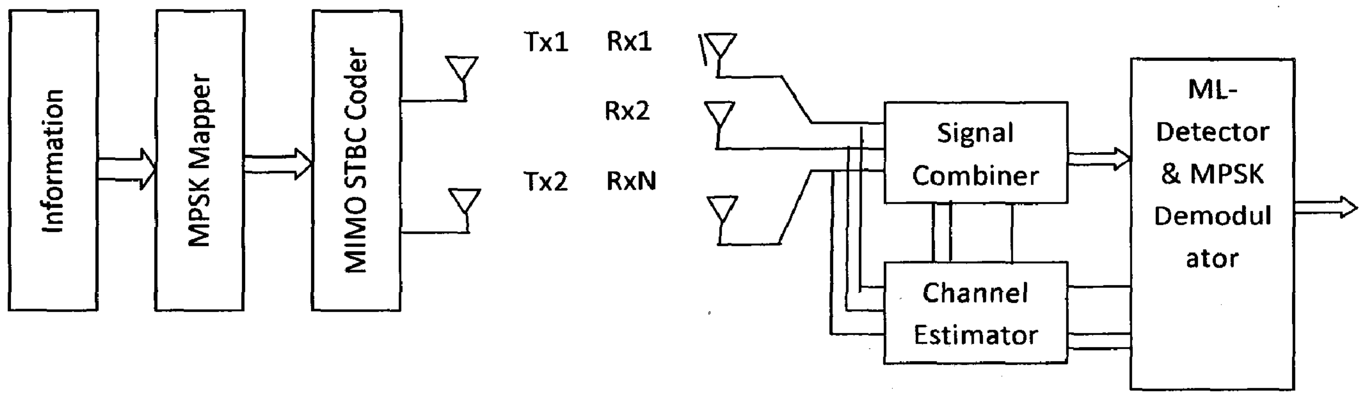


Fig.: - 3.4 Alamouti scheme model for $2\times N_R$ configuration

Then by applying maximum likelihood detector s_0 and s_1 can be estimated. In the similar way for 2 by 4 system following matrix form in equation (3.20) for the received signal in time interval t1 and t2 can be written

$$\begin{pmatrix} r_0 & r_1 \\ r_2 & r_3 \\ r_4 & r_5 \\ r_6 & r_7 \end{pmatrix} = \begin{pmatrix} h_0 & h_1 \\ h_2 & h_3 \\ h_4 & h_5 \\ h_6 & h_7 \end{pmatrix} \times \begin{pmatrix} s_0 & -s_1^* \\ s_1 & s_0^* \end{pmatrix} + \begin{pmatrix} n_0 & n_1 \\ n_2 & n_3 \\ n_4 & n_5 \\ n_4 & n_7 \end{pmatrix} \tag{3.20}$$

For the combiner, following simple matrix form in equation (3.21) can be defined

$$\begin{pmatrix} \bar{s}_0 \\ \bar{s}_1 \end{pmatrix} = \begin{pmatrix} h_0^* & h_1 & h_2^* & h_3 & h_4^* & h_3 & h_6^* & h_7 \\ h_1^* & -h_0 & h_3^* & -h_2 & h_5^* & -h_4 & h_7^* & -h_6 \end{pmatrix} \begin{pmatrix} r_0 \\ r_1^* \\ r_2 \\ r_3^* \\ r_4 \\ r_5^* \\ r_6 \\ r_7^* \end{pmatrix} \tag{3.21}$$

Then by applying maximum likelihood detector s_0 and s_1 can be estimated. In the similar way for 2 by N_R system following matrix form in equation (3.22) for the received signal in time interval t1 and t2 can be written.

$$\begin{pmatrix} r_0 & r_1 \\ r_2 & r_3 \\ \vdots & \vdots \\ r_{2m-2} & r_{2m-1} \end{pmatrix} = \begin{pmatrix} h_0 & h_1 \\ h_2 & h_3 \\ \vdots & \vdots \\ h_{2m-2} & h_{2m-1} \end{pmatrix} \times \begin{pmatrix} s_0 & -s_1^* \\ s_1 & s_0^* \end{pmatrix}$$

$$+ \begin{pmatrix} n_0 & n_1 \\ n_2 & n_3 \\ \vdots & \vdots \\ n_{2m-2} & n_{2m-1} \end{pmatrix} \qquad (3.22)$$

For the combiner for 2 by N_R antenna configuration, following simple matrix form in equation (3.23) can be written

$$\begin{pmatrix} \bar{s}_0 \\ \bar{s}_1 \end{pmatrix} = \begin{pmatrix} h_0^* & h_1 & h_2^* & h_3 & \cdots\cdots & h_{2m-2}^* & h_{2m-1} \\ h_1^* & -h_0 & h_3^* & -h_2 & \cdots\cdots & h_{2m-1}^* & -h_{2m-2} \end{pmatrix} \begin{pmatrix} r_0 \\ r_1^* \\ r_2 \\ r_3^* \\ \vdots \\ r_{2m-2}^* \\ r_{2m-1}^* \end{pmatrix} \qquad (3.23)$$

And finally in the same way symbols s_0 and s_1 can be detected using maximum likelihood detector.

3.5 Space-Time Block Codes

The Alamouti scheme brought in a revolution of sorts in multi antenna systems by providing full diversity of two without CSI at the transmitter and a very simple maximum likelihood decoding system at the receiver. Maximum likelihood decoders provide full diversity gain of N_R at the receiver. Hence, such a system provides a guaranteed overall diversity gain of $2N_R$, without CSI at the transmitter. This is achieved by the key feature of orthogonality between the sequences generated by the two transmit antennas. Due to these reasons, the scheme was generalized to an arbitrary number of transmit antennas by applying the theory of orthogonal designs. The generalized schemes are referred to as space-time block codes (STBCs) [7]. These codes can achieve the full transmit diversity of N_TN_R , while allowing a very simple maximum likelihood decoding algorithm, based only on linear processing of received signals [10].

Let N_T represent the number of transmit antennas and p represent the number of time periods for transmission of one block of coded symbols. Let us also assume that the signal constellation consists of 2m points. Then each encoding operation maps a block of km information bits into the signal constellation to select k modulated signals s_1, s_2,……, s_k, where each group of m bits selects a constellation signal. These k modulated signals are then encoded in a space-time block encoder to generate N_T parallel signal sequences of length p, as shown in Figure 3.5. This gives rise to a transmission matrix S of size $N_T \times p$. These sequences are transmitted through N_T transmit antennas simultaneously in p time periods. Therefore, the number of symbols the encoder takes as its input in each encoding operation is k. The number of transmission periods required to transmit the entire S matrix is p. The rate of the space-time block code is defined as the ratio between the number of symbols the encoder takes as its input and the number of space-time coded symbols transmitted from each antenna. It is given by

$$R = \frac{k}{p} \tag{3.24}$$

The spectral efficiency of the space-time block code is given by

$$\eta = \frac{r_b}{B} = \frac{r_s mR}{r_s} = \frac{km}{p} \frac{\frac{bits}{s}}{Hz} \tag{3.25}$$

where r_b and r_s are the bit and symbol rate, respectively, and B is the bandwidth.

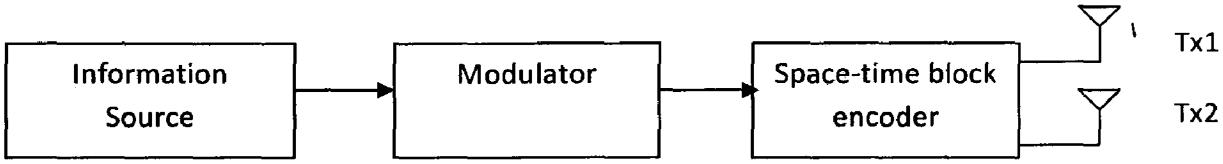


Fig.:- 3.5 Encoder for STBC

The entries of the transmission matrix S are so chosen that they are linear combinations of the k modulated symbols s_1 , s_2, . . . , s_k and their conjugates s_1^*, s_2^*, . . . , s_k^*. The matrix itself is so constructed based on orthogonal designs such that [7]

$$S.S^H = c(|s_1|^2 + |s_2|^2 + \dots\dots\dots + |s_k|^2)I_{N_T} \tag{3.26}$$

where c is a constant, N_T is the number of transmit antennas, S^H is the Hermitian of S, and I_{N_T} is an $N_T \times N_T$ identity matrix. This approach yields a diversity of N_T. These code transmission matrixes are cleverly constructed such that the rows and columns of each matrix are orthogonal to each other (i.e., the dot product of each row with another row is zero). If this condition is satisfied, (3.26) will be satisfied, yielding the full transmit diversity of N_T.

If the rows of a matrix are orthogonal (i.e., their dot product is zero), then the rows of that matrix are deemed independent. This implies that each row contributes an eigenvalue (i.e., the matrix is of full rank). Hence, full transmit diversity s is achieved as each transmit antenna contributes to one row in that matrix. The code rates will, however, vary depending on how the matrix is constructed. Based on (3.24), we can have R = 1, which is a full rate. This implies that there is no bandwidth expansion involved, whereas a code with rate R < 1 implies a bandwidth expansion factor of 1/R. It will be shown in this chapter that code rates of unity (i.e., full rates) are relatively easily achievable if the matrix is real, but the choice for full-rate codes is more restricted if the matrix is complex. Using (3.26), the orthogonality achieved in all cases enables us to achieve full transmit diversity, irrespective of the code rate and additionally allows the receiver to decouple the signals transmitted from different antennas. Consequently, a simple maximum likelihood decoding, based only on linear processing of the received signals, can be employed at the receiver.

3.5.1 STBC for Real Signal Constellations

Let us consider square transmission matrixes. Such matrixes exist if the number of transmit antennas N_T = 2, 4, or 8 [7]. These codes are full rate, since the matrix is square, and also full transmit diversity of N_T. The transmission matrixes are given by

$$S_2 = \begin{bmatrix} s_1 & -s_1 \\ s_1 & s_1 \end{bmatrix} \tag{3.27}$$

for $N_T = 2$ transmit antennas.

$$S_4 = \begin{bmatrix} s_1 & -s_2 & -s_3 & -s_4 \\ s_2 & s_1 & s_4 & -s_3 \\ s_3 & -s_4 & s_1 & s_2 \\ s_4 & s_3 & -s_2 & s_1 \end{bmatrix} \tag{3.28}$$

for $N_T = 4$ transmit antennas.

All the preceding matrixes have independent rows in that their dot product is zero for any real constellation and the code rate for all these matrixes is unity.

For example, equation (3.28), there are four transmit antennas (i.e., dealing with a space-time block code of size 4, corresponding to four rows). There are also four transmission periods p corresponding to each column of the matrix. There are also four symbols (i.e., $k = 4$, s_1 , s_2 , s_3 , and s_4). Hence, during the first transmission interval, s_1 , s_2 , s_3 , and s_4 are transmitted. During the next transmission interval, $-s_2$, s_1 , $-s_4$, and s_3 are transmitted. So,

$$R = \frac{k}{p} = \frac{4}{4} = 1 \tag{3.29}$$

To construct full code rate R = 1 transmission schemes for any number of antennas (N_T transmit antennas), the minimum value of transmission periods p to achieve the full rate is given by

$$p = \min(2^{4c+d}) \tag{3.30}$$

where the minimization is taken over the set

$$c, d | 0 \le c, \qquad 0 \le d \le 4, \qquad and \;\; 8c + 2^d \ge N_T \tag{3.31}$$

Non-square matrixes of size 3 and 5 for real numbers, yielding full diversity and full rate [7]

$$S_3 = \begin{bmatrix} s_1 & -s_2 & -s_3 & -s_4 \\ s_2 & s_1 & s_4 & -s_3 \\ s_3 & -s_4 & s_1 & s_2 \end{bmatrix} \quad (3.32)$$

$$S_5 = \begin{bmatrix} s_1 & -s_2 & -s_3 & -s_4 & -s_5 & -s_6 & -s_7 & -s_8 \\ s_2 & s_1 & -s_4 & s_3 & -s_6 & s_5 & s_8 & -s_7 \\ s_3 & s_4 & s_1 & -s_2 & -s_7 & -s_8 & s_5 & s_6 \\ s_4 & -s_3 & s_2 & s_1 & -s_8 & s_7 & -s_6 & s_5 \\ s_5 & s_6 & s_7 & s_8 & s_1 & -s_2 & -s_3 & -s_4 \end{bmatrix} \quad (3.33)$$

3.5.2 STBC for Complex Signal Constellations

Complex orthogonal design matrixes are defined as matrixes of size $N_T \times p$ with complex entries of $s_1, s_2, \ldots, s_k$ and their conjugates. The Alamouti scheme is itself one such matrix with complex entries for two transmit antennas. This is represented by

$$G_2 = \begin{bmatrix} s_1 & -s_2^* \\ s_1 & s_1^* \end{bmatrix} \quad (3.34)$$

This scheme provides the full diversity of 2 with a full code rate of 1. The Alamouti scheme is unique in that for complex entries, it is the only such matrix with a code rate of unity. The following complex transmission matrixes of size $N_T \times 3$ and $N_T \times 4$ incorporating a code rate of 1/2 [10]

$$G_3 = \begin{bmatrix} s_1 & -s_2 & -s_3 & -s_4 & s_1^* & -s_2^* & -s_3^* & -s_4^* \\ s_2 & s_1 & s_4 & -s_3 & s_2^* & s_1^* & s_4^* & -s_3^* \\ s_3 & -s_4 & s_1 & s_2 & s_3^* & -s_4^* & s_1^* & s_2^* \end{bmatrix} \quad (3.35)$$

$$G_5 = \begin{bmatrix} s_1 & -s_2 & -s_3 & -s_4 & s_1^* & -s_2^* & -s_3^* & -s_4^* \\ s_2 & s_1 & s_4 & -s_3 & s_2^* & s_1^* & s_4^* & -s_3^* \\ s_3 & -s_4 & s_1 & s_2 & s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & s_3 & -s_2 & s_1 & s_4^* & s_3^* & -s_2^* & s_1^* \end{bmatrix} \quad (3.36)$$

The inner product of any two rows of these matrixes is zero. This proves that the matrix is orthogonal and of full rank yielding full diversity of $N_T = 3$ and $N_T = 4$, respectively. In the case of G_3, for example, there are four symbols, s_1, s_2, s_3, and s_4 and their complex conjugates, yielding $k = 4$, and there are eight transmission periods, yielding p = 8. This gives us a code rate of R = k/p = 4/8 = 1/2. Similarly, G_4 has a code rate of R = k/p = 4/8 = 1/2, but with a diversity of $N_T = 4$.

The desire for higher code rates leads us to more complex linear processing. The following are size 3 and 4 codes with rate 3/4 [10]

$$H_3 = \begin{bmatrix} s_1 & -s_2^* & s_3^*/\sqrt{2} & s_3^*/\sqrt{2} \\ s_2 & s_1^* & s_3^*/\sqrt{2} & -s_3^*/\sqrt{2} \\ s_3/\sqrt{2} & s_3/\sqrt{2} & (-s_1 - s_1^* + s_2 - s_2^*)/2 & (s_2 + s_2^* + s_1 - s_1^*)/2 \end{bmatrix} \quad (3.37)$$

$$H_4 = \begin{bmatrix} s_1 & -s_2 & s_3^*/\sqrt{2} & s_3^*/\sqrt{2} \\ s_2 & s_1 & s_3^*/\sqrt{2} & -s_3^*/\sqrt{2} \\ s_3/\sqrt{2} & s_3/\sqrt{2} & (-s_1 - s_1^* + s_2 - s_2^*)/2 & (s_2 + s_2^* + s_1 - s_1^*)/2 \\ s_3/\sqrt{2} & -s_3/\sqrt{2} & (-s_2 - s_2^* + s_1 - s_1^*)/2 & -(s_1 + s_1^* + s_2 - s_2^*)/2 \end{bmatrix}$$

(3.38)

3.6 ST trellis coding scheme (STTC)

Space-Time block codes can achieve a maximum possible diversity advantage with a simple decoding algorithm. They are very attractive because of their simplicity. However, the coding gain provided by space-time block codes is non-full rate and limited, and space-time block codes can introduce bandwidth expansion.

STTCM (space-time trellis coded modulation) is another coding scheme has been proposed by Tarokh [11], can simultaneously offer a Substantial coding gain, spectral efficiency, and diversity improvement on flat fading Channels. A joint design of error control coding, modulation, transmit and receive diversity to develop an effective signaling scheme, space-time trellis codes (STTC), which is able to combat the effects of fading. STTC can simultaneously offer a substantial coding gain, spectral efficiency, and diversity improvement on flat fading channels [11].

3.7 Spatial –multiplexing

Spatial multiplexing scheme exploits the rich scattering wireless channel allowing the receiver antennas to detect the different signals simultaneously transmitted by the transmit antennas. Spatial multiplexing method uses multiple antennas at the transmitter and the receiver in conjunction with rich scattering environment within the same frequency band to provide a linearly increasing capacity gain in the number of antennas [3]. Hence, the concept of spatial multiplexing is different from space-time coding method, which permits to efficiently introduce a space-time correlation among transmitted signals to improve information protection and increase diversity gain. There are three well known algorithm in spatial multiplexing namely D-BLAST, VBLAST and HBLAST. The more popular one is VBLAST transmission which has optimum tradeoff between performance efficiency and computational complexity.

3.8 Simulation Results: -

The simulation is carried out for MIMO-STBC system with Rayleigh flat fading channel with additive white Gaussian noise. The M PSK modulation scheme is used and complex block of data are transmitted

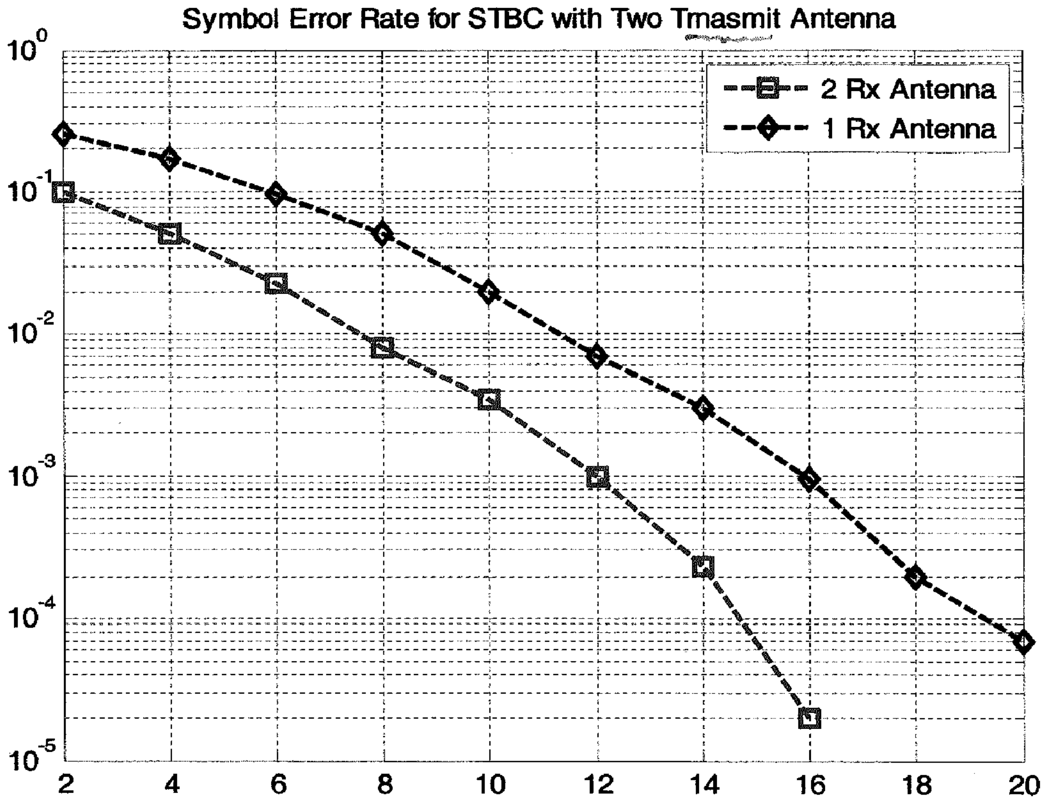


Fig. : - 3.6 Symbol error rate comparison for two transmit and different receive antenna with 4 PSK constellation.

As it can be seen from the graph, the performance of the system is highly depend on the number of transmit and receive antenna. Fig. 3.6 shows the symbol error rate comparison of one and two receive antenna with fixed two transmit antenna.

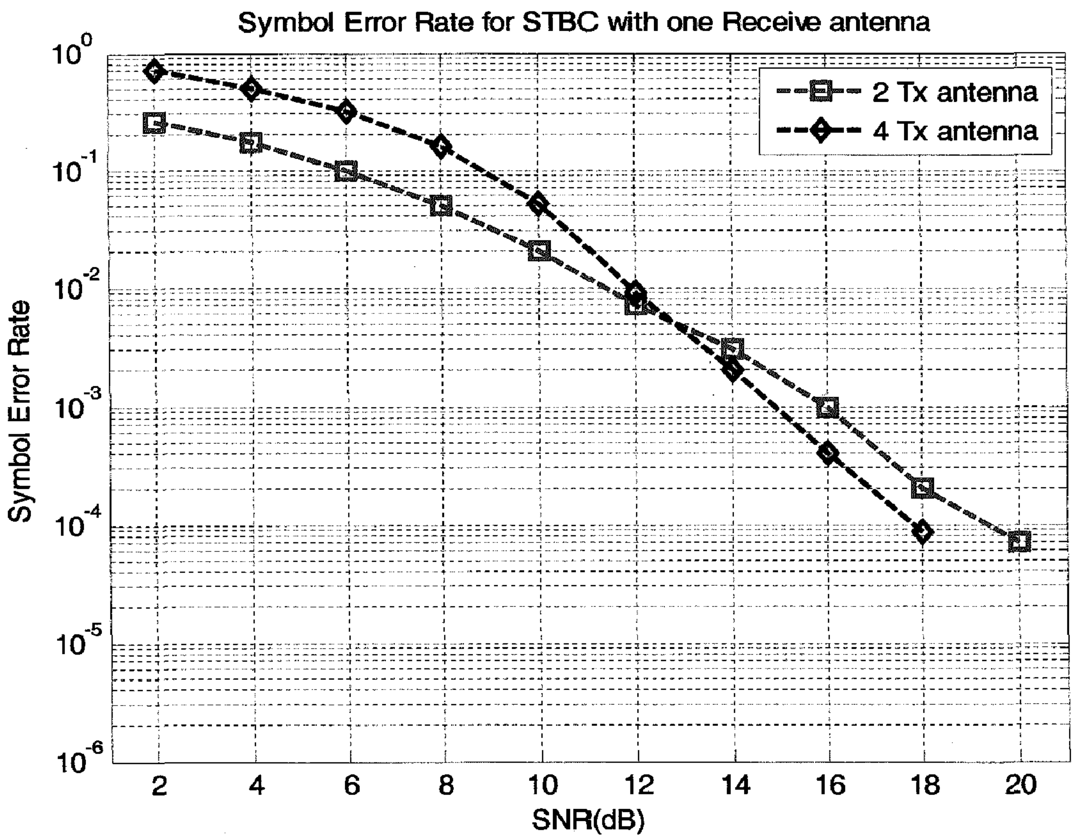


Fig.: - 3.7 Symbol error rate comparison for one receive and different transmit antenna with 4 PSK constellation.

Similarly fig. 3.7 shows the symbol error rate comparison of two and four transmit antenna with fixed one receive antenna. These graphs are significantly comparable with the standard results. By fig. 3.6 we can see that the symbol error rate is significantly better in case 2 receive antenna. Similarly in fig. 3.7 the symbol error rate is better in case of four transmit antenna.

CHAPTER 4

OFDM SYSTEMS

4.1 Introduction

Orthogonal frequency-division multiplexing (OFDM) has received increased attention due to its capability of supporting high-data-rate communication in frequency selective fading environments which cause inter-symbol interference (ISI). OFDM is a special form of multicarrier modulation (MCM), where a single data stream is transmitted over a number of lower rate subcarriers. The basic principle of OFDM is to split a high-data-rate sequence into a number of low-rate sequences that are transmitted simultaneously over a number of subcarriers [12].

Because the symbol duration is increased for the low rate parallel subcarriers, the relative amount of dispersion in time caused by multipath delay spread is decreased. Inter symbol interference (ISI) is eliminated almost completely by introducing a guard interval at the start of each OFDM symbol. In the guard interval, an OFDM symbol is cyclically extended to avoid inter-carrier interference (ICI). Thus, a highly frequency selective channel is transformed into a large set of individual flat fading, non-frequency selective, narrowband channels. The sub-carriers have the minimum frequency separation required to maintain orthogonality of their corresponding time domain waveforms, yet the signal spectra corresponding to the different sub-carriers overlap in frequency. The spectral overlap results in a waveform that uses the available bandwidth with very high bandwidth efficiency. OFDM is simple to use on channels that exhibit time delay spread or, equivalently, frequency selectivity. Frequency selective channels are characterized by either their delay spread or their channel coherence bandwidth which measures the channel decorrelation in frequency. The coherence bandwidth is inversely proportional to the root-mean-square (rms) delay spread. By choosing the sub-carrier spacing properly in relation to the channel coherence bandwidth, OFDM can be used to convert a frequency selective channel into a parallel collection of frequency flat subchannels [8].

Recent advances in digital signal processing promise a more efficient method for implementing OFDM. An integrated circuit implementation of a discrete Fourier transform removes the need for the entire bank of separate transmitters and receivers. The modulation of the set of K OFDM subcarriers using an inverse fast Fourier transform (IFFT) is equivalent to modulating each subcarrier individually with a rectangular baseband pulse shaper. The receiver samples the transmitted waveform to Obtain K samples on which a fast Fourier transform (FFT) is performed then. The FFT modulation is equivalent to performing an integral and dump on each subcarrier using a matched filter of the rectangular baseband waveform [2]. The OFDM system plays prime role to transform frequency selective channel to narrow band flat fading channel and generally OFDM make optimum use of frequency selective channel and eliminate the need for high complexity rake receiver. Therefore because of the significant role of OFDM in MIMO-OFDM system in this dissertation one chapter is assigned to explain the concept of OFDM from modulation to detection part [13].

4.2 OFDM

Let $\{x_k\}_{k=0}^{N-1}$ can be the complex symbols to be transmitted by OFDM modulation; the OFDM (modulated) signal can be expressed as

$$x(k) = \sum_{k=0}^{N-1} X_K\, e^{j2\pi f_k t} = \sum_{k=0}^{N-1} X_k \emptyset_k(t)\ ,\ for\ \ 0 \le t \le T_s\ , \tag{4.1}$$

Where $f_k = f_0 + k\Delta f$ and

$$\emptyset_k(t) = \begin{cases} e^{j2\pi f_k t} & if\ 0 \le t \le T_s, \\ 0 & Otherwise \end{cases} \tag{4.2}$$

for k = 0, 1,. . . , N - 1. T_s and Δf are called the symbol duration and sub channel space of OFDM, respectively. In order for receiver to demodulate OFDM signal, the symbol duration must be long enough such that $T_s\ \Delta f = 1$, which is also called orthogonality condition [14].

Because of the orthogonality condition, we have

$$\frac{1}{T_S}\int_0^{T_S} \emptyset_k(t)\emptyset_l^*(l)dt$$
$$= \frac{1}{T_s}\int_0^{T_s} e^{j2\pi(f_k-f_l)t}dt$$
$$= \frac{1}{T_s}\int_0^{T_s} e^{j2\pi(k-l)t}dt$$
$$= \delta[k-l] \tag{4.3}$$

Where $\delta[k\text{-}l]$ is the delta function defined as

$$\delta[n] = \{ \begin{matrix} 1, & if\ n = 0, \\ 0, & \text{otherwise} \end{matrix}$$

Equation (4.3) shows that $\{x_k\}_{k=0}^{N-1}$ is a set of orthogonal functions. Using this property, the OFDM signal can be demodulated bys

$$\frac{1}{T_s}\int_0^{T_s} x(t)e^{-j2\pi f_k t}dt$$
$$= \frac{1}{T_s}\int_0^{T_s}\left(\sum_{l=0}^{N-1} x_l\emptyset_l(t)\right)\emptyset_k^*(t)dt$$
$$= \sum_{l=0}^{N-1} x_l\ \delta[l-k]$$
$$= s_k \tag{4.4}$$

4.3 FFT Implementation

An OFDM signal can be expressed as

$$x(k) = \sum_{k=0}^{N-1} X_K\ e^{j2\pi f_k t}$$

If s(t) is sampled at an interval of $T_{sa} = T_s/N$, then

$$X_n = x(n\Delta_s) = \sum_{k=0}^{N-1} X_K\, e^{j2\pi f_k \frac{nT_s}{N}} \tag{4.5}$$

Without loss of generality, setting $f_0 = 0$, then $f_k\, Ts = k$ and (4.5) becomes

$$x_n = \sum_{k=0}^{N-1} X_K\, e^{j2\pi \frac{nk}{N}} = IDFT(x_k),$$

where IDFT denotes the inverse discrete Fourier transform. Therefore, the OFDM transmitter can be implemented using the IDFT. For the same reason, the receiver can be also implemented using DFT. The FFT algorithm provides an efficient way to implement the DFT and the IDFT. It reduces the number of complex multiplications from N^2 to $\frac{N}{2}\log_2 N$ for an N-point DFT or IDFT. Hence, with the help of the FFT algorithm, the implementation of OFDM is very simple [15].

4.4 Time Domain OFDM System Model

In order to understand the mathematical principles of OFDM systems, first take an overview of the time domain uncoded OFDM system model, which is shown in Fig. 4.1.

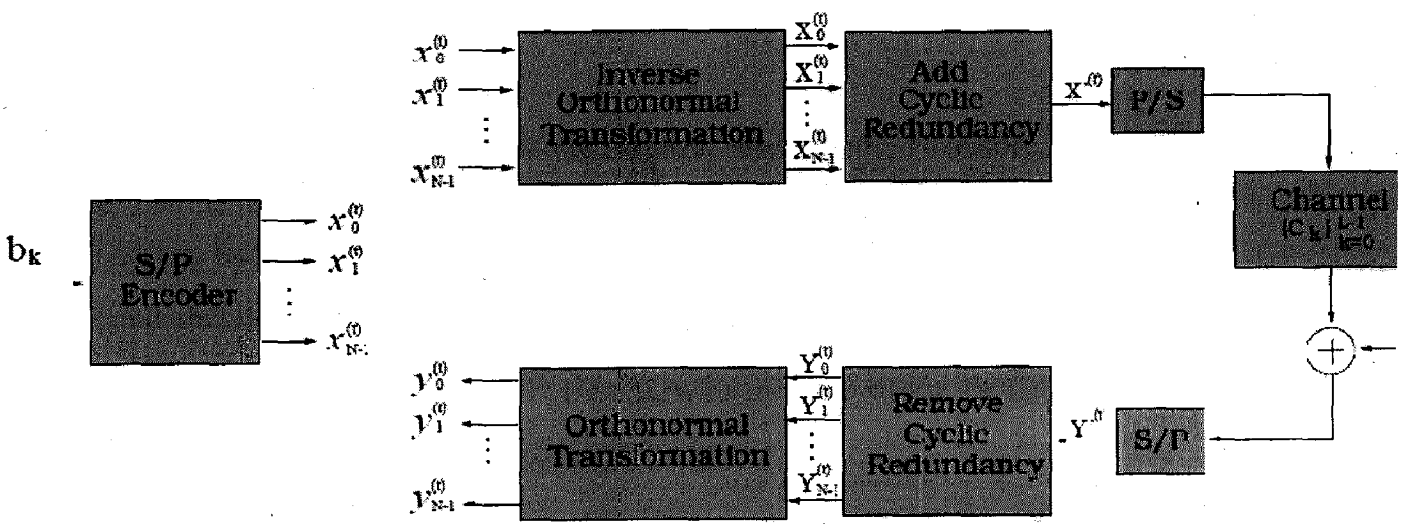


Fig.: - 4.1 Time domain OFDM system model

Let b_k represent the binary data sequence to be transmitted over the channel. This data is divided into non-overlapping blocks of n = N . $\log_2 M$ bits. The n bits of data are partitioned into N groups, with each $\log_2 M$ bits mapped into a complex symbol of constellation size M. Symbol $x_k^{(t)}$ is the signal transmitted over the k-th subcarrier during the t-th OFDM frame. At the transmitter, an inverse discrete Fourier transform (IDFT) is performed as a method of modulation, which results in samples $X_k^{(t)}$ given by

$$X_k^{(t)} = \frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} x_k^{(t)} . \exp\,(j\frac{2\pi k i}{N}), \qquad 0 \leq i \leq N-1. \tag{4.6}$$

Assume that the channel is frequency-selective, and hence the implementation of a cyclic redundancy of sufficient length to the N-point OFDM frame is an effective method to counter inter symbol interference (ISI) caused by the channel. The cyclic prefix causes the sequence $\{X_k^{(t)}\}$ to appear periodic to the channel and clears the channel memory at the end of each OFDM frame. This action makes successive transmissions independent. The output from the channel, with additive noise $N_k^{(t)}$, may be written as

$$Y_k^{(t)} = c_k^{(t)} * X_k^{(t)} + N_k^{(t)}, \tag{4.7}$$

where $c_k^{(t)}$ is the discretized fading channel coefficient .

At the receiver, the cyclic prefix is discarded to obtain a frame of N symbols $Y_k^{(t)}$. Taking the N-point discrete Fourier transform of the received samples $Y_k^{(t)}$, we have the output samples given by

$$y_k^{(t)} = \frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} Y_k^{(t)} . \exp\,(-j\frac{2\pi k i}{N}), \qquad 0 \leq i \leq N-1. \tag{4.8}$$

4.5 Equivalent Frequency Domain OFDM System Model

Although the time domain model provided in the previous section is conceptually straightforward, it is much more insightful to analyze the OFDM system in the frequency domain since the information symbols modulate different subcarriers in the frequency domain. Hence, let us consider the frequency model for a coded OFDM system illustrated in Fig. 4.2. A block of k information bits, denoted as $b = (b_1, b_2, \ldots\ldots, b_k)$, is encoded into a codeword $x = (x_1, x_2, \ldots\ldots, x_n)$ of length n, where each symbol x_i is an element from a complex alphabet χ. There are a total of m codewords in the code book and the code rate is defined to be $R = (\log_2 m)/n$. Note that here we combine encoder, modulator, and interleaver together to form one super encoder E.

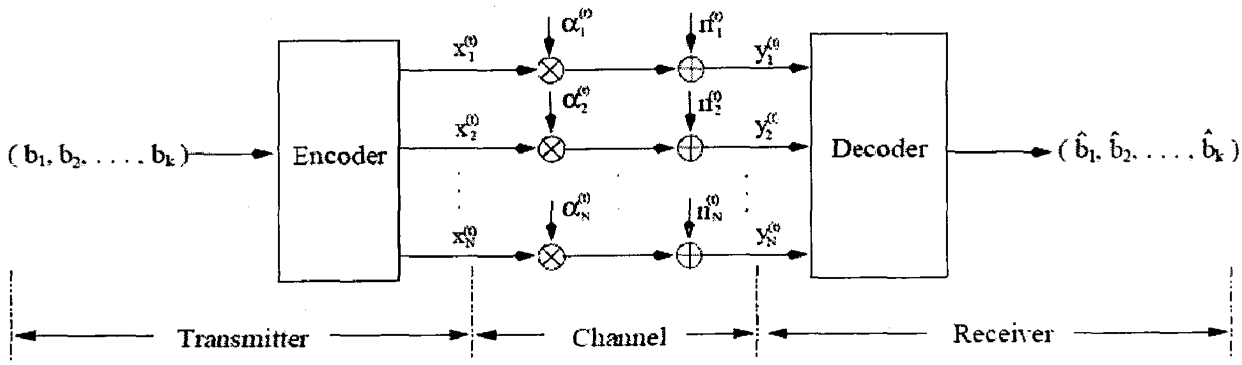


Fig.: - 4.2 Frequency domain OFDM system model

The encoded block $x = (x_1, x_2, \ldots\ldots, x_n)$ is segmented into *l* frames, each of length N ($lN = n$). The individual frame is transmitted by N dependent parallel sub- channels, each representing a different OFDM sub-carrier. According to the tapped- delay-line model, the fading coefficients $\alpha_k^{(t)}$ of the t^{th} OFDM frame are related to the fading envelopes $c_k^{(t)}$ through

$$c^{(t)} = [c_1^{(t)}, c_2^{(t)}, \ldots\ldots\ldots, c_L^{(t)}, 0, \ldots\ldots\ldots, 0]^T \quad \in \mathbb{C}^{N\times 1}, \tag{4.9}$$

$$\alpha^{(t)} = [\alpha_1^{(t)}, \alpha_2^{(t)}, \ldots\ldots\ldots, \alpha_N^{(t)}]^T \quad \in \mathbb{C}^{N\times 1}, \tag{4.10}$$

$$\alpha^{(t)} = W_{N\times N} c^{(t)}, \tag{4.11}$$

where the Fourier transformation $W_{N\times N}$ is given by

$$W_{N\times N} = \begin{pmatrix} \omega^0 & \omega^0 & & \cdots & \omega^0 \\ \omega^0 & \omega^1 & & \cdots & \omega^{N-1} \\ \vdots & \vdots & \vdots & & \ddots \\ \omega^0 & \omega^{N-1} & \cdots & & \omega^{(N-1)(N-1)} \end{pmatrix}, \quad \omega = e^{-j\frac{2\pi}{N}}. \tag{4.12}$$

We can further stack l consecutive fading coefficients and envelopes into a compact vector form as

$$c = c^{(1)^T}, c^{(2)^T}, \dots \dots \dots, c^{(l)^T}]^T \quad \in C^{n\times 1}, \tag{4.13}$$

$$\alpha = [\alpha^{(1)^T}, \alpha^{(2)^T}, \dots \dots \dots, \alpha^{(l)^T}]^T \quad \in C^{n\times 1}, \tag{4.14}$$

where the fading envelope $c^{(t)}$ is assumed to be constant within one OFDM frame, but is a function of time t from frame to frame. Each component $c_i^{(t)}$ of the fading envelope is assumed to be independent from tap to tap.

The received output vectors $y^{(t)}$ are given by

$$y^{(t)} = [y_1^{(t)}, y_2^{(t)}, \dots \dots \dots, y_N^{(t)}]^T, \qquad (1 \le t \le l) \tag{4.15}$$

$$y_i^{(t)} = \alpha_i^{(t)} x_i^{(t)} + n_i^{(t)}, \qquad (1 \le i \le N) \tag{4.16}$$

where additive complex Gaussian noise $n_i^{(t)}$ is white with variance N_0. The receiver is assumed to have perfect knowledge of the channel state information (CSI) and makes decisions based on the observation of the received vector $y^{(t)}$, $1 \le t \le l$, and the channel state information c.

Chapter 5

STBC-OFDM SYSTEMS

In wideband wireless communications, the symbol period becomes smaller relative to the channel delay spread, and consequently, the transmitted signals experience frequency-selective fading. Space-time coding techniques could be used to achieve very high data rates in wideband systems. Therefore, it is desirable to investigate the effect of frequency-selective fading on space-time code performance.

Space-time coding provides diversity gain by coding over spatial that is multiple antennas, and time dimensions. Transmitting data over MIMO-OFDM systems is possible by applying space-time codes to each sub-carrier [19].

5.1 Space-Time Coding on Frequency-Selective Fading Channels

It is assumed that the delay spread τ_d is relatively small compared with the symbol duration. In order to investigate the effect of frequency-selective fading on the code performance, we assume that no equalization is used at the receiver. Consider a system with n_T transmit and n_R receiver antennas. Let $h_{j,i}$ (t, τ) denote the channel impulse response between the i-th transmit antenna and j-th receive antenna. The transmitted signal from antenna i

$$u^i(t) = \sum_{k=-\infty}^{\infty} x_k^i g(t - kT_s) \tag{5.1}$$

where T_s is the symbol period, where x^i_k is the message for the i-th antenna at the k-th symbol period and g(t) is the pulse shaping function. The received signal can be decomposed into the following three terms [16] [17].

$$r_t^j = \alpha \sum_{i=1}^{n_T} \sum_{l=1}^{L_p} h_{j,i}^{t,l} x_t^i + I_t^j + n_t^j \tag{5.2}$$

where I^j_t is a term representing the inter symbol interference (ISI), and α is a constant dependent on the channel power delay profile, which can be computed as

$$\alpha = \frac{1}{T_s} \int_{-T_s}^{T_s} P(\tau)\,(T_s - |\tau|) d\tau \tag{5.3}$$

For simplicity, the ISI term is approximated by a Gaussian random variable with a zero-mean and single-sided power spectral density $N_I = \sigma^2_I T_s$. The sum of the additive noise and the ISI is denoted by $\bar{n}^j_t$.

$$\bar{n}_t^j = I_t^j + n_t^j \tag{5.4}$$

The received signal can be rewritten as

$$r_t^j = \alpha \sum_{i=1}^{n_T} \sum_{l=1}^{L_p} h_{j,i}^{t,l} x_t^i + \bar{n}_t^j \tag{5.5}$$

where $\bar{n}^j_t$ is a complex Gaussian random variable with a zero mean and the single-sided power spectral density $N_I + N_0$. The additive noise and the ISI are uncorrelated with the signal term. The pair wise error probability under this approximation is given by [17]

$$P(X, \hat{X}) \leq \left[\prod_{i=1}^{n_T} (1 + \lambda_i \frac{\alpha^2 E_s}{4(N_0 + N_I)}) \right]^{-n_R}$$

$$\leq \left[\prod_{i=1}^{r} \left(\lambda_i \frac{\alpha^2}{\frac{N_I}{N_0} + 1} i \right) \right]^{-n_R} (\frac{E_s}{4N_0})^{-rn_r} \tag{5.6}$$

where r is the rank of the codeword distance matrix, and λ_i , i = 1, 2, . . . , r, are the nonzero eigen values of the matrix. From the above upper bound, we can observe that the diversity gain achieved by the space-time code on multipath and frequency-selective fading channels is rn_R, which is the same as that on frequency-nonselective fading channels. The coding gain is

$$G_{coding} = \frac{(\prod_{i=1}^{r} \lambda_i)^{1/r} \frac{\alpha^2}{N_I/N_0 + 1}}{d_u^2} \tag{5.7}$$

The coding gain is reduced by a factor of $\frac{\alpha^2}{N_I/N_0+1}$ compared to the one on frequency flat fading channels. Furthermore, at high SNRs, there exists an irreducible error rate floor [16] [17]. The above performance analysis is performed under the assumptions that the time delay spread is small and no equalizer is used at the receiver. When the delay spread becomes relatively high, the coding gain will decrease considerably due to ISI, and cause a high performance degradation. In order to improve the code performance over frequency selective fading channels, additional processing is required to remove or prevent ISI.

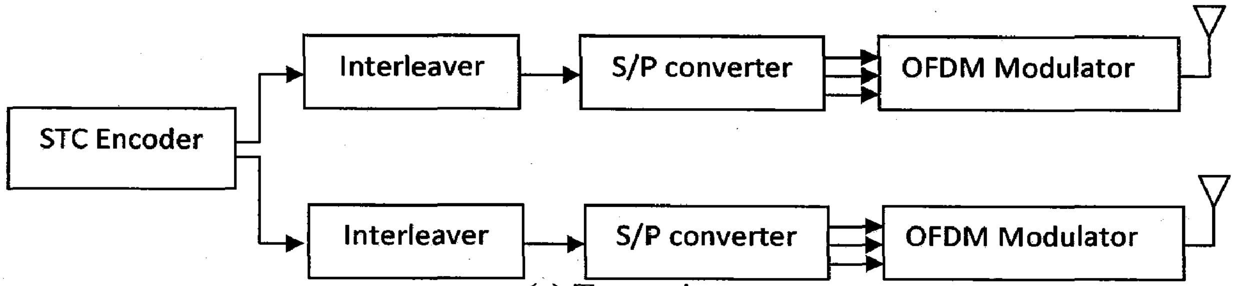
Space-time code on frequency-selective fading channels can achieve at least the same diversity gain as that on frequency-nonselective fading channels provided that maximum likelihood decoding is performed at the receiver. An optimal space-time code on frequency-selective fading channels may achieve a higher diversity gain than on frequency-nonselective fading channels. As the maximum likelihood decoding on frequency-selective channels is prohibitively complex, a reasonable solution to improve the performance of space-time codes on frequency-selective fading channels is to mitigate ISI. By mitigating ISI, one can convert frequency-selective channels into frequency-nonselective channels. Then, good space-time codes for frequency-nonselective fading channels can be applied [9] [4].

A conventional approach to mitigate ISI is to use an adaptive equalizer at the receiver. An optimum space-time equalizer can suppress ISI, and therefore, the frequency-selective fading channels become inter symbol interference free. The main drawback of this approach is a high receiver complexity because a multiple-input/multiple-output equalizer (MIMO-EQ) has to be used at the receiver [18]. An alternative approach is to use orthogonal frequency division multiplexing (OFDM) techniques [18]. In OFDM, the entire channel is divided into many narrow parallel subchannels, thereby increasing the symbol duration and reducing or eliminating the ISI caused by the multipath environments [15]. Since MIMO-EQ is not required in OFDM systems, this approach is less complex.

5.2 STC-OFDM Systems

We consider a baseband STC-OFDM communication system with K OFDM sub-carriers, n_T transmit and n_R receive antennas. The total available bandwidth of the system is W Hz. It is divided into K overlapping sub-bands. The system block diagram is shown in Fig. 5.1.

A parallel data system can alleviate ISI even without equalization. In such a system the high-rate data stream is demultiplexed into a large number of sub-channels with the spectrum of an individual data element occupying only a small part of the total available bandwidth. A parallel system employing conventional frequency division multiplexing (FDM) without sub-channel overlapping is bandwidth inefficient. A much more efficient use of bandwidth can be obtained with an OFDM system in which the spectra of the individual sub-channels are permitted to overlap and the carriers are orthogonal [4].



(a) Transmitter

(b) Receiver

Fig.: - 5.1 An STC-OFDM system block diagram

At each time t, a block of information bits is encoded to generate a space-time codeword which consists of $n_T L$ modulated symbols. The space-time codeword is given by

$$X_t = \begin{bmatrix} x_{t,1}^1 & x_{t,2}^1 & \cdots & x_{t,L}^1 \\ x_{t,1}^2 & x_{t,2}^2 & \cdots & x_{t,L}^2 \\ \vdots & \vdots & \ddots & \vdots \\ x_{t,1}^{n_T} & x_{t,2}^{n_T} & \cdots & x_{t,L}^{n_T} \end{bmatrix} \quad (5.8)$$

where the i-th row $x^i_t = x^i_{t,1}, x^i_{t,2}, \ldots, x^i_{t,L}$, $i = 1, 2, \ldots, n_T$ is the data sequence for the i-th transmit antenna. For the sake of simplicity, we assume that the codeword length is equal to the number of OFDM sub-carriers, L = K. Signals $x^i_{t,1}, x^i_{t,2}, \ldots, x^i_{t,L}$ are OFDM modulated on K different OFDM sub-carriers and transmitted from the i-th antenna simultaneously during one OFDM frame, where $x^i_{t,k}$ is sent on the k-th OFDM sub-carrier.

For Alamouti scheme the l-th (out of L) OFDM frame transmitted from the first and second antenna, respectively, are given by

$$\left(X_l^1(0) \ldots\ldots\ldots X_l^1(M-1)\right) \xrightarrow{IDFT} \left(x_{l,0}^1 \ldots\ldots\ldots x_{l,M-1}^1\right)$$

$$\xrightarrow{CP} \left(x_{l,M-v}^1 \ldots\ldots\ldots x_{l,M-1}^1, x_{l,0}^1 \ldots\ldots\ldots x_{l,M-1}^1\right)$$

$$\left(X_l^2(0) \ldots\ldots\ldots X_l^2(M-1)\right) \xrightarrow{IDFT} \left(x_{l,0}^2 \ldots\ldots\ldots x_{l,M-1}^2\right)$$

$$\xrightarrow{CP} \left(x_{l,M-v}^2 \ldots\ldots\ldots x_{l,M-1}^2, x_{l,0}^2 \ldots\ldots\ldots x_{l,M-1}^2\right)$$

For the next frame interval, the OFDM frame

$$\left(x_{l,M-v}^{1\ *} \ldots\ldots\ldots x_{l,M-1}^{1\ *}, x_{l,0}^{1\ *} \ldots\ldots\ldots x_{l,M-1}^{1\ *}\right)$$

is transmitted from second antenna and

$$\left(x_{l,M-v}^{2\ *} \ldots\ldots\ldots x_{l,M-1}^{2\ *}, x_{l,0}^{2\ *} \ldots\ldots\ldots x_{l,M-1}^{2\ *}\right)$$

is transmitted from the first antenna according to the signaling structure of Alamouti's scheme, where (.)* denotes complex conjugation.

In OFDM systems, in order to avoid ISI due to the delay spread of the channel, a cyclic prefix is appended to each OFDM frame during the guard time interval. The cyclic prefix is a copy of the last L_p samples of the OFDM frame, so that the overall OFDM frame length is $L + L_p$, where L_p is the number of multipath in fading channels.

In the performance analysis, we assume ideal frame and symbol synchronization between the transmitter and the receiver. A sub-channel is modeled by quasi-static Rayleigh fading. The fading process remains constant during each OFDM frame. It is also assumed that channels between different antennas are uncorrelated.

At the receiver, after matched filtering, the signal from each receive antenna is sampled at a rate of W Hz and the cyclic prefix is discarded from each frame. Then these samples are applied to an OFDM demodulator. The output of the OFDM demodulator for the k-th OFDM sub-carrier, k = 1, 2. . . K, at receive antenna j, j = 1, 2. . . n_R, is given by [5]

$$R_{t,k}^{j} = \sum_{i=1}^{n_T} H_{j,i}^{t,k}\, x_{t,k}^{i} + N_{t,k}^{j} \tag{5.9}$$

where $H^{t,k}_{j,i}$ is the channel frequency response for the path from the i-th transmit antenna to the j-th receive antenna on the k-th OFDM sub-channel, and $N^{j}_{t,k}$ is the OFDM demodulation output for the noise sample at the j-th receive antenna and the k-th sub-channel with power spectral density N_0. Assuming that perfect channel state information is available at the receiver, the maximum likelihood decoding rule is given by

$$\hat{X}_t = arg\min_{\hat{X}} \sum_{j=1}^{n_R} \sum_{k=1}^{K} \left| R_{t,k}^{j} - \sum_{i=1}^{n_T} H_{j,i}^{t,k}\, x_{t,k}^{i} \right|^2 \tag{5.10}$$

where the minimization is performed over all possible space-time code words. The channel impulse response in the time domain is modeled as a tapped-delay line. The channel impulse response between the i-th transmit antenna to the j-th receive antenna is given by

$$h_{j,i}(t;\tau) = \sum_{l=1}^{L_p} h_{j,i}^{t,l}\,\delta(\tau - \tau_l) \tag{5.11}$$

where L_p is the number of multipaths, τ_l is the time delay of the l-th path and $h^{t,l}_{j,i}$ is the complex amplitude of the l-th path. Let us denote by T_f the time duration of each OFDM frame and by Δf the separation between the OFDM sub-carriers.

$$T_f = KT_s$$

$$T_s = \frac{1}{W} = \frac{1}{K\Delta f} \tag{5.12}$$

Now the delay of the *l*-th path can be represented as

$$\tau_l = n_l T_s = \frac{n_l}{K\Delta f} \tag{5.13}$$

where n_l is an integer. Performing the Fourier transform of the channel impulse response, we can get the channel frequency response at time t as

$$\begin{aligned} H_{j.i}^{t,k} &= H_{j,i}(tT_f, k\Delta f) \\ &= \int_{-\infty}^{+\infty} h_{j,i}(tT_f, \tau) e^{-j2\pi k\Delta f \tau} d\tau \\ &= \sum_{l=1}^{L_p} h_{j,i}(tT_f, n_l T_s) e^{-j2\pi k n_l / K} \\ &= \sum_{l=1}^{L_p} h_{j,i}(t, n_l) e^{-j2\pi k n_l / K} \end{aligned} \tag{5.14}$$

Let

$$h_{j,i}^{t} = [h_{j,i}^{t,1}, h_{j,i}^{t,2}, \dots \dots \dots, h_{j,i}^{t,L_p}]^H$$

$$w_k = [e^{-\frac{j2\pi k n_1}{K}}, e^{-\frac{j2\pi k n_2}{K}}, \dots \dots \dots, e^{-\frac{j2\pi k n_{L_p}}{K}}]^T \tag{5.15}$$

The equation (5.14) can be rewritten as

$$H_{j,i}^{t,k} = (h_{j,i}^{t})^H . w_k \tag{5.16}$$

From (5.14), we can see that the channel frequency response $H^{t,k}_{j,i}$ is the digital Fourier transform of the channel impulse response $h^{t}_{j,i}$. The transform is specified by the vector w_k for the k-th OFDM sub-carrier, k = 1, 2…… K.

5.3 Space-frequency coded MIMO-OFDM

Space-time coding provides diversity gain by coding over spatial, that is multiple antennas, and time dimensions. Transmitting data over MIMO-OFDM systems is possible by applying space-time codes to each sub-carrier. It does not provide the maximum possible diversity gain. In fact, the frequency diversity and the correlation among different sub-carriers are ignored in such a system.

Another approach for transmission over MIMO channels using OFDM is to replace the "time" dimension with the "frequency" dimension [22]. In other words, different sub-carriers of OFDM can be used as a replacement for the time dimension of a space-time code. The result is called space-frequency coding and Figure 5.2 depicts its block diagram. The space-time encoder in Figure 5.2 generates $L \times N$ symbols at each OFDM time slot. The interleaver transmits the symbol (l, n) through the l-th sub-carrier of the n-th antenna.

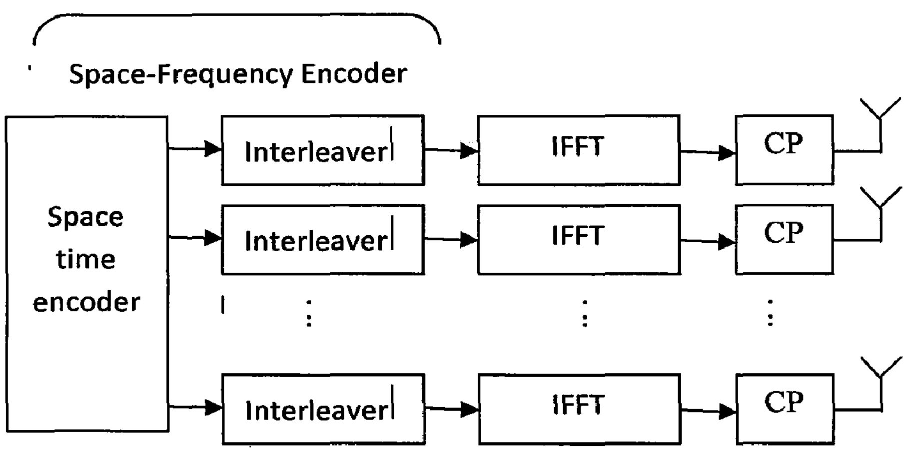


Fig.: - 5.2 Block diagram of space-frequency coding

A simple example of such a code using the Alamouti code. Assume L input symbols $s_1, s_2, \ldots\ldots, s_L$ at each time slot. The Alamouti code generates **C**(1) = (s1 s2) and **C**(2) = $(-s^*_2 \;\; s^*_1)$ from the first two symbols, **C**(3) = (s3 s4) and **C**(4) = $(-s^*_4 \;\; s^*_3)$ from the third and fourth symbols, and so on. Therefore, at each time slot, the first antenna sub-carriers, L, is relatively large. It is difficult to design codes for a large number of transmit antennas. Therefore, it is desirable to group the sub-carriers and design codes only across the sub-carriers in the same group [21]. A space-time-frequency code over three dimensions of space, time, and frequency is designed for each group. This reduces the frequency dimension of the code and decreases the complexity. For groups with at least J sub-carriers, if designed carefully, such a grouping will not degrade the diversity gain of the space-time-frequency codes.

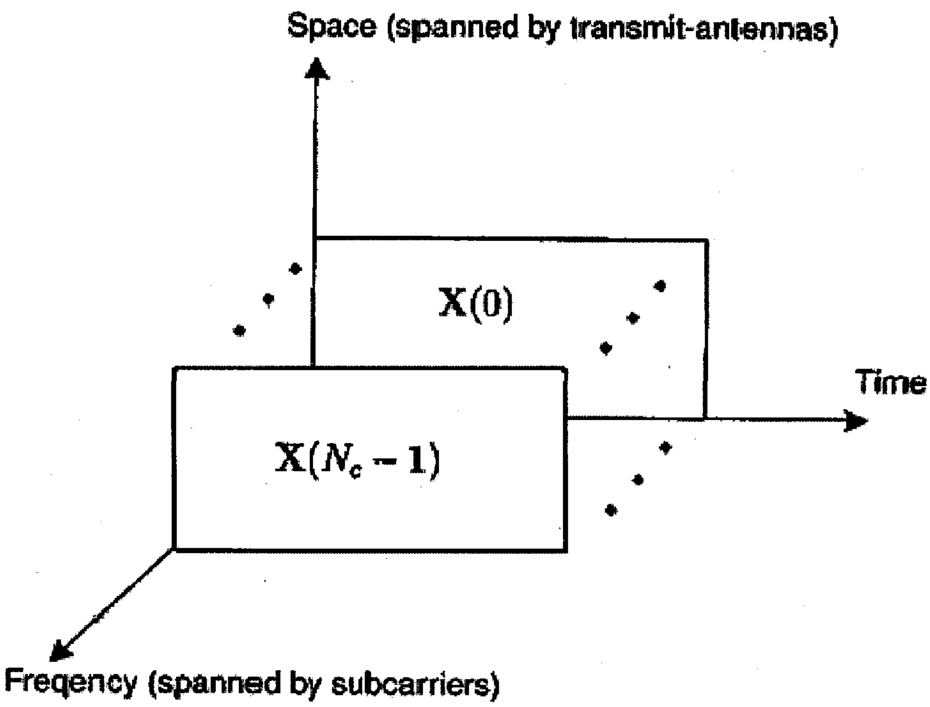


Fig.: - 5.4 Illustration of space-time-frequency code

The most popular coding technique in MIMO-OFDM exploits the independent fading of the parallel channels $H_{(k)}$ across the frequency tones. By coding across space and frequency, both space and frequency diversity gains can be extracted. If the coherence bandwidth of the channel is small (i.e. if the delay-spread and the number of taps are large), the channel gains $H_{(k)}$ vary significantly from tone to tone. The channel in the frequency domain can then be considered as fast fading. Swapping frequency with time, one expects that a code designed for flat fast fading channels will perform well if used as a space-frequency code in a highly frequency selective channel. This also suggests that certain codes designed for slow fading channels might not perform well if they are used as space-frequency codes in highly selective channels. For instance, O-STBCs require the channel to remain static over the codeword duration. As an example, if the Alamouti scheme is used across two tones, one should make sure that the channel coherence bandwidth is larger than the width of these tones. Otherwise, the matched filtering operation will not decouple the streams at the receiver, causing performance degradation.

Analogous to SC transmissions, a maximum diversity gain of $n_r n_t L$ could theoretically be extracted. MIMO-OFDM does not reduce the number of degrees of freedom. A slight loss in spectral efficiency is caused by the cyclic prefix, but for long frame lengths T this loss is negligible.

By contrast to SC transmissions, OFDM modulation generally requires a better synchronization and is more susceptible to phase noise. Furthermore, it is subject to large peak-to-average power ratios. Despite those drawbacks, SF-MIMO OFDM is often preferred to MIMO-SC and is now part of an increasing number of standards [19].

This scheme is not very practical for large codeword lengths T, as it requires the channel to remain constant over T OFDM symbols, i.e. over a usually large period. Furthermore, no coding is performed across the frequency tones, implying that no frequency diversity is extracted.

5.4 Simulation Result

The simulation is carried out for MIMO-OFDM STBC system with additive white Gaussian noise. The M PSK modulation scheme is used and complex block of data are transmitted.

As it can be seen from the graph, the performance of the system is highly depend on the number of transmit and receive antenna. Fig. 5.5 shows the symbol error rate comparison of one and two receive antenna with fixed two transmit antenna. For this simulation following parameter was considered FFT length 64, cyclic prefix 16, and frame length 50. For each SNR the signal was transmitted 1000 times. 4-PSK Modulation scheme was used.

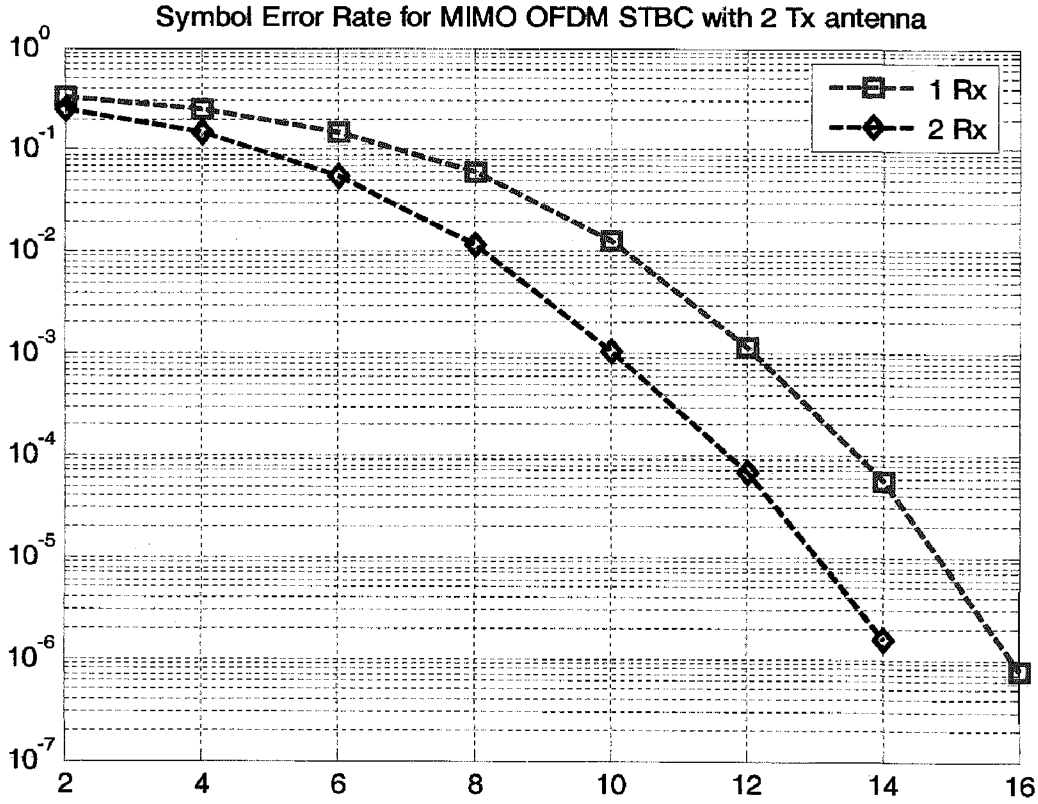


Fig.: - 5.5 Symbol error rate comparison of one and two receive antennas with Two transmit antennas of STBC-OFDM.

CHAPTER 6

SPACE-TIME CODED OFDM SYSTEMS WITH FREQUENCY OFFSET

OFDM, also known as multicarrier modulation, incorporates a large number of orthogonally selected subcarriers to transmit a high data-rate stream in parallel in the frequency domain. Spectral efficiency and multipath immunity are two major advantages of the OFDM technique. On the other hand, space-time (ST) communication systems incorporating multiple transmit and/or receive antennas (MIMO systems) provide large gains in achievable data rates (channel capacity). Thus, combining MIMO techniques with OFDM, i.e. MIMO-OFDM, is a promising for future fourth generation (4G) wireless communication systems, e.g. Gigabit wireless.

A major drawback of OFDM is its relatively high sensitivity to carrier frequency offset (CFO) errors, compared to a single carrier system. The frequency offset error is caused by the misalignment in carrier frequencies at the receiver due to fluctuations in receiver RF oscillators or a channel's Doppler frequency. This frequency offset can destroy the subcarrier orthogonality of the OFDM signal, introducing inter-carrier interference (ICI). The ICI results in severe degradation of the bit-error-rate (BER) performance of OFDM systems. Although frequency synchronization techniques using baseband signal processing can largely compensate for CFO, any residual error in frequency synchronization (CFO estimation error) contributes to the degradation of receiver performance [24].

Since OFDM subcarriers are closely packed compared to the system bandwidth the amount of tolerable frequency offset is a small fraction of the OFDM bandwidth. CFO introduces both attenuation/rotation of the useful signal and ICI which destroys the orthogonality of demodulated subcarriers. ICI and attenuation/phase rotation effects if not compensated, increase the system error rate and reduce the overall throughput. Alamouti coded OFDM systems are more sensitive to CFO errors than conventional OFDM [25].

6.1 Interference Due to frequency offset

6.1.1 SISO-OFDM Signal Model

For a single-input-single-output (SISO) OFDM system, the post-FFT signal $y(k)$ at the kth subcarrier, where $0 \leq k \leq N - 1$, can be given as

$$y(k) = s_0 h(k)x(k) + \sum_{l=0, l\neq k}^{N-1} s_{l-k} h(l)x(l) + w(k) \tag{6.1}$$

where, $x(k)$ and $h(k)$ are the transmit data symbol and frequency response for the kth subcarrier, respectively. $w(k)$ is additive white Gaussian noise (AWGN). The sequence s_k (ICI coefficients) depends on the CFO and is given by

$$s_k = \frac{\sin \pi(k+\epsilon)}{N \sin \frac{\pi}{N}(k+\epsilon)} \exp\left[j\pi\left(1-\frac{1}{N}\right)(k+\epsilon)\right] \tag{6.2}$$

where, ε is the normalized frequency offset, which is the ratio between the CFO and the adjacent subcarrier spacing.

For an SISO-OFDM system, the decision variable $\hat{x}\,(k)$ after equalization can be formed as$\hat{x}\,(k) = \bar{s}_0 \bar{h}(k) y(k)$, where it is assumed that the effective channel ~$\tilde{h}(k) = s_0 h(k)$ is known, i.e. perfectly estimated.

$$\hat{x}(k) = |s_0|^2 |h(k)|^2 x(k) + \bar{s}_0 \bar{h}(k) i(k) + \bar{s}_0 \bar{h}(k) w(k) \tag{6.3}$$

where, $i(k) = \sum_{l=0, l\neq k}^{N-1} s_{l-k} h(l)x(l)$ is the ICI interference term. In (6.3), the three terms show the signal, interference, and channel noise, respectively.

The power associated with the three terms in (6.3) for a given channel realization $h(k)$ for the kth subcarrier can be calculated as follows. The signal power $P_x(k)$ for the kth subcarrier becomes

$$P_X(k) = |s_0|^4|h(k)|^4E\{|x(k)|^4\} = |s_0|^4|h(k)|^4\sigma_X^2 \tag{6.4}$$

where $\sigma_X^2 = E\{|x(k)|^2\}$ is the average symbol power, and the notation E{.} depicts the expected value of a random variable. The interference power $P_i(k)$ for a given $h(k)$ becomes

$$P_i(k) = |s_0|^2|h(k)|^2E\{|i(k)|^2|h(k)\} \tag{6.5}$$

where, $E\{|i(k)|^2|h(k)\}$, the conditional power of $i(k)$ for a given $h(k)$, $r_{lk} = E\{h(l)\bar{h}(k)\}$ is the channel correlation coefficient between the lth and the kth ($k \neq l$) subcarriers. The constant average power of each subcarrier is given by $\sigma_h^2 = E\{|h(l)|^2\}$. Therefore, by solving it further the SINR providing the SER lower-bound performance can be given as [25]

$$SINR(k) = \frac{\rho|s_0|^2|h(k)|^2}{\rho(1-|s_0|^2)|h(k)|^2+1} \tag{6.6}$$

6.1.2 MIMO-OFDM Signal Model

For an MIMO-OFDM system, the post-FFT signal $y_n(k)$ at the nth receive antenna for the kth OFDM subcarrier becomes

$$y_n(k) = \frac{1}{\sqrt{N_t}}\sum_{m=0}^{N_t-1} s_0h_{mn}(k)x_m(k) + \frac{1}{\sqrt{N_t}}\sum_{m=0}^{N_t-1}\left(\sum_{l=0,l\neq k}^{N-1} s_{l-k}h_{mn}(l)x_m(l)\right) + w_n(k) \tag{6.7}$$

for $0 \leq n \leq N_r - 1$ and $0 \leq k \leq N - 1$. $x_m(k)$ is the transmit data symbol from the mth transmit antenna on the kth subcarrier. The channel response from the mth transmit antenna to nth receive antenna for the k the subcarrier is given by $h_{mn}(k)$. The factor $1/\sqrt{N_t}$ scales the total transmit power from the N_t antennas to σ^2_X. The above (6.7) can be written as

$$y_n(k) = \frac{1}{\sqrt{N_t}}\sum_{m=0}^{N_t-1} s_0h_{mn}(k)x_m(k) + z_n(k) \tag{6.8}$$

where, $z_n(k) = i_n(k)+w_n(k)$ is the total noise term consisting of the interference term $i_n(k)$ and the AWGN term $w_n(k)$. The interference term $i_n(k)$ that accounts for both inter-antenna and inter-carrier interference is given by [26]

$$i_n(k) = \frac{1}{\sqrt{N_t}} \sum_{m=0}^{N_t-1} \left(\sum_{l=0, l\neq k}^{N-1} s_{l-k} h_{mn}(l) x_m(l) \right) \tag{6.9}$$

6.1.3 Space-Time Coded OFDM

Consider an Alamouti coded OFDM system incorporating two transmit antennas and a single receive antenna as shown in Fig.6.1. An MIMO-OFDM system (i.e. wideband MIMO) can be considered as an aggregation of narrowband MIMO systems, where each subcarrier frequency provides a narrowband (flat fading) MIMO channel. Thus, the Alamouti coding can be applied to each subcarrier independently to achieve space-time diversity. Let the two symbols transmitted simultaneously from the antennas 0 and 1 on the kth subcarrier in the symbol time $t = 0$ to be $x_0(k)$ and $x_1(k)$, respectively. According to Alamouti coding, in the next symbol time $t = 1$, antennas 0 and 1 transmit the symbols $-\bar{x}_1(k)$ and $\bar{x}_0(k)$, respectively. Let the frequency response of the channel 0 and 1 for the kth subcarrier at the symbol time t to be $h^{(t)}{}_0(k)$ and $h^{(t)}{}_1(k)$, respectively. An assumption made in Alamouti coding is that $h^{(t)}{}_0(k)$ and $h^{(t)}{}_1(k)$ are constant over two consecutive symbol periods, that is

$$h_m^{(2t)}(k) = h_m^{(2t+1)}(k), \qquad m = 0, 1 \tag{6.10}$$

We assume a receiver with a single receive antenna. The post FFT receive signals $y^{(0)}(k)$ and $y^{(1)}(k)$ over the two consecutive symbol periods $t = 0$ and $t = 1$ for the kth subcarrier can be obtained using (6.7) as

$$\begin{bmatrix} y^{(0)}(k) \\ \bar{y}^{(1)}(k) \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} s_0 h_0(k) & s_0 h_1(k) \\ \bar{s}_0 \bar{h}_1(k) & -\bar{s}_0 \bar{h}_0(k) \end{bmatrix} \begin{bmatrix} x_0(k) \\ x_1(k) \end{bmatrix} + \begin{bmatrix} z^{(0)}(k) \\ \bar{z}^{(1)}(k) \end{bmatrix} \tag{6.11}$$

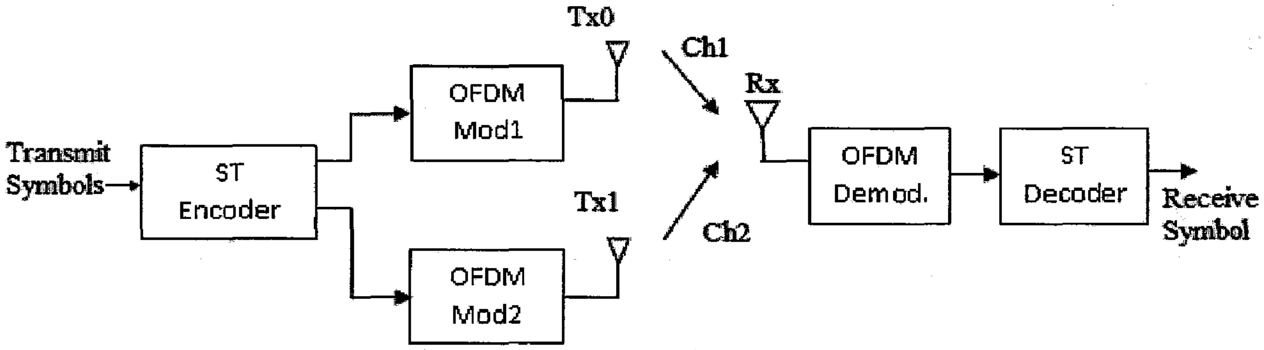


Fig.: -6.1 Block schematic of an Alamouti coded OFDM systems incorporating two transmit antennas and one receive antenna.

where, $h^{(0)}{}_0(k) = h^{(1)}{}_0(k) = h_0(k)$ and $h^{(0)}{}_1(k) = h^{(1)}{}_1(k) = h_1(k)$. The total noise terms including both ICI interference noise and AWGN channel noise are given by $z^{(0)}(k) = i^{(0)}(k) + w^{(0)}(k)$ and $z^{(1)}(k) = i^{(1)}(k) + w^{(1)}(k)$. The decision variables $\hat{x}_0(k)$ and $\hat{x}_1(k)$ derivable by combing the receive signals $y^{(0)}(k)$ and $y^{(1)}(k)$ (Alamouti decoding) becomes

$$\hat{x}_0(k) = \bar{s}_0\bar{h}_0(k)y^0(k) + s_0h_1(k)\bar{y}^{(1)}(k) \tag{6.12}$$

$$\hat{x}_1(k) = \bar{s}_0\bar{h}_1(k)y^0(k) - s_0h_0(k)\bar{y}^{(1)}(k) \tag{6.13}$$

We choose the symbol $x_0(k)$ to analyze the effect of ICI noise on the performance of the ST-coded OFDM system. Substitution of (6.11) in (6.12) gives

$$\hat{x}_0(k) = \frac{1}{\sqrt{2}}|s_0|^2[|h_0(k)|^2 + |h_1(k)|^2]x_0(k) + \bar{s}_0\bar{h}_0(k)\left[i^{(0)}(k) + w^{(0)}(k)\right]$$

$$+ s_0h_1(k)\left[\bar{i}^{(1)}(k) + \bar{w}^{(1)}(k)\right] \tag{6.14}$$

Therefore, the signal power $P_x(k)$ in the decision variable $\hat{x}_0(k)$ given $h_0(k)$ and $h_1(k)$ becomes

$$P_X(k) = \frac{1}{2}|s_0|^4[|h_0(k)|^2 + |h_1(k)|^2]^2\,\sigma_X^2 \tag{6.15}$$

The interference noise power $P_i(k)$ due to ICI becomes

$$P_i(k) = |s_0|^2|h_0(k)|^2E\left\{\left|i^{(0)}(k)\right|^2|h_0(k),h_1(k)\right\}$$

$$+ |s_0|^2|h_1(k)|^2E\left\{\left|i^{(1)}(k)\right|^2|h_0(k),h_1(k)\right\} \quad (6.16)$$

The noise power $P_w(k)$ due to channel's AWGN can be given by

$$P_w(k) = |s_0|^2[|h_0(k)|^2 + |h_1(k)|^2]\,\sigma_w^2 \quad (6.17)$$

where, $\sigma_w^2 = E\{|w|^2\}$ is the power of any noise term $w^{(i)}(k)$. Using (20), (21) and (23), the signal-to-interference-and-noise ratio (SINR) can be obtained as

$$SINR(k) = \frac{P_X(k)}{P_w(k) + P_i(k)}$$

$$= \frac{\frac{1}{2}|s_0|^2[|h_0(k)|^2 + |h_1(k)|^2]^2\,\sigma_X^2}{E\{|i(k)|^2|h_0(k),h_1(k)\} + \sigma_w^2} \quad (6.18)$$

The best SINR(k) (minimizing the SER performance) w.r.t. the channel correlation coefficients can be derived as the one that maximizes the noise terms simultaneously fading with the channel ($h_0(k)$ and $h_1(k)$) and minimizes noise terms independent of the channel. This is achieved when $q_{lk} = 1$ and $r_{lk} = 1$ for all l and k, i.e. when both channels of the Alamouti-coded OFDM system are frequency-flat fading. Under this condition

$$SINR(k) = \frac{\frac{1}{2}\rho|s_0|^2[|h_0(k)|^2 + |h_1(k)|^2]^2}{\frac{1}{2}\rho(1 - |s_0|^2)[|h_0(k)|^2 + |h_1(k)|^2] + 1} \quad (6.19)$$

where, $\sigma = \sigma_X^2/\sigma_w^2$ is the average channel SNR [26].

6.2 Simulation

The simulation is carried out for MIMO-OFDM STBC system with additive white Gaussian noise. The M PSK modulation scheme is used and complex block of data are transmitted.

As it can be seen from the graph, the performance of the system is highly depending on the Frequency offset of the carrier. Fig. 6.2 and 6.3 shows the symbol error rate comparison of different frequency offset provided. For this simulation following parameter was considered FFT length 64, cyclic prefix 16, and frame length 50. For each SNR the signal was transmitted 1000 times. 4-PSK Modulation scheme was used.

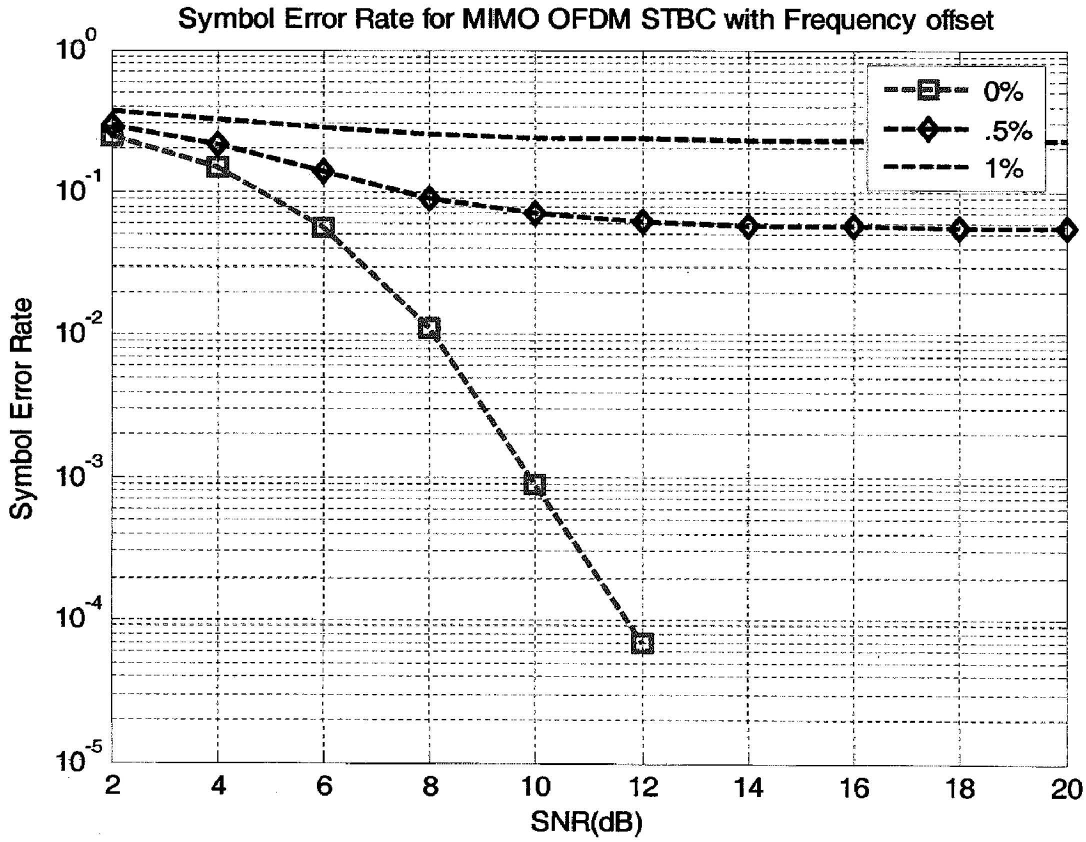


Fig.: -6.2 Symbol error rate comparison for different frequency offset given for STBC-OFDM systems.

By graph 6.2 and 6.3 it observed that the Symbol error rate increases very rapidly if there is any frequency offset in carrier at receiver side from transmitter side. Due to frequency offset the orthogonality is destroyed, and this leads to a severe performance degradation.

The frequency offset was given by generating a FFT matrix to calculate IDFT of data at transmitter side and then at receiver side a fix amount of offset is given to FFT matrix to DFT of the received data.

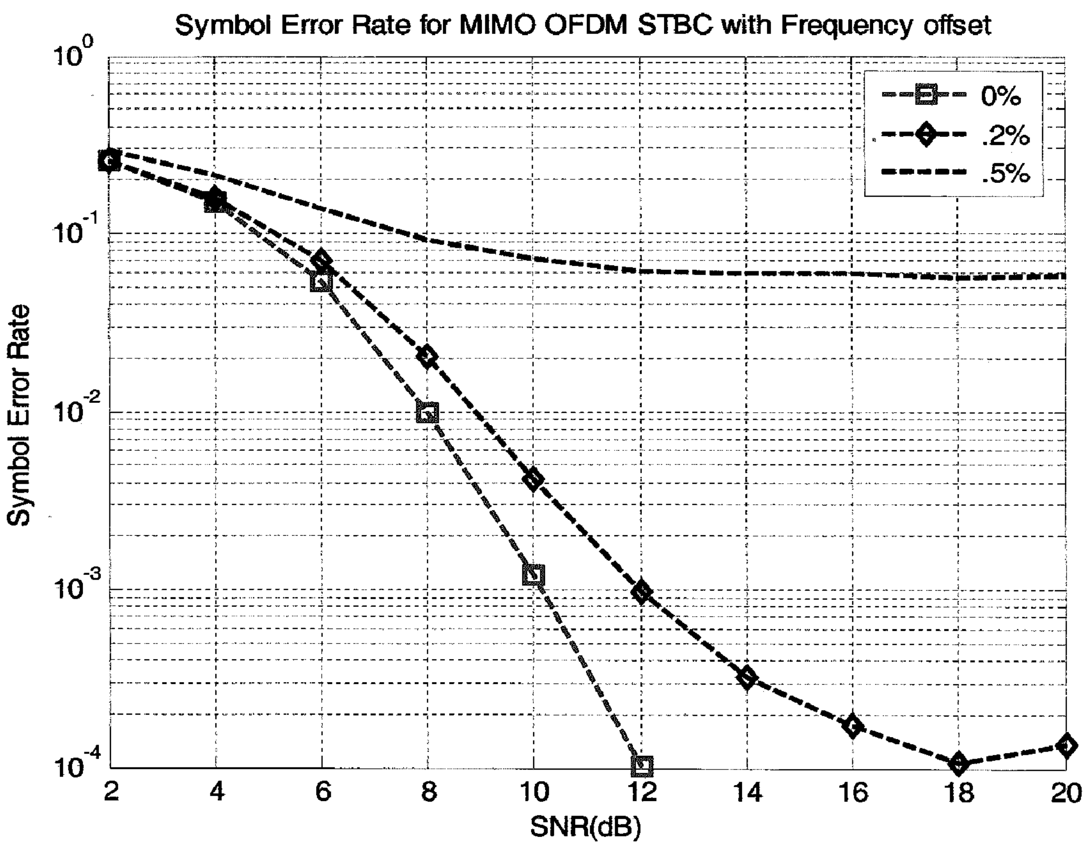


Fig.: -6.2 Symbol error rate comparison for different frequency offset given for STBC-OFDM systems.

CONCLUSIONS

The reliability of the wireless link can improve using STBC- MIMO and STBC-OFDM system. Increasing the number of antenna at both ends can enhance the reliability of the system proportional with diversity gain. This result can be achieved with no increase in transmitted power and with no cost of extra bandwidth. The robustness of this system in fading channel environment made it as a possible candidate technology for new generation of wireless system.

In Chapter 2, Different diversity schemes are discussed. Among these diversity schemes the space and time diversity becomes the base to the Space Time Code. In Chapter 3, analytically performance is analyzed for basic principle of MIMO scheme, Basic space time code (Alamouti scheme), and general space time block. And simulation results are given for Basic Alamouti scheme and Space time block code for different transmit and receive antenna.

In Chapter 4-5, performance of Space time code for frequency selective channel is analyzed. The literature review of the SISO OFDM, STBC OFDM and Space time frequency systems is given. The performance for STBC OFDM is evaluated by simulation.

In Chapter 6, an exact method of analytically evaluating the error performance of SISO-OFDM and ST-coded OFDM systems with carrier frequency offset in frequency-flat channels has presented. CFO effects for Alamouti coded OFDM systems have been analyzed. Orthogonality among the demodulated subcarriers is lost and the receiver combining introduces additional interference terms for the decision variables. Simulation results showed that STC OFDM systems are more sensitive to CFO errors. This is because the channel faded ICI is also significant in the case of space-time OFDM.

In particular, we studied following schemes (i) space-time coding for MIMO (ii) OFDM systems (iii) STBC for OFDM systems (iv) Effect of frequency offset on performance of STBC OFDM systems.

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