ANALYTICAL PERFORMANCE EVALUATION OF SPACE TIME CODED MIMO OFDM SYSTEMS IMPAIRED BY TIMING JITTER

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Abstract

Orthogonal frequency division multiplexing (OFDM) is very attractive for high bit rate wireless communications, and is effective in avoiding intersymbol interference caused by multipath delay. However, it is sensitive to time-selective fading which destroys the orthogonality among different subcarriers in one OFDM symbol leading to intercarrier interference. OFDM signal is also degraded by timing jitter caused by time synchronization errors at the receiver. An analytical approach to determine the impact of time selective fading, timing jitter and AWGN on OFDM systems with DQPSK, QPSK and DPSK modulation has been presented in this dissertation. The BER performance results are evaluated for different values of fading and jitter variance. The performance of the OFDM system in Rayleigh and Rician fading channels is also compared.

Multiple antennas can be combined with OFDM to increase the diversity gain and to improve the spectral efficiency through spatial multiplexing and space-time coding. Similar to single-antenna OFDM, Space time block coded (STBC) OFDM suffers from significant performance degradation due to noise, jitter and time-selective fading. Analytical approach is also developed to evaluate the BER performance of a quasi-orthogonal STBC-OFDM having multiple transmitting and single receiving antennas. The analysis is extended for a MIMO-OFDM system using the "selection method" for combining multiple receiving antennas, which offers significant improvement in the system performance. The effects of increase in number of OFDM subcarriers and increase in Doppler frequency are also investigated. Performance improvement in all the above mentioned cases is also observed when appropriate convolution coding is applied.

The computed results show that for both OFDM and STBC-OFDM, DPSK and QPSK systems suffer higher amount of power penalty than DQPSK system due to the effect of jitter. It is also noticed that the QPSK system suffers almost the same amount of power penalty as DPSK system for lower values of jitter variance and at higher values; DPSK suffers more penalty than QPSK. Numerical computations show the improvement in the BER performance in STBC-OFDM from the stand alone OFDM system. It is also found that the power penalty due to jitter is reduced in the STBC-OFDM system. For QPSK modulation, at a jitter variance of 0.2, the penalty at a BER of 10^{-8} is 8.5 dB for OFDM and 6dB for STBC-OFDM.

Chapter 1

Introduction

1.1 Introduction to Communication Systems

The exchange of thoughts, messages, or information, as by speech, signals, writing, or behavior is known as communication. Any transmission, emission, or reception of signal by wire, radio, visual, optical or other electromagnetic systems are known as electrical communication. Electrical communication systems are designed to send messages or information from a source that generates the messages to one or more destinations. The heart of the communication system consists of three basic parts, namely, the transmitter, the channel, and the receiver as shown in Fig. 1.1.

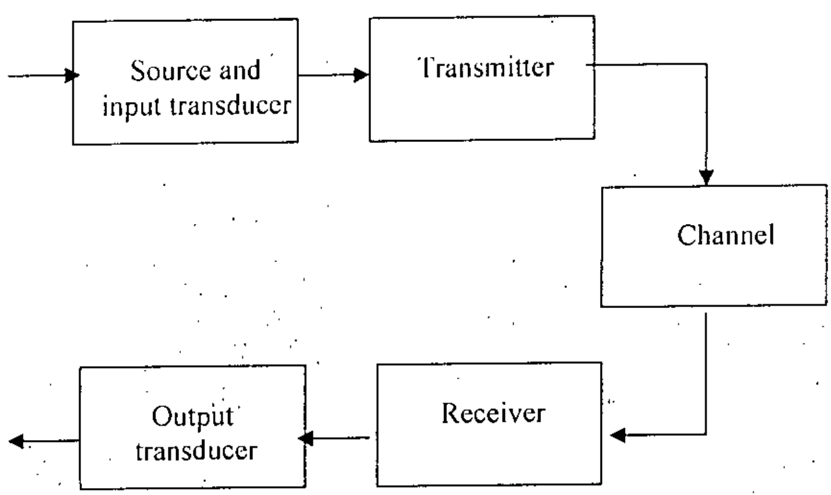

Fig. 1.1 Functional block diagram of a communication system

The transmitter converts the electrical signal into a format that is suitable for transmission through the physical channel or transmission medium. The transmitter must translate the information signal into the appropriate frequency range that matches the frequency allocation assigned to the transmitter. In general, it performs the matching of the message to the channel by a process called modulation. In addition to modulation the other functions that are performed by the transmitter are filtering of the information-bearing signal, amplification of the modulated signal, and in the case of wireless transmission, radiation of the signal by means of a transmitting antenna.

The function of the receiver is to recover the message signal contained in the received signal. If the message signal is transmitted by carrier modulation, the receiver performs carrier demodulation in order to extract the message from the sinusoidal carrier. Since the signal demodulation is performed in the presence of additive noise and possibly other signal distortion, the demodulated message signal is generally degraded to some extent by the presence of these distortions in the received signal.

The communication channel is the physical medium that is used to send the signal from the transmitter to the receiver. The telephone network makes extensive use of wire lines for voice transmission as well as data and video transmission. Twisted pair wirelines and coaxial cable are basically guided electromagnetic channels that provide relatively modest bandwidths. Telephone wire has a bandwidth of several hundred kilohertz, whereas coaxial cable has a bandwidth of several megahertz. Fiber-optic communication is a method of transmitting information from one place to another by sending light through an optical fiber. The light forms an electromagnetic carrier wave that is modulated to carry information. Optical fibers offer a channel bandwidth that is several orders of magnitude larger than coaxial cable channels. Fiber-optic cable is used by many telecommunications companies to transmit telephone signals, Internet communication, and cable television signals, sometimes all on the same optical fiber. Due to much lower attenuation and interference, optical fiber has large advantages over existing copper wire in long-distance and high-demand applications. However, infrastructure development within cities is relatively difficult and time-consuming, and fiber-optic systems are complex and expensive to install and operate. Due to these difficulties, fiber-optic communication systems have primarily been installed in long-distance applications, where they can be used to their full transmission capacity, offsetting the increased cost.

There are also underwater ocean channels in which the information-bearing signal is transmitted acoustically. Electromagnetic waves do not propagate over long distances underwater except at extremely low frequencies. Although acoustics has been used effectively for point-to-point communications in vertical deep-water channels, acoustics has had limited success in shallow water. Effects such as time-varying multi-path propagation and non-Gaussian noise are two of the major factors that limit acoustic communications in shallow water. Power line communication is a system for using electric power lines to carry radio signals for communication purposes.

Some media that can be characterized as special communication channels are data storage media, such as magnetic tape, magnetic disks, and optical disks. In wireless transmission the channel is usually the free space over which the signal is radiated by the use of an antenna [1].

1.2 Historical Review

One of the earliest inventions of major significance to communications was the invention of the electric battery by Alessandro Volta in 1799. This invention made it possible for Samuel Morse to develop the electric telegraph, which he demonstrated in 1837. Telephony came into being with the invention of the telephone in the 1870s. Alexander Graham Bell patented his invention in 1876. The development of wireless communications stems from the work of Oersted, faraday, Gauss, Maxwell, and Hertz. James C. Maxwell in 1864 predicted the existence of electromagnetic radiation and formulated the basic that has been in use for over a century. In 1894 a sensitive device that could detect radio signals, was used by its inventor Oliver Lodge to demonstrate wireless communication over a distance of 150 yards at Oxford, England. Guglielmo Marconi is credited with the development of wireless telegraphy. Marconi demonstrated the transmission of radio signals at a distance of approximately 2kms in 1895. The invention of vacuum tube was especially instrumental in the development of radio communication systems. Fleming invented the vacuum tube in 1904 and the vacuum triode amplifier was invented by De Forest in 1906. The invention of triode made radio broadcast possible in the early part of the twentieth century [1].

1.3 General Features of Wireless Communication

The term wireless is normally used to refer to any type of electrical or electronic operation that is accomplished without the use of a "hard wired" connection. Wireless communication is the transfer of information over a distance without the use of electrical conductors or "wires". The distances involved may be short (a few meters as in television remote control) or very long (thousands or even millions of kilometers for radio communications). When the context is clear the term is often simply shortened to "wireless". Wireless communications is generally considered to be a branch of telecommunications. Wireless communication may be via:

- radio frequency communication,
- microwave communication, for example long-range line-of-sight via highly directional antennas, or short-range communication, or
- infrared (IR) short-range communication, for example from remote controls.

Applications may involve point-to-point communication, point-to-multipoint communication, broadcasting, cellular networks and other wireless networks. The following situations justify the use of wireless technology:

- To span a distance beyond the capabilities of typical cabling,
- To avoid obstacles such as physical structures, EMI, or RFI,
- To provide a backup communications link in case of normal network failure,
- To link portable or temporary workstations,
- To overcome situations where normal cabling is difficult or financially impractical, or
- To remotely connect mobile users or networks.

In wireless communication an important feature is that the transmitted signal is corrupted in a random manner by a variety of possible mechanisms. The most common form of signal degradation comes in the form of additive noise, which is generated at the front end of the receiver where signal amplification is performed. This noise is often called thermal noise. In wireless transmission, additional additive disturbance are man-made noise, the atmospheric noise picked by a receiving antenna. Interference from other users of the channel is another form of additive noise that often arises in both wireless and wireline communication systems. However, in the mobile radio environment, signals are usually impaired by fading and multipath delay spread phenomenon. In such channels, severe fading of the signal amplitude and ISI due to the frequency selectivity of the channel lead to unacceptable degradation of the system error performance. Channel coding and adaptive equalization techniques have been widely used in the single carrier mobile communication systems to combat fading and multipath propagation. However, due to the inherent delay in the coding and equalization process and high cost of the hardware, there are practical difficulties to use these techniques in systems operating at high bit rates.

1.4 OFDM and Other Multiplexing Techniques

Modern telephone networks allow bandwidths in their channels that are much larger than those needed for a digitalized telephone channel. Basically, a number of channels share a common transmission medium with the aim of reducing costs and complexity in the network. Multiplexing is defined as the process by which several signals from different channels share a channel with greater capacity. When the sharing is carried out with respect to a remote resource, such as a satellite, this is referred to as multiple access rather than multiplexing. There are various ways of performing this sharing such as:

- FDM/FDMA (Frequency Division Multiplexing/Frequency Division Multiple Access)
- TDM/TDMA (Time Division Multiplexing/Time Division Multiple Access)
- CDMA (Code Division Multiple Access)
- PDMA (Polarization Division Multiple Access)
- SDMA (Space Division Multiple Access)
- CSMA (Carrier sense multiple access)

There is another multiplexing technique known as Orthogonal Frequency Division Multiplexing (OFDM) which is in some respect similar to conventional FDM. The difference lies in the way in which the signals are modulated and demodulated. Priority is given to minimizing the interference, or crosstalk, among the channels and symbols comprising the data stream. OFDM spread spectrum technique distributes the data over a large number of carriers that are spaced apart at precise frequencies. This spacing provides the "orthogonality" in this technique which prevents the demodulators from seeing frequencies other than their own.

1.5 Review of Previous Works

Several research work have been carried out during the last few years on the performance evaluation of an OFDM system both analytically and also by simulations [12]-[22]. In this section, a partial review of these works is presented.

The BER performance results for OFDM system over fading channels are reported in [12]-[16]. The performance of non-coherently detected BFSK/OFDM over multipath fading channels with noise is investigated in [12]. This analysis demonstrates that it is possible to transmit information in selective channels with no symbol interference. Expressions of the Bit Error Probability (BEP) are derived in the context of frequency selective Rayleigh and frequency selective Rician fading channels with and without convolutional coding.

Ref. [13] proposes an approximate derivation method of the bit error rate (BER) in DQPSK/OFDM systems over frequency non-selective Nakagami-Rice and Rayleigh fading channels. The validity of this equation is confirmed from the fact that the BER derived from this approximate equation coincides with that from the computer simulation, even when the system parameters, for example Doppler frequency, Rician parameter and so on, are varied.

The performance of OFDM has been evaluated in fading channels exhibiting both time-selectivity and frequency-selectivity in [14]. Investigation is also performed to find how various parameters, such as the number of carriers and the guard time length affect the system performance. Further, the optimum values of the above parameters, which minimize the degradation of the signal-to-noise ratio at the input of the decision device, are determined.

The maximum a posteriori probability (MAP) receiver is derived for OFDM signals in a fading channel in [15]. As the complexity of the MAP receiver is high, a low-complexity, suboptimal receiver is obtained and its performance is also evaluated.

Ref. [16] proposes a simple calculation method, that is, an approximate closed-form equation of the bit error rate (BER) in DPSK/OFDM systems mentioned above over both time and frequency selective Rician fading channels. In orthogonal frequency division multiplexing (OFDM) systems with differential phase shift keying (DPSK), it is possible to apply differential modulation either in the time or frequency domain depending on the condition of fading channels, such as the Doppler frequency shift and the delay spread. The validity of the proposed method is demonstrated by the fact that the BER performances given by the derived equation coincide with those by Monte Carlo simulation.

The effect of timing jitter on the performance of an OFDM system is also reported in [17], [18]. The bit error rate (BER) degradation is evaluated in OFDM systems where the sampling instant is affected by timing jitter in [17]. The timing jitter is modeled by a stationary random process with a known statistic and the error caused by it is described as additive noise. This decomposition is useful because it allows for the analytical determination of the BER. The method may be applied for input data sequences with and without code.

The effect of random jitter in the sampling circuit at the receiver in an OFDM communication system is studied in [18]. It is shown that the effect of jitter can be looked at as interference between different sub-carriers and derive a lower bound for the variance of the interference in terms of the eigenvalues of the covariance matrix of the jitter process. The obtained results are compared with simulation results. The effect of over sampling on the interference is described and explanation is provided for the way interference varies with sub-carrier index for different sampling rates.

A simple two-branch transmit diversity scheme is presented in [19]. Using two transmit antennas and one receive antenna the scheme provides the same diversity order as maximal-ratio receiver combining (MRRC) with one transmit antenna, and two receive antennas. It is also shown that the scheme may easily be generalized to two transmit antennas and M receive antennas to provide a diversity order of 2M. The new scheme does not require any bandwidth expansion, any feedback from the receiver to the transmitter and its computation complexity is similar to MRRC.

The theory of orthogonal space-time block code (OSTBC) for wireless fading environment is developed in [20]. Data is encoded using a space-time block code and the encoded data is split into n streams which are simultaneously transmitted using n transmit antennas. The received signal at each receive antenna is a linear superposition of the n transmitted signals perturbed by noise. Maximum-likelihood decoding is achieved in a simple way through decoupling of the signals transmitted from different antennas rather than joint detection.

In Ref. [21], it has been shown that a complex orthogonal design that provides full diversity and full transmission rate for a space-time block code is not possible for more than two antennas. Previous attempts have been concentrated in generalizing orthogonal designs, which provide space-time block codes with full diversity and a high transmission rate. This paper modifies the complex orthogonal codes to quasi-orthogonal codes, which have rate one and provide partial diversity. The decoder of the proposed codes works with pairs of transmitted symbols instead of single symbols. However the work is for single-carrier system and the analytical BER performance results are not reported.

The performance of QOSTC OFDM system over fast fading channel is investigated in terms of carrier to interference (C/I) and signal to interference and noise ratios in [22]. Comparison is also made on five different detection schemes and their SER floors are also evaluated through simulations. However the jitter effect is ignored and only a single receiving antenna is used.

1.6 Objectives of the Thesis

The objectives of this research work are:

- To carryout the bit error rate (BER) performance analysis of an OFDM system considering all three channel impairments like AWGN, fading and timing jitter in Rayleigh and Rician fading environments.
- To develop an analytical approach in order to find the BER of a QOSTC OFDM system considering all those channel impairments.
- To extend the analysis for a MIMO-OFDM system with switching/selection method for combining multiple receiving antennas and to evaluate the improvement in QOSTC OFDM system performance.
- To extend the analysis for the OFDM and MIMO-OFDM systems with convolution coding to evaluate the effect of coding on system performance in presence of channel impairments.

- To evaluate the performance results by numerical computation and to find the optimum system design parameters.

1.7 Contribution of this Work

The mathematical expression of signal-to-noise plus interference ratio (SNIR) for an OFDM system is derived considering the combined effects of AWGN, fading and timing jitter. In the SNIR expression, both the channel attenuation and Doppler frequency shift are considered as fading effect. With the expression of SNIR, the BER of the OFDM system affected by AWGN, fading and jitter is found. Similarly the SNIR and BER are evaluated for STBC-OFDM and MIMO-OFDM systems.

1.8 Organization of the Thesis

Chapter 1 gives a brief overview of a communication system and description of different impairments in wireless communication link. The background and objective of the thesis are also presented in chapter 1. Chapter 2 gives a comparative description of OFDM with different multiplexing/multiple access technologies. The concept of OFDM, STBC and convolution coding is also described in this chapter.

In chapter 3 the performance of OFDM and MIMO-OFDM is evaluated. In this chapter we present the block diagrams of the systems under consideration. We evaluate the different interferences caused by fading in an OFDM system. Then we examine the BER performance of the system in presence of AWGN, fading and timing jitter. We also perform the same analysis for STBC-OFDM system having only transmit diversity. We extend our analysis by applying diversity combining into the receiving side.

In chapter 4 we perform numerical computations on the expressions of the systems described in chapter 3. Different BER plots are shown and results are presented and compared. Chapter 5 concludes the thesis by discussing some of the expected problems and scope for further research.

Chapter 2

Overview of OFDM and MIMO Technology

This chapter highlights different multiplexing techniques and technical details of OFDM technology. The overview of diversity, STBC and MIMO technology are also presented in this chapter.

2.1 Different Multiplexing and Multiple Access Techniques

The various ways of performing multiplexing is already mentioned in Chapter 1. The comparative description of these schemes are mentioned below:

FDM/FDMA (Frequency Division Multiplexing/Frequency Division Multiple Access): Assigns a portion of the total bandwidth to each of the channels.

TDM/TDMA (Time Division Multiplexing/Time Division Multiple Access): Assigns all of the transport capacity sequentially to each of the channels.

CDMA (Code Division Multiple Access): In certain circumstances it is possible to transmit multiple signals in the same frequency and at the same time, with the receiver being responsible for separating them. This technique has been used for years in military technology, and is based on extending the spectrum of the signal and reducing the transmission power.

PDMA (Polarization Division Multiple Access): Given that polarization can be maintained, the polarization direction can be used as a multiple access technique, although when there are many obstacles noise can make it unsuitable, which is why it is not usually used in indoor installations. Outside, however, it is widely used to increase transmission rates in installations that use microwaves.

SDMA (Space Division Multiple Access): With directional aerials, the same frequency can be re-used provided the alignment of the aerials is correctly adjusted. There is a great deal of interference but this system lets frequencies obtain a high degree of reusability.

CSMA (Carrier sense multiple access): Carrier Sense Multiple Access (CSMA) is a probabilistic Media Access Control (MAC) protocol in which a node verifies the absence of other traffic before transmitting on a shared physical medium, such as an electrical bus, or a band of electromagnetic spectrum. In CSMA multiple nodes send and receive on the medium. Transmissions by one node are generally received by all other nodes using the medium.

2.1.1 Code Division Multiple Access (CDMA) describes a communication channel access principle that employs spread-spectrum technology and a special coding scheme (where each transmitter is assigned a code). By contrast, time division multiple access (TDMA) divides access by time, while frequency-division multiple access (FDMA) divides it by frequency. CDMA is a form of "spread-spectrum" signaling, since the modulated coded signal has a much higher bandwidth than the data being communicated.

In radio CDMA, each group of users is given a shared code. Many codes occupy the same channel, but only users associated with a particular code can understand each other. CDMA is characterized by high capacity and small cell radius. CDMA has been used in many communications and navigation systems, including the Global Positioning System and the OmniTRACS satellite system for transportation logistics [2].

Features:

- Narrowband message signal multiplied by wideband spreading signal or pseudonoise code
- Each user has his own pseudonoise (PN) code
- Soft capacity limit: system performance degrades for all users as number of users increases
- Cell frequency reuse: no frequency planning needed
- Soft handoff increases capacity
- Near-far problem
- Interference limited: power control is required

2.1.2 Time Division Multiple Access (TDMA) is a channel access method for shared medium (usually radio) networks. It allows several users to share the same frequency channel by dividing the signal into different timeslots. The users transmit in rapid succession, one after the other, each using his own timeslot. This allows multiple stations to share the same transmission medium (e.g. radio frequency channel) while using only the part of its bandwidth they require. TDMA is used in the digital 2G cellular systems such as Global System for Mobile Communications (GSM), in satellite systems, and combat-net radio systems.

TDMA is a type of Time-division multiplexing, with the special point that instead of having one transmitter connected to one receiver, there are multiple transmitters. In the case of the uplink from a mobile phone to a base station this becomes particularly difficult because the mobile phone can move around and vary the timing advance required to make its transmission match the gap in transmission from its peers. The difference between time-division multiplexing (TDM) and time-division multiple access is that time-division multiplexing requires users to be collocated to be multiplexed into the channel. In that regard, time-division multiple access can be considered as a remote multiplexing technology [3].

TDMA Features:

- Shares single carrier frequency with multiple users
- Non-continuous transmission makes handoff simpler
- Slots can be assigned on demand in dynamic TDMA
- Less stringent power control than CDMA due to reduced intra cell interference
- Higher synchronization overhead than CDMA
- Advanced equalization is necessary for high data rates
- Cell breathing (borrowing resources from adjacent cells) is more complicated than in CDMA

Comparison of TDMA with other multiple-access schemes:

In radio systems, TDMA is usually used alongside Frequency-division multiple access (FDMA) and Frequency division duplex (FDD); the combination is referred to as FDMA/TDMA/FDD. This is the case in both GSM and IS-136 for example. A major advantage of TDMA is that the radio part of the mobile only needs to listen and broadcast for its own timeslot. For the rest of the time, the mobile can carry out measurements on the network, detecting surrounding transmitters on different frequencies. This allows safe inter frequency handovers, something which is difficult in CDMA systems.

CDMA, by comparison, supports "soft hand-off" which allows a mobile phone to be in communication with up to 6 base stations simultaneously, a type of "same-frequency handover". The incoming packets are compared for quality, and the best one is selected. CDMA's "cell breathing" characteristic, where a terminal on the boundary of two congested cells will be unable to receive a clear signal, can often negate this advantage during peak periods.

A disadvantage of TDMA systems is that they create interference at a frequency which is directly connected to the timeslot length. This is the irritating buzz which can sometimes be heard if a GSM phone is left next to a radio or speakers. Another disadvantage is that the "dead time" between timeslots limits the potential bandwidth of a TDMA channel. These are implemented in part because of the difficulty ensuring that different terminals transmit at exactly the times required. Handsets that are moving will need to constantly adjust their timings to ensure their transmission is received at precisely the right time, because as they move further from the base station, their signal will take longer to arrive. This also means that the major TDMA systems have hard limits on cell sizes in terms of range, though in practice the power levels required to receive and transmit over distances greater than the supported range would be mostly impractical anyway [3].

2.1.3 Frequency Division Multiple Access or FDMA is an access technology that is used by radio systems to share the radio spectrum. The terminology "multiple access" implies the sharing of the resource amongst users, and the "frequency division" describes how the sharing is done: by allocating users with different carrier frequencies of the radio spectrum. In an FDMA scheme, the given Radio Frequency (RF) bandwidth is divided into adjacent frequency segments. Each segment is provided with bandwidth to enable an associated communications signal to pass through a transmission environment with an acceptable level of interference from communications signals in adjacent frequency segments [4].

2.2 Fundamentals of OFDM:

OFDM Overview

OFDM is a modulation technique where multiple low data rate carriers are combined by a transmitter to form a composite high data rate transmission. Digital signal processing makes OFDM possible. The basic idea behind this scheme is to spread out the effect of a fade over many bits. Rather than having a few adjacent bits completely destroyed, we now have all the bits only slightly affected by a fade.

Each carrier in an OFDM system is a sinusoid with a frequency that is an integer multiple of a base or fundamental sinusoid frequency. Therefore, each carrier is like a Fourier series component of the composite signal. An OFDM signal is created in the frequency domain, and then transformed into the time domain via the Discrete Fourier Transform (DFT). Two periodic signals are orthogonal when the integral of their product, over one period, is equal to zero. The carriers of an OFDM system are sinusoids that meet this requirement because each one is a multiple of a fundamental frequency. Each one has an integer number of cycles in the fundamental period [5].

Preliminary Concepts

When the DFT (Discrete Fourier Transform) of a time signal is taken, the frequency domain results are a function of the time sampling period and the number of samples. The fundamental frequency of the DFT is equal to 1/NT (1/total sample time). Each frequency represented in the DFT is an integer multiple of the fundamental frequency. The maximum frequency that can be represented by a time signal sampled at rate 1/T is fmax = 1/2T as given by the Nyquist sampling theorem. This frequency is located in the center of the DFT points. All frequencies beyond that point are images of the representative frequencies. The maximum frequency bin of the DFT is equal to the sampling frequency (1/T) minus one fundamental (1/NT).

The IDFT (Inverse Discrete Fourier Transform) performs the opposite operation to the DFT. It takes a signal defined by frequency components and converts them to a time signal. The parameter mapping is the same as for the DFT. The time duration of the IDFT time signal is equal to the number of DFT bins (N) times the sampling period (T). It is perfectly valid to generate a signal in the frequency domain, and convert it to a time domain equivalent for practical use. This is how modulation is applied in OFDM. The frequency domain is a mathematical tool used for analysis. Anything usable by the real world must be converted into a real, time domain signal. In practice the Fast Fourier Transform (FFT) and IFFT are used in place of the DFT and IDFT [5].

Orthogonality

In OFDM, the sub-carrier frequencies are chosen so that the sub-carriers are orthogonal to each other, meaning that cross-talk between the sub-channels is eliminated and inter-carrier guard bands are not required. This greatly simplifies the design of both the transmitter and the receiver; unlike conventional FDM, a separate filter for each sub-channel is not required. The orthogonality also allows high spectral efficiency, near the Nyquist rate. Almost the whole available frequency band can be utilized. OFDM generally has a nearly 'white' spectrum, giving it benign electromagnetic interference properties with respect to other co-channel users.

The orthogonality allows for efficient modulator and demodulator implementation using the FFT algorithm. OFDM requires very accurate frequency synchronization between the receiver and the transmitter; any deviation and the sub-carriers are no longer orthogonal, causing inter-carrier interference (ICI), i.e. cross-talk between the sub-carriers. Frequency offsets are typically caused by mismatched transmitter and receiver oscillators, or by Doppler shift due to movement. Whilst Doppler shift alone may be compensated for by the receiver, the situation is worsened when combined with multipath, as reflections will appear at various frequency offsets, which is much harder to correct. This effect typically worsens as speed increases, and is an important factor limiting the use of OFDM in high-speed vehicles. Several techniques for ICI suppression are suggested, but they may increase the receiver complexity [5].

Guard Period

One key principle of OFDM is that since low symbol rate modulation schemes (i.e. where the symbols are relatively long compared to the channel time characteristics) suffer less from intersymbol interference caused by multipath, it is advantageous to transmit a number of low-rate streams in parallel instead of a single high-rate stream. Since the duration of each symbol is long, it is feasible to insert a guard interval between the OFDM symbols, thus eliminating the intersymbol interference. The guard interval also eliminates the need for a pulse-shaping filter, and it reduces the sensitivity to time synchronization problems.

The cyclic prefix, which is transmitted during the guard interval, consists of the end of the OFDM symbol copied into the guard interval, and the guard interval is transmitted followed by the OFDM symbol. The reason that the guard interval consists of a copy of the end of the OFDM symbol is so that the receiver will integrate over an integer number of sinusoid cycles for each of the multipaths when it performs OFDM demodulation with the FFT [5].

Multipath Characteristics

OFDM avoids frequency selective fading and ISI by providing relatively long symbol periods for a given data rate. For a given transmission channel and a given source data rate, OFDM can provide better multipath characteristics than a single carrier. However, since the OFDM carriers are spread over a frequency range, there still may be some frequency selective attenuation on a time-varying basis. A deep fade on a particular frequency may cause the loss of data on that frequency for a given time, but the use of Forward Error Coding can fix it. If a single carrier experienced a deep fade, too many consecutive symbols may be lost and correction coding may be ineffective.

OFDM is more bandwidth efficient than a single carrier. Another efficient aspect of OFDM is that a single transmitter's bandwidth can be increased incrementally by addition of more adjacent carriers. In addition, no bandwidth buffers are needed between transmit bandwidths of separate transmitters as long as orthogonality can be maintained between all the carriers [5].

Simplified equalization

The effects of frequency-selective channel conditions, for example fading caused by multipath propagation, can be considered as constant (flat) over an OFDM sub-channel if the sub-channel is sufficiently narrow-banded, i.e. if the number of sub-channels is sufficiently large. This makes equalization far simpler at the receiver in OFDM in comparison to conventional single-carrier modulation. The equalizer only has to multiply each sub-carrier by a constant value, or a rarely changed value. Some of the sub-carriers in some of the OFDM symbols may carry pilot signals for measurement of the channel conditions, i.e. the equalizer gain for each sub-carrier. Pilot signals may also be used for synchronization. If differential modulation such as DPSK or DQPSK is applied to each sub-carrier, equalization can be completely omitted, since these schemes are insensitive to slowly changing amplitude and phase distortion [5].

Physical Implementation

Since OFDM is carried out in the digital domain, there are many ways it can be implemented. Some options are provided in the following list. Each of these options should be viable given current technology [5]:

1. *ASIC (Application Specific Integrated Circuit)*

- ASICs are the fastest, smallest, and lowest power way to implement OFDM
- Cannot change the ASIC after it is built without designing a new chip

2. *General-purpose Microprocessor or MicroController*

- PowerPC 7400 or other processor capable of fast vector operations
- Highly programmable
- Needs memory and other peripheral chips
- Uses the most power and space, and would be the slowest

3. *Field-Programmable Gate Array (FPGA)*

- An FPGA combines the speed, power, and density attributes of an ASIC with the programmability of a general-purpose processor.
- An FPGA could be reprogrammed for new functions by a base station to meet future (currently unknown requirements).
- This should be the best choice

Orthogonal Frequency Division Multiple Access (OFDMA) is a multi-user version of the popular OFDM digital modulation scheme. Multiple access is achieved in OFDMA by assigning subsets of subcarriers to individual users. This allows simultaneous low data rate transmission from several users. Based on feedback information about the channel conditions, adaptive user-to-subcarrier assignment can be achieved. If the assignment is done sufficiently fast, this further improves the OFDM robustness to fast fading and narrow-band cochannel interference, and makes it possible to achieve even better system spectral efficiency. Different number of sub-carriers can be assigned to different users, in view to support differentiated Quality of Service (QoS), i.e. to control the data rate and error probability individually for each user.

OFDMA resembles code division multiple access (CDMA) spread spectrum, where users can achieve different data rates by assigning a different code spreading factor or a different number of spreading codes to each user. OFDMA can also be described as a combination of frequency domain and time domain multiple access, where the resources are partitioned in the time-frequency space, and slots are assigned along the OFDM symbol index as well as OFDM sub-carrier index [6].

Applications

OFDM applications include the following:

- Digital Audio Broadcasting (DAB), wireless CD-quality sound transmission
- Digital Video Broadcasting (DVB), specifically, Digital Terrestrial Television Broadcasting (DTTB)
- Wireless LAN: (IEEE 802.11a) and Hiper LAN/2
- Wireless MAN: (IEEE 802.16) WiMAX
- ADSL (Asymmetric Digital Subscriber Line), also called DMT (Digital Multi-Tone)

Advantages

- Can easily adapt to severe channel conditions without complex equalization
- Robust against narrow-band co-channel interference
- Robust against intersymbol interference (ISI) and fading caused by multipath propagation
- High spectral efficiency
- Efficient implementation using FFT
- Low sensitivity to time synchronization errors
- Tuned sub-channel receiver filters are not required (unlike conventional FDM)
- Facilitates Single Frequency Networks, i.e. transmitter macrodiversity.

Disadvantages

- Sensitive to Doppler shift.
- Sensitive to frequency synchronization problems.
- Inefficient transmitter power consumption, due to linear power amplifier requirement.

2.3 Bluetooth

2.3.1 Bluetooth Overview

Bluetooth is an industrial specification for wireless personal area networks (PANs). Bluetooth provides a way to connect and exchange information between devices such as mobile phones, laptops, PCs, printers, digital cameras, and video game consoles over a secure, globally unlicensed short-range radio frequency. The Bluetooth specifications are developed and licensed by the Bluetooth Special Interest Group (SIG).

Bluetooth is analogous to USB, and is acceptable for situations when two or more devices are in proximity to each other and don't require high bandwidth. Bluetooth also simplifies the discovery and setup of services. Bluetooth devices advertise all services they provide. This makes the utility of the service that much more accessible, without the need to worry about network addresses, permissions and all the other considerations that go with typical networks.

Bluetooth is a radio standard and communications protocol primarily designed for low power consumption, with a short range (power-class-dependent: 1 meter, 10 meters, 100 meters) based on low-cost transceiver microchips in each device. Bluetooth lets these devices communicate with each other when they are in range. The devices use a radio communications system, so they do not have to be in line of sight of each other, and can even be in other rooms, as long as the received transmission is powerful enough. The data rate achieved is 1 Mbps, 3Mbps, 53-480 Mbps. In order to use Bluetooth, a device must be compatible with certain Bluetooth profiles. These define the possible applications and uses [7].

Applications:

More prevalent applications of Bluetooth include:

- Wireless control of and communication between a mobile phone and a hands-free headset or car kit. This was one of the earliest applications to become popular.
- Wireless networking between PCs in a confined space and where little bandwidth is required.
- Wireless communications with PC input and output devices, the most common being the mouse, keyboard and printer.
- Replacement of traditional wired serial communications in test equipment, GPS receivers, medical equipment, bar code scanners, and traffic control devices.
- For controls where infrared was traditionally used.
- Sending small advertisements from Bluetooth enabled advertising hoardings to other, discoverable, Bluetooth devices.

2.3.2 Bluetooth vs Wi-Fi in Networking

Bluetooth differs from Wi-Fi in that the latter provides higher throughput and covers greater distances, but requires more expensive hardware and higher power consumption. They use the same frequency range, but employ different multiplexing schemes. While Bluetooth is a cable replacement for a variety of applications, Wi-Fi is a cable replacement only for local area network access. Bluetooth is often thought of as wireless USB, whereas Wi-Fi is wireless Ethernet, both operating at much lower bandwidth than the cable systems they are trying to replace [7].

2.4 Diversity Techniques in Wireless Communication

2.4.1 Diversity Schemes

In telecommunications, a diversity scheme refers to a method for improving the reliability of a message signal by utilizing two or more communication channels with different characteristics. Diversity plays an important role in combating fading and co-channel interference and avoiding error bursts. It is based on the fact that individual channels experience different levels of fading and interference. Multiple versions of the same signal may be transmitted and/or received and combined in the receiver. Alternatively, a redundant forward error correction code may be added and different parts of the message transmitted over different channels. Diversity techniques may exploit the multipath propagation, resulting in a diversity gain, often measured in decibels. The following classes of diversity schemes can be identified [8]:

Time diversity: Multiple versions of the same signal are transmitted at different time instants. Alternatively, a redundant forward error correction code is added and the message is spread in time by means of bit-interleaving before it is transmitted. Thus, error bursts are avoided, which simplifies the error correction.

Frequency diversity: The signal is transferred using several frequency channels or spread over a wide spectrum that is affected by frequency-selective fading. Examples are:

- OFDM modulation in combination with subcarrier interleaving and forward error correction
- Spread spectrum, for example frequency hopping or DS-CDMA.

Space diversity: The signal is transferred over several different propagation paths. In the case of wired transmission, this can be achieved by transmitting via multiple wires. In the case of wireless transmission, it can be achieved by antenna diversity using multiple transmitter antennas (transmit diversity) and/or multiple receiving antennas (diversity reception). In the latter case, a diversity combining technique is applied before further signal processing takes place. If the antennas are at far distance, for example at different cellular base station sites or WLAN access points, this is called macrodiversity. If the antennas are at a distance in the order of one wavelength, this is called microdiversity. A special case is phased antenna arrays, which also can be utilized for beamforming, MIMO channels and Space–time coding (STC).

Polarization diversity: Multiple versions of a signal are transmitted and received via antennas with different polarization. A diversity combining technique is applied on the receiver side.

Multiuser diversity: Multiuser diversity is obtained by opportunistic user scheduling at either the transmitter or the receiver. Opportunistic user scheduling is as follows that the transmitter selects the best user among candidate receivers according to qualities of each channel between the transmitter and each receiver. In FDD systems, a receiver must feed back the channel quality information to the transmitter with the limited level of resolution.

Antenna diversity: transmitted along different propagation paths.

2.4.2 STBC Overview

Space–time block coding is a technique used in wireless communications to transmit multiple copies of a data stream across a number of antennas and to exploit the various received versions of the data to improve the reliability of data-transfer. The fact that transmitted data must traverse a potentially difficult environment with scattering, reflection, refraction and so on as well as can be corrupted by thermal noise in the receiver means that some of the received copies of the data will be better than others. This redundancy results in a higher chance of being able to use one or more of the received copies of the data to correctly decode the received signal. In fact, space–time coding combines all the copies of the received signal in an optimal way to extract as much information from each of them as possible.

STC involves the transmission of multiple redundant copies of data to compensate for fading and thermal noise in the hope that some of them may arrive at the receiver in a better state than others. In the case of STBC in particular, the data stream to be transmitted is encoded in blocks, which are distributed among spaced antennas and across time. While it is necessary to have multiple transmit antennas, it is not necessary to have multiple receive antennas, although to do so improves performance. This process of receiving diverse copies of the data is known as diversity reception.

STBCs as originally introduced, and as usually studied, are orthogonal. This means that the STBC is designed such that the vectors representing any pair of columns taken from the coding matrix is orthogonal. The result of this is simple, linear, optimal decoding at the receiver. Its most serious disadvantage is that all but one of the codes that satisfy this criterion must sacrifice some proportion of their data rate.

Apart from there being no full-rate, complex, orthogonal STBC for more than 2 antennas, it has been derived that, for more than three antennas, the maximum possible rate is 3/4. Codes have been designed which achieve a good proportion of this, but they have very long block-length and are unsuitable for practical use. This is because decoding cannot proceed until *all* transmissions in a block have been received, so a longer block-length, T results in a longer decoding delay.

Quasi-orthogonal STBCs

These codes exhibit partial orthogonality and provide only part of the diversity gain mentioned above. An example reported by Hamid Jafarkhani is:

$$C_{4,1} = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & -s_3 & -s_2 & s_1 \end{bmatrix}.$$

The orthogonality criterion only holds for columns (1 and 2), (1 and 3), (2 and 4) and (3 and 4). Crucially, however, the code is full-rate and still only requires linear processing at the receiver, although decoding is slightly more complex than for orthogonal STBCs. Results show that this Q-STBC outperforms (in a bit-error rate sense) the fully-orthogonal 4-antenna STBC over a good range of signal-to-noise ratios (SNRs). At high SNRs, though (above about 22dB in this particular case), the increased diversity offered by orthogonal STBCs yields a better BER. Beyond this point, the relative merits of the schemes have to be considered in terms of useful data throughput [11].

2.4.3 Diversity Combining

Diversity combining is the technique applied to combine the multiple received signals of a diversity reception device into a single improved signal. Various diversity combining techniques can be distinguished [12]:

Maximal-ratio combining: The received signals are weighted with respect to their SNR and then summed. Assuming ideal operation, predetection stage maximal-ratio combining achieves the best performance improvement compared with the other methods. However it requires cophasing, weighting, and summing circuits resulting in the most complicated implementation.

Equal gain combining: All the received signals are summed coherently. It is equal to maximal-ratio combining, except that the weighting circuits are ommited. The performance improvement by an equal-gain combiner is slightly inferior to that of a maximal-ratio combiner, since interference and noise corrupted signals may combined with high qaulity (interference and noise free) signals.

Selection combining: Out of the N received signals, the strongest signal is selected. For VHF, UHF, and microwave communications, both the maximal-ratio and equal-gain combining methods are unsuitable. It is difficult to realize a cophasing circuit having precise and stable tracking performance in a rapidly changing random phase, multipath fading environment. Compared with other two menthods the selection method is more suitable for mobile radio applications because of its simple implementations. Stable opeation is easily achieved and its performance is slightly inferior to the one obtained by maximal-ratio combinng method.

Switched combining: The receiver switches to another signal when current signal drops below a predefined threshold. This is a less efficient technique than selection combining.

2.4.4 MIMO Technology

Multiple-input multiple-output, or MIMO, refers to the use of multiple antennas both at the transmitter and receiver. Another common term for this technology is smart antennas, which performs spatial information processing with multiple antennas. MIMO technology has attracted attention in wireless communications, since it offers significant increases in data throughput and link range without additional bandwidth or transmit power. It achieves this by higher spectral efficiency (more bits per second per Hertz of bandwidth) and link reliability or diversity (reduced fading).

Spatial multiplexing requires MIMO antenna configuration. In spatial multiplexing, a high rate signal is split into multiple lower rate streams and each stream is transmitted from a different transmit antenna in the same frequency channel. If these signals arrive at the receiver antenna array with sufficiently different spatial signatures, the receiver can separate these streams, creating parallel channels for free. Spatial multiplexing is very powerful technique for increasing channel capacity at higher Signal to Noise Ratio (SNR). The maximum number of spatial streams is limited by the lesser in the number of antennas at the transmitter or receiver. Spatial multiplexing can be used with or without transmit channel knowledge.

Diversity coding techniques are used when there is no channel knowledge at the transmitter. In diversity methods a single stream (unlike multiple streams in spatial multiplexing) is transmitted, but the signal is coded using techniques called space-time coding. The signal is emitted from each of the transmit antennas using certain principles of full or near orthogonal coding. Diversity exploits the independent fading in the multiple antenna links to enhance signal diversity. Because there is no channel knowledge, there is no beamforming or array gain from diversity coding. [10]

2.5 Convolutional Coding

In telecommunication, a convolutional code is a type of error-correcting code in which (a) each m-bit information symbol (each m-bit string) to be encoded is transformed into an n-bit symbol, where m/n is the code rate ($n \geq m$) and (b) the transformation is a function of the last k information symbols, where k is the constraint length of the code. Convolutional codes are often used to improve the performance of digital radio, mobile phones, satellite links, and Bluetooth implementation.

A free distance d is a minimal Hamming distance between different encoded sequences. A correcting capability t of a convolutional code is a number of errors that can be corrected by the code. Since a convolutional code doesn't use blocks, processing instead a continuous bitstream, the value of t applies to a quantity of errors located relatively near to each other. That is, multiple groups of t errors can usually be fixed when they are relatively far. Free distance can be interpreted as a minimal length of an erroneous "burst" at the output of a convolutional decoder. Several algorithms exist for decoding convolutional codes. For relatively small values of k, the Viterbi algorithm is universally used as it provides maximum likelihood performance and is highly parallelizable. [9]

This chapter has provided an overview of different multiplexing techniques along with OFDM, fundamentals of diversity, and convolutional coding. Analysis of OFDM and STBC-OFDM is the subject of the next chapter.

Chapter 3

System Description and Modeling

3.1 Performance Analysis of an OFDM System:

3.1.1 System Model

The model of the OFDM system considered for analysis is shown in Fig. 3.1.

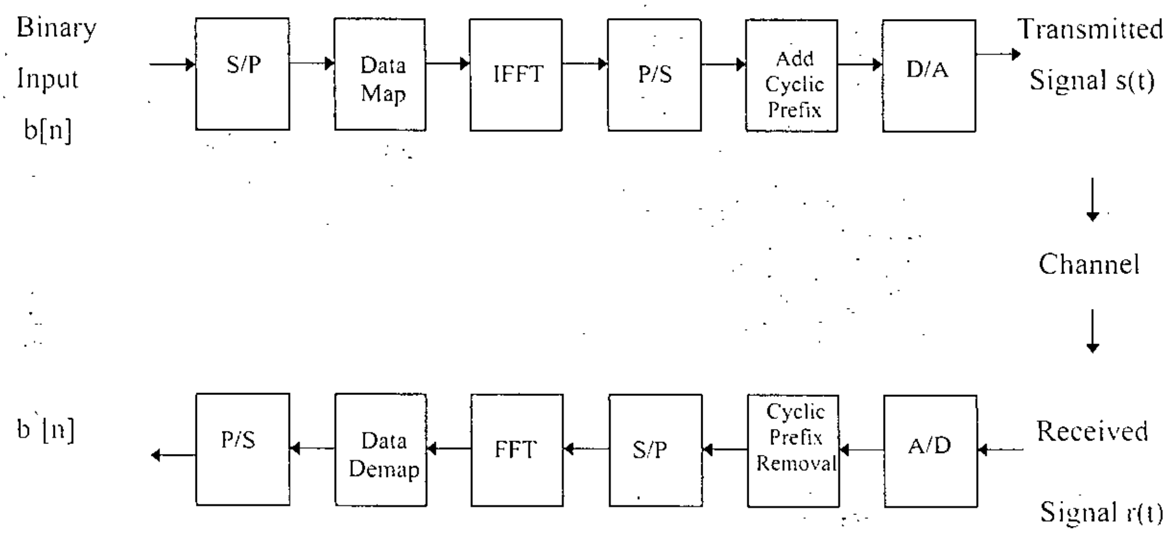

Fig. 3.1 Block diagram of an OFDM system

Serial to Parallel Conversion

The input serial data stream b[n] is formatted into the word size required for transmission, e.g. 2 bits/word for QPSK, and shifted into a parallel format. The data is then transmitted in parallel by assigning each data word to one carrier in the transmission.

Modulation of Data

The data to be transmitted on each carrier is mapped into a Phase Shift Keying (PSK) format. The data on each symbol is then mapped to a phase angle based on the modulation method.

For example, for QPSK the phase angles used are 0, 90, 180, and 270 degrees. For DQPSK and DPSK (DBPSK) modulation, differential coding is performed in the time domain.

Inverse Fourier Transform

Using IFFT, OFDM modulation is computed on each set of symbols, resulting in time-domain samples.

Guard Period

The guard period is a cyclic extension of the symbol to be transmitted. After the guard has been added, the symbols are then converted back to a serial time waveform. This is then the base band signal for the OFDM transmission.

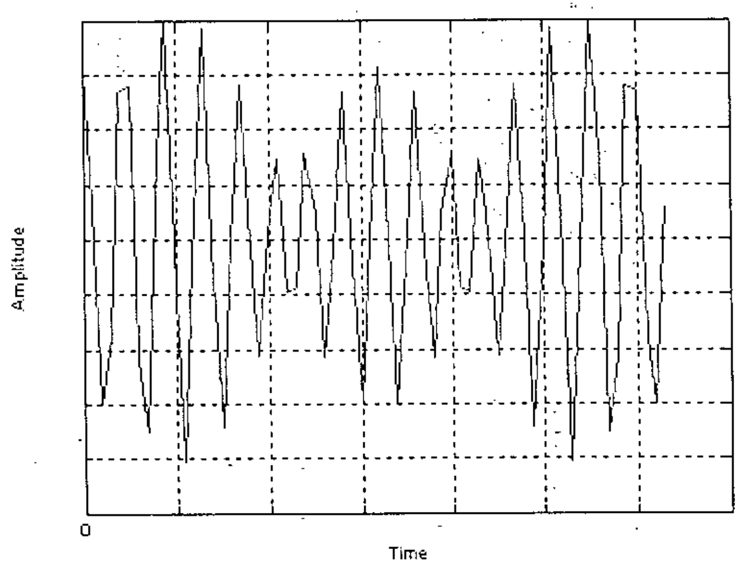

Fig. 3.2 OFDM signal in time domain

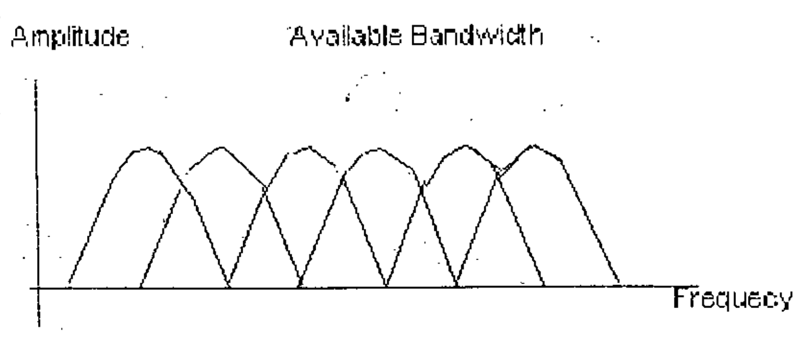

Fig. 3.3 OFDM signal in frequency domain

Channel

The channel is time-selective Rayleigh/Rician fading with AWGN.

Receiver

The receiver basically does the reverse operation to the transmitter. The guard period is removed. The FFT of each symbol is then taken to find the original transmitted spectrum. This returns N parallel streams. The phase angle of each transmission carrier is then evaluated and converted back to binary stream by demodulating the received phase. These streams are then re-combined into a serial stream, $\hat{b}[n]$ which is an estimate of the original binary stream at the transmitter.

3.1.2 Effect of Fading

Rayleigh fading is caused by multipath reception. The mobile antenna receives a large number of reflected and scattered waves. Because of wave cancellation effects, the instantaneous received power seen by a moving antenna becomes a random variable, dependent on the location of the antenna. All waves experience their own phase rotation. The resulting vector may significantly change in amplitude if individual components undergo different phase shifts. In mobile radio channels with high terminal speeds, such changes occur rapidly. Rayleigh fading then causes the signal amplitude and phase to fluctuate rapidly. Rician fading is similar to that for Rayleigh fading, except that in Rician fading a strong dominant component is present. This dominant component can for instance be the line-of-sight wave. The Rician *K*-factor is defined as the ratio of signal power in dominant component over the (local-mean) scattered power.

Multipath fading degrades the signal energy in single carrier systems but it causes additional interference in OFDM systems. We will analyze the interference based on the concept of complex Fourier series. The complex Fourier coefficients c_k can be expressed as

$$c_k = \frac{1}{T}\int_0^T f(x) e^{-jk\frac{2\pi}{T}x} dx, \qquad (2.1)$$

where T is the Nyquist interval. Now f(x) is assumed to be the interference which occurs at a certain subcarrier, the complex Fourier coefficients of f(x) are classified into two cases:

1. c_k ($k \neq 0$) is an alternating current (AC) component of f(x), and $|c_k|^2$ can be considered as the inter-carrier interference (ICI) power,

2. c_0 is a direct current (DC) component of f(x), and $|c_0|^2$ can be considered as the intra-symbol interference power.

Inter-Carrier Interference:

When a carrier $\cos(2\pi f_c t)$ passes through the fading channels, the received carrier frequency is affected by Doppler frequency $f_d \cos\theta$, where θ is a random phase with the uniform distribution of $(0, 2\pi)$. Then the received carrier r(t) can be expressed as [16]

$$
\begin{aligned}
r(t) &= \cos\{2\pi(f_c + f_d \cos\theta)t\} \\
&= \cos(2\pi f_c t) - (2\pi f_d t \cos\theta)\sin(2\pi f_c t), \\
&\qquad \text{for } |2\pi f_d t \cos\theta| \ll 1.
\end{aligned} \tag{3.2}
$$

The second term of Eq. (3.2) means the ICI from the loss of orthogonality among subcarriers. From Eq. (3.1), the interference power to a subcarrier k ($k \neq 0$) can be obtained as [16]

$$
\begin{aligned}
|c_k|^2 &= \left| \frac{1}{T_d} \int_0^{T_d} 2\pi f_d t(\cos\theta) e^{-jk\frac{2\pi}{T_d}t} dt \right|^2 \\
&= \frac{(f_d T_d \cos\theta)^2}{k^2} \quad \text{for } k \neq 0
\end{aligned} \tag{3.3}
$$

It is found from Eq. (3.3) that the interference power from a remote subcarrier k_l (large number of k) is kept to be small. Therefore, by using the symmetricity of the complex Fourier coefficients and Euler zeta function, the total power of ICI I_a^{fd} from subcarriers k ($-k_l \leq k \leq k_l$, $k \neq 0$) can be calculated as [16]

$$
\begin{aligned}
I_a^{fd} &= 2 \sum_{k=1}^{k_l} |c_k|^2 \\
&= 2 \sum_{k=1}^{k_l} \frac{(f_d T_d \cos\theta)^2}{k^2} \\
&\cong \frac{(\pi f_d T_d \cos\theta)^2}{3}, \quad k_l \gg 1
\end{aligned} \tag{3.4}
$$

where it is assumed that the number of subcarriers in OFDM systems is very large. Finally, the average power I_a of ICI caused by the Doppler frequency shift can be derived by averaging Eq. (3.4) by uniformly distributed θ as [16]

$$I_a = \frac{1}{2\pi}\int_0^{2\pi}\frac{(\pi f_d T_d \cos\theta)^2}{3}d\theta$$

$$= \frac{(\pi f_d T_d)^2}{6}$$

$$= \frac{(\pi f_d T_s)^2}{6(1+\delta_c)^2} \qquad (3.5)$$

Intra-Symbol Interference

Intra-symbol interference is created because of phase rotation angle of the demodulating symbol caused by the fading. This rotation gives rise to the degradation of the symbol energy and the interference to the quadrature channel component. The phase rotation angle ψ in terms of time can be expressed as [16]

$$\psi(t) = 2\pi f_d t \cos\theta, \qquad (3.6)$$

which is caused by the Doppler frequency shift. Therefore, the received carrier r(t) with the phase rotation ψ becomes as [16]

$$r(t) = \cos(2\pi fct + \psi)$$
$$= \cos\psi \cos(2\pi fct) - \sin\psi \sin(2\pi fct). \qquad (3.7)$$

In Eq. (3.7), cos ψ and sin ψ mean the degradation of the symbol energy and the interference to the quadrature channel component, respectively. Assuming $|2\pi f_d t \cos\theta| << 1$, cosψ and sinψ in Eq. (3.7) can be approximated as [16]

$$\cos\psi = \cos(2\pi f_d t(\cos\theta)) \cong 1$$
$$\sin\psi = \sin(2\pi f_d t(\cos\theta)) \cong 2\pi f_d t(\cos\theta) \qquad (3.8)$$

Consequently, the degradation of the symbol energy can be neglected, while the interference to the quadrature channel component must be considered as the intra-symbol interference. The average interference power to the quadrature channel component can be calculated by taking account of a DC component $|c_0|^2$ of complex Fourier coefficient in Eq. (3.1). From Eq. (3.8), the interference power I_d^{fd} and its average power $\overline{I_d^{fd}}$ caused by the Doppler frequency shift become as [16]

$$I_d^{fd} = (\frac{1}{T_s}\int_0^{T_s} 2\pi f_d t \ \cos\theta \, dt)^2$$

$$= (\pi f_d T_s \cos\theta)^2$$

$$\overline{I_d^{fd}} = \frac{1}{2\pi}\int_0^{2\pi} (\pi f_d T_s (\cos\theta)^2 d\theta$$

$$= \frac{(\pi f_d T_s)^2}{2}$$

(3.9)

3.1.3 SNIR in Presence of Fading and AWGN

Due to the inter-carrier and intra-symbol interferences mentioned above, the BER is more degraded than single carrier systems. The derivation of the signal-to-noise plus interference power ratio (SNIR) involves the interference power estimated as the Gaussian noise. It is a point to notice that the two interference powers should be included in the SNIR of the received Rayleigh wave since those interferences are caused by fading.

Considering P_s and P_n as the signal power and noise power respectively the signal-to-noise ratio is expressed as

$$\text{SNR} \ \gamma_N = \frac{P_s}{P_n}$$

$$= \frac{\frac{E_b}{(K+1)(1+\delta_c)}}{\frac{N_0}{2}} \qquad (3.10)$$

where $1/(1+\delta_c)$ means the energy loss caused by the removal of the GI and K is the Rician factor. For the special case of Rayleigh fading channel and negligible GI the SNR expression is changed as follows

$$\text{SNR} \ \gamma_N = \frac{2E_b}{N_0}$$

Considering P_I as the total interference power, the expression for Signal-to-Noise Plus Interference Power Ratio can be expressed as follows

$$\text{SNIR} \ \gamma_{NI} = \frac{P_s}{P_n + P_I}$$

$$= \frac{1}{\frac{P_n}{P_s} + \frac{P_I}{P_s}}$$

$$= \frac{1}{\frac{1}{\gamma_N} + (I_a + I_d)} \tag{3.11}$$

Finally substituting Eq. (3.5), Eq. (3.9) and Eq. (3.10) into Eq. (3.11) leads the SNIR equation as

$$\text{SNIR } \gamma_{NI} = \frac{1}{\frac{(k+1)(1+\delta_c)}{\frac{2E_b}{N_0}} + \frac{(\pi F_d T_s)^2}{2} + \frac{(\pi F_d T_d)^2}{6}} \tag{3.12}$$

3.1.4 Modified SNIR in Presence of Fading, AWGN and Jitter

We modify the equation of intercarrier interference in Eq. (3.4) to express it in terms of number of OFDM subcarriers. The total interference power from N_s number of subcarriers can be expressed as

$$I_a^{Id} = 2\sum_{k=1}^{N_s} |c_k|^2$$

$$= 2\sum_{k=1}^{N_s} \frac{(f_d T_d \cos\theta)^2}{k^2}$$

$$= 2\ (f_d T_d \cos\theta)^2 \sum_{k=1}^{N_s} \frac{1}{k^2}$$

The average ICI power is expressed as

$$I_a = \frac{1}{2\pi} \int_0^{2\pi} 2(f_d T_d)^2 (\cos\theta)^2 \sum_{k=1}^{N_s} \frac{1}{k^2} d\theta$$

$$= \sum_{k=1}^{N_s} \frac{1}{k^2} (f_d T_d)^2\ \frac{1}{2\pi} \int_0^{2\pi} 2(\cos\theta)^2 d\theta$$

$$= \frac{(f_d T_d)^2}{2\pi} [\theta + \frac{\sin 2\theta}{2}]_0^{2\pi} \sum_{k=1}^{N_s} \frac{1}{k^2}$$

$$= \frac{(f_d T_d)^2}{2\pi} (2\pi + 0) \sum_{k=1}^{N_s} \frac{1}{k^2}$$

$$= \sum_{k=1}^{N_s} \frac{1}{k^2} (f_d T_d)^2$$

(3.13)

Substituting Eq. (3.9), Eq. (3.10) and Eq. (3.13) into Eq. (3.11) we derive the equation of SNIR in presence of only fading and AWGN as

$$\text{SNIR} = \frac{1}{\dfrac{(k+1)(1+\delta_c)}{\dfrac{2E_b}{N_0}} + \dfrac{(\pi F_d T_s)^2}{2} + \sum_{k_1=1}^{N_s} \dfrac{1}{k_1^2} (F_d T_f)^2}$$

(3.14)

Now we will derive the SNIR expression considering jitter along with fading and AWGN. In any communication system timing jitter causes the signal power to degrade by a factor of (1-ϵ) over a time slot. The jitter also causes interference, which is expressed as $I_\epsilon = E_b\,\epsilon$. Here ϵ is the timing error normalized by symbol duration T_s i.e. ($\epsilon = \Delta / T_s$). So the total interference power becomes $I = I_a + I_d + I_\epsilon$. To incorporate the jitter effect along with fading and AWGN, the equation of SNIR in Eq. (3.14) is modified as follows:

$$\text{SNIR}(\epsilon) = \frac{1}{\dfrac{(k+1)(1+\delta_c)}{\dfrac{2E_b(1-\varepsilon)}{N_0}} + \dfrac{(\pi F_d T_s)^2}{2} + \sum_{k1=1}^{N_s} \dfrac{1}{k_1^2} (F_d T_f)^2 + E_b \varepsilon}$$

(3.15)

In fading environment, the received signal energy is random depending on random fade but the noise energy remains unchanged. The received bit energy E_b is the product of input bit energy E_{in} and the multiplicative fade α^2, $E_b = E_{in} \times \alpha^2$. So with this modification the equation of SNIR in Eq. (3.15) becomes:

$$\text{SNIR}(\epsilon, \alpha) = \frac{1}{\dfrac{(k+1)(1+\delta_c)}{\dfrac{2(E_{in}\alpha^2)(1-\varepsilon)}{N_0}} + \dfrac{(\pi F_d T_s)^2}{2} + (F_d T_f)^2 \sum_{k1=1}^{N_s} \dfrac{1}{k_1^2} + (E_{in}\alpha^2)\varepsilon}$$

(3.16)

3.1.5 Expressions of BER

With DQPSK modulation the expression of BER in fading channel is evaluated as [16]

$$\mathrm{Pe} = \frac{(1-\rho)\dfrac{\Gamma}{K+1}+1}{2(\dfrac{\Gamma}{K+1}+1)}\exp(-\frac{K\dfrac{\Gamma}{K+1}}{\dfrac{\Gamma}{K+1}+1}),$$

where $\rho = J_o(2\pi F_d T_s)$ is the time correlation function and $\Gamma = E_b/N_0$ is the bit SNR. (3.17)

The terms $\Gamma/(K+1)$ and γ_{NI} (SNIR) are related in [16] as following

$$\Gamma/(K+1) = \frac{\gamma_{NI}}{2} \quad \text{for DQPSK} \tag{3.18}$$

Substituting Eq. (3.18) in Eq. (3.17) we find the BER equations for DQPSK modulation as

$$\mathrm{Pe} = \frac{\{1-J_0(2\pi f_d T_s)\}(\dfrac{SNIR}{2})+1}{2(\dfrac{SNIR}{2}+1)}\exp(-\frac{K.\dfrac{SNIR}{2}}{\dfrac{SNIR}{2}+1}) \tag{3.19}$$

To incorporate the jitter effect along with fading and AWGN, we modify the expression of BER in Eq. (3.19) as follows

$$\mathrm{Pe}(\epsilon,\alpha) = \frac{\{1-J_0(2\pi f_d T_s)\}\{\dfrac{SNIR(\varepsilon,\alpha)}{2}\}+1}{2\{\dfrac{SNIR(\varepsilon,\alpha)}{2}+1\}}\exp\{-\frac{K.\dfrac{SNIR(\varepsilon,\alpha)}{2}}{\dfrac{SNIR(\varepsilon,\alpha)}{2}+1}\}, \quad \text{for DQPSK} \tag{3.20}$$

The equation of BER of QPSK system in AWGN channel is found as [23]

$$\mathrm{Pe} = 0.5\,\mathrm{erfc}\sqrt{\frac{E_b}{N_0}}$$

$$=0.5\,\mathrm{erfc}\sqrt{\mathrm{SNIR}} \tag{3.21}$$

To find the BER performance of QPSK/OFDM system in presence of fading, jitter and AWGN we modify Eq. (3.21) as follows:

$$\mathrm{Pe}(\epsilon,\alpha) = 0.5\,\mathrm{erfc}\sqrt{\mathrm{SNIR}(\varepsilon,\alpha)} \tag{3.22}$$

The equation of BER of DPSK (DBPSK) system in AWGN channel is found as [23]

$$Pe = 0.5 \exp\left(-\frac{E_b}{N_0}\right)$$

$$= 0.5 \exp(-SNIR)$$

(3.23)

To find the BER performance of DPSK/OFDM system in presence of fading, jitter and AWGN we modify Eq. (3.23) as follows:

$$Pe(\epsilon, \alpha) = 0.5\, e^{-SINR(\epsilon,\alpha)}$$

or

$$Pe(\epsilon, \alpha) = 0.5 \exp\{-SINR(\epsilon, \alpha)\}$$

(3.24)

For a fixed value of α the probability of bit error is $Pe(\alpha)$. This $Pe(\alpha)$ is considered as a conditional probability of error for a given value of channel attenuation/gain α. For Rayleigh fading channel, α is Rayleigh distributed. So the pdf of α is:

$$p(\alpha) = \frac{\alpha}{\sigma} \times \exp\left(-\frac{\alpha^2}{2\sigma^2}\right)$$

(3.25)

Here σ is Rayleigh parameter. The mean and variance value of the Rayleigh density function is a function of σ. Mean $=1.25\times\sigma$ and variance $=0.655\times\sigma$. For Rician fading channel, α is Rician distributed. So the pdf of α is:

$$p(\alpha) = \frac{\alpha}{\sigma} \times \exp\left(-(\alpha^2 + m^2)/2\sigma^2\right) \times I_0((\alpha\times m)/\sigma^2)$$

(3.26)

Here I_0 is the zeroth order Bessel function. $m^2 = k \times 2\sigma^2$. The pdf of zero mean Gaussian distributed jitter is:

$$p(\epsilon) = \frac{1}{\sqrt{2\pi\sigma_\epsilon^2}} \times \exp\left(-\epsilon^2/2\sigma_\epsilon^2\right)$$

(3.27)

The unconditional BER can be calculated by averaging the conditional BER, over all possible values of α and ϵ

$$Pe = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} Pe(\alpha, \epsilon)\; p(\alpha)\; p(\epsilon)\; d\alpha\; d\epsilon$$

(3.28)

Substituting Eq. (3.20), Eq. (3.25) and Eq. (3.27) in Eq. (3.28) we can calculate the unconditional BER for DQPSK modulation in Rayleigh fading channel. In the same way by substituting Eq. (3.20), Eq. (3.25) and Eq. (3.26) in Eq. (3.28) we can calculate the unconditional BER for DQPSK modulation in Rician fading channel. Similarly unconditional BER of OFDM with QPSK and DPSK modulations can be find in Rayleigh and Ricain fading channels.

3.1.6 Convolutional Coding:

To improve transmission performance, channel coding is added. To optimize the use of the correction capacity of a particular code, soft decision is always a good solution. For this reason convolutional coding is chosen for the system. In the presence of channel coding, an expression of the bit error probability P_e cannot be worked out exactly, showing the need for a good upper bound. It is well known that using a rate R= K/N convolutional coding and a Viterbi algorithm decoding, the bit error probability for an information symbol is bounded by as [12]

$$P_e \leq \frac{1}{K} \sum_{d=d_f}^{\infty} W(d)P(d) \tag{3.29}$$

where P(d) is the probability for the decoding algorithm to choose a path at distance d from the correct path in the decoding trellis, d_f is the free distance of the encoder and W(d), a characteristic coefficient of the encoder, is defined as [12]:

$$W(d) = \sum_{i=1}^{\infty} ia(d,i) \tag{3.30a}$$

where a(d, i) is the number of paths at distance d from the correct path and corresponding to i information symbols equal to '1'. In general, the a(d, i) are deduced from the transfer function of the encoder. Considering the uncoded BER to be P_{un} the value of $P(d)$ can be expressed as follows:

$$P(d) = \{4\ P_{un}\ (1 - P_{un})\}^{d/2} \tag{3.30b}$$

W(d) is obtained from the code weights in Table 3.1 as [12]. Substituting the unconditional BER (Pe) from Eq. (3.28) in Eq. (3.30b), we can calculate the coded BER from Eq. (3.29).

Table 3.1: Weigh Spectrum of convolutional encoders

Hamming Weight d	**W(d) for R=1/2**	**W(d) for R=1/3**
10	3.6×10^{01}	-
11	0	-
12	2.11×10^{02}	-
13	0	-
14	1.404×10^{03}	-
15	0	1.1
16	1.633×10^{04}	1.6
17	0	1.9
18	7.7433×10^{04}	2.8
19	0	5.5
20	5.0269×10^{05}	9.6
21	0	1.69×10^{02}
22	3.322763×10^{06}	3.38×10^{02}
23	0	6.36×10^{02}
24	2.129291×10^{07}	1.276×10^{03}
25	0	2.172×10^{03}
26	1.3436491×10^{08}	-

3.2 Performance Analysis of a STBC-OFDM System:

3.2.1 System Model

The model of a STBC-OFDM system considered for analysis is shown in Fig 3.4.

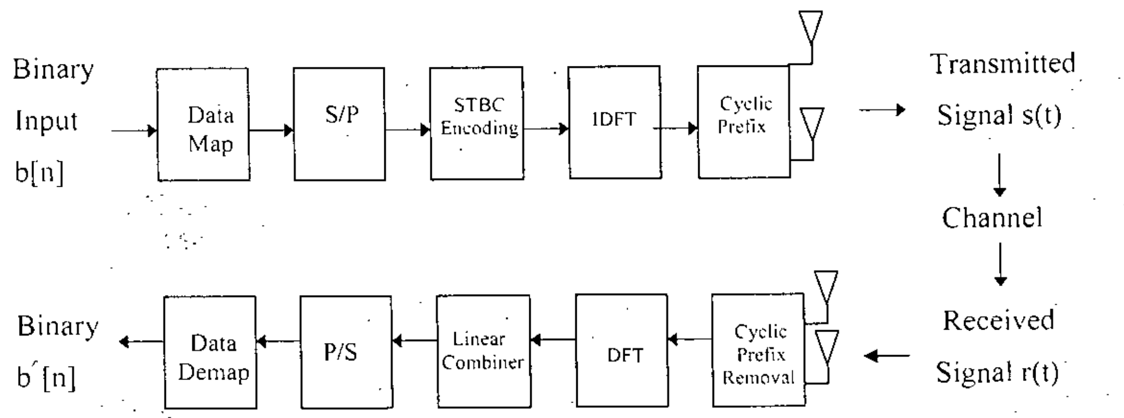

Fig 3.4 Block diagram of a STBC-OFDM system

We consider MIMO-OFDM system with P number of transmiting and two receiving antennas. There are N_s number of OFDM subcarriers. We also consider time-selective Rayleigh and Rician fading channel. Binary input data is mapped to a modulation symbols $\{a(i)\}$ that are assumed to have the following properties:

$$E[a(i)] = 0$$

$$E[a(i)\, a^*(j)] = 1,\ i=j$$
$$= 0,\ i \neq j$$

The input sequence $\{a(i),\ i=0,\ 1,\ 2\ \ldots\ldots\ (N_sP-1)\}$ is serial-to-parallel converted into P sequences each of length N_s, as

$a_p(k) = a(k+(p-1)N_s)$ where $p=1, 2\ldots\ldots P$ and $k=0, 1, 2\ldots\ldots(N_s-1)$

Each of the N_s sequences $\{a_1(k)\ \ldots\ldots\ a_P(k)\}$, $k=0,1,2\ldots\ldots,(N_s-1)$ is mapped to a matrix Ψ_k of size P×P by using a quasi orthogonal STBC with constellation rotation. For P=4, the 4×4 quasi orthogonal scheme is given as [22]:

$$\Psi_k = \begin{bmatrix} a_1(k) & -a_2^*(k) & e^{j\Phi}a_3(k) & -e^{-j\Phi}a_4^*(k) \\ a_2(k) & a_1^*(k) & e^{j\Phi}a_4(k) & e^{-j\Phi}a_3^*(k) \\ e^{j\Phi}a_3(k) & -e^{-j\Phi}a_4^*(k) & a_1(k) & -a_2^*(k) \\ e^{j\Phi}a_4(k) & e^{-j\Phi}a_3^*(k) & a_2(k) & a_1^*(k) \end{bmatrix} \quad (3.31)$$

where the rotation angle Φ depends on the signal constellation. Then we take the IDFT of Ψ_1, Ψ_2, ... Ψ_k in order to form the transmitted signals as

$$S_m = \frac{1}{\sqrt{N_s}} \sum \Psi_k . e^{j(2\pi/Ns)mk}, \quad m=0,1......(Ns-1)$$

(3.32)

S_m is a P×P matrix, which represents the transmitted signals on the mth subcarrier. We define

$$\Psi=[\Psi_0^T, ...\Psi_{Ns-1}^T]^T, (N_sP\times P)$$

(3.33)

$$S=[S_0^T, ...S_{Ns-1}^T]^T, (N_sP\times P)$$

(3.34)

where $(.)^T$ denotes transpose, then S can be written as

$$S=(U\otimes I_p)^H\Psi$$

(3.35)

Here $(.)^H$ denotes complex conjugate transpose, $\otimes$ denotes kronecker product, I_p is the P×P identity matrix, and U is the $N_s\times N_s$ unitary discrete Fourier transform (DFT) matrix. In frequency-selective fading channels with L resolvable paths, there exists interblock interference (IBI). To minimize this IBI, a cyclic prefix of length c_p ($c_p \geq L$) is added to each OFDM symbol. At the receiver, the cyclic prefix is discarded, leaving IBI-free, information-bearing signals.

The model of the channel with L resolvable multipath components can be expressed as [22]

$$h(\tau) = \sum_{l=0}^{L-1} \rho_l \delta(\tau - \tau_l T_s)$$

(3.36)

where ρ_l is the zero-mean complex Gaussian random variable, and τ_l is the delay of the lth path normalized with respect to T_s. The delays $\{\tau_l\}$ are assumed to be uniformly distributed over the cyclic prefix c_p. The channel has an exponential power-delay profile $\theta(\tau_l) = e^{-\frac{\tau_l}{\tau_{rms}}}$, where τ_{rms} represents the rms delay spread, which is also normalized with respect to T_s. The P symbols in each column of Ψ_k are transmitted from the P transmit antennas simultaneously during every OFDM symbol period. Considering the channel matrix H the received signals is expressed in an $N_s \times P$ matrix as:

$$R = HS + V, \tag{3.37}$$

where $V=[v_0,\ldots\ldots v_{N_s-1}]^T$ ($N_s \times P$) is the additive white Gaussian noise (AWGN) matrix whose elements are independent and identically distributed. Hence

$$E[\mathrm{vec}(V).\mathrm{vec}(V)^H] = \sigma^2 I_{N_sP} \tag{3.38}$$

where σ^2 is the variance of the zero-mean noise samples when the transmitted symbol energy is normalized to unity. For OFDM systems over fast fading channels, channel estimation is generally carried out by transmitting pilot symbols in given positions of the frequency-time grid. We assume hereafter that channel state information (CSI) is known at the receiver. In the presence of time-selective fading, H is no longer a block-circulant matrix. Consequently, $G = UH(U \otimes I_P)^H$ is not a block diagonal matrix. This shows that time-selective fading causes ICI, which is represented by the off-diagonal blocks of G.

The received signal R is processed by multiplying it with U, forming $N_s \times P$ matrix X as $X=[x_0^T,\ldots\ldots\ldots x_{N_s-1}^T]^T = UR$,

Now from (3.37) we get

$$X = U(HS+V) = G\Psi + W \tag{3.39}$$

where $x_k = [x_1(k),\ \ldots\ldots x_P(k)]^T$, $w_k = [w_1(k),\ldots\ldots w_P(k)]^T$ and $W=UV=[w_0,\ \ldots w_{N_s-1}]^T$ ($N_s \times P$)

$$x_k^T = g_{k,k}^T \Psi_k + \sum_{k'=0,k'\neq k}^{N_s-1} g_{k,k'}^T \Psi_{k'} + w_k^T, \qquad k=0,\ldots\ldots,N_s-1$$

$$\text{and } g_{k,k'} = [g_{k,k'}^{(1)},\ldots\ldots\ldots\ldots, g_{k,k'}^{(P)}]^T, \quad k, k' = 0,\ldots\ldots\ldots, N_s-1 \tag{3.40}$$

$g_{k,k'}$ is the (k,k')th block of G. The signal X is received in two receiving antennas. Diversity combiner at the receiver selects the best instantaneous signal of the two antennas. The received signal can be detected through differential or coherent scheme.

It is shown in the Appendix that in the presence of time-varying fading, SINR for quasi-orthogonal STBC-OFDM system has an expression as:

SINR=

$$=\frac{4\sum_{l=0}^{L-1}[N_s+2\sum_{i=1}^{N_s-1}(N_s-1)J_0(2\pi i f_d T_s)]e^{-\frac{\tau_l}{\tau_{rms}}}}{4\sum_{k'=1}^{N_s-1}\sum_{l=0}^{L-1}[N_s+2\sum_{i=1}^{N_s-1}(N_s-1)J_0(2\pi i f_d T_s)\cos(\frac{2\pi}{N_s}k'i)\,e^{-\frac{\tau_l}{\tau_{rms}}}]+\sigma^2} \tag{3.41}$$

3.2.2 Modified SNIR in Presence of Fading, AWGN and Jitter

We Modify Eq. (3.41) and express the signal to noise and interference ratio (SNIR) with N_s OFDM subcarriers without timing error for STBC-OFDM as follows:

$$\text{SINR}=\frac{N_T\sum_{l=0}^{L-1}[[N_s+2\sum_{i=1}^{N_s-1}(N_s-i)J_0(2\pi i f_d T_s)]]e^{-\frac{\tau_l}{\tau_{rms}}}}{N_T\;[\sum_{k'=1}^{N_s-1}\sum_{l=0}^{L-1}\{N_s+2\sum_{i=1}^{N_s-1}(N_s-i)J_0(2\pi i f_d T_s)\cos(\frac{2\pi}{N_s}k'i)\}e^{-\frac{\tau_l}{\tau_{rms}}}]+\sigma^2} \tag{3.42}$$

where N_T is the number of transmitting antennas.

Both Eq. (3.41) and Eq. (3.42) are for unit symbol energy. We consider symbol energy Es, GI ratio δ_c and Rician parameter K. For DQPSK and QPSK symbol energy is twice the bit energy E_b and for DPSK it is equal to E_b. However Jitter causes the signal power to degrade by a factor of $(1-\epsilon)$ over a time slot. The jitter also causes interference, which is added as $E_b\,\epsilon$. Here ϵ is the timing error normalized by symbol duration T_s i.e. ($\epsilon = \Delta / T_s$). To incorporate the jitter effect along with fading and AWGN, the equation of SNIR for DPSK is modified as follows:

SINR(ϵ) =

$$\frac{N_T \sum_{l=0}^{L-1}[\{\frac{E_b(1-\varepsilon)}{(K+1)(1+\delta_c)}\}\ [N_s + 2\sum_{i=1}^{N_s-1}(N_s-i)J_0(2\pi i f_d T_s)]\ ]e^{-\frac{\tau_l}{\tau_{rms}}}}{N_T\{\frac{E_b(1-\varepsilon)}{(K+1)(1+\delta_c)}\}[\sum_{k'=1}^{N_s-1}\sum_{l=0}^{L-1}\{N_s+2\sum_{i=1}^{N_s-1}(N_s-i)J_0(2\pi i f_d T_s)\cos(\frac{2\pi}{N_s}k'i)\}e^{-\frac{\tau_l}{\tau_{rms}}}]+(\sigma^2+E_b\varepsilon)} \tag{3.43}$$

In fading environment, the received signal energy is random depending on random fade but the noise energy remains unchanged. The received bit energy Eb is the product of input bit energy E_{in} and the multiplicative fade α^2, $Eb = E_{in} \times \alpha^2$. So with this modification the equation of SNIR for DPSK becomes:

SINR(ϵ, α) =

$$\frac{N_T \sum_{l=0}^{L-1}[\{\frac{(E_{in}\alpha^2)(1-\varepsilon)}{(K+1)(1+\delta_c)}\}\ [N_s + 2\sum_{i=1}^{N_s-1}(N_s-i)J_0(2\pi i f_d T_s)]\ ]e^{-\frac{\tau_l}{\tau_{rms}}}}{N_T\{\frac{(E_{in}\alpha^2)(1-\varepsilon)}{(K+1)(1+\delta_c)}\}[\sum_{k'=1}^{N_s-1}\sum_{l=0}^{L-1}\{N_s+2\sum_{i=1}^{N_s-1}(N_s-i)J_0(2\pi i f_d T_s)\cos(\frac{2\pi}{N_s}k'i)\}e^{-\frac{\tau_l}{\tau_{rms}}}]+(\sigma^2+E_{in}\alpha^2\varepsilon)} \tag{3.44}$$

Similarly for DQPSK and QPSK the equation of SNIR is derived as follows:

SINR(ϵ, α) =

$$\frac{N_T \sum_{l=0}^{L-1}[\{\frac{(2E_{in}\alpha^2)(1-\varepsilon)}{(K+1)(1+\delta_c)}\}\ [N_s + 2\sum_{i=1}^{N_s-1}(N_s-i)J_0(2\pi i f_d T_s)]\ ]e^{-\frac{\tau_l}{\tau_{rms}}}}{N_T\{\frac{(2E_{in}\alpha^2)(1-\varepsilon)}{(K+1)(1+\delta_c)}\}[\sum_{k'=1}^{N_s-1}\sum_{l=0}^{L-1}\{N_s+2\sum_{i=1}^{N_s-1}(N_s-i)J_0(2\pi i f_d T_s)\cos(\frac{2\pi}{N_s}k'i)\}e^{-\frac{\tau_l}{\tau_{rms}}}]+(\sigma^2+E_{in}\alpha^2\varepsilon)} \tag{3.45}$$

Substituting Eq. (3.45) in Eq. (3.20) and Eq. (3.22) we obtain the conditional BER of STBC-OFDM system for DQPSK and QPSK modulation respectively. To obtain the the conditional BER of STBC-OFDM system for DPSK, Eq. (3.44) is substituted in Eq. (3.24). The unconditional BER of STBC-OFDM is calculated in the same approach as described in section 3.1.5 for OFDM systems. The expressions mentioned in section 3.1.6 to calculate the convoultional coded BER also holds for STBC-OFDM systems.

3.2.3 MIMO-OFDM:

STBC-OFDM with transmitting diversity is transformed into MIMO-OFDM by addition of receiving diversity. The block diagram of a MIMO system is shown in Fig 3.5.

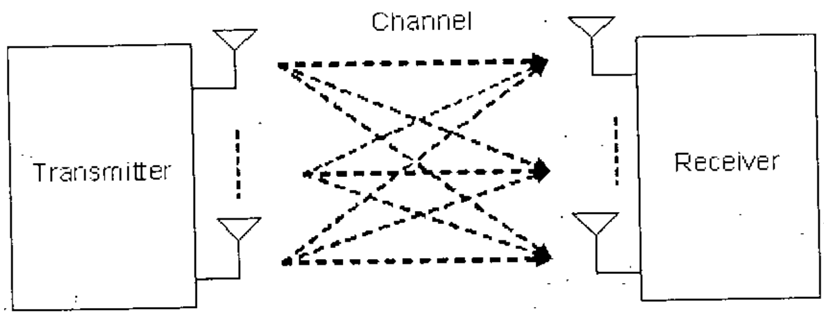

Fig. 3.5 Block diagram of a MIMO system

To utilize the receiving diversity scheme diversity combining is required. For selective diversity combining the instantaneous processed bit SNR/SIR at the output of the combiner is given by [23]

$$\gamma = \max\{ \lambda 1, \lambda 2 \} \tag{3.46}$$

Here λ is the instantaneous SIR at each receiving antenna. γ is the the instantaneous SIR of the combined branch. For a single antenna the pdf of λ is obtained as follows

$$P_1(\lambda)=\frac{A}{(\lambda+A)^2} \tag{3.47}$$

Where A denotes the average value λ. The pdf of γ for the selection combining method is

$$P_2(\gamma)=\frac{2A}{(\lambda+A)^2}-\frac{2A}{(2\lambda+A)^2} \tag{3.48}$$

The average Bit error probability without diversity is $Pe_1(A)$ and with diversity is Pe2(A)

$$Pe_1(A) = \int_{-\infty}^{\infty} Pe(\gamma) \quad P_1(\gamma) \quad d\gamma$$

$$= \int_{-\infty}^{\infty} Pe(\gamma) \quad \frac{A}{(\lambda+A)^2} \quad d\gamma$$

(3.49)

$$Pe_2(A) = \int Pe(\gamma) \quad P_2(\gamma) \quad d\gamma$$

$$= \int Pe(\gamma) \quad \left\{ \frac{2A}{(\lambda+A)^2} - \frac{2A}{(2\lambda+A)^2} \right\} \quad d\gamma$$

(3.50)

$$Pe2(A) = \int Pe(\gamma) \quad \left\{ \frac{2A}{(\lambda+A)^2} - \frac{2A}{(2\lambda+A)^2} \right\} \quad d\gamma$$

$$= \int Pe(\gamma) \quad \frac{2A}{(\lambda+A)^2} \quad d\gamma \quad - \quad \int Pe(\gamma) \quad \frac{2A}{(2\lambda+A)^2} \quad d\gamma$$

$$= 2 \int Pe(\gamma) \frac{A}{(\lambda+A)^2} \quad d\gamma \quad - \quad \int Pe(\gamma) \quad \frac{(A/2)}{(\lambda+\frac{A}{2})^2} \quad d\gamma$$

$$= 2 \; Pe_1(A) \; - \; Pe_1(A/2)$$

So for SIR=A, the relationship becomes:

$Pe_2(A)=2Pe_1(A) - Pe_1(A/2)$

(3.51)

For SNR=A the same result is found: $Pe_2(A)=2Pe_1(A) - Pe_1(A/2)$

From the expressions we can conclude that for SINR=A the same relationship will hold. In this case, $Pe_1(A)$ represents BER for average SINR=A with single receiving antenna and $Pe_2(A)$ represents BER for average SINR=A with diversity combining of two receiving antennas. Substituting the unconditional BER of STBC-OFDM with single receiving antenna from section 3.2.2, in Eq. (3.51) we obtain the unconditional BER for MIMO-OFDM with two receiving antenna. The expressions mentioned in section 3.1.6 can be used to calculate the convoultional coded BER for MIMO-OFDM systems.

In this chapter we have presented the analytical method for evaluating OFDM, STBC-OFDM and MIMO-OFDM in presents of three impairments. The next chapter shows the numerical results of the presented analysis.

Chapter 4

Results and Discussion

In this chapter we evaluate the performance of OFDM and STBC-OFDM systems considering different channel impairments. For numerical calculation we have considered some practical values of the system parameters shown in Table 4.1.

Table 4.1: System constants and parameters

Parameter	Description	Values considered
T	Temperature	300
R_l	Receiver resistance	50
k_l	Boltzman's Constant	1.38×10^{-23}
B	Data Rate per OFDM subcarrier	$1\text{Mbps}=10^{6}$ bps
N_0	AWGN	3.312×10^{-16}
T_s	Symbol Period	10^{-6}
δ_c	GI ratio	0.25
T_d	Effective Symbol period deducting guard period	0.75×10^{-6}

$N_0 = 4\times k_l\times T\times B/R_l = 3.312\times10^{-16}$

$T_d = T_s - T_G$

$= T_s - 0.25\,T_s$

$= 0.75\times T_s$

4.1 Performance Results of an OFDM System:

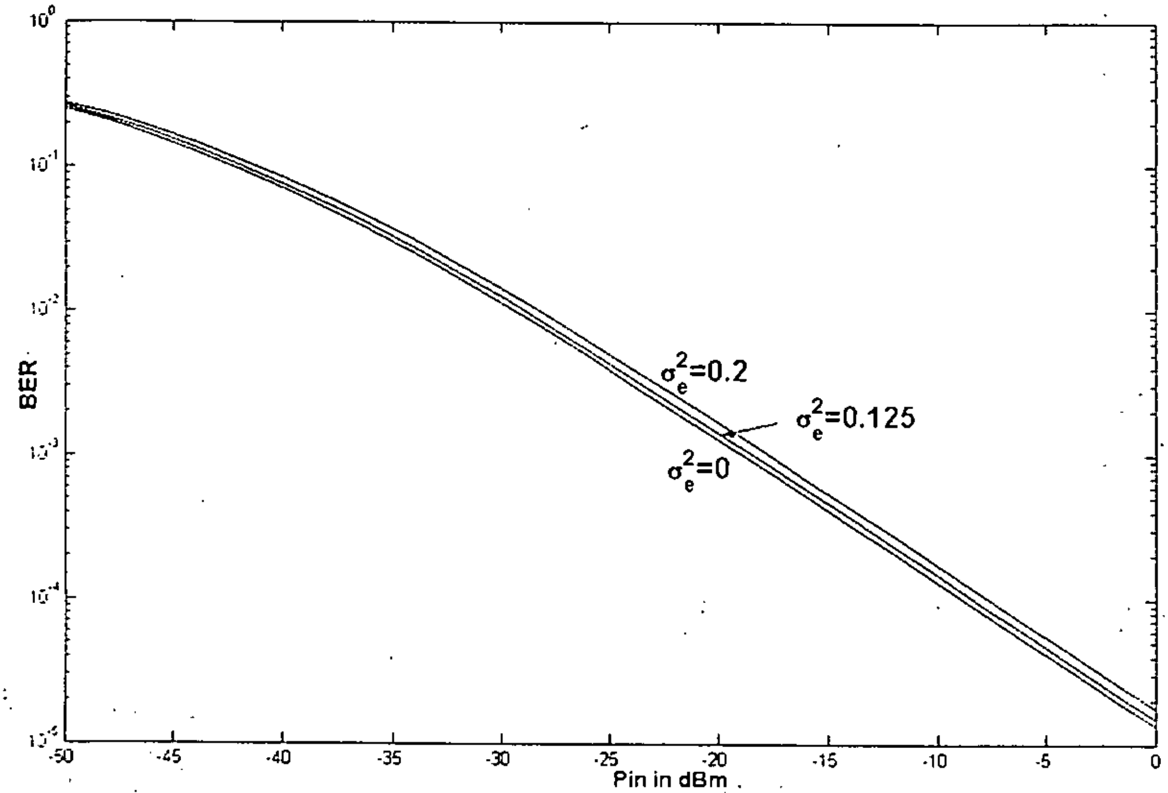

Fig 4.1.1: BER vs. Pin (dBm) in presence of timing jitter for DQPSK-OFDM
(N_s=16, F_d=60 Hz, σ_f^2=0.1, $T_s=10^{-6}$s)

Following the analytical approach presented in section 3.1, the bit error rate performance results are evaluated at a data rate of 1 Mbps per OFDM subcarrier with maximum Doppler frequency of 60 Hz in Rayleigh fading channel. Keeping number of subcarriers to sixteen and fading variance to 0.1 we compare the performance of the system with and without jitter. Fig. 4.1.1 shows the plots of BER vs. Pin (dBm) for DQPSK-OFDM system. From Fig. 4.1.1 it is noticed that the BER slightly degrades with increase in jitter.

Fig. 4.1.2 and Fig. 4.1.3 show the plots of BER vs. Pin (dBm) for QPSK and DPSK systems respectively. The plots show that OFDM is robust against lower values of jitter variance but the BER is highly degraded when jitter is substantially high and results in BER floor. At a BER of 10^{-9}, the jitter effect is negligible for jitter variance σ_e^2 <0.125 for both QPSK and DPSK modulation. It is found that the jitter effect is more pronounced in QPSK and DPSK

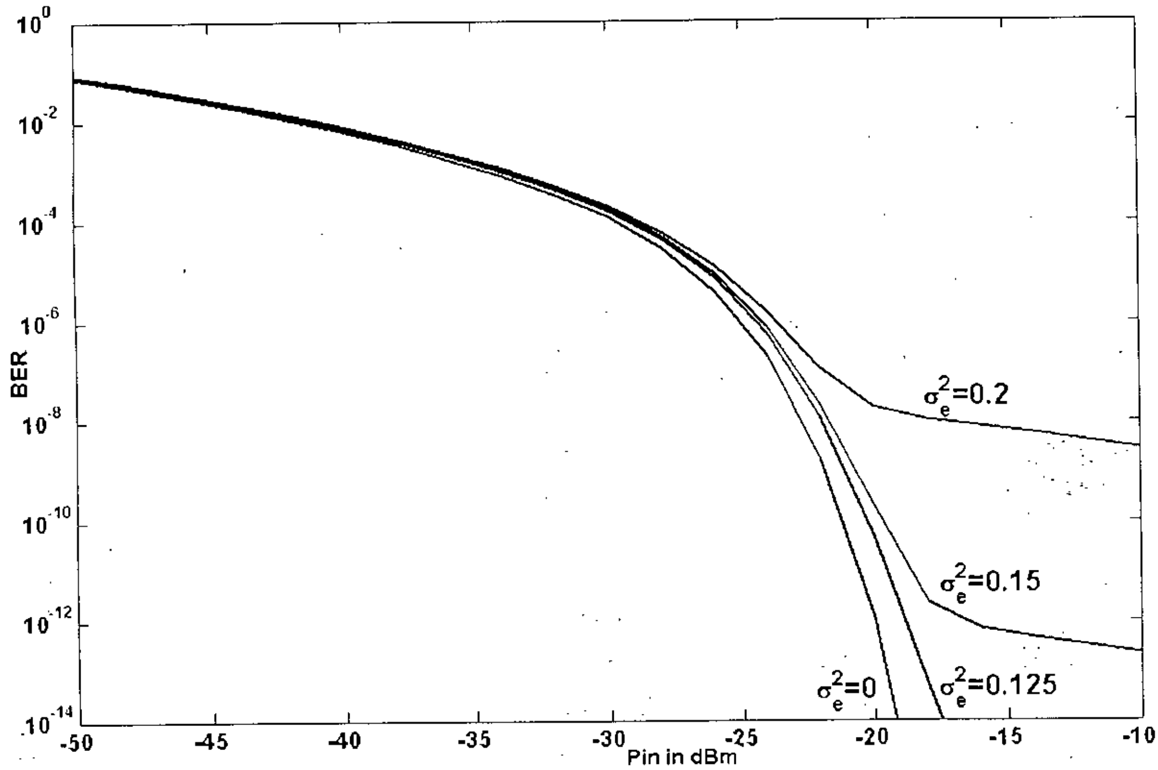

Fig. 4.1.2 BER vs. P_{in} (dBm) in presence of timing jitter for QPSK-OFDM
(N_s=16, F_d=60 Hz, σ_f^2=0.1, T_s=10^{-6}s)

than in DQPSK system. The amounts of penalty suffered by the systems due to timing jitter are shown in Table 4.1.1. It is noticed that the QPSK system suffers almost the same amount of power penalty as DPSK system for lower values of jitter variance and at higher values of jitter, DPSK suffers more penalty than QPSK For example, At a jitter variance of 0.2, the penalty at a BER of 10-8 is 10.7dB for DPSK and 8.5dB for QPSK.

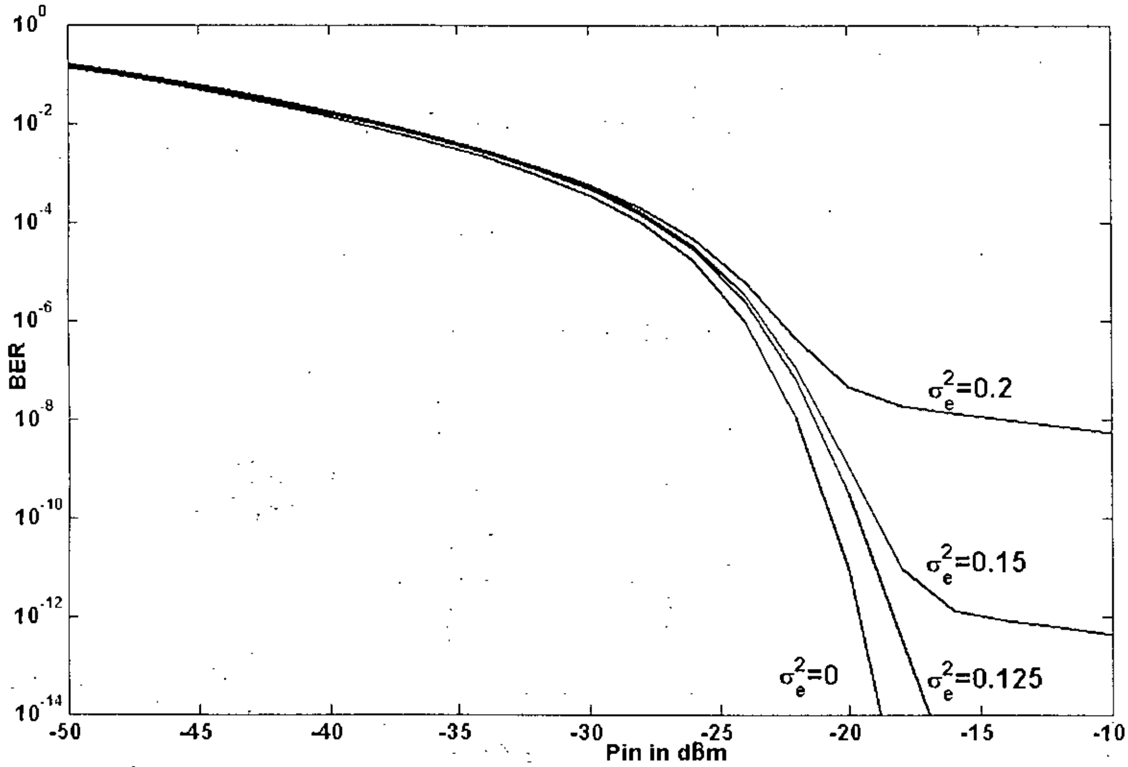

Fig. 4.1.3: BER vs. P_{in} (dBm) in presence of timing jitter for DPSK-OFDM (N_s=16, F_d=60 Hz, σ_f^2=0.1, $T_s=10^{-6}$s)

Table 4.1.1: Power penalty (in dB) due to jitter at BER=10^{-8} and BER=10^{-6}

Type	BER=10^{-8}			BER=10^{-6}		
	σ_ϵ^2 =0.2	σ_ϵ^2 =0.15	σ_ϵ^2 =0.125	σ_ϵ^2 =0.2	σ_ϵ^2 =0.15	σ_ϵ^2 =0.125
QPSK	8.5	1.2	0.8	1.5	0.8	0.6
DPSK	10.7	1.3	0.85	1.7	0.85	0.65
DQPSK	-	-	-	0.85	0.65	0.5

Fig. 4.1.4, Fig. 4.1.5, Fig. 4.1.6 show the plots of the system with and without convolutional coding in presence of jitter for DQPSK, QPSK and DPSK respectively. From Table 4.1.2 it is found that, significant improvement of the BER performance is achieved by applying convolutional coding. For convolution code of rate ½, the coding gain is 12dB for constraint length K=6 and 13 dB for K=7 at an uncoded BER of 10^{-9} with QPSK modulation. It is also noticed that for higher amounts of input power, the coding gain is substantially higher in K=7 than in K=6.

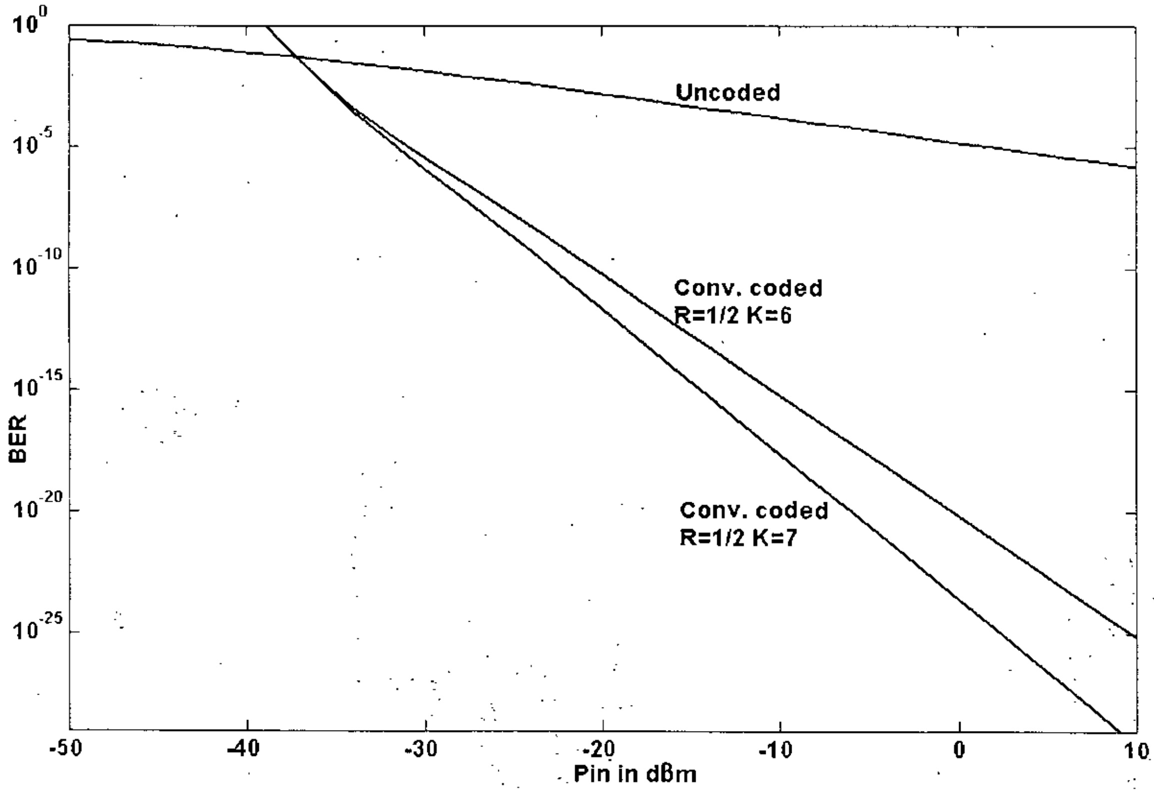

Fig. 4.1.4: BER vs. P_{in} (dBm) with and without coding for DQPSK-OFDM

(N_s=16, F_d=60 Hz, σ_e^2=0.2, σ_f^2=0.1, T_s=10^{-6}s)

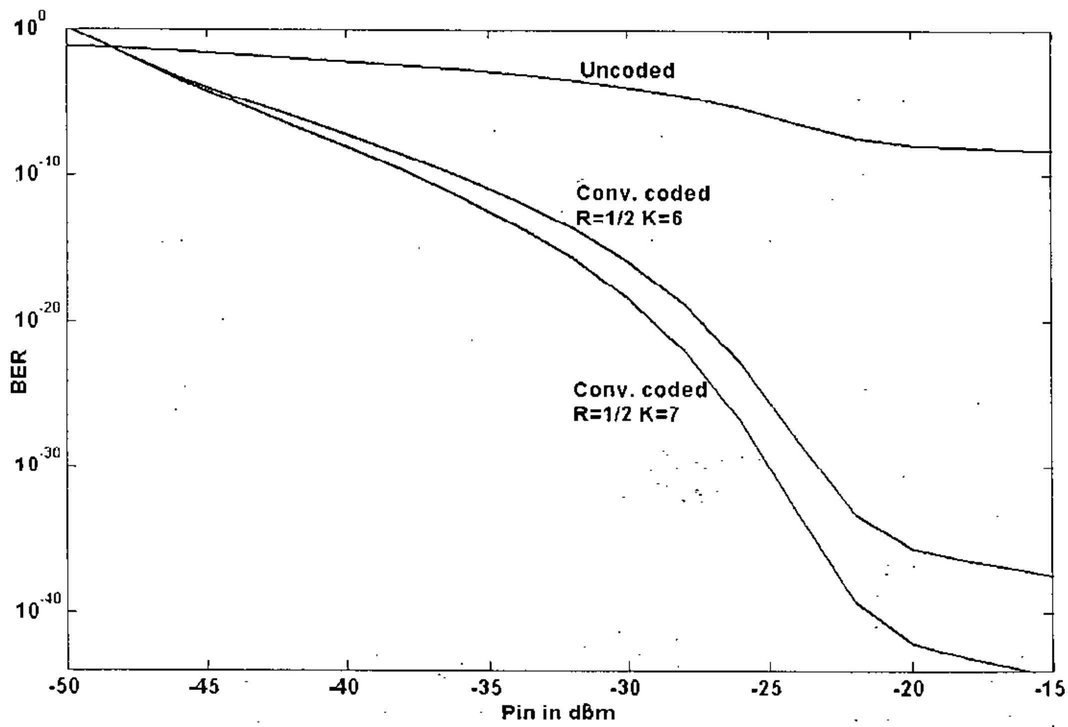

Fig. 4.1.5: BER vs. P_{in} (dBm) with and without coding for QPSK-OFDM
(N_s=16, F_d=60 Hz, σ_c^2=0.2, σ_f^2=0.1, T_s=10^{-6}s)

Table 4.1.2: BER Improvement due to coding for OFDM

Modulation	Pin (dBm)	Uncoded BER	BER (R=1/2 K=6)	BER (R=1/2 K=7)
DQPSK	-20	10^{-3}	10^{-11}	10^{-13}
QPSK	-20	10^{-10}	10^{-36}	10^{-42}
DPSK	-20	10^{-10}	10^{-34}	10^{-40}

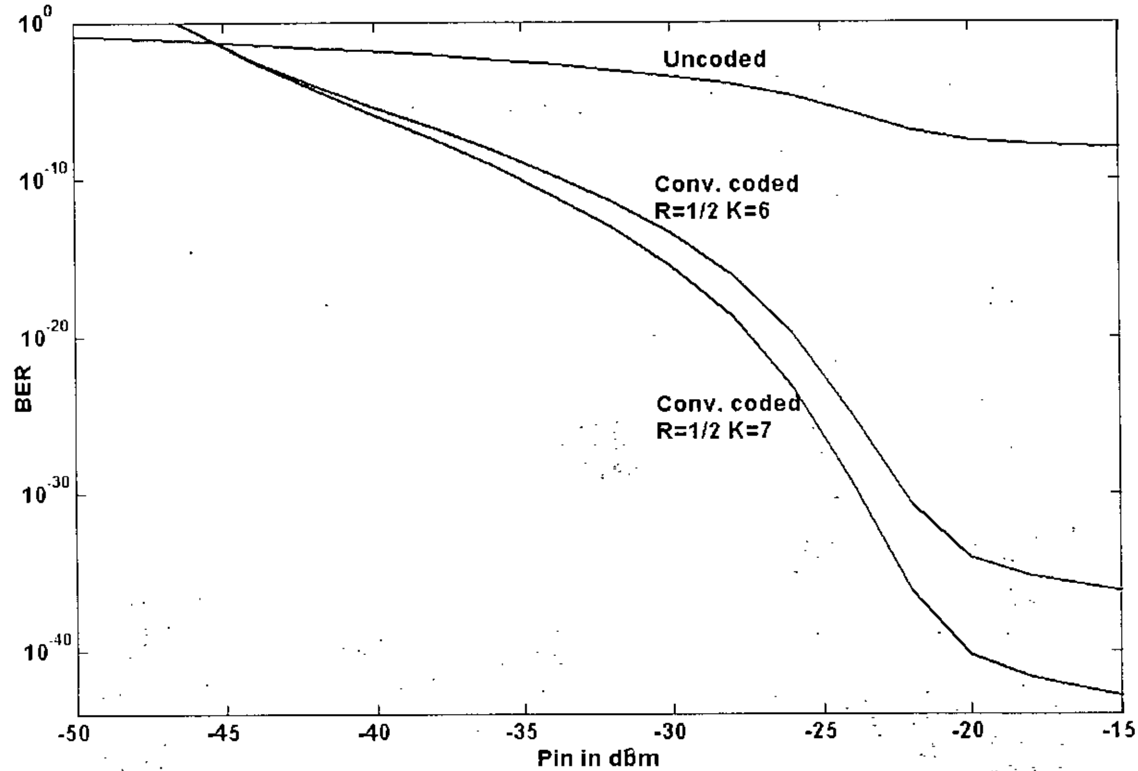

Fig. 4.1.6: BER vs. P_{in} (dBm) with and without coding for DPSK-OFDM

(N_s=16, F_d=60 Hz, σ_c^2=0.2, σ_f^2=0.1, $T_s=10^{-6}$s)

Fig. 4.1.7 shows the plots of BER vs. Pin (dBm) for DQPSK system in Rayleigh and Rician channels. It is noticed that BER performance is better in Rician fading channel. This is because in Rician channel there exists one dominant line-of-sight path that is absent in Rayleigh channel. From Fig. 4.1.7 it is also revealed that BER performance improves with increase in Rician factor. For example, at 0dB input power and DQPSK muduation, the BER of a Rayleigh fading channel is 10^{-4} while Rician channels have a BER in the order of 10^{-5}, 10^{-6} and 10^{-7} for K=0dB, 3dB and 6dB respectively. This result can be explained from the fact that higher Rician factor indicates more dominant line-of-sight path.

Fig. 4.1.8 shows the plots of BER vs. Pin (dBm) for the analysis presented by Sasamori in [16] and the proposed analysis of this thesis. The plots are considered for DQPSK system in Rayleigh channel with F_d=60 Hz and $T_s=10^{-6}$s. The proposed analysis shows more BER than the previous analysis presented by Sasamori [16], because the proposed analysis considers additional impairments of timing jitter (σ_c^2=0.1) and channel attenuation (σ_f^2=0.1).

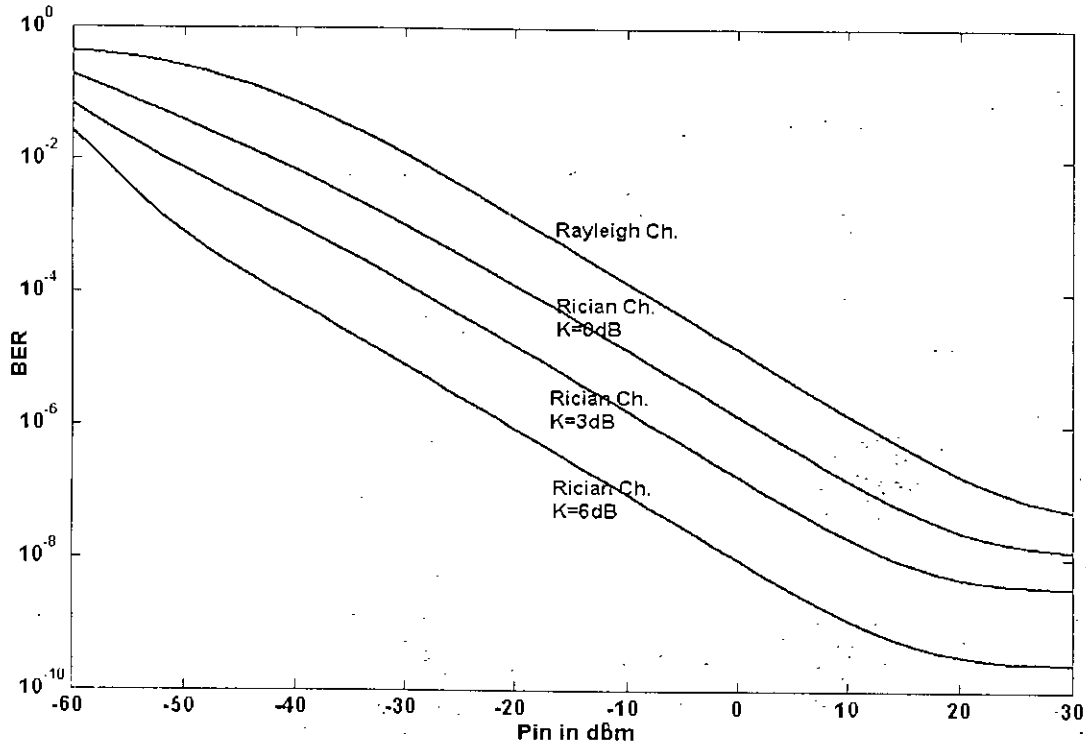

Fig. 4.1.7: BER vs. P_{in} (dBm) for Rayleigh and Rician channels for DQPSK-OFDM (N_s=16, F_d=60 Hz, σ_ϵ^2=0.2, σ_f^2=0.1, T_s=10^{-6}s)

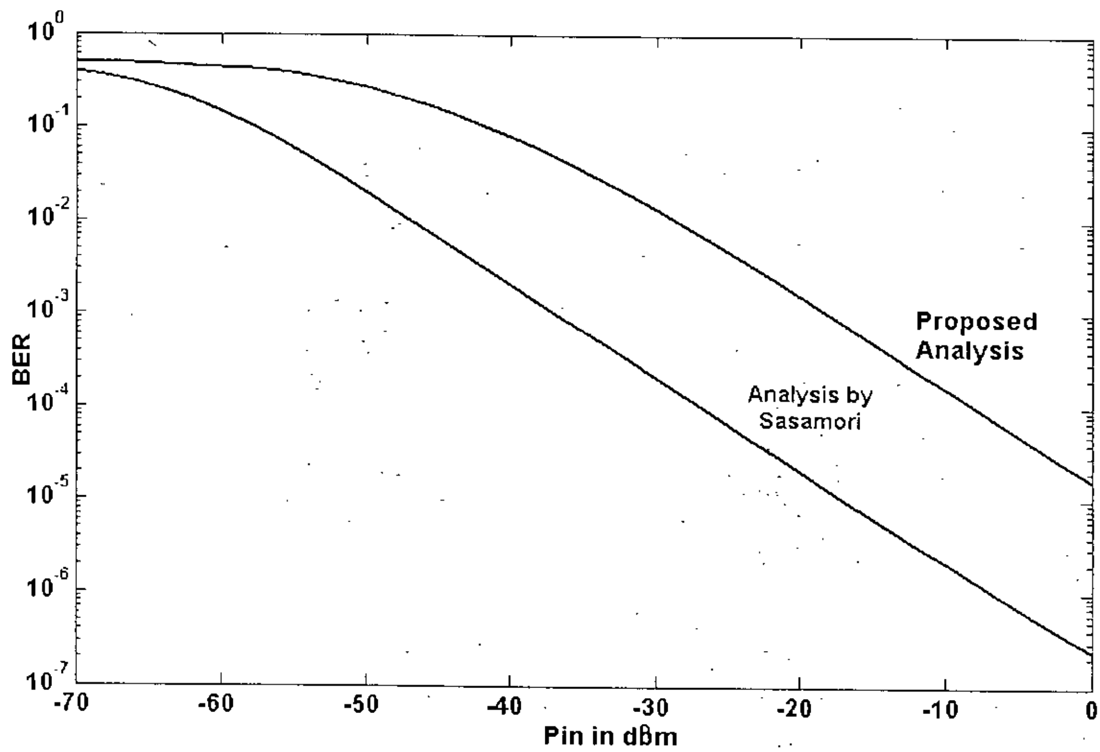

Figure 4.1.8: BER vs. P_{in} (dBm) comparison for previous and proposed analysis

4.2 Performance Results of a STBC-OFDM System:

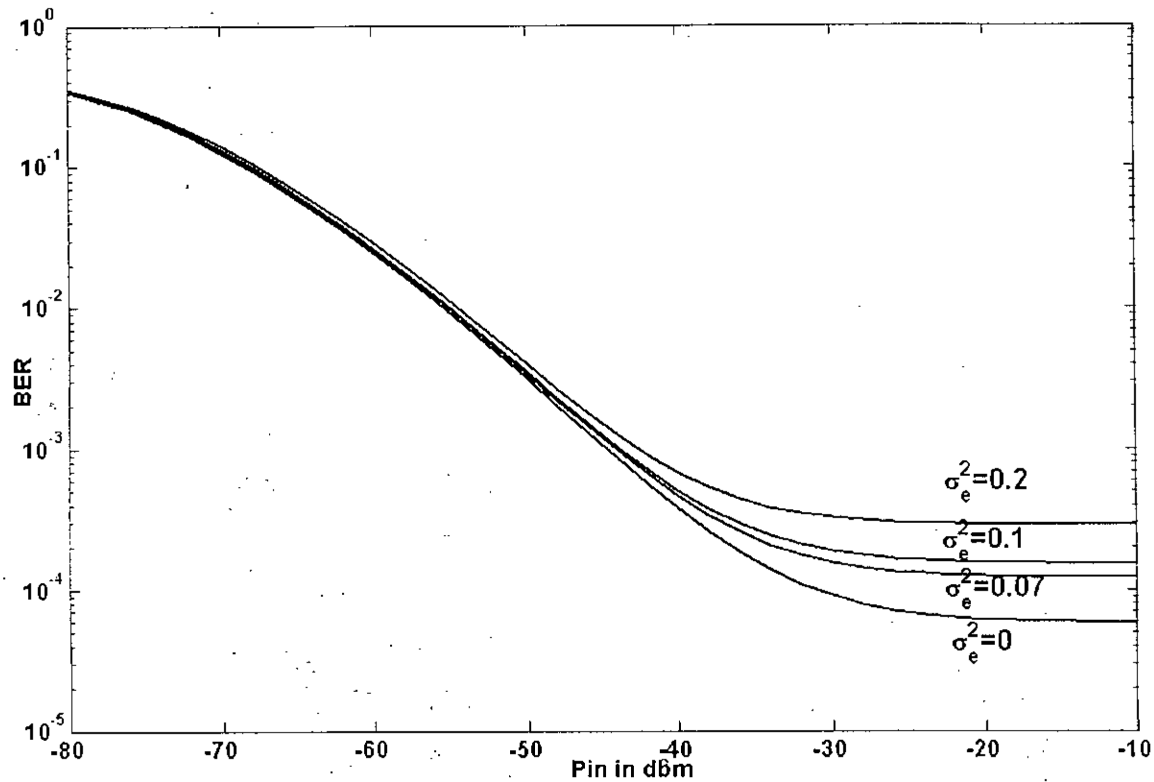

Figure 4.2.1: BER vs. P_{in} (dBm) in presence of timing jitter for STBC-OFDM (DQPSK)
(N_s=16, F_d=60 Hz, σ_f^2=0.1, $T_s=10^{-6}$s, T_x=4, R_x=1)

Following the theoretical analysis presented in section 3.2, we evaluate the BER performance of a STBC-OFDM wireless communication system considering the effect of timing jitter. We consider four transmit antennas, sixteen OFDM subcarriers, maximum Doppler frequency of 60 Hz and 1 Mbps data rate. Fig. 4.2.1 shows the effect of timing jitter in a DQPSK OFDM system with one receiving antenna and fading variance of 0.1. From Fig 4.2.1. it is revealed that with the increase in jitter the BER performance degrades and results in BER floor for higher values of jitter variance. At a jitter variance σ_e^2 of 0.2, the BER floor occurs at about 10^{-3} compared to 10^{-4} for the case of without timing error.

We also evaluate the similar performance of the system employing QPSK and DPSK modulation as shown in Fig. 4.2.2 and Fig. 4.2.3 respectively. It is noticed that the jitter causes BER floor in both the cases. It is found that the jitter effect is more pronounced in

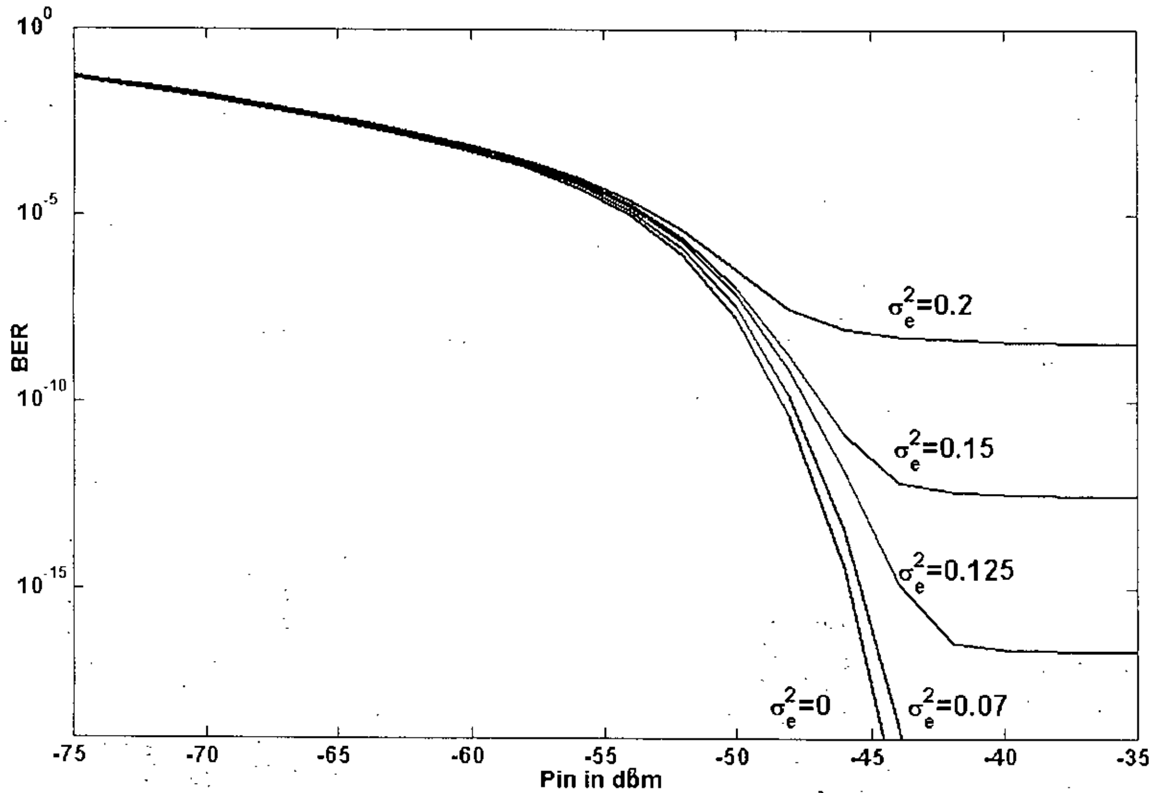

Figure 4.2.2: BER vs. P_{in} (dBm) in presence of timing jitter for STBC-OFDM (QPSK) (N_s=16, F_d=60 Hz, σ_f^2=0.1, T_s=10^{-6}s, T_x=4, R_x=1)

QPSK and DPSK than in DQPSK in terms of increase in BER floor. The amounts of penalty suffered by the systems due to timing jitter at BER=10^{-8} are shown in Table 4.2.1. It is noticed that the QPSK system suffers almost the same amount of power penalty as DPSK system for lower values of jitter variance and at higher values of jitter, DPSK suffers more penalty than QPSK For example, At a jitter variance of 0.2, the penalty at a BER of 10^{-8} is 9dB for DPSK and 6dB for QPSK.

Table 4.2.1: Power penalty (in dB) due to jitter at BER=10^{-8}

Type	BER=10^{-8}			
	σ_ϵ^2 =0.07	σ_ϵ^2 =0.125	σ_ϵ^2 =0.15	σ_ϵ^2 =0.2
QPSK	0.3	0.8	1.1	6
DPSK	0.3	0.8	1.2	9

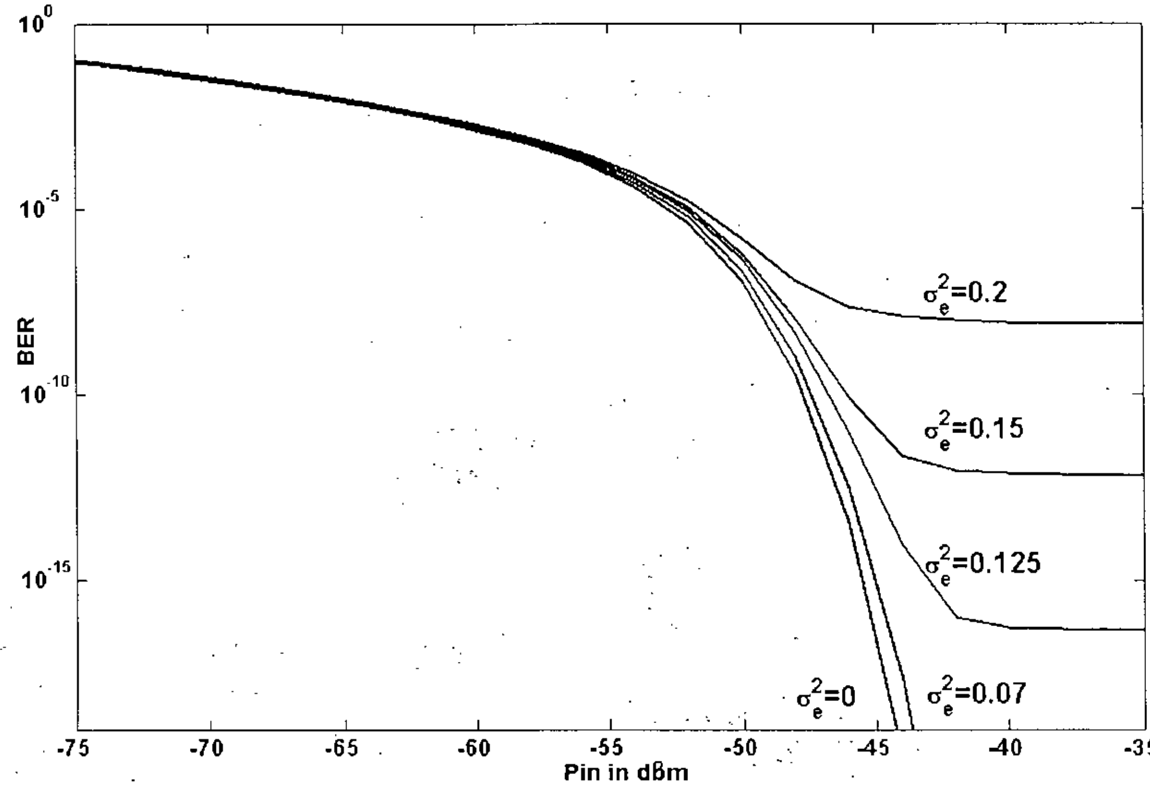

Figure 4.2.3: BER vs. P_{in} (dBm) in presence of timing jitter for STBC-OFDM (DPSK) (N_s=16, F_d=60 Hz, σ_f^2=0.1, $T_s=10^{-6}$s, T_x=4, R_x=1)

Fig. 4.2.4, Fig. 4.2.5, Fig. 4.2.6 show the plots of the system with and without convolutional coding in presence of jitter for DQPSK, QPSK and DPSK respectively. From Table 4.2.2, it is found that, significant improvement of the BER performance is achieved by applying convolutional coding. For convolution code of rate ½, the coding gain is 19dB for constraint length K=6 and 20dB for K=7 at an uncoded BER of 10^{-9} with QPSK modulation. It is also noticed that for higher amounts of input power, the coding gain is substantially higher in K=7 than in K=6.

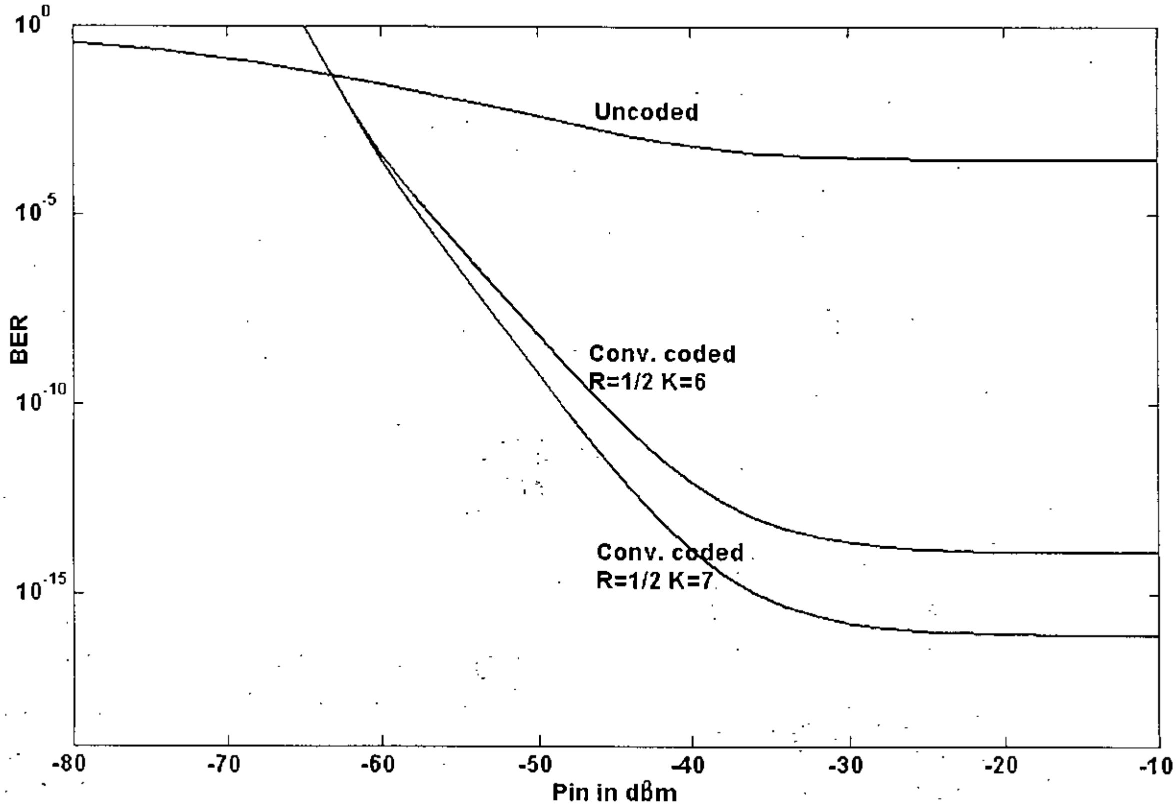

Fig. 4.2.4: BER vs. P_{in} (dBm) with and without coding for DQPSK-OFDM (N_s=16, F_d=60 Hz, σ_ϵ^2=0.2, σ_f^2=0.1, T_s=10^{-6}s, T_x=4, R_x=1)

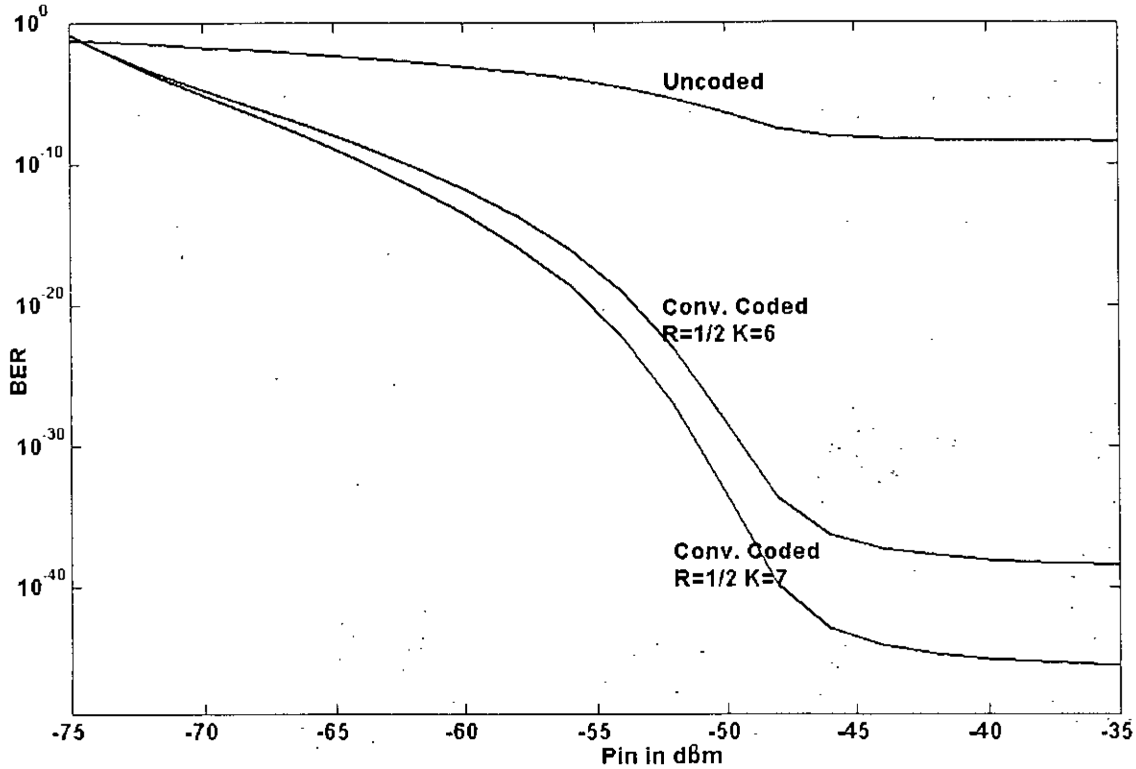

Fig. 4.2.5: BER vs. P_{in} (dBm) with and without coding for QPSK-OFDM
(N_s=16, F_d=60 Hz, σ_c^2=0.2, σ_f^2=0.1, $T_s=10^{-6}$s, T_x=4, R_x=1)

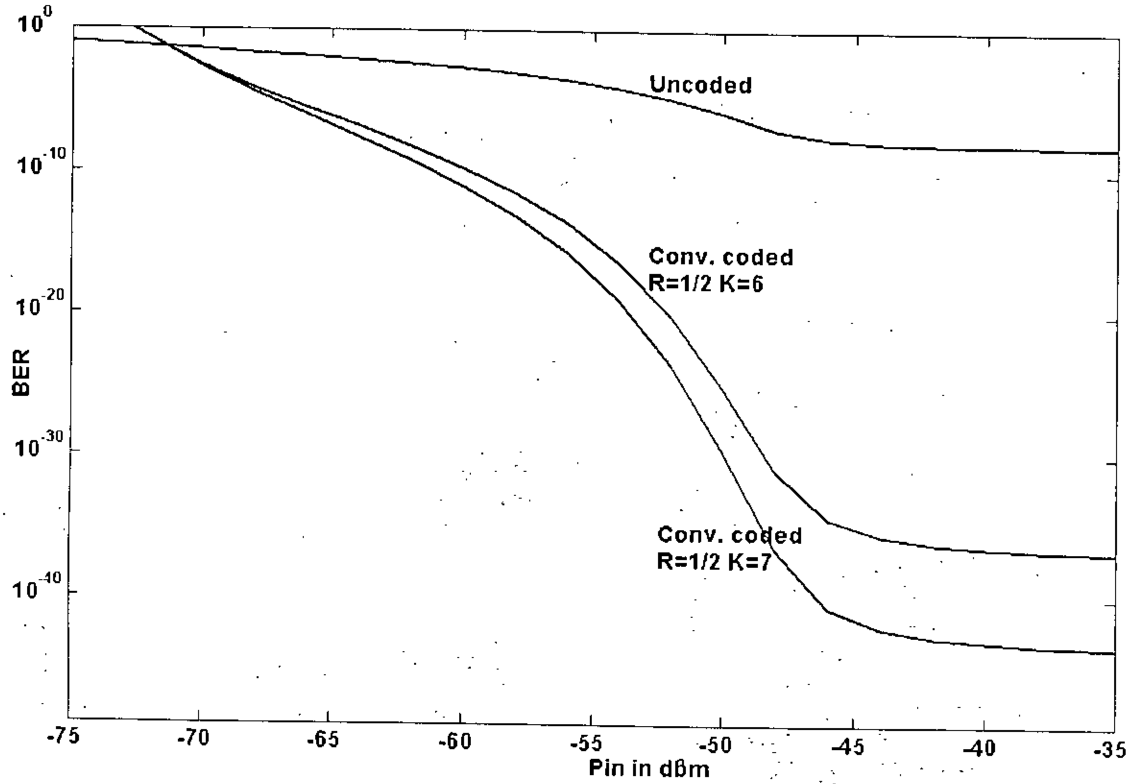

Fig. 4.2.6: BER vs. P_{in} (dBm) with and without coding for DPSK-OFDM

(N_s=16, F_d=60 Hz, σ_ϵ^2=0.2, σ_f^2=0.1, T_s=10^{-6}s, T_x=4, R_x=1)

Table 4.2.2: BER Improvement due to coding for STBC-OFDM

Modulation	Pin (dBm)	Uncoded BER	BER (R=1/2 K=6)	BER (R=1/2 K=7)
QPSK	-40	10^{-9}	10^{-39}	10^{-46}
DPSK	-40	10^{-9}	10^{-37}	10^{-43}
DQPSK	-40	10^{-3}	10^{-13}	10^{-15}
DQPSK	-20	10^{-4}	10^{-14}	10^{-16}

Fig. 4.2.7 shows the effect of fading variance on the BER performance of QPSK OFDM system with one receiving antenna and four transmitting antenna. It is found that with decrease in fading variance there is degradation in system BER performance. Keeping jitter variance fixed to 0.1, we find that at fading variance σ_f^2 =0.25 the BER is 10^{-5} whereas at σ_f^2 =0.05 the BER increases to 10^{-3} at -60dBm input power.

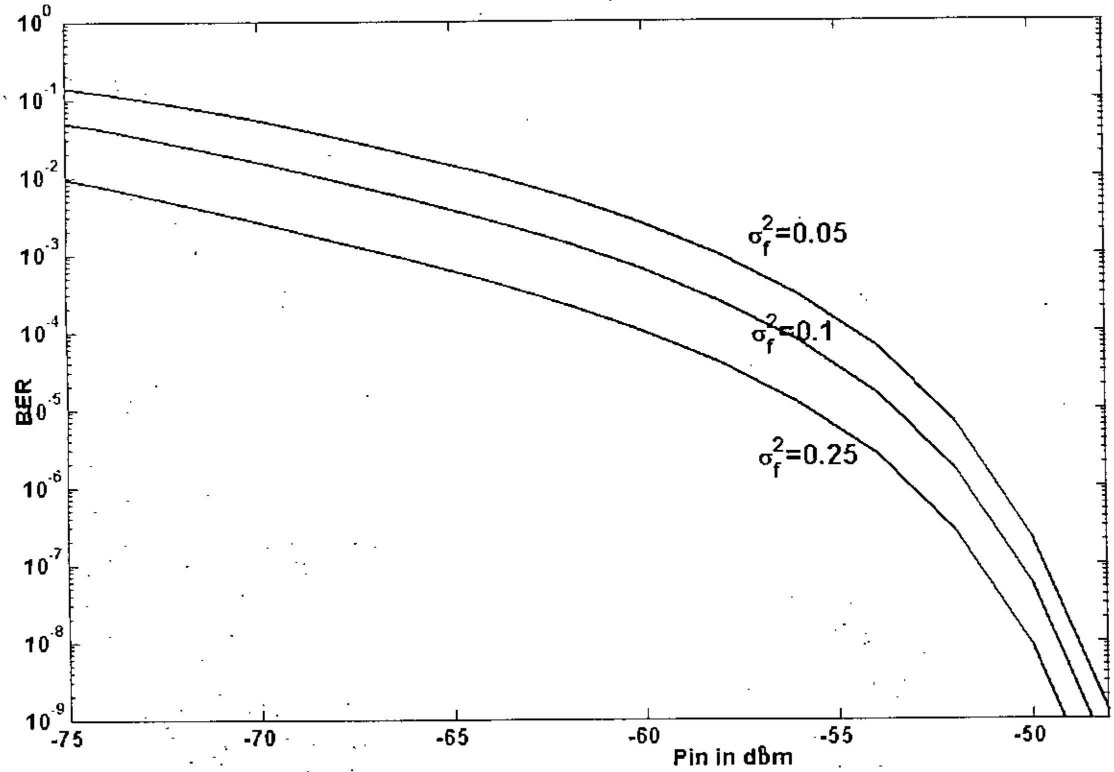

Figure 4.2.7: BER vs. P_{in} (dBm) with variation in fading for STBC-OFDM (QPSK) (N_s=16, F_d=60 Hz, σ_ϵ^2=0.1, T_s=10^{-6}s, T_x=4, R_x=1)

Fig. 4.2.8 illustrates the performance comparison of STBC-OFDM with and without receiver diversity. The plot shows that the BER reduces from 10^{-3} to 10^{-6} in presence of jitter σ_ϵ^2=0.2 by deploying diversity combining in receiving side with two receiving antenna. It shows two pair of curves, one without jitter and another with jitter variance of 0.2.

Fig. 4.2.9 shows the plots of Space time block coded MIMO-OFDM with DQPSK modulation system with and without convolutional coding in presence of jitter considering four transmitting and two receiving antenna. For convolution code of rate ½, the coding gain is 26dB for constraint length K=6 and 28dB for K=7 at an uncoded BER of 10^{-7}. It is also noticed that for higher amounts of input power, the coding gain is substantially higher in K=7 than in K=6.

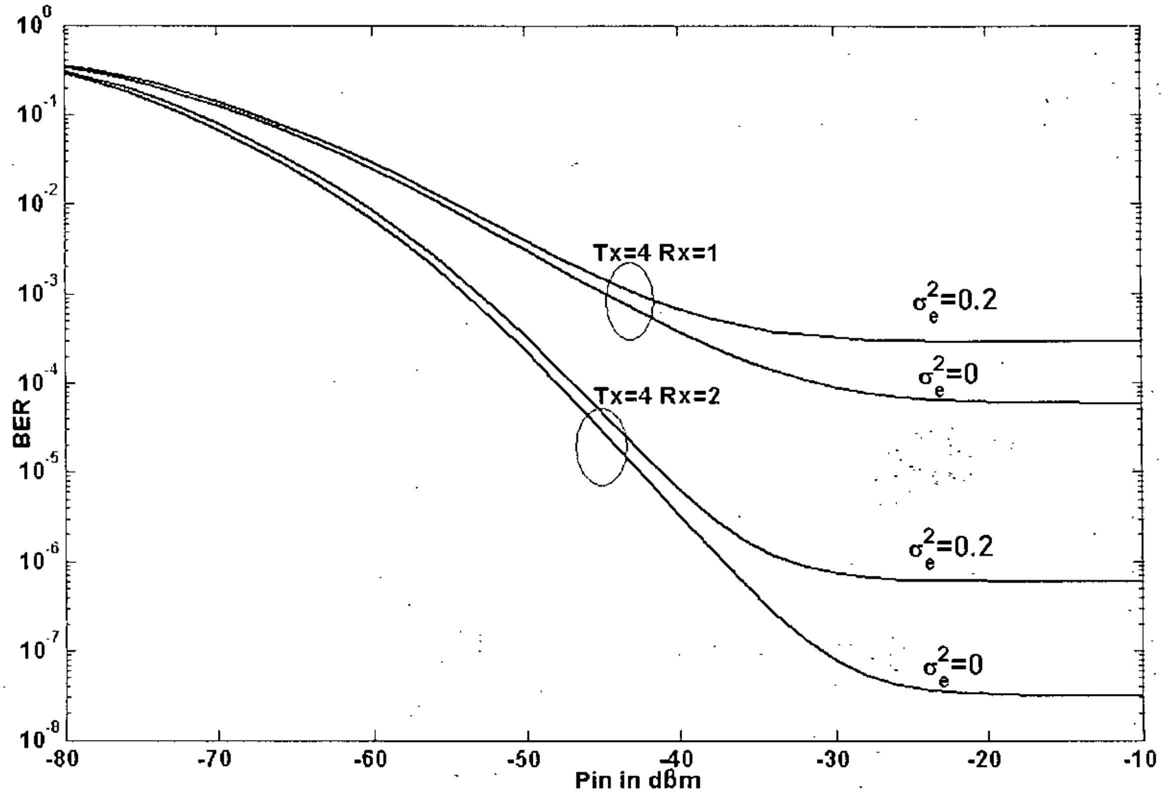

Figure 4.2.8: Plots of BER vs. P_{in} (dBm) with & without receiving diversity for STBC-OFDM (DQPSK)

(N_s=16, F_d=60 Hz, σ_f^2=0.1, T_s=10^{-6}s, T_x=4)

Fig. 4.2.10 and Fig. 4.2.11 show the plots of BER vs. number of subcarriers (N_s) for different values of Doppler frequency normalized by symbol period. It is noticed that with the increase in Doppler frequency the BER increases. From Fig. 4.2.10 it is also revealed that with increase in N_s, the BER decreases and becomes minimal and then increases for a particular value of Doppler frequency. From Fig. 4.2.11 it is observed that BER is minimal when number of subcarriers (N_s) is 12.

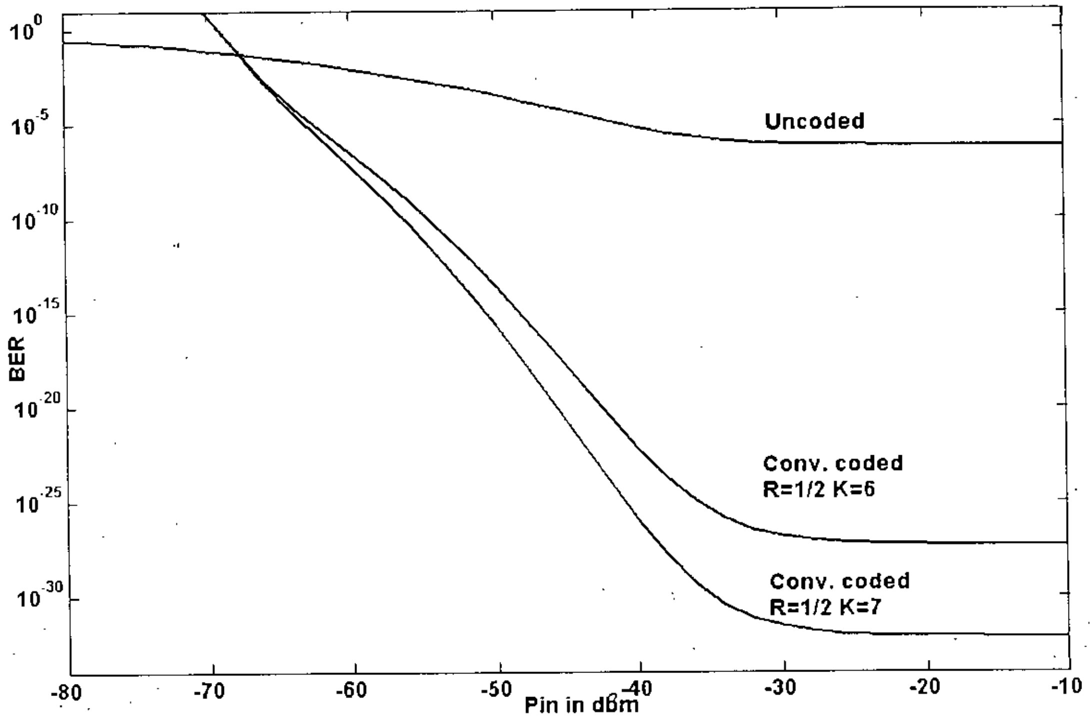

Figure 4.2.9: BER vs. P_{in} (dBm) with and without coding for MIMO-OFDM (DQPSK)
(N_s=16, F_d=60 Hz, σ_ϵ^2=0.2, σ_f^2=0.1, T_s=10^{-6}s, T_x=4, R_x=2)

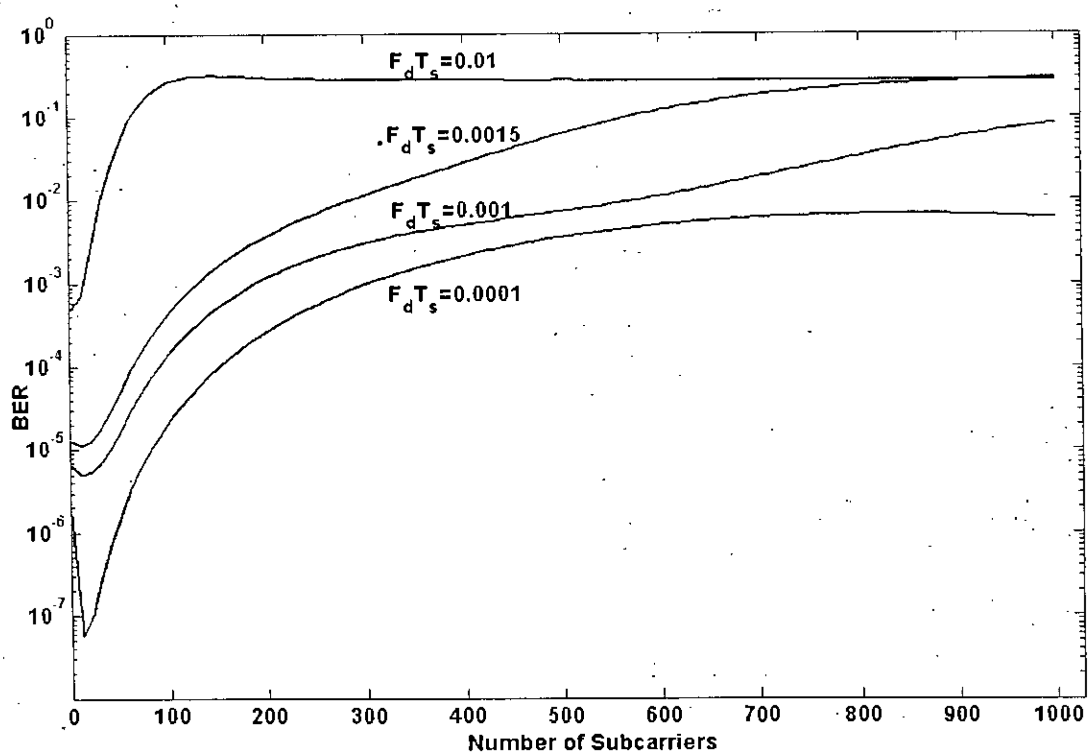

Figure 4.2.10: BER vs. N_s (number of subcarriers) for MIMO-OFDM (DQPSK)
(F_d=60 Hz, σ_ϵ^2=0.01, σ_f^2=0.1, T_s=10^{-6}s, T_x=4, R_x=2)

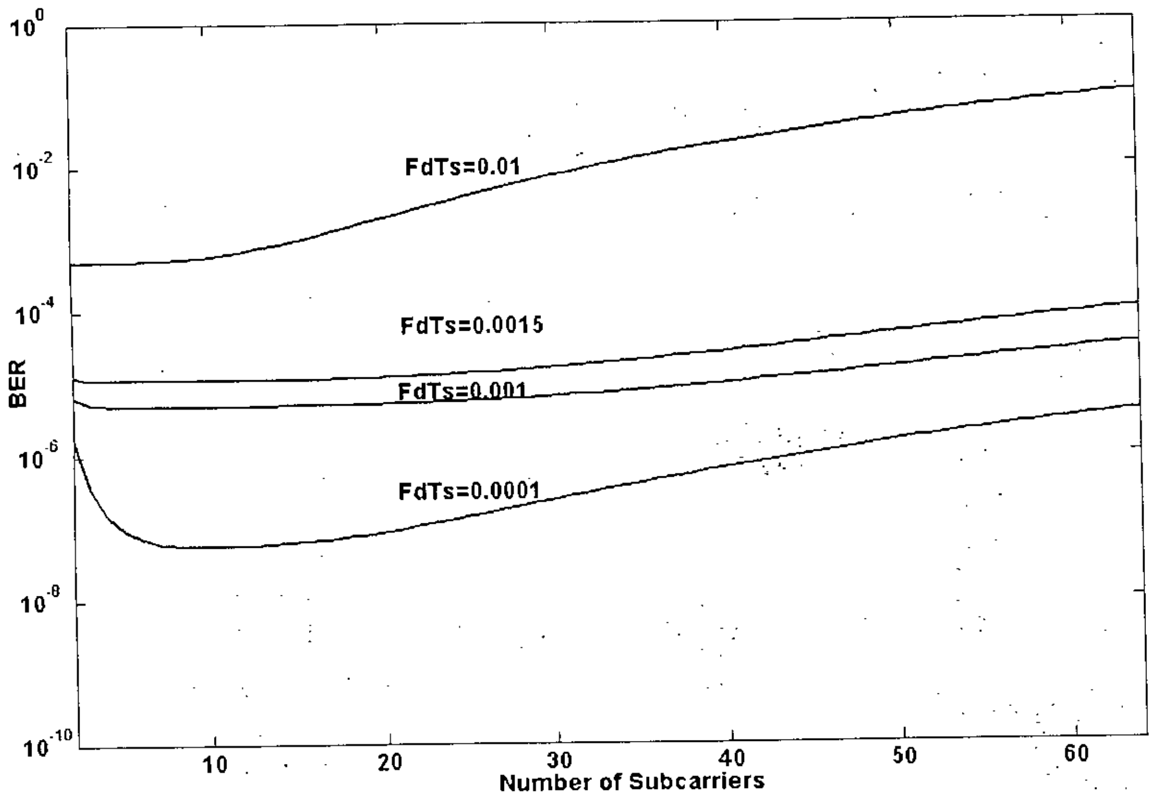

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Figure 4.2.11: BER vs. N_s for MIMO-OFDM (DQPSK) with values of N_s upto 64
(F_d=60 Hz, σ_ε^2=0.01, σ_f^2=0.1, T_s=10^{-6}s, T_x=4, R_x=2)

Fig. 4.1.12 shows the plots of BER vs. Pin (dBm) for the analysis presented by Zhang in [22] and the proposed analysis of this thesis. The plots are considered for Rayleigh channel with F_d=60 Hz, T_s=10^{-6}s, four number of transmitting and single receiving antennas. The proposed analysis shows more BER than the previous analysis presented by Zhang [22], because the proposed analysis considers additional impairments of timing jitter (σ_ε^2=0.1) and channel attenuation (σ_f^2=0.1).

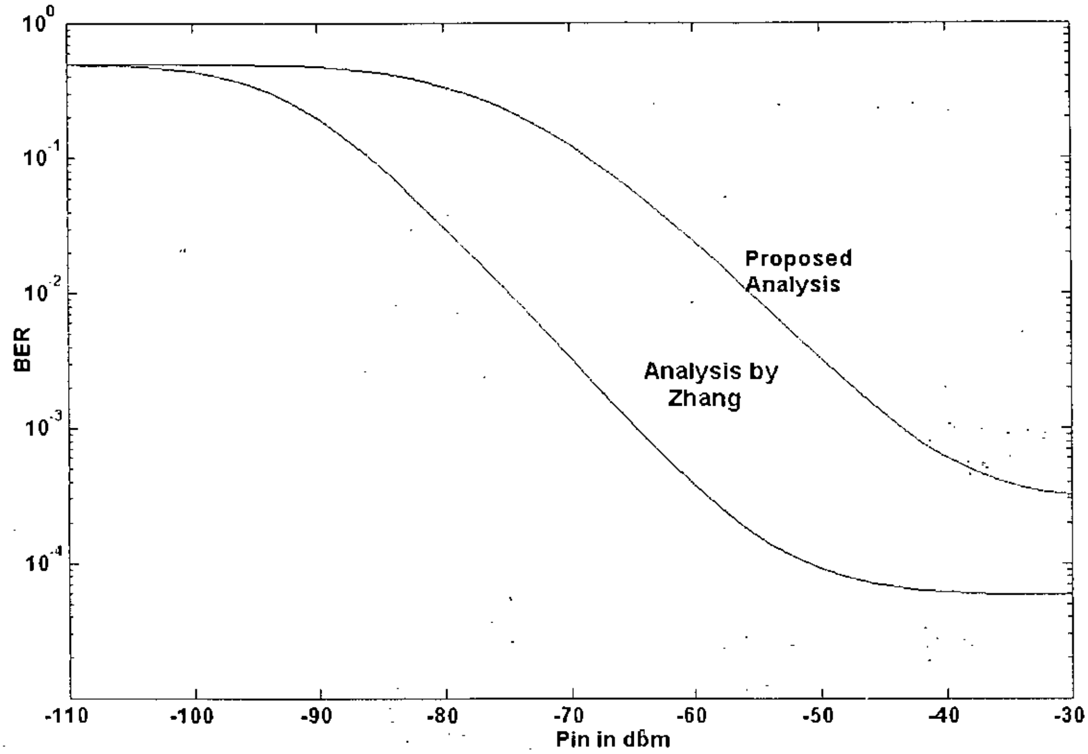

Figure 4.2.12: BER vs. P_{in} (dBm) for previous and proposed analysis (F_d=60 Hz, $T_s=10^{-6}$s, T_x=4, R_x=1)

From the findings of the evaluation we can propose an optimum system with following design parameters:

1. Convolutional coding of rate ½ and K=7
2. STBC-OFDM with Tx=4, Rx=2
3. Number of OFDM subcarriers 12
4. DQPSK as modulation scheme as it is most robust against jitter.

Chapter 5

Conclusion and Future Work

5.1 Conclusion

OFDM, which is effective in avoiding ISI that multipath delay might cause, is very vulnerable to time-selectivity of the channel. However, it is also sensitive to AWGN and timing jitter. In this thesis, we have developed an analytical approach to evaluate the combined effects of timing jitter, time selective fading and AWGN on the BER performance of an OFDM system. We consider both the Doppler frequency and channel attenuation as the outcome of multipath fading in our BER calculation. We compare the performance for DQPSK, QPSK and DPSK modulation schemes and find that none of them can completely eliminate the channel impairments. We also notice that for DQPSK the jitter effect is less pronounced than coherently detected QPSK and DPSK. For example, at a jitter variance of 0.2, the penalty at a BER of 10^{-6} is 1.7dB for DPSK, 1.5dB for QPSK and 0.85dB for DQPSK. It is also noticed that the QPSK system suffers almost the same amount of power penalty as DPSK system for lower values of jitter variance and at higher values of jitter, DPSK suffers more penalty than QPSK. For example, at a jitter variance of 0.2, the penalty at a BER of 10^{-8}, is 10.7dB for DPSK and 8.5dB for QPSK whereas at a jitter variance of 0.125, the penalty is 0.83dB for DPSK and 0.80dB for QPSK. We also observe that the BER of an OFDM system is much higher in Rayleigh fading channel than Rician channel. We compute the BER of OFDM in Rician fading channels with different Rician parameters and numerical results show that the performance is improved with higher values of Rician parameters.

Multi-antenna OFDM including STBC-OFDM and MIMO-OFDM are capable of achieving spatial diversity and/or increasing spectral efficiency. However, similar to single-antenna OFDM, multi-antenna OFDM system is also sensitive to those three channel impairments. We evaluate the performance of quasi-orthogonal STBC-OFDM system having multiple transmit and single receiving antenna. It is noticed that STBC-OFDM shows better BER performance than OFDM system. For QPSK modulation, at a jitter variance of 0.2, the penalty at a BER of 10^{-8} is 8.5 dB for OFDM and 6dB for STBC-OFDM. Numerical computations indicate that BER becomes higher with increase in Doppler frequency and increase in number of subcarirers. We extend the analysis for a MIMO-OFDM system with selection method for combining multiple receiving antennas and find substantial improvement in the system performance. For example, the BER reduces from 10^{-3} to 10^{-6} in presence of jitter σ_{ϵ}^{2}=0.2 by deploying diversity combining in receiving side with two receiving antennas.

We apply convolutional coding to mitigate the effect of channel impairments in the performance of single-antenna and multi-antenna OFDM system. It is observed that at –20dBm input power the uncoded MIMO-OFDM system results in BER of 10^{-6} whereas coded system shows BER of 10^{-28}. As coded MIMO-OFDM successfully achieves outstanding performance, it can be used in practical applications where fading, jitter and AWGN strongly exist.

5.2 Suggestions for Future Work

In our analysis we have considered the combined effects of fading, jitter and Gaussian noise in Rayleigh and Rician fading channels. However a few suggestions for future work are given below:

OFDM is very sensitive to phase synchronization errors, one of them being Wiener phase noise. Phase noise reflects imperfections of the local oscillator (LO), i.e. random drift of the LO phase from its reference. So to take into account this noise effect the expressions for SNIR should be modified.

We have considered MIMO-OFDM with four transmitting and only two receiving antennas. The theoretical analysis can be done for four or more number receiving antennas. Bluetooth systems produce unintentional interference in orthogonal frequency division multiplexing (OFDM) systems using the industry, science and medical (ISM) band. The performance of OFDM systems in the presence of Bluetooth interference can be numerically analyzed in fading channels in addition to AWGN and jitter.

Convolutional codes are now giving way to turbo codes, a new class of iterated short convolutional codes that closely approach the theoretical limits imposed by Shannon's theorem with much less decoding complexity. Turbo coding can be used as error correcting channel code instead of convolutional coding in our analysis.

Appendix:

Derivation of SNIR in an STBC-OFDM Systems

The expression of SNIR of a STBC-OFDM system in presence of multipath fading and AWGN is derived as [22].

From Eq. (3.39) the processed received signal X has an expression as follows:

$$X=U(HS+V)=G\Psi+W \tag{A.1}$$

where $x_k=[\, x_1(k), \ldots\ldots x_P(k)\,]^T$, $w_k=[\, w_1(k),\ldots\ldots w_P(k)\,]^T$ and $W=UV=[w_0, \ldots w_{N_s-1}\,]^T$ $(N_s\times P)$

$$x_k^T = g_{k,k}^T\Psi_k + \sum_{k'=0,k'\neq k}^{N_s-1} g_{k,k'}^T\Psi_{k'} + w_k^T, \qquad k=0,\ldots\ldots,N_s-1$$

$$\text{and } g_{k,k'} = [g_{k,k'}^{(1)},\ldots\ldots\ldots\ldots, g_{k,k'}^{(P)}]^T, \quad k, k'=0,\ldots\ldots\ldots, N_s-1 \tag{A.2}$$

$g_{k,k'}$ is the (k,k')th block of G. Let us define three vectors:

$$y_k = [x_1(k),\ x_2^*(k),\ x_3(k),\ x_4^*(k)]^T$$

$$z_k = [w_1(k),\ w_2^*(k),\ w_3(k),\ w_4^*(k)]^T$$

$$\psi_k = [a_1(k),\ a_2(k),\ e^{j\Phi}a_3(k),\ e^{j\Phi}a_4(k)]^T \tag{A.3}$$

From Eq. (A.1), Eq. (A.2) and Eq. (A.3) y_k can be expressed as

$$y_k = M_{k,k}\psi_k + \sum_{k'=0,k'\neq k}^{N_s-1} M_{k,k'}\psi_{k'} + z_k \tag{A.4}$$

where

$$M_{k,k'} = \begin{bmatrix} M_{k,k'}^{(1,2)}(0) & M_{k,k'}^{(3,4)}(0) \\ M_{k,k'}^{(3,4)}(2(N_s+c_p)) & M_{k,k'}^{(1,2)}(2(N_s+c_p)) \end{bmatrix}. \tag{A.5}$$

$k,k'=0,\ldots\ldots N_s-1$ with

$$M_{k,k'}^{(i,j)}(n) = \begin{bmatrix} g_{k,k'}^{(i)}(n) & g_{k,k'}^{(j)}(n) \\ g_{k,k'}^{(j)*}(n+N_s+c_p) & -g_{k,k'}^{(i)*}(n+N_s+c_p) \end{bmatrix}$$

(A.6)

In the presence of time-varying fading, SINR for quasi-orthogonal STBC-OFDM system is obtained as [22]

$$\text{SINR} = \frac{P_s}{P_n + P_I}$$

$$= \frac{E[\left\| M_{k,k}\psi_k \right\|_F^2]}{E[\left\| \sum_{k'=0,k'\neq k}^{N_s-1} M_{k,k'}\psi_k + z_k \right\|_F^2]}$$

$$= \frac{tr\{E[M_{k,k}\psi_k\psi_k^H M_{k,k}^H]\}}{tr\{E[(\sum_{k'=0,k'\neq k}^{N_s-1} M_{k,k'}\psi_k + z_k)(\sum_{k'=0,k'\neq k}^{N_s-1} M_{k,k'}\psi_k + z_k)^H + z_k z_k^H]\}}$$

$$= \frac{E[\sum_{q=0}^{3}\sum_{p=1}^{4}\left| g_{k,k}^{(p)}(q(N_s+c_p))^2 \right|]}{E[\sum_{k'=0,k'\neq k}^{N_s-1}\sum_{q=0}^{3}\sum_{p=1}^{4}\left| g_{k,k'}^{(p)}(q(N_s+c_p))^2 \right|] + \sigma^2}$$

$$= \frac{4\sum_{l=0}^{L-1}[N_s + 2\sum_{i=1}^{N_s-1}(N_s-1)J_0(2\pi i f_d T_s)]e^{-\frac{\tau_l}{\tau_{rms}}}}{4\sum_{k'=1}^{N_s-1}\sum_{l=0}^{L-1}[N_s + 2\sum_{i=1}^{N_s-1}(N_s-1)J_0(2\pi i f_d T_s)\cos(\frac{2\pi}{N_s}k'i)\, e^{-\frac{\tau_l}{\tau_{rms}}}] + \sigma^2}$$

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