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Project Report on

"Investigating immunity to Fading in MIMO communication with concatenated coding techniques"

Submitted in partial fulfilment for the award of degree of

Master of Technology
in
Digital Communication & Networking

Submitted by

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Under the Guidance
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Certificate

Certified that this project work entitled "**Investigating immunity to Fading in MIMO communication with concatenated coding techniques"** was carried out by **Parvathi.C (1RE11LDN16)** a bonafide student of **Reva Institute of Technology & Management** in partial fulfilment for the award of Master of Technology in **Digital Communication & Networking** of the Visvesvaraya Technological University, Belgaum, during the year **2012-2013.** It is certified that all corrections/suggestions indicated in internal reviews have been incorporated in the report. The project report has been approved as it satisfies the academic requirements in respect of project work prescribed for the said degree.

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UNDERTAKING

I, **PARVATHI.C (1RE11LDN16)** a bonafide student of IV semester, **Master of Technology** in **Digital Communication & Networking**, **Reva Institute of Technology & Management**, Bangalore-64, hereby declare that project work entitled "**Investigating immunity to Fading in MIMO communication with concatenated coding techniques"** has been independently carried out by me under the supervision of my Internal Guide **Mr.Seetha Rama Raju Sanapala,** Sr.Associate Professor, Department of ECE, Reva ITM and External Guide **Dr. Ramesh Balasubramanyam, Associate Professor, Raman Research Institute, Bangalore** and I submitted my report in partial fulfilment of the requirements for the award of degree in Master of Technology in Digital Communication & Networking by **Visvesvaraya Technological University** during the academic year 2012-2013.

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CERTIFICATE

Certified that the project work entitled "**Investigating immunity to Fading in MIMO communication with concatenated coding techniques"**, was carried out by Ms. **PARVATHI.C, USN:1RE11LDN16,** a bonafide student of **REVA Institute of Technology & Management**, Bangalore-64, in partial fulfilment for the award of **Master of Technology** in **Digital Communication & Networking**, of the Visveswaraiah Technological University, Belgaum during the year 2012 – 2013. It is certified that, she has completed the project satisfactorily.

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Abstract

The present third generation wireless cellular communication technology employs single antenna both at the transmitter and at the receiver, which results in point-to-point communication. This configuration is known as Single Input Single Output (SISO). The reliable communication depends on the signal strength. If the signal path is in deep fade, there is no guarantee for reliable communication. Since the channel capacity depends on the signal-to-noise ratio and the number of antennas used at both the transmitter and at the receiver, the capacity provided by SISO is limited. If the transmitted signal, passes through multiple signal paths, each path fades independently, there is a possibility that atleast one path is strong, which ensures reliable communication. This technique is called “diversity”. To achieve this, we need to use multiple antennas at the transmitter side and at the receiver side. This technology is called as Multiple Input Multiple Output (MIMO). Hence the multipath fading channels are considered as an advantage scenario in MIMO. This is the most preferred technique in fourth generation cellular communication, since higher data rates can be achieved. Since fading due to multipath propagation in wireless channel can be modeled as burst mode error, the error control channel coding techniques developed for wired channel can be efficiently implemented for wireless fading channel. Our goal is to assess concatenating various error control codes in the MIMO setting using bit error rate (BER) and effective data rate (EDR) as performance indicators.

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Glossary

AWGN	Additive White Gaussian Noise
BCH code	Bose, Chaudhuri, Hocquenghem code
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CC	Convolutional code
D-BLAST	Diagonal Bell Labs Layered Space-Time Architecture
Eb/No	Bit Energy to Noise ratio
ISI	Intersymbol Interference
LDPC	Low Density Parity Check code
LOS	Line of Sight
MIMO	Multi-Input Multiple-Output
MISO	Multi-Input Single-Output
MLSE	Minimum Least Square Error
MMSE	Minimum Mean Square Error
MRRC	Maximal-ratio receiver combiner
NLOS	Non Line of Sight
OFDM	Orthogonal Frequency Division Multiplexing
OSTBC	Orthogonal Space-Time Block code
QPSK	Quadrature Phase Shift Keying
QAM	Quadrature Amplitude Modulation
RF	Radio Frequency
RMS	Root Mean Square
RS code	Reed-Solomon code
SIMO	Single-Input Multiple-Output
SISO	Single-Input Single-Output
SM	Spatial Multiplexing
SNR	Signal-to-Noise ratio
TCM	Trellis Coded Modulation
V-BLAST	Vertical Bell Labs Layered Space-Time Architecture

Chapter 1

INTRODUCTION

1.1 Introduction to Wireless Communication

In wired communication, the communication takes place over a stable medium like copper wires or optical fibers. The properties of the medium are well defined and nearly time in-variant. Interference from other users either does not happen or the properties of the interference are stationary. The Bit Error Rate (BER) decreases exponentially with increasing Signal-to-Noise Ratio (SNR). This means that a relatively small increase in transmit power can greatly decrease the error rate. In wireless communication, air is the medium. Due to user mobility as well as multipath propagation, the channel varies strongly with time. Interference due to noisy environment and from other users are inherent and the channel is also time-variant due to the mobility of either the transmitter or the receiver or the reflecting components in the channel. The BER decreases only linearly with increasing SNR [1]. Increasing the transmit power does not help in reducing the BER. Hence more sophisticated error control coding techniques are required to decrease the BER.

1.1.1 Multipath Propagation :

In wireless communications, the medium is the radio channel between the transmitter and the receiver. The transmitted signal can propagate through a number of different paths. In some cases, a Line Of Sight (LOS) path might exist between the transmitter and the receiver. The transmitted signal is reflected and scattered by different interacting objects in the environment such as buildings, people, or vehicles. The number of these possible propagation paths is very large. Each path can have a distinct amplitude, delay, and direction of arrival. Each multipath component has different phase shift with respect to each other. As shown in Fig.1.1, each of the paths has different amplitude, time delay, and phase shifts with respect to each other [1].

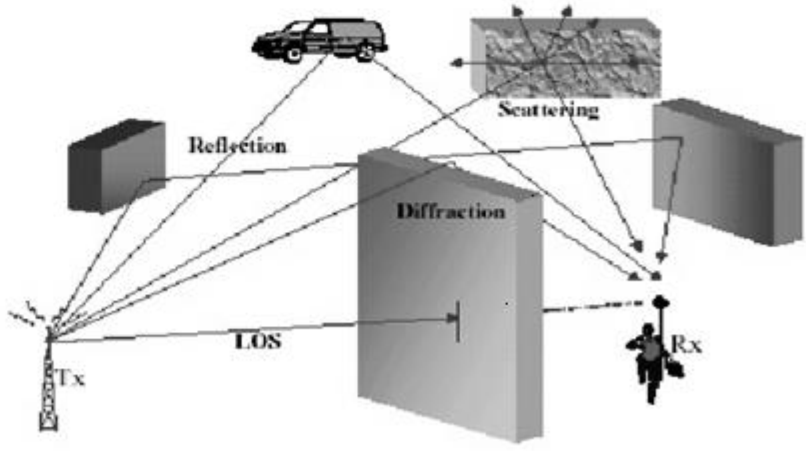


Fig.1.1 Multipath Propagation in Wireless Communication

1.1.2 Impulse Response of a Multipath Channel :

The small scale variations of a mobile radio signal can be directly related to the impulse response of mobile radio channel. The impulse response is a wideband channel characterization. It contains all the information necessary to simulate and analyze the channel. This is because the mobile radio channel can be modeled as a linear filter with a time varying impulse response. Multipath fading is characterized by the channel impulse response which includes the information of relative time, signal power and signal phase, when the delayed signals arrive at the receiver as compared to the direct wave. If the receiver is in motion, then the relative lengths and attenuations of the various reception paths will change with time, that is, the channel is time varying [2]. The channel impulse response in time and frequency domain are shown in Fig.1.2.

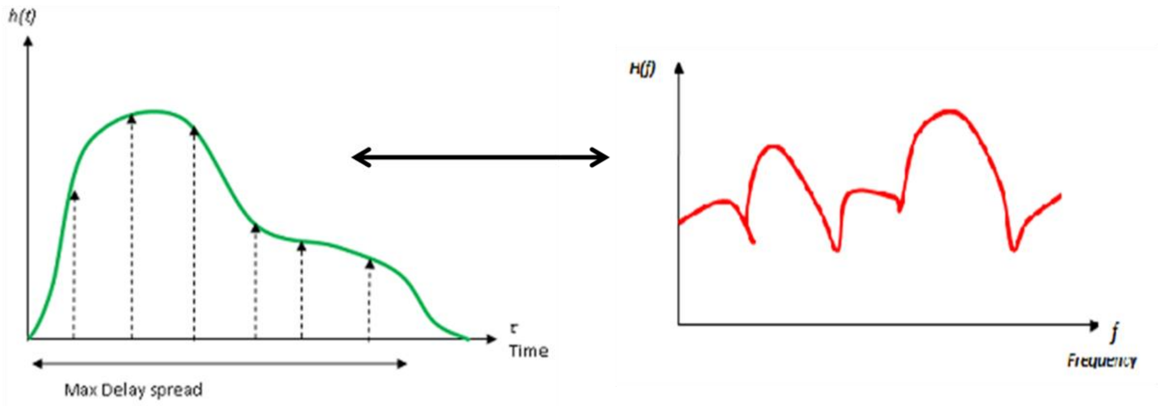


Fig.1.2 Channel Impulse response in time and frequency

When we observe a multipath-fading environment from the perspective of frequency-domain, a characteristic of multipath-fading is that some frequencies are enhanced whereas others are attenuated. This means that the fading can be flat or frequency selective. The channel characteristics or the impulse response function depends on the distance between the transmitter and receiver. Each multipath signal travels through different path. The delay time, at which each signal reaches the receiver, depends on the path length. The total time taken for the arrival of signal copies from different directions decides the delay spread. The impulse response of a multipath channel is measured in the field using channel sounding techniques.

1.1.3 Power Delay profile of Multipath channels:

Power delay profile can be measured in time domain and frequency domain. Important parameters are RMS Delay spread, Coherence bandwidth, Doppler spread and Coherence time [2].

RMS Delay Spread:

Each multipath signal travels different path length, so the time of arrival for each path is different. A single transmitted pulse will be spread in time when it reaches the receiver. This effect which spreads out the signal is called Delay Spread. Delay spread leads to increase in the signal bandwidth. RMS Delay Spread characterizes time-dispersiveness of the channel. It indicates the delay, during which the power of the received signal is above a certain value.

Coherence Bandwidth:

Coherence Bandwidth is a measure of the maximum frequency difference for which the signals are strongly correlated in amplitude. This implies that fading, if it happens, will be constant across this bandwidth. This is known as flat fading. If the transmission bandwidth is larger than the coherence bandwidth, the signals can be decorrelated in amplitude which overcomes the fading effect. Coherence bandwidth is inversely proportional to RMS delay spread. Two frequencies that are larger than coherence bandwidth fade independently. This concept is useful in frequency diversity. In frequency diversity, multiple copies of same message are sent using different carrier frequencies.

These frequencies are separated by more than the coherence bandwidth of the channel. Coherence Bandwidth indicates frequency selectivity during transmission. Coherence Bandwidth and delay spread describes only the time dispersive nature of the multipath channel in a local area.

Doppler Effect :

When a wave source (transmitter) and/or receiver is/are moving, the frequency of the received signal will not be the same as that of the transmitted signal. When they are moving toward each other, the frequency of the received signal is higher than the source. When they are moving away from each other, the frequency decreases. Thus the frequency of the received signal is $f_R = f_C - f_D$. Where f_C is the frequency of source carrier f_D is the Doppler shift in frequency.

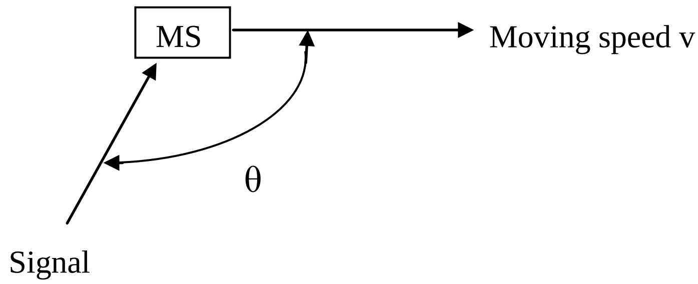


Fig.1.3 Doppler Effect

Doppler Shift :

$f_D = \left(\frac{v}{\lambda}\right)\cos\theta$, where v is the moving speed, λ is the wavelength of carrier. Doppler shift (f_D) depends on the relative velocity of the receiver with respect to the transmitter, the frequency (or wavelength) of transmission, and the direction of travelling with respect to the direction of the arriving signal. The Doppler shift is positive, if the mobile is moving toward the direction of arrival of the wave, and it is negative, if the mobile is moving away from the direction of arrival of the wave.

Doppler Spread:

Doppler Spread and Coherence Time describe the time-varying nature of the channel in a small-scale region caused by relative motion of the transmitter and the receiver. Doppler Spread B_D is the Maximum Doppler shift. Doppler spread decides frequency-dispersiveness or the spreading of transmitted frequency due to different doppler shifts. It is obtained from the doppler spectrum. It indicates the range of frequencies over which the received doppler spectrum is above a certain value. If the baseband signal bandwidth is much greater than B_D, then effects of doppler spread are negligible at the receiver. This is a slow fading channel.

Coherence Time:

Coherence time is a statistical measure of the time duration over which the channel impulse response is essentially time-invariant. If the symbol period of the baseband signal (reciprocal of the baseband signal bandwidth) is greater than the Coherence time of the channel, then the channel will change during the transmission of the signal, hence there will be distortion at the receiver. Coherence time is approximately equal to the reciprocal of Doppler spread. The definition of coherence time implies that two signals arriving with a time separation greater than the coherence time are affected differently by the channel. A large coherence time implies that the channel changes slowly.

1.1.4 Inter symbol Interference (ISI):

If the delay spread of the channel is comparable with the symbol length, we get ISI. It occurs when the second multipath is delayed and is received during next symbol. ISI has an impact on burst error rate of the channel. Pulse shaping and equalization techniques are adopted to overcome ISI under fading channels. The effect of ISI is shown in Fig.1.4 [3].

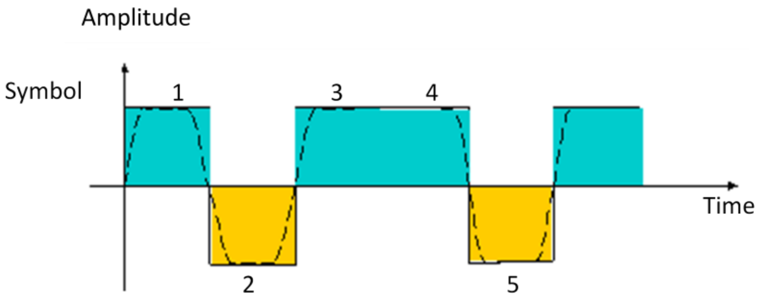


Fig 1.4.1 – Sequence 101101 to be sent, the dashed line is the shape that is actually sent.

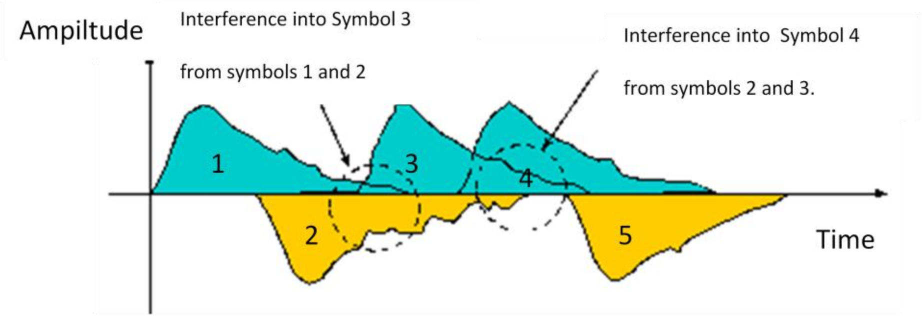


Fig 1.4.2 - Each symbol is spread by the medium

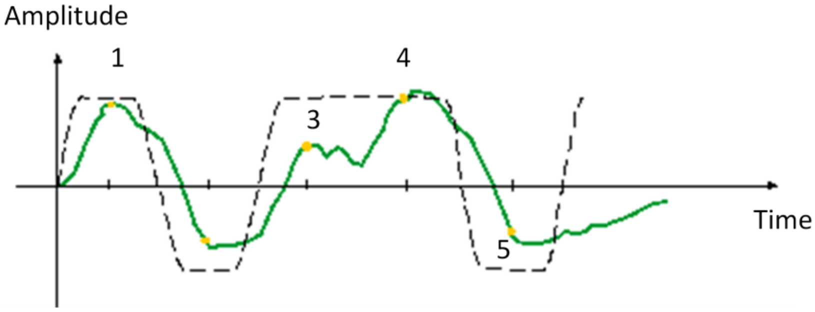


Fig 1.4.3 - Received signal vs. the transmitted signal

Fig.1.4 Effect of Inter symbol Interference

1.2 Importance of Error detection and correction

Error detection and correction are techniques that enable reliable delivery of digital data over unreliable communication channels. Many communication channels are subject to channel noise, and thus errors may be introduced during the transmission from the source to a receiver. Error detection techniques allow detecting such errors, while error correction enables reconstruction of the original data. The general idea for achieving error detection and correction is to add some redundancy (i.e., some extra data) to a message. The parity bits are added in a known fashion, which is to be known both to the encoder and to the decoder. The Shannon theorem states that given a noisy channel with channel capacity C and information transmitted at a rate R, where R < C, then if there exists a coding technique, that allow the probability of error at the receiver to be made arbitrarily small [4]. This means that, theoretically, it is possible to transmit information nearly without error at any rate below a limiting rate, C.

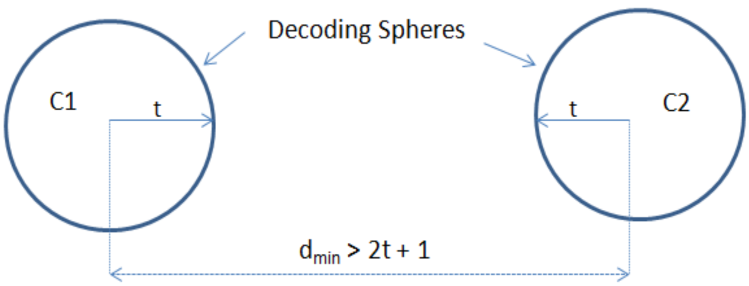


Fig 1.5 Error Detection & Correction using Decoding Spheres

Error-correcting codes are usually distinguished between convolutional codes and block codes: Convolutional codes are processed on a bit-by-bit basis and the Viterbi decoder allows optimal decoding. Block codes are processed on a block-by-block basis. Examples of block codes are Hamming codes, Reed–Solomon codes and Low-Density Parity-Check codes (LDPC). A code having a minimum distance ‘ d_{min} ‘ is capable of correcting all patterns of errors $t = (d_{min} - 1)/2$ or fewer errors in a code word, and is referred to as a random error correcting code.

In block codes [5], to detect t errors per block, the minimum distance of the block code should be $d_{min} \geq t+1$. *i.e.*, $t \leq d_{min} - 1$ errors can be detected. If the number of errors are equal to *dmin,* then one of the two codewords which form the pair for minimum distance might get changed into the other codeword. Hence the number of errors less than d_{min}, can be detected, because it will not make any codeword into another codeword. It takes d_{min} changes to transform from one codeword to another codeword. Similarly, to correct t errors per block, we must have $d_{min} \geq 2t+1$. Considering the above fig.1.5, the codewords C1 and C2 are spatially separated by the minimum distance d_{min}. i.e., if the number of errors occurs in C2 is equal to d_{min}, then C2 is transformed to C1. If t errors make the decoding sphere boundary around the codewords C1 and C2, and if the circumference of these two spheres touch each other, then the d_{min} is equal to $2t$. In this case, if any t errors occur, the erroneous codeword can be mapped to the codeword corresponding to the decoding sphere. If the erroneous codeword is exactly t distance from C1 and t distance from C2, the correct decision cannot be made. To avoid this ambiguity, the condition for correcting t errors per block should be $d_{min} \geq 2t + 1$.When these two decoding spheres intersects, the decoding will be incorrect. In order to increase the error correcting capability of a block code, the minimum distance should be larger.

Bit interleaving is a well-known technique for dispersing the errors that occur in burst when the received signal level fades, and which are likely to exceed the error correcting capability of a code [6]. Before a message is transmitted, the entire bit stream is interleaved. Hence the burst errors will be shared among the interleaved code words and only a simple code is required to correct them. Note that the interleaving process does not involve adding redundancy. Concatenated coding schemes are used to provide even more protection against bit errors than is possible with a single coding scheme.

1.3 Need for MIMO communication

In wireless communication, multiple antennas can be utilized in order to enhance the bit rate, and signal-to-noise-plus-interference ratio of wireless systems. Increased capacity is achieved by introducing multiple spatial channels by diversity at both the transmitter and at the receiver. Sensitivity to fading is reduced by the spatial diversity provided by space-time coding. In a MIMO system, given total transmit power can be divided among multiple spatial paths. For a given fixed bandwidth, there is always a fundamental tradeoff between bandwidth efficiency (high bit rates) and power efficiency (small error rates). Conventional single-antenna transmission techniques aiming at an optimal wireless system performance operate in the time domain and/or in the frequency domain. With MIMO, spatial domain is exploited by using space-time coding to overcome the detrimental effects of multipath fading [7].

For a Single-Input Single-Output antenna system, the capacity is given by $C = log (1+SNR)$. With MIMO, the capacity is $C \approx m. log (1+SNR)$, where m is the minimum number of antennas in the transmitter and receiver sides. Spatial multiplexing techniques simultaneously transmit independent information sequences, over multiple antennas. Spatial Multiplexing takes the high rate signal and breaks it down to lower rate streams. Using M transmit antennas, the overall bit rate compared to a single-antenna system is thus enhanced by a factor of M without requiring extra bandwidth or extra transmission power. Channel coding is often employed, in order to guarantee a certain error performance. Since the individual data streams are superimposed during transmission, they have to be separated at the receiver using an interference cancellation type of algorithm (typically in conjunction with multiple receive antennas). A well-known spatial multiplexing scheme is the Bell-Labs Layered Space-Time Architecture (BLAST). Similar to channel coding, multiple antennas can also be used to improve the error rate of a system, by transmitting and/or receiving redundant signals representing the same information sequence. By means of two-dimensional coding in time and space, commonly referred to as space-time coding, the information sequence is spread out over multiple transmit antennas. At the receiver, an appropriate combining of the redundant signals has to be performed. Optionally, multiple receive antennas can be used, in order to further improve the error performance (diversity reception). The advantage over conventional channel coding is that redundancy can be accommodated in the spatial domain, rather than in the time domain. Correspondingly, a diversity gain and a coding gain can be achieved without lowering the effective bit rate compared to single-antenna transmission. Well-known spatial diversity technique for systems with multiple transmit antennas is Alamouti’s transmit diversity scheme [8].

1.4 Purpose

Concatenated codes for MIMO system is a methodology to increase reliability by reducing bit error rate and improving effective data rate in the presence of fading environment. There has been a tremendous effort to develop the coded MIMO systems, which include code-design, the invention of soft-output detectors, joint detection and decoding techniques and so forth, with the ultimate goal to approach MIMO capacity. Most of the previous works have studied the coded MISO system with BER analysis. Therefore, lacking in the literature is a performance study of coded MIMO system with BER and effective data rate. In this project, our aim is to develop an efficient concatenated coding scheme to reduce the bit error rate and to achieve effective data rate under fading channel condition. The concatenated code will be constructed from coding schemes applicable to both wired and wireless communication. The performance of bit error rate and effective data rate by considering various coding schemes, fading channels and modulation schemes will be analysed.

1.5 Scope of work

- Fading can be approximated as a channel selective burst mode error.
- Different coding techniques can help immunising against burst mode error.
- MIMO allows multiple ways of implementing coding.
- A relative performance analysis of them using BER and effective data rate is essential. This is explored in Matlab simulation.

1.6 Motivation

Under rich scattering environment, Multiple Input Multiple Output (MIMO) systems have the potential to achieve greater reliability than by Single Input Single Output (SISO) systems under fading environment. This makes it one of the most exciting developments to have occurred in wireless communications. Within a short duration, it has matured from a research topic into a technology to find a place in upcoming wireless communication standards. This project focuses on the error control coding aspects of MIMO. The performance analysis of BER over SNR of 2×2 MIMO with concatenation of channel coding with space time coding schemes will be carried out. Since fading due to multipath propagation can be modeled as burst mode error, channel coding schemes developed for wired communication can also be employed for wireless fading MIMO channels.

1.7 Methodology

- To repeat some of the work done previously to gain experience.
- To implement and explore
    - Different coding schemes in Matlab.
    - Different fading models.
    - Different modulation schemes.
    - A MIMO link.
    - Different ways of implementing concatenated coding techniques and assess their relative performance.

1.8 Software Requirements

The complete MIMO link was implemented in MATLAB version R2012a using Communication toolbox.

Chapter 2

LITERATURE SURVEY

2.1 Literature Survey

We have reviewed some relevant work that report on the performance analysis of concatenation of channel coding scheme with SISO/MISO/MIMO-STBC in the presence of fading. The different types of coding scheme considered are Convolutional code with different code rates, Turbo code, Reed-Solomon code, Low-Density Parity check code (LDPC), and Trellis coded Modulation (TCM). The modulation schemes considered are binary phase-shift keying, Quadrature phase-shift keying, and quadrature-amplitude modulation (QAM). The analysis have been carried out in the presence of Rayleigh and Rician fading channels. The type of channel fading is flat.

In [6], performance analysis of orthogonal space time block codes concatenated with channel coding is done by considering quasi-static and block-fading Rayleigh as well as Rician fading. The channel codes considered are convolutional codes, turbo codes, trellis coded modulation (TCM), and multiple trellis coded modulation (MTCM). Union bounds evaluated for convolutional and turbo codes with BPSK modulation, and for 4-state, 8-PSK TCM and MTCM codes. The OSTBC for two-transmit antenna used here is the Alamouti scheme [8]. A random interleaver is placed between the channel code and OSTBC. Convolutional codes, with rate–1/2 is concatenated with Alamouti OSTBC with one receive antenna. With BPSK, the coding gain suffers about 1 dB (at BER= 10^{-5}) with a transmit correlation of $\rho_t = 0.7$. For a turbo coded Alamouti system, containing a rate-1/3 code with four-state constituent recursive convolutional codes the degradation due to spatial correlation of $\rho_t = 0.7$ is about 0.8 dB at BER = 10^{-7} . With TCM, 2-Tx and 1-Rx antennas and Rayleigh block fading channel, the performance loss due to spatial correlation of $\rho_t = 0.7$ is about 1.2 dB at BER = 10^{-5}. Under Rician fading, with parameter K = 5dB, and two transmit and one receive antennas the loss due to a spatial correlation of $\rho_t = 0.7$ is around 1dB.

In [9], a study on turbo-space-time coded modulation (turbo – STCM) scheme utilizing parallel concatenated systematic space-time codes (STC) with multilevel modulation and multiple transmit/receive antennas is carried out. The turbo-STCM encoder consists of two systematic recursive space–time component codes and it features full rate. Simulation results are provided for 4 state 4-PSK over the block-fading channel. It is shown that at frame error rate (FER) = 10^{-2}, and for a two transmit-one receive antenna configuration, the performance with recursive codes has an advantage of about 1.3 dB over the configuration with non recursive codes. For the same FER and antenna configuration, turbo-STCM provides an advantage of 2.7 dB over conventional 4-state STCs and 0.6 dB over conventional 32-state space–time codes.

In [10], the performance of convolutional codes (CC), Reed-Solomon (RS) block code as well as concatenated coding schemes that are used to encode the data stream in wireless communications are investigated. It is shown that by concatenating two different codes we can improve the total bit-error rate (BER). RS codes are preferred in correcting burst errors while convolutional codes are good for correcting random errors that are caused due to a fading channel. The simulation results confirm the outperformance of the concatenated codes especially RS-CC when compared to CC and RS codes. Due to a good burst error-correcting capability of RS codes, total BER of RS-CC has significant coding gain, and it increases as E_b/N_o increases.

In [11], a concatenated code that achieves full system diversity by appropriately selecting the outer convolutional code (CC) with an inner reduced-rank space-time block code (STBC) using Trellis diagram is proposed. The advantage of the lower rank STBC is that the number of RF chains can be reduced. The number of RF chains considered is 8. For simulation, considered BPSK transmission, and the channel is modeled as i.i.d quasi-static Rayleigh flat-fading channel. The coding gain is largest when the full rank STBC is employed, while it is smallest when no STBC is involved. From a practical point of view, the concatenation of rank four STBC and rate 1/2 CC offers the best trade off since it requires only half the number of RF chains with a small coding gain loss. It is clear that the coding gain increases as the rank of the STBC gets higher. The concatenation of rate

1/2 CC and Alamouti scheme offers the best trade off between coding gain and system complexity (i.e., number of RF chains).

In [12], upper bound on the error probability of low-density parity-check (LDPC) coded modulation schemes operating on Rayleigh and Rician MIMO fading channels are obtained. LDPC code is concatenated with the orthogonal space-time block code (OSTBC) as an inner code. Over single-input single output (SISO) fading channel, the SNR difference between the bit error probabilities and the corresponding upper bound is about 2 - 4 dB. The error probability decreases faster as the Rician factor ‘K’ increases from zero, five to twenty. This is because a Rician channel converges to an AWGN channel as K goes to infinity. The final SNR difference, as K goes up is thus expected to be about 1.5 dB. In the case of concatenated MIMO system with 4PSK and 8QAM modulation, the derived bounds are about 2.5 dB away from the simulation results. The difference decreases to 1.5 dB for the system with four transmit and four receive antennas, where the orthogonal space-time block code is adopted.

2.2 Literature survey – Summary

In summary, it is proved that the use of interleaving [6] between outer and inner code improves the error performance and to achieve coding gain, outer code must have maximum hamming distance. Here the focus was to consider only transmit diversity (MISO) in the presence of flat fading channels with single channel code rate and only BPSK is considered. No detailed study on MIMO is done. With turbo code under slow rayleigh fading SISO link [13], the performance depends on the code rate at low SNRs. No concatenation scheme is considered. In [11], the author talks about the performance of concatenation of Convolutional code (CC) with STBC under MISO system in the presence of Rayleigh flat fading. Different code rates of CC considered here are below 0.5, which provides more redundancy. It is shown that as the code rate of concatenation scheme increases, coding gain increases. In [10], the concatenation of CC with RS is been considered for SISO under AWGN channel. It is shown that concatenation improves the error performance, and the outer code should be a block code which can handle burst error and the inner code should be a convolutional code which can correct random errors.

In [12], the work evaluates the performance of concatenating LDPC with STBC and [9] describes parallel concatenation of turbo with stbc. From these two works, it is understood, that as the number of transmit and receive antenna increases, the error performance increases. There is no clear idea on the effect of code rate. In summary, concatenation of various channel codes like Convolutional, LDPC, Turbo with MIMO-STBC provides better error performance under different fading channels such as Rayleigh and Rician. The type of fading considered in all above cases are flat fading. The performance analysis work does not emphasis the effect of different code rates on bit error rate. Hence it is necessary to concentrate on the effect of code rates in the error performance.

In the following work, we undertake performance analysis in the presence of frequency-selective fading channel since there is no literature available on frequency selective fading. It is also shown, the single error correcting Hamming code with different code rates can be used to correct burst errors by concatenating with interleaver followed by STBC in MIMO fading channel. Convolutional codes with different higher code rates are also in consideration.

Chapter 3

ERROR CONTROL CODING TECHNIQUES

3.1 Linear Block Codes

3.1.1 Definition of Linear Block Codes

The output of an information source is a sequence of the binary digits 0 and 1. In block coding, this binary information sequence is segmented into message blocks of fixed length; each message block, denoted by *u*, consists of *k* information digits. There are a total of 2^k distinct messages. The encoder according to certain rules, transforms each input message *u* into a binary *n*-tuple *v* with $n > k$. This binary *n*-tuple *v* is referred to as the codeword of the message *u*. Therefore for 2^k possible messages, there are 2^k codewords. This set of 2^k codewords is called a block code. Therefore there should be a one-to-one correspondence between a message *u* and its codeword *v*. An important parameter of a block code is called the minimum distance d_{min}. This parameter determines the random error detecting and error correcting capabilities of a code. A block code with minimum distance d_{min} is capable of detecting all the error patterns of d_{min} -1 or fewer errors. A block code with minimum distance d_{min} guarantees correction of all the error patterns of t = [d_{min} -1/2] or fewer errors. The parameter t is called the random error correcting capability of the code. The code is referred to as a *t*-error correcting code. A desirable structure for a block code to posses is linearity, which greatly reduces the encoding complexity. The (*n*,*k*) block code B(*n*,*k*,*d*) with minimum Hamming distance *d* over the finite field Fq is called linear, if B(*n*,*k*,*d*) is a subspace of the vector space F_q^n of dimension k. Important linear block codes are Hamming codes, Reed-Muller code, Cyclic codes and Interleaved codes.

3.1.2 Generator Matrix

The linearity property of linear block code B(*n*,*k*,*d*) can be exploited for efficiently encoding a given information word $u = (u_{o,} u_{1,} u_{2,} \dots, u_{k-1})$. To this end, a basis $(g_{o,} g_{1,} g_{2,} \dots, g_{k-1})$ of the subspace spanned by the linear block code is chosen, consisting of *k* linearly independent n-dimensional vectors $g_i = (g_{i,o,} g_{i,1,} \cdots, g_{i,n-1})$

with $0 \leq I \leq k-1$. The corresponding code word $b = (b_{o,} b_{1,} \ldots, b_{n-1})$ is then given by $b = u_0 g_0 + u_1 g_1 + \cdots + u_{k-1} g_{k-1}$ with the q-nary information symbols $u_i \epsilon F_q$. For each linear block code B(n,k,d) an equivalent linear block code can be found that is defined by the k x n generator matrix

$$G = (I_k \mid A_{k,n-k}).$$

Owing to the k x k identity matrix I_k and the encoding rule $b = uG = (u \mid u\, A_{k,n-k})$, the first k code symbols b_i are identical to the k information symbols u_i. Such an encoding scheme is called *systematic*. The remaining $m = n - k$ symbols within the vector $u\, A_{k,n-k}$ correspond to m parity check symbols which are attached to the information vector u for the purpose of error detection and error correction.

3.1.3 Parity-Check Matrix

With the help of generator matrix $G = (I_k \mid A_{k,n-k})$, the following (n-k) x n matrix, the so-called *parity-check matrix* can be defined as $H = (\, B_{n-k,k} \mid I_{n-k}\,)$ (with the (n - k) x (n- k) identity matrix I_{n-k}. The (n - k) x k matrix $B_{n-k,k}$ is given by $B_{n-k,k} = -A^T_{k,n-k}$. For the matrices G and H the following property can be derived $HG^T = B_{n-k,k} + A^T_{k,n-k} = 0_{n-k,k}$ with the (n - k) x k zero matrix $0_{n-k,k}$. The generator matrix G and the parity-check matrix H are orthogonal. i.e., all row vectors of G are orthogonal to all row vectors of H.

3.1.4 Syndrome and Cosets

A vector r corresponds to a valid code word of a given linear block code B(n,k,d) with parity-check matrix H if and only if the parity-check equation $Hr^T = 0$ is true. Otherwise, r is not a valid code word of B(n,k,d). We can interpret r as the received vector which is obtained from $r = b + e$, with the transmitted code vector b and the error vector e. The jth component of the error vector e is $e_j = 0$ if no error has occurred at this particular position; otherwise the jth component is $e_j \neq 0$. In view of the parity-check equation, we define the so-called syndrome $s^T = Hr^T$ which is used to check whether the received vector r belongs to the channel code B(n,k,d). Inserting the received vector

r into this definition, we obtain, $s^T = Hr^T = H(b+e)^T = Hb^T + He^T = He^T$. Here, we have taken into account that for each code vector $b \epsilon B(n,k,d)$ the condition $Hb^T = 0$ holds. Finally, we recognize that the syndrome does exclusively depend on the error vector e, i.e., $s^T = He^T$. Thus, for the purpose of error detection the syndrome can be evaluated. If s is zero, the received vector *r* is equal to a valid code vector, i.e., $r \epsilon B(n,k,d)$. In this case no error can be detected and it is assumed that the received vector corresponds to the transmitted code vector. If *e* is zero, the received vector *r* = *b* delivers the transmitted code vector *b*. However, all non-zero error vectors *e* that fulfill the condition $He^T = 0$ also lead to a valid code word. These errors cannot be detected. In general, the (*n*-*k*) dimensional syndrome $s^T = He^T$ of a linear (*n*,*k*) block code B(*n*,*k*,*d*) corresponds to *n*-*k* scalar equations for the determination of the *n*-dimensional error vector *e*. The matrix equation $s^T = He^T$ does not uniquely define the error vector *e*. All vectors e with $He^T = s^T$ form a set, the so-called coset of the *k*-dimensional subspace B(*n*,*k*,*d*) in the finite vector space F_q^n. This coset has q_k elements. For an error-correcting *q*-nary block code B(*n*,*k*,*d*) and a given syndrome *s*, the decoder has to choose one out of these q_k possible error vectors. An optimal decoder implementing the minimum distance decoding rule chooses the error vector out of the coset that has the smallest number of non-zero components.

3.2 Hamming Codes

Hamming codes can be defined as binary block codes, defined by parity-check matrix. If we want to add *m* parity-check symbols to an information word *u* of length *k*, the parity-check matrix is obtained by writing down all *m*-dimensional non-zero column vectors. Since there are *n = 2m-1* binary column vectors of length *m*, the parity-check matrix H is of dimension *m* x (*2m-1*), H = (*B n-k, k* | *I n-k*). The corresponding k x n generator matrix is given by $G = (I_k | - B^T_{(n-k,k)})$. The codeword length of Hamming code is given by $n = 2^m-1$, $m = n - k$, with the number of information symbols $k = n - m$. Since the columns of parity-check matrix are pairwise linearly independent and there exist three columns which sum up to the all-zero vector, the minimum hamming distance of the binary Hamming code is $d_{min} = 3$. The binary Hamming code can be characterized as a linear block code *B(n,k,3)*, that is able to detect 2 errors or to correct 1 error. The decoding of an erroneously received vector $r = b + e$, can be carried out by first calculating the syndrome. The syndrome equals the position number in which the error occurred. The error vector has n elements with one non-zero element in the position of the syndrome number. This vector is then added to the received vector to get the correctly decoded vector. The code rate is given by k / n. Hamming codes are used in the short-range wireless communication system Bluetooth [14].

3.3 Convolutional code

Convolutional codes differ from block codes in that the encoder contains memory, and the encoder outputs at any given time unit depend not only on the inputs at that time but also on some number of previous inputs. A rate R = k/n convolutional encoder with memory order m can be realized as a k-input, n-output linear sequential circuit with input memory m; that is, inputs remain in the encoder for an additional m time units after entering. Typically, n and k are small integers, and the codeword is divided in to blocks of length n. In the important special case when $k = 1$, the information sequence is not divided into blocks and is processed continuously. Unlike with block codes, large minimum distances and low error probabilities are achieved not by increasing k and n but by increasing the memory order m. An important decoding algorithm for convolutional codes are Viterbi algorithm. Convolutional codes process information serially or continuously in short block lengths. A rate k/n convolutional encoder has m shift registers, k input bits, n output bits that are given by linear combinations of the content of the registers and the input information bit. To achieve higher code rates, puncturing techniques can be applied. State diagram representation helps in determining whether a convolutional encoder is catastrophic. A convolutional encoder is said to be catastrophic, if a finite number of error produces an infinite number of errors after decoding. An efficient solution to the decoding problem is the Viterbi algorithm based decoder. This is a maximum likelihood decoder, which tracks the states of the trellis, to find the closest coded sequence v’ to the received sequence r’.

3.3.1 Convolutional Encoder

Consider the convolutional encoder depicted in Figure 3.1. Information bits are shifted into a register of length *m*=2, i.e. a register with two binary memory elements. The output sequence results from multiplexing the two sequences denoted by $\mathbf{b}^{(1)}$ and $\mathbf{b}^{(2)}$. Each output bit is generated by modulo 2 addition of the current input bit and some symbols of the register contents. For instance, the information sequence $\mathbf{u}$ = (1,1,0,1,0,0,…….) will be encoded to $\mathbf{b}^{(1)}$ = (1,0,0,0,1,1,…..) and $\mathbf{b}^{(2)}$ = (1,1,1,0,0,1,….). The generated code sequence after multiplexing of $\mathbf{b}^{(1)}$ and $\mathbf{b}^{(2)}$ is $\mathbf{b}$ = (1,1,0,1,0,1,0,0,1,0,1,1,….).

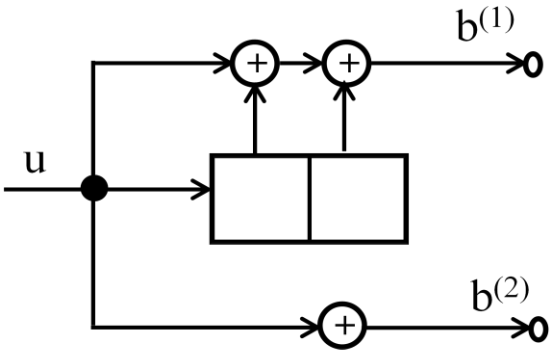


Fig.3.1 A rate R = ½ convolutional encoder with memory *m* = 2

A convolutional encoder is a linear sequential circuit and therefore a Linear Time-Invariant (LTI) system. It is well known that an LTI system is completely characterized by its impulse response. Let us therefore investigate the two impulse responses of this particular encoder. The information sequence $\mathbf{u}$ = 1,0,0,0,…. results in the output of $\mathbf{b}^{(1)}$ = (1,1,1,0…) and $\mathbf{b}^{(2)}$ = (1,0,1,0,….) i.e. we obtain the generator impulse responses $\mathbf{g}^{(1)}$ = (1,1,1,0,…) and $\mathbf{g}^{(2)}$ = (1,0,1,0,…) respectively. These generator impulse responses are helpful for calculating the output sequences for an arbitrary input sequence

$$b_i^{(1)} = \sum_{l=0}^{m} u_{i-1} g_l^{(1)} \leftrightarrow \boldsymbol{b}^{(1)} = \boldsymbol{u} * \boldsymbol{g}^{(1)}$$

$$b_i^{(2)} = \sum_{l=0}^{m} u_{i-1} g_l^{(2)} \leftrightarrow \boldsymbol{b}^{(2)} = \boldsymbol{u} * \boldsymbol{g}^{(2)}$$

The generating equations for $\mathbf{b}^{(1)}$ and $\mathbf{b}^{(2)}$ can be regarded as convolutions of the input sequence with the generator impulse responses $\mathbf{g}^{(1)}$ and $\mathbf{g}^{(2)}$. The code B generated by this encoder is the set of all output sequences $\mathbf{b}$ that can be produced by convolution of arbitrary input sequence $\mathbf{u}$ with the generator impulse responses. This explains the name *convolutional codes*.

3.3.2 State Diagram of a Convolutional Encoder

The state diagram of a convolutional encoder describes the operation of the encoder. From this graphical representation, we observe that the encoder is a finite-state machine. For the construction of the state diagram we consider the contents of the encoder registers as *encoder state σ*. The set S of encoder states is called the *encoder state space*. Each memory element contains only 1 bit of information. Therefore, the number of encoder states is 2^v. We will use the symbol $\boldsymbol{\sigma_i}$ to denote the encoder state at time *i*. the state diagram is a graph that has 2^v nodes representing the encoder states. An example of the state diagram is given in Fig.3.2.

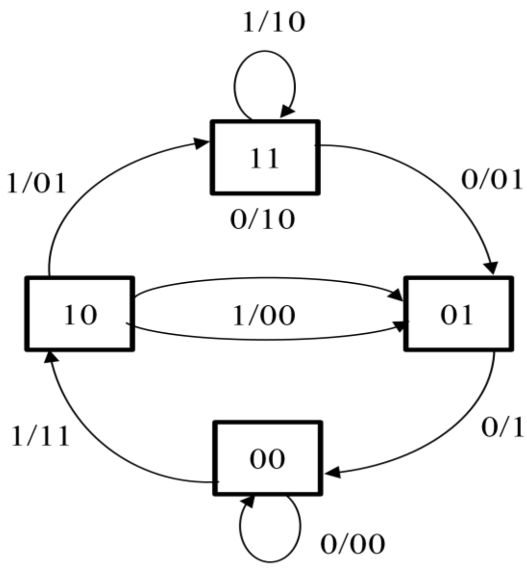


Fig. 3.2 State diagram of the encoder in Fig.3.1

The branches in the state diagram represent possible state transitions, e.g. if the encoder in Fig.3.1 has the register contents $\boldsymbol{\sigma_i}$ = (00) and the input u_i at time *I* is a 1, then the state of the encoder changes from $\boldsymbol{\sigma_i}$ =(00) to $\boldsymbol{\sigma_{i+1}}$ = (10). Along with this transition, the two output bits $\mathbf{b}_i$ = (11) are (00) generated. Similarly, the information sequence **u** = (1,1,0,1,0,0,…….) corresponds to the state sequence $\boldsymbol{\sigma}$ = **(**00,10,,11,01,10,01,00,…….**)** subject to the encoder starting in the all-zero state. The code sequence is again **b** = (11,01,01,00,10,11,00,…). In general, the output bits only depend on the current input and the encoder state. Therefore, we label each transition with the k input bits and the n output bits (input/output).

3.3.3 Trellises

A trellis is a directed graph, where the nodes represent encoder states. In contrast to the state diagram, the trellis has a time axis, i.e. the *i*th level of the trellis corresponds to all possible encoder states at time *i*. The trellis of a convolutional code has a very regular structure. It can be constructed from the trellis module, which is simply a state diagram, where the states $\boldsymbol{\sigma_i}$ at each time *i* and $\boldsymbol{\sigma_{i+1}}$ at time *i* + 1 are depicted separately. The transitions leaving a state are labeled with the corresponding input and output bits (input/output).As we usually assume that the encoder starts in the all-zero state, the trellis also starts in the zero state. Hence, the first m levels differ from the trellis module. There are only initial state transitions that depart from the all-zero state.

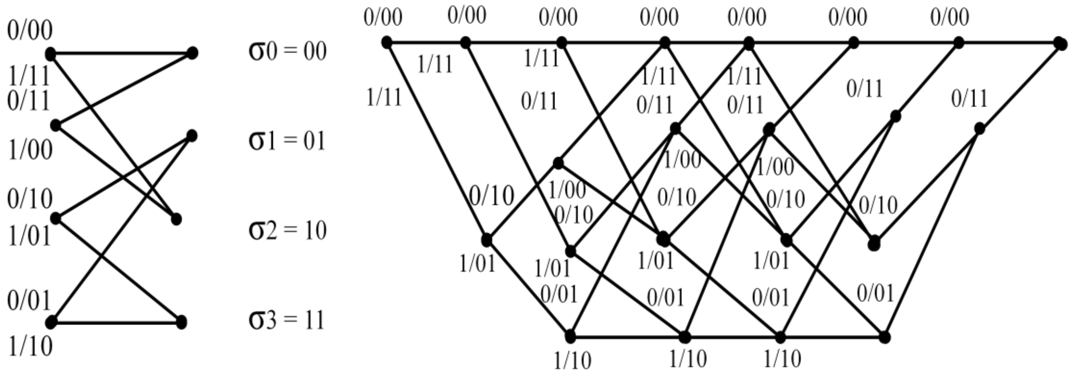


Fig.3.3 Trellis for the encoder in Fig.3.1

In Figure 3.3 we consider a terminated convolutional code with five information bits and two bits for termination. Therefore, the final encoder state is always the all-zero state. All code words begin and end with the encoder state $\boldsymbol{\sigma}$ = 00. Moreover, each non-zero code word corresponds to a sequence of state transitions that depart from all-zero state some number of times and return to the all-zero state. The trellis is a representation of the terminated convolutional codes, i.e. there is a one-to-one correspondence between the set of all paths through the trellis and the set of code sequences, i.e. the code B. Each path from the initial all-zero state to the final all-zero state corresponds to a code word, and vice versa. Moreover, there is a one-to-one correspondence between the set of all paths through the trellis and the set of information sequences. Note that this one-to-one correspondence exists only if the encoder non-catastrophic.

3.3.4 Viterbi Algorithm

Step 1: Assign metric zero the initial node □ ($\boldsymbol{\sigma}_{0,0}$) = 0 and set time $i = 1$.

Step 2: Each node $\boldsymbol{\sigma}_{j,i}$ of the time instant i is processed as follows(forward pass):

a. **ADD:** Calculate the metrics of all branches $\boldsymbol{\sigma}_{j`,i-1} \rightarrow \boldsymbol{\sigma}_{j,i}$ that enter the node $\boldsymbol{\sigma}_{j,I}$ by □ ($\boldsymbol{\sigma}_{j`,i-1}$)+dist($\mathbf{r}_i$,$\mathbf{b'}_i$),where □ ($\boldsymbol{\sigma}_{j`,i-1}$) is the node metric of this branch's predecessor node $\boldsymbol{\sigma}_{j`,i-1}$ and $\mathbf{b'}_i$ is the code block corresponding to the branch $\boldsymbol{\sigma}_{j`,i-1} \rightarrow \boldsymbol{\sigma}_{j,i}$.

b. **COMPARE:** Assign the smallest branch metric among all branches merging at node $\boldsymbol{\sigma}_{j,i}$ as the node metric □ ($\boldsymbol{\sigma}_{j,i}$).

c. **SELECT:** Store the branch corresponding to the smallest branch metric as the survivor.

Step 3: If $i \leq L+m$, then continue with the next level, i.e. increment i by 1 and go to step 2. Otherwise go to step 4.

Step 4: To find the best path, we have to start from the terminating node $\boldsymbol{\sigma}_{0,L+m}$ and go to the initial node $\boldsymbol{\sigma}_{0,0}$, following the survivors(backward pass).

Example of the Viterbi algorithm – backward pass

Fig.3.4 shows the Viterbi algorithm with backward pass mode. Starting from the terminating node and going back to the initial node using the labeled survivors, we obtain the estimated code word b' = (11 01 01 00 01 01 11).

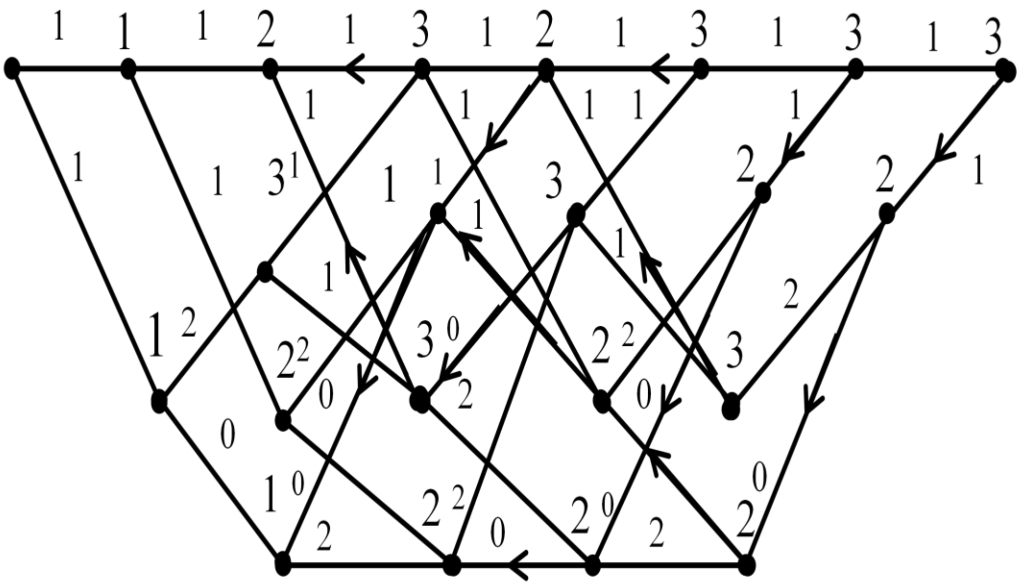


Fig.3.4 Example of the Viterbi algorithm – backward pass

Chapter 4

MODULATION SCHEMES

4.1 Introduction

In digital communication, to transmit data over a band-pass channel, it is necessary to modulate the incoming data onto a carrier wave (usually sinusoidal) with fixed frequency limits imposed by the channel. The modulation process involves switching or keying the amplitude, frequency, or phase of the carrier in accordance with the incoming data. There are three basic modulation techniques for the transmission of digital data; they are known as Amplitude-Shift Keying (ASK), Frequency-Shift Keying (FSK), and Phase-Shift Keying (PSK). Modulation is defined as the process by which some characteristics of a carrier is varied in accordance with a modulating wave. In M-ary signaling, the modulator produces one of an available set of M = 2^m distinct signals in response to *m* bits of source data at a time. The demodulation at the receiver can be either *coherent* or *noncoherent*. In coherent detection, the receiver has the knowledge of the carrier wave's phase reference. Coherent detection is performed by cross-correlating the received signal with each one of the replicas, and then making a decision based on comparisons with preselected thresholds. In noncoherent detection, knowledge of the carrier wave's phase is not required. The complexity of the receiver is reduced but at the expense of an inferior error performance, compared to a coherent system. In our work, we have considered coherent binary phase shift keying (BPSK) and coherent quadrature phase shift keying (QPSK).

4.2 Binary Phase shift keying (BPSK)

In a coherent BPSK system, the pair of signals, $s_1(t)$ and $s_2(t)$, used to represent binary symbols 1 and 0, respectively are defined by

$$\boldsymbol{s_1(t) = \sqrt{\frac{2E_b}{T_b}} cos(2\pi f_c t)} \quad (4.1)$$

$$\boldsymbol{s_1(t) = \sqrt{\frac{2E_b}{T_b}} cos(2\pi f_c t + \pi) = -\sqrt{\frac{2E_b}{T_b}} cos(2\pi f_c t)} \quad (4.2)$$

Where $0 \leq t < T_b$, and E_b is the transmitted signal energy per bit. In order to ensure that each transmitted bit contains an integral number of cycles of the carrier wave, the carrier frequency f_c is chosen equal to n_c / T_b for some fixed integer n_c. A pair of sinusoidal waves that differ only in a relative phase-shift of 180 degrees, hence these sine waves are referred to as antipodal signals.

From Eqs.4.1 and 4.2, it is clear that there is only one basis function of unit energy, namely

$$\boldsymbol{\emptyset_1(t) = \sqrt{\frac{2}{T_b}} cos(2\pi f_c t)} \qquad \boldsymbol{0 \leq t < T_b} \qquad (4.3)$$

The transmitted signals $s_1(t)$ and $s_2(t)$ in terms of $\emptyset_1(t)$ as follows

$$\boldsymbol{s_1(t) = \sqrt{E_b}\emptyset_1(t)} \qquad \boldsymbol{0 \leq t < T_b} \qquad (4.4)$$

$$\boldsymbol{s_2(t) = -\sqrt{E_b}\emptyset_1(t)} \qquad \boldsymbol{0 \leq t < T_b} \qquad (4.5)$$

A coherent BPSK system is therefore characterized by having a signal space that is one-dimensional (i.e., N = 1) and with two message points (i.e., M = 2), as shown in Fig.4.1. The coordinates of the message points equal

$$s_{11} = \int_0^{T_b} s_1(t)\, \emptyset_1(t) dt = +\sqrt{E_b} \qquad (4.6)$$

$$s_{21} = \int_0^{T_b} s_2(t)\, \emptyset_1(t) dt = -\sqrt{E_b} \qquad (4.7)$$

The message point corresponding to $s_1(t)$ is located at $s_{11} = +\sqrt{E_b}$, and the message point corresponding to $s_2(t)$ is located at $s_{21} = -\sqrt{E_b}$.

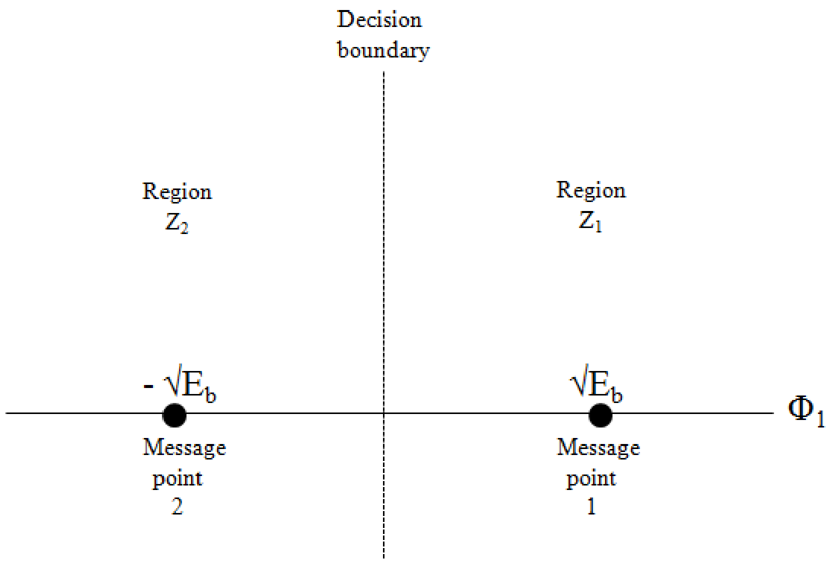


Fig. 4.1 Signal space diagram for coherent BPSK system

To realize a rule for making a decision in favour of symbol 1 or symbol 0, the signal space of Fig.4.1 must be partitioned into two regions:

1. The set of points closest to the message point at $+\sqrt{E_b}$.
2. The set of points closest to the message point at $-\sqrt{E_b}$.

This is accomplished by constructing the midpoint of the line joining these two message points, and then marking off the appropriate decision regions. In Fig.4.1, these decision regions are marked Z_1 and Z_2, according to the message point around which they are constructed. The decision rule is now simply to guess signal $s_1(t)$ or binary symbol 1 was transmitted if the received signal point falls in region Z_1 and guess signal $s_2(t)$ or binary symbol 0 was transmitted if the received signal point falls in region Z_2.

The average probability of symbol error coherent BPSK is given by

$$\boldsymbol{P_e = \frac{1}{2}\ erfc\left(\sqrt{\frac{E_b}{N_o}}\right)} \tag{4.8}$$

To generate a BPSK wave, we have to represent the input binary sequence in polar form with symbols 1 and 0 represented by constant amplitude levels of $+\sqrt{E_b}$ and $-\sqrt{E_b}$, respectively. This binary wave and a sinusoidal carrier wave $\emptyset_1(t)$ are applied to product modulator, as in Fig.4.2a. The carrier and the timing pulses used to generate the binary wave are usually extracted from a common master clock. The desired PSK wave is obtained at the modulator output. To detect the original binary sequence of 1s and 0s, we apply the noisy PSK wave *x(t)* (at the channel output) to a correlator, which is also applied supplied with a locally generated coherent reference signal $\emptyset_1(t)$, as in Fig.4.2b.

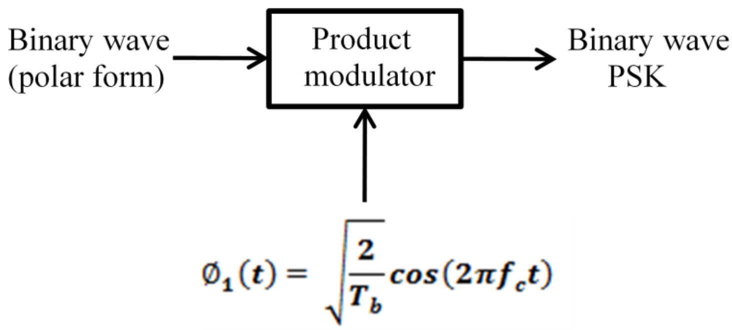


Fig.4.2a Block diagram for BPSK Transmitter

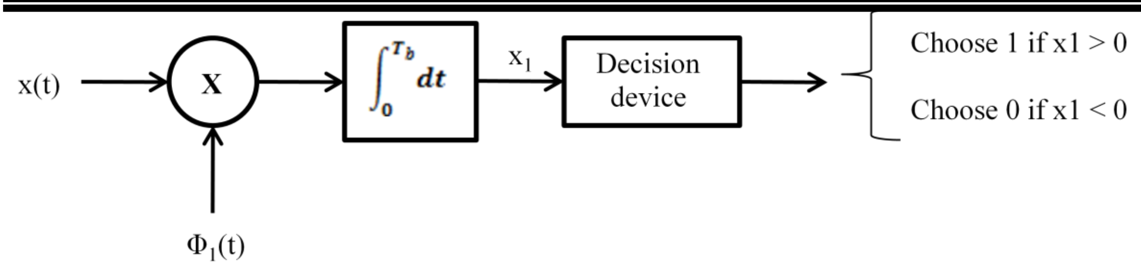


Fig.4.2b Block diagram for coherent BPSK Receiver

The correlator output x_1 is compared with a threshold of zero volts. If $x_1 > 0$, the receiver decides in favor of symbol. On the other hand, if $x_1 < 0$, it decides in favor of symbol 0.

4.3 Quadriphase - shift keying (QPSK)

As with BPSK, this modulation scheme is characterized by the fact that the information carried by the transmitted wave is contained in the phase. In QPSK, the phase of the carrier takes on one of four equally spaced values, such as π/4, 3π/4, 5π/4, and 7π/4, as shown by

$$S_i(t) = \begin{cases} \sqrt{\frac{2E}{T}} \cos\left[2\pi f_c t + (2i-1)\frac{\pi}{4}\right] & 0 \le t \le T \\ 0 & \text{elsewhere} \end{cases} \qquad (4.9)$$

where *i* = 1, 2, 3, 4 and ***E*** is the transmitted signal energy per symbol, ***T*** is the symbol duration, and the carrier frequency f_c equals n_c / T for some fixed integer n_c. Each possible value of the phase corresponds to unique pair of bits called a *dibit*.

Using a well-known trigonometric identity, Eq.4.9 can be written as

$$S_i(t) = \begin{cases} \sqrt{\frac{2E}{T}} \cos\left[(2i-1)\frac{\pi}{4}\right] \cos(2\pi f_c t) \\ \quad - \sqrt{\frac{2E}{T}} \sin\left[(2i-1)\frac{\pi}{4}\right] \sin(2\pi f_c t) & 0 \le t \le T \\ 0 & \text{elsewhere} \end{cases} \qquad (4.10)$$

where *i* = 1, 2, 3, 4. Based on this representation, we can make the following observations:

1. There are only two orthonormal basis functions, $\emptyset_1(t)$ and $\emptyset_2(t)$, contained in the expansion of $S_i(t)$. The appropriate forms for $\emptyset_1(t)$ and $\emptyset_2(t)$ are defined by

$$\boldsymbol{\emptyset_1(t) = \sqrt{\frac{2}{T}}\, cos(2\pi f_c t) \qquad 0 \le t \le T} \tag{4.11}$$

and

$$\boldsymbol{\emptyset_2(t) = \sqrt{\frac{2}{T}}\, sin(2\pi f_c t) \qquad 0 \le t \le T} \tag{4.12}$$

2. There are four message points, and the associated signal vectors are defined by

$$S_i = \begin{bmatrix} \sqrt{E}\cos(2i-1)\frac{\pi}{4} \\ -\sqrt{E}\sin(2i-1)\frac{\pi}{4} \end{bmatrix} \qquad i = 1, 2, 3, 4 \tag{4.13}$$

QPSK signal is characterized by having a two-dimensional signal constellation (i.e., N = 2) and four messages points (i.e., M = 4) as illustrated in Fig.4.3.

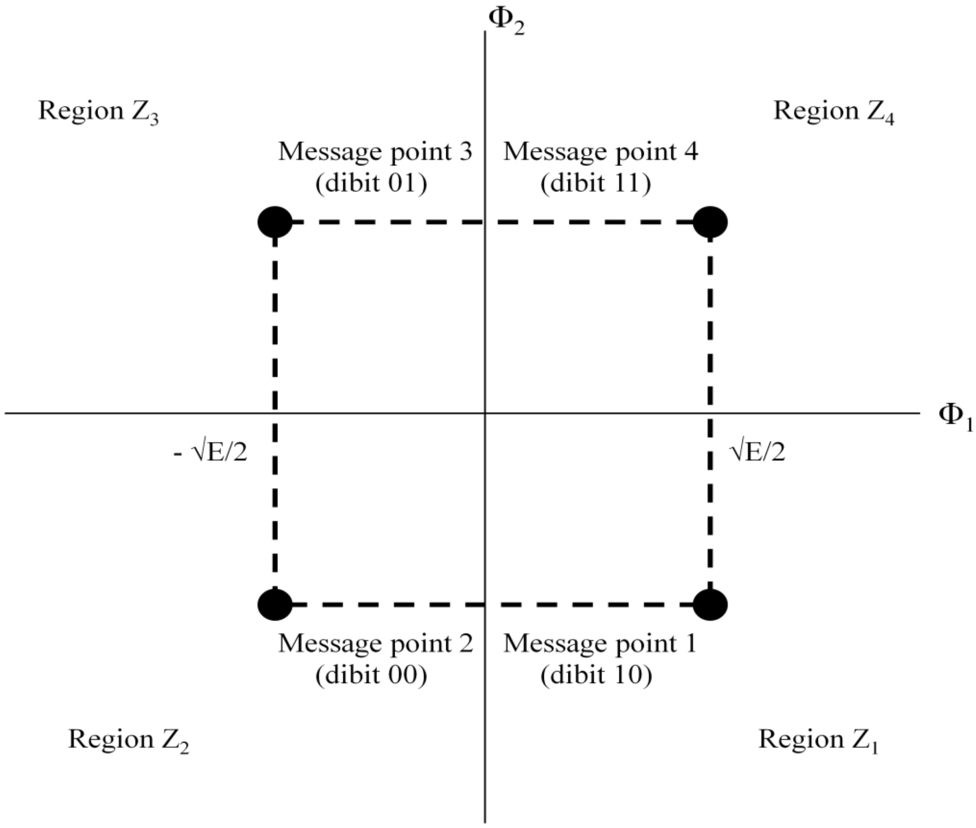


Fig.4.3 Signal space diagram for coherent QPSK system

The average probability of symbol error in terms of the ratio E_b / N_o is given by

$$P_e \approx erfc\left(\sqrt{\frac{E_b}{N_o}}\right) \tag{4.14}$$

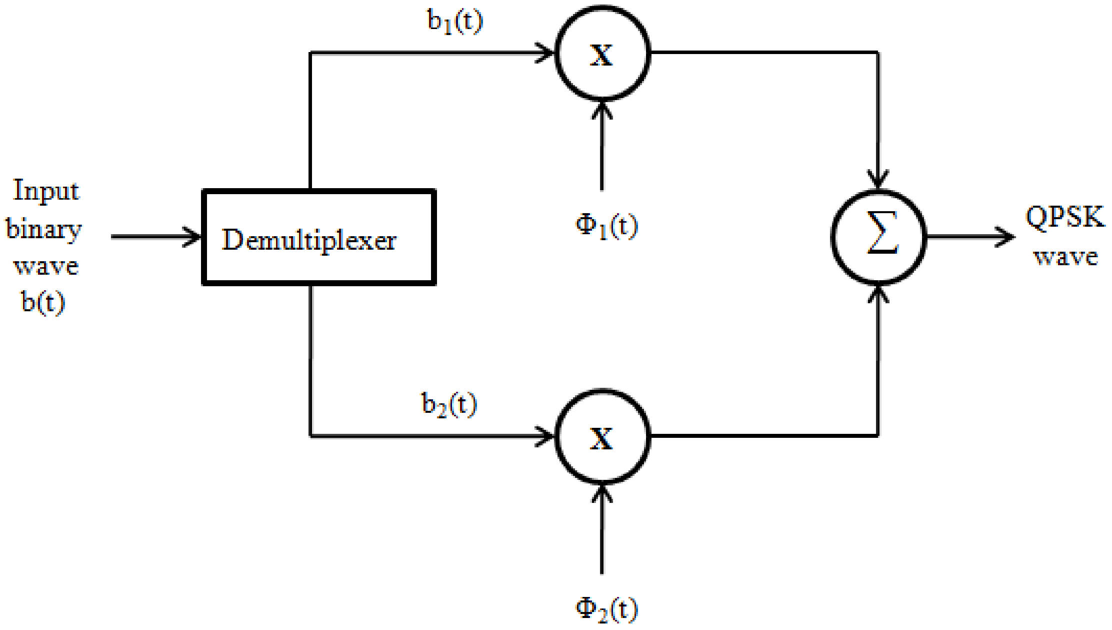


Fig .4.4a Block diagram for QPSK Transmitter

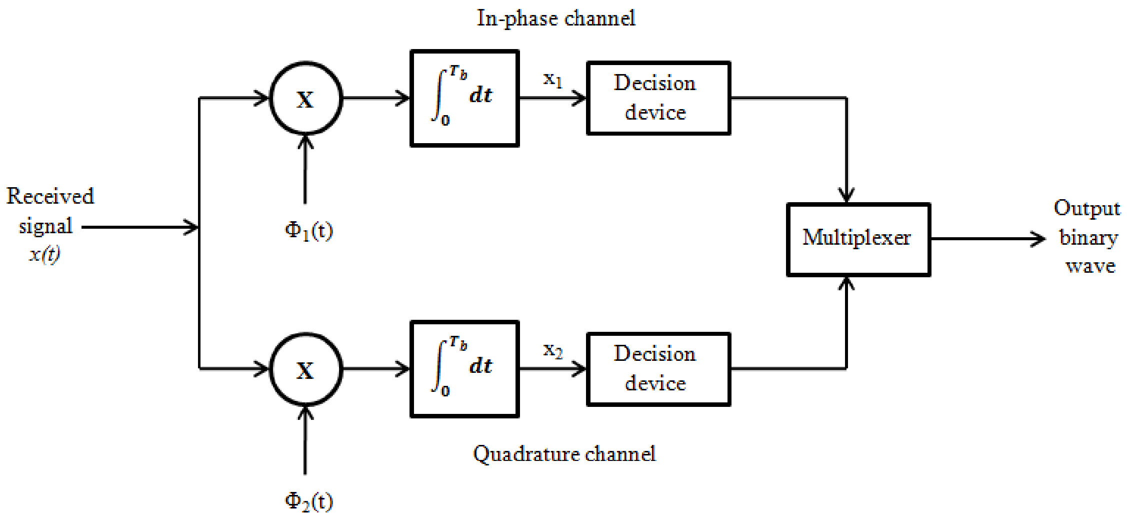


Fig .4.4b Block diagram for QPSK Receiver

Fig.4.4a shows the block diagram of a typical QPSK transmitter. The input binary sequence b(t) is represented in polar form, with symbols 1 and 0 represented by $+\sqrt{E_b}$ and $-\sqrt{E_b}$ volts, respectively. This binary wave is divided by means of a Demultiplexer into two separate binary waves consisting of odd and even numbered input bits. These two binary waves are denoted by $b_1(t)$ and $b_2(t)$. In any signaling interval, the amplitudes of $b_1(t)$ and $b_2(t)$ equals s_{i1} and s_{i2}, respectively, depending on the particular dibit that is being transmitted. The two binary waves $b_1(t)$ and $b_2(t)$ are used to modulate a pair of quadrature carriers or orthonormal basis functions : $\Phi_1(t)$ equal to $\sqrt{2/T}\cos(2\pi f_c\, t)$ and $\Phi_2(t)$ equal to $\sqrt{2/T}\sin(2\pi f_c\, t)$. The result is a pair of BPSK waves, which may be detected independently due to the orthogonality of $\Phi_1(t)$ and $\Phi_2(t)$. Finally, the two BPSK waves are added to produce the desired QPSK wave. The symbol duration T of a QPSK wave is twice as long as the bit duration, T_b, of the input binary wave. That is, for a given bit rate $1/T_b$, a QPSK wave requires half the transmission bandwidth of the corresponding BPSK. Equivalently, for a given transmission bandwidth, a QPSK wave carries twice as many bits of information as the corresponding BPSK wave.

The QPSK receiver consists of a pair of correlators with a common input and supplied with a locally generated pair of coherent reference signals $\Phi_1(t)$ and $\Phi_2(t)$, as in Fig.4.4b. The correlator outputs, x_1 and x_2 are each compared with a threshold of zero volts. If $x_1 > 0$, a decision is made in favor of symbol 1 for the upper or in-phase channel output, but if $x_1 < 0$, a decision is made in favor of symbol 0. Similarly, if $x_2 > 0$, a decision is made in favor of symbol 1 for the lower or quadrature channel output, but if $x_2 < 0$, a decision is made in favor of symbol 0. Finally, these two binary sequences at the in-phase and quadrature channel outputs are combined in a multiplexer to reproduce the original binary sequence at the transmitter input with the minimum probability of symbol error [15].

Chapter 5

FADING IN WIRELESS CHANNEL

5.1 Introduction to wireless fading channel

In wired communication, each transmitter-receiver pair is connected by a point-to-point shielded link which provides substantial immunity against interference. But in wireless communication, there is no physical link between the transmitter and receiver and the transmitter emits broadly in the direction of the receiver over the air. In propagating over the air, the signal may suffer absorption or spurious additions coming from other over the air transmissions leading to quality degradation when received. Multipath propagation can cause fading in wireless communication. If the signal strength goes below a threshold signal-to-noise ratio level, the signal is lost. This is known as deep fading [16]. Fading can be both time and frequency dependent. If we consider that the channel fading corrupts substantial amount of data, this can be considered as equivalent to the burst mode error occurs in wired channel. Hence the error control coding techniques developed for wired channel can be efficiently implemented for wireless channel to combat fading. Concatenated codes are error-correcting codes, that are constructed from concatenating two different codes to achieve good error control capability at all data rates less than capacity, with less decoding complexity. The concatenation can be either serial or parallel. The inner code can be used to treat the distributed error and the outer code is to treat the burst mode error. In wireless communication, with single transmitter and a receiver, the communication will not be reliable under fading. A recent advancement in wireless communications is to use multiple antennas both at the transmitter and at the receiver. This provides diversity. The principle of diversity is to provide the receiver with multiple copies of the same signal. Hence the probability that all the signal paths will be affected at the same time is less. This technology is known as Multi-input Multi-output (MIMO). With MIMO, the data can be coded across space and time. The error control coding technique developed for MIMO is known as space-time block codes. MIMO with STBC offers multiple ways of implementing error correction codes. Our aim here is to develop an efficient concatenated coding scheme to reduce the bit error rate and to achieve effective data rate under fading. Performance of various coding schemes in different fading environments and using different modulation schemes will be assessed using bit error rate and effective data rate.

A receiver cannot distinguish between the different multipath components, it just adds them up, so that they interfere with each other. The interference between them can be constructive or destructive, depending on the phases of the multipath components. The amplitude of the total signal changes with time if either transmitter or receiver is moving. If this signal amplitude is above a threshold level, this can be detected by receiver. If this signal amplitude goes below the threshold, this cannot be detected at the receiver. Then it is said that, the signal is faded. Fading due to multipath is known as small scale fading or fading. The following figure shows the fading as a function of signal strength against time.

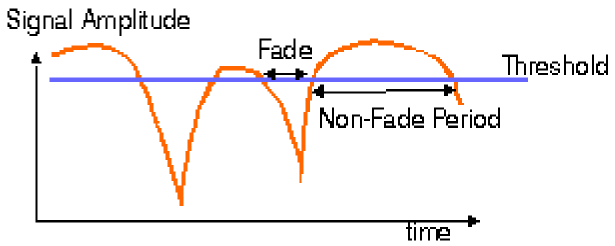


Fig.5.1 Fading as a function of signal amplitude and time

5.2 Fading Terminology

- Level crossing Rate :

  Average number of times per second that the signal envelope crosses the level in positive going direction.
- Fading Rate :

  Number of times signal envelope crosses middle value in positive going direction per unit time.
- Depth of Fading :

  Ration of mean square value and minimum value of fading signal.
- Fading Duration

  Time for which signal is below given threshold.

5.3 Fading Classifications

Flat Fading :

Occurs due to fluctuations in the gain of the multipath channel which leads to change in amplitude of the received signal with time. It occurs when symbol period of the transmitted signal is much larger than the Delay spread of the channel. Flat fading may cause deep fades.

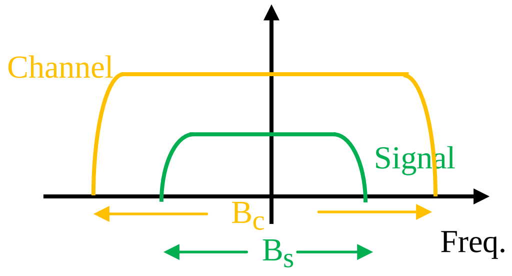


Fig. 5.2 Flat Fading as a function of Delay spread

Frequency selective fading :

Occurs when the channel multipath delay spread is greater than the symbol period. Symbols face time dispersion. Channel will induce ISI. Bandwidth of the signal s(t) is wider than the channel impulse response. Frequency selective fading also cause distortion of the received baseband signal.

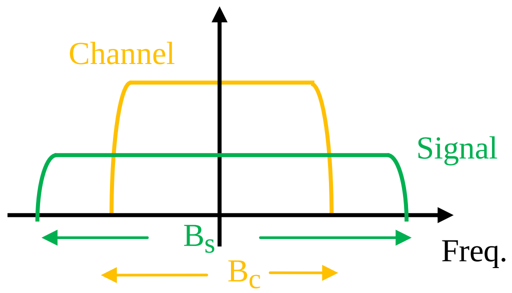


Fig. 5.3 Frequency Selective Fading as a function of Delay spread

Fast fading :

Occurs due to Doppler spread. The rate of change of the channel characteristics is larger than the rate of change of the transmitted signal. As s result, the channel changes during a symbol period. The channel changes because of relative motion between the receiver and the baseband signalling. Coherence time (Tc) of the channel is smaller than the symbol period (Ts) of the transmitted signal.

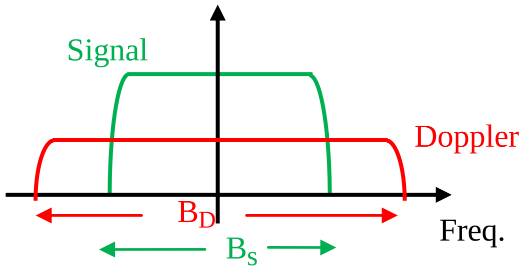


Fig.5.4 Fast Fading as a function of Doppler spread

Slow fading:

Rate of change of the channel characteristics is much smaller than the rate of change of the transmitted signal. The channel may be assumed static over one or several reciprocal bandwidth intervals. In frequency domain, this means that Doppler Spread of channel is much smaller than the bandwidth of baseband signal.Velocity of the mobile (or the velocity of objects in the channel) and the baseband signalling determines whether a signal undergoes fast fading or slow fading.

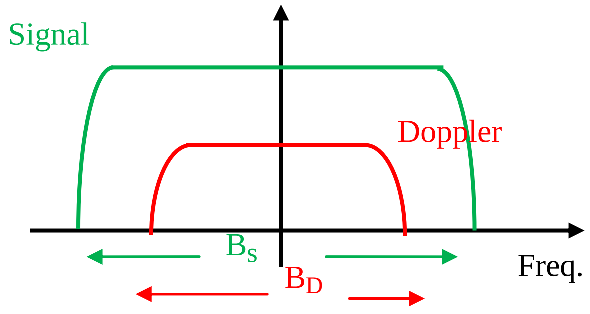


Fig. 5.5 Slow Fading as a function of Doppler spread

Fading Classification-Summary:

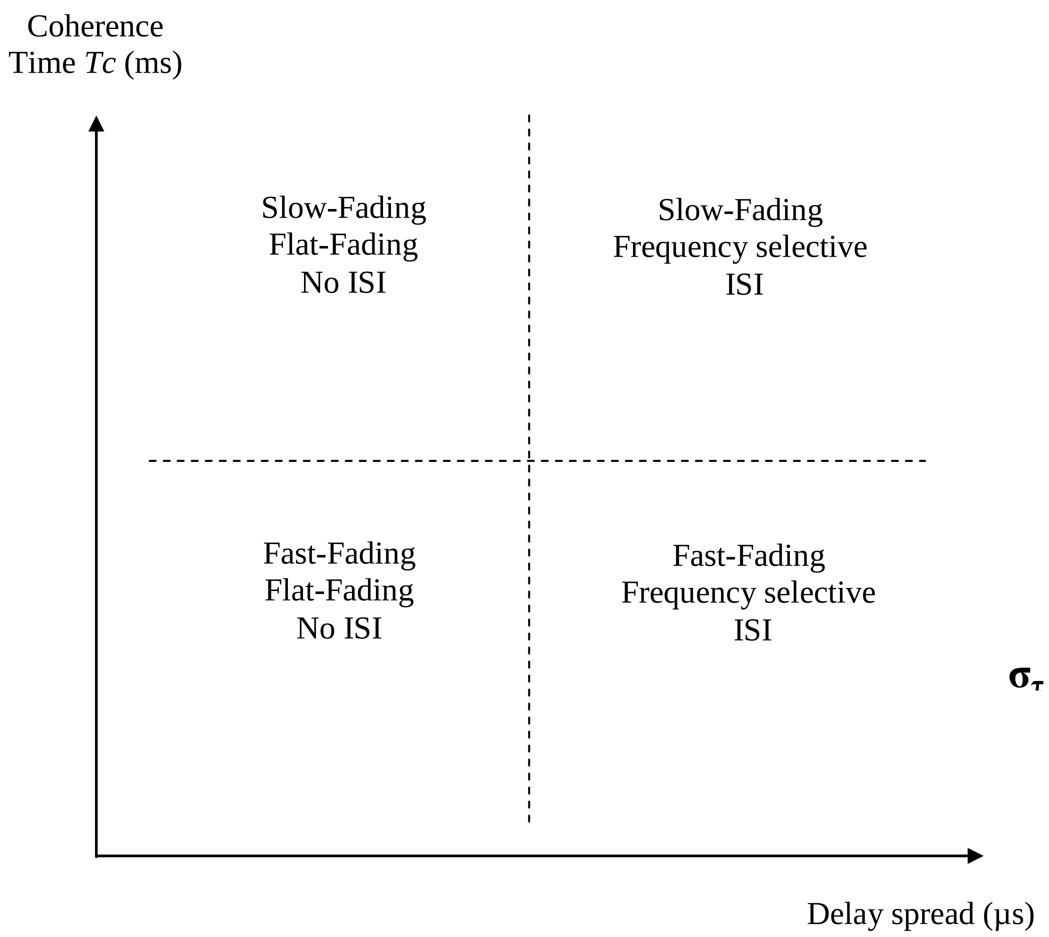


Fig. 5.6 Fading classification : Summary

Fig.5.6 shows the summary of fading classifications. At lower values of coherence time and delay spread, the channel variation is fast and the type of fading flat, hence no ISI. At lower value of coherence time and as the delay spread increases, the fading becomes frequency-selective, which leads to ISI. At higher values of coherence time and lower values of delay spread, the channel variation is slow and the fading is flat, hence there is no ISI. At higher values of coherence time and delay spread, the fading becomes frequency-selective, which causes ISI [2].

5.4 Fading Distributions

Statistical characterization of the variation of the envelope of the received signal over time. Two most common distributions, Rayleigh Fading, Rician Fading.

Rayleigh Fading:

If all the multipath components have approximately the same amplitude (that is when mobile station is far from base station and there are several reflectors) the envelope of the received signal is Rayleigh distributed. This distribution must not have dominant signal component such as the line of sight (LOS) component. A sample of a Rayleigh fading signal is shown in the following figure. Signal amplitude (in dB) versus time for an antenna moving at constant velocity. Notice the deep fades that occur occasionally. Although fading is a random process, deep fades have a tendency to occur approximately every half a wavelength of motion.Fig.5.7 shows the fading distribution for the carrier frequency 900MHz and the receiver speed is 120km/hr [16].

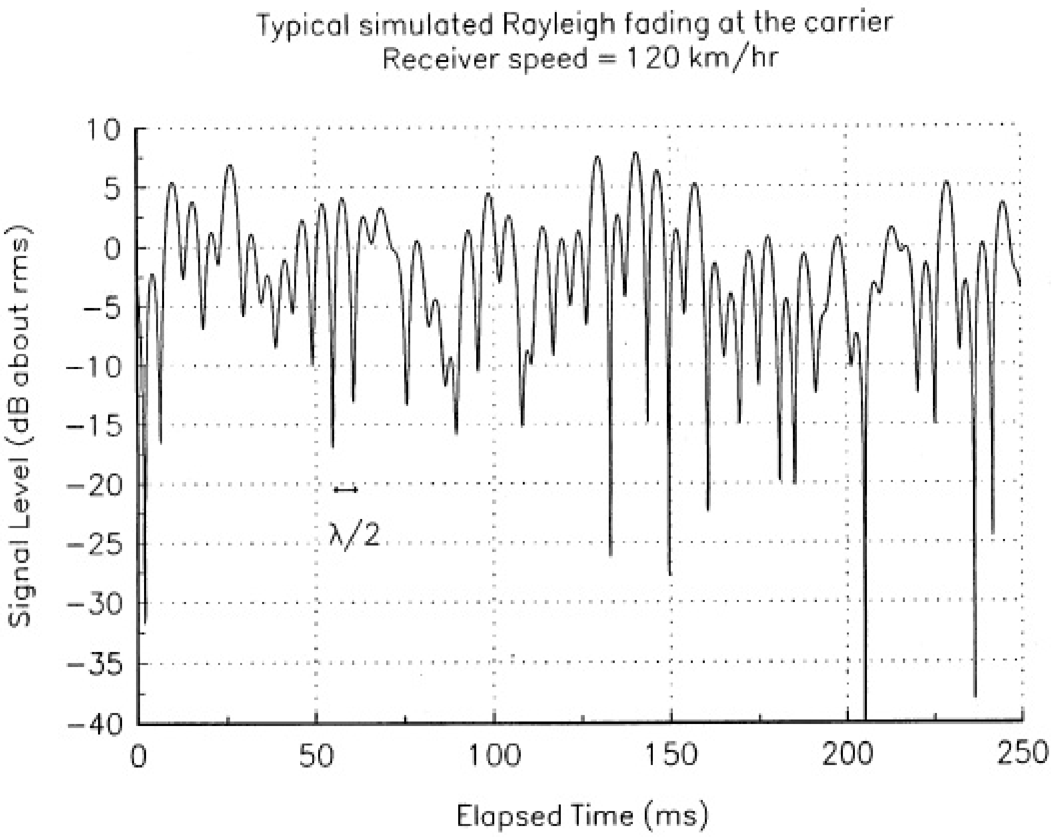


Fig. 5.7 Rayleigh Fading

The Rayleigh distribution has a probability density function (pdf) given by

$$f_{Rayleigh}\,(r) = \frac{r}{\sigma_o^2}\exp\left[-\frac{r^2}{2\sigma_o^2}\right]$$

Ricean Fading :

When there is a dominant, stationary (non-fading) signal component present such as LOS, which is usually possible when MS and BS are close to each other, the fading envelope is Ricean. The Ricean distribution degenerates to Rayeigh when the dominant component fades away. The fading amplitude r_i at the *i*th time instant can be represented as

$$r_i \;= \sqrt{(x_i \,+\, \beta)^2 + {y_i}^2},$$

Where β is the amplitude of the specular component and xi, yi are samples of zero-mean stationary Gaussian random process with variance σ_o^2. The ratio of specular to diffuse energy defines the so-called Rician K-factor, which is given by $K \;=\; \beta^2/2\sigma_o^2$. The Rician PDF is given by

$$f_{Rice}\,(r) = \frac{r}{\sigma_o^2}\exp\left[-\frac{r^2 + \beta^2}{2\sigma_o^2}\right] I_o \,\left[\frac{r\beta}{\sigma_o^2}\right]$$

Where $I_o\;[.]$ is the zero-order modified Bessel function of the first kind and r ≥ 0 . Now, if there is no dominant propagation path, K = 0 and $I_o\;[.] = 1$ yielding Rayleigh PDF. Fig.5.8 shows the Rician PDF with various K-factors [17].

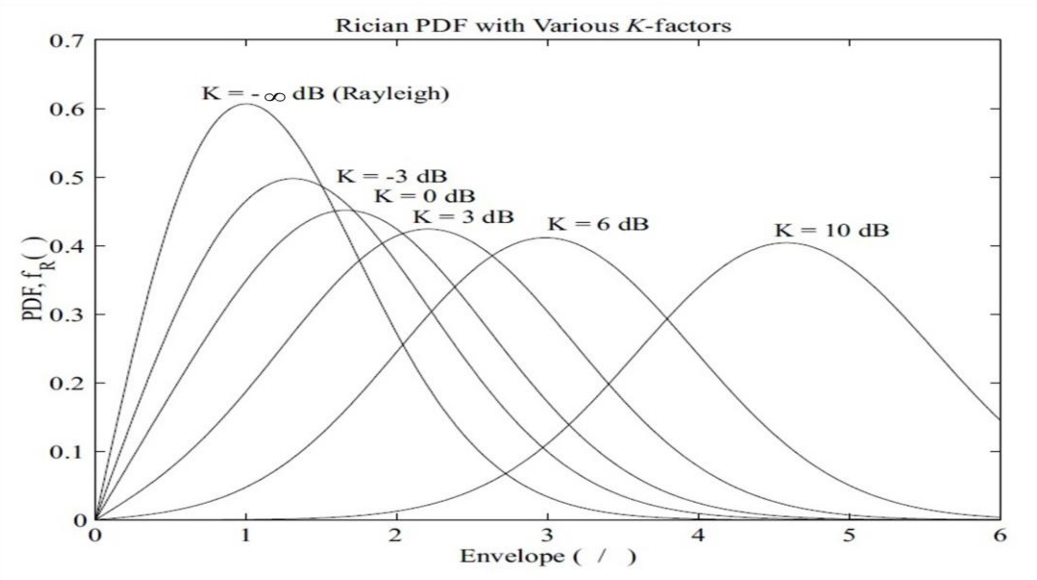


Fig. 5.8 Rician PDF with various K-factors

Chapter 6

MULTI-INPUT MULTI-OUTPUT (MIMO)

6.1 MIMO Link

6.1.1 MIMO Configurations

Several different diversity modes are used to make radio communications more robust, even with time varying channels. These include time diversity (different timeslots and channel coding), frequency diversity (different channels, spread spectrum, and OFDM), and also spatial diversity. Spatial diversity requires the use of multiple antennas at the transmitter or the receiver end. Multiple antenna systems are typically known as Multiple Input Multiple Output systems (MIMO). Multiple antenna technology can also be used to increase the data rate by spatial multiplexing technique. In practice, both diversity and spatial multiplexing methods are used separately or in combination, depending on the channel state information available at the transmitter and at the receiver. The different configurations of multiple antennas are shown in Fig.6.1 [18].

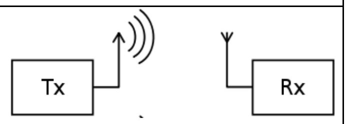
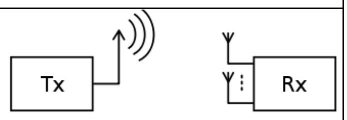
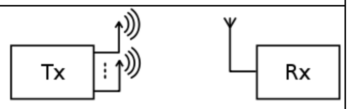
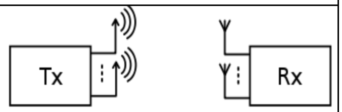
Multi-antenna types		
SISO	Single-input-single-output means that the transmitter and receiver of the radio system have only one antenna.	
SIMO	Single-input-multiple-output means that the receiver has multiple antennas while the transmitter has one antenna.	
MISO	Multiple-input-single-output means that the transmitter has multiple antennas while the receiver has one antenna.	
MIMO	Multiple-input-multiple-output means that the both the transmitter and receiver have multiple antennas.	

Fig. 6.1 MIMO Configurations

6.1.2 MIMO Channel

Consider a MIMO system with M_T transmit antennas and M_R receive antennas as shown in Fig.5.2 [8]. Denoting the impulse response between the *j* th (*j* = 1, 2, ... , M_T) transmit antenna and *i* th (*i* = 1, 2, ... , M_R) receive antenna by $h_{(i,j)}(\tau, t)$,

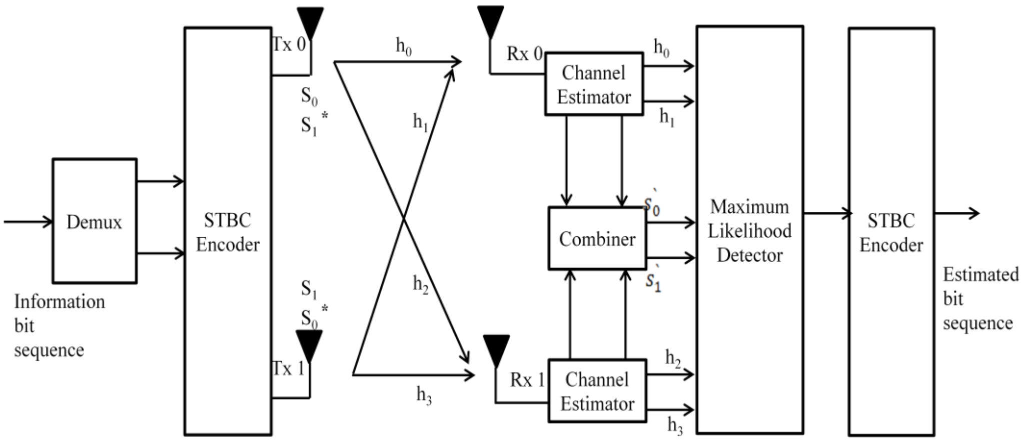


Fig 6.2 : 2x2 MIMO – STBC Channel

the MIMO channel is given by the $M_R \times M_T$ matrix $H(\tau, t) =$

$$\begin{bmatrix} h_{1,1}\ (\tau,t) & h_{1,2}\ (\tau,t) & \cdots & h_{1,M_T}\ (\tau,t) \\ h_{2,1}\ (\tau,t) & h_{2,1}\ (\tau,t) & \cdots & h_{2,M_T}\ (\tau,t) \\ \vdots & \vdots & \ddots & \vdots \\ h_{M_R,1}\ (\tau,t) & h_{M_R,2}\ (\tau,t) & \cdots & h_{M_R,M_T}\ (\tau,t) \end{bmatrix}$$

The vector $[h_{1,j}\ (\tau,t) \quad h_{2,j}\ (\tau,t)\ \ldots\ h_{M_R,j}\ (\tau,t)]^T$ is the spatio-temporal signature or channel induced by the *j* th transmit antenna across the receive antenna array. Given that the signal S_j (t) is launched from the *j* th transmit antenna, the signal received at the *i* th receive antenna, Y_i (t), is given by

$$Yi\ (t) = \sum_{j=1}^{M_T} h_{i,j}\ (\tau,t) * S\,j\ (\text{t}),$$

where i = 1,2,..,M_R. The input-output relation for the MIMO channel [12]. may be expressed in matrix notation as $y(t) = H(\tau,t) * S(t)$, where $S(t) = [s_1(t)\ \ s_2(t) \ldots\ s_{M_T}(t)]^T$ is an $M_T \times 1$ vector and

$Y(t) = [y_1(t)\ \ y_2(t) \ldots\ \ y_{M_T}(t)]^T$ is a vector of dimension M_R *x 1*.

The following transmission formula results from receive vector ***y***, transmit vector ***x***, and noise ***n***: ***y = Hx + n*** . The channel capacity of a SISO system is given by *C = log* $_2$ *(1 + SNR).* For a MIMO system, the channel capacity is given by *C = min (n, m) * log* $_2$ *(1 + SNR)* [19].

6.1.3 Spatial Diversity

MIMO exploits the space dimension to improve wireless systems capacity, range and reliability. It offers significant increase in data throughput and link range without additional bandwidth or increased transmit power. MIMO achieves this goal by spreading the same total transmit power over the antennas to achieve an array gain that improves the spectral efficiency (more bits per second per hertz of bandwidth) or to achieve a diversity gain that improves the link reliability (reduced fading). As the number of antenna element increases, the channel capacity also increases.

6.1.4 Receive Diversity

Receive diversity uses more antennas on the receiver side than on the transmitter side. The simplest scenario consists of two receive and one transmit antenna (SIMO, 1x2).

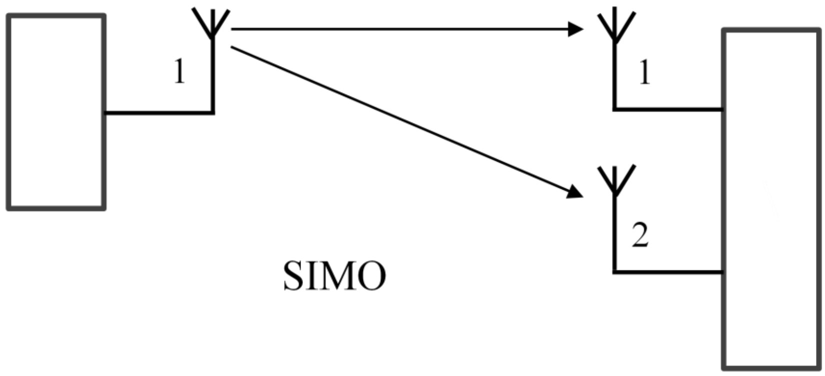


Fig. 6.3 SIMO antenna configuration

Because special coding methods are not needed, this scenario is very easy to implement. Only two RF paths are needed for the receiver. Because of the different transmission paths, the receiver gets two differently faded signals. By using the appropriate method at the receiver end, the signal-to-noise ratio can now be increased. Switched diversity always uses the stronger signal, while maximum ratio combining uses the summation of signals from the two receivers.

6.1.5 Transmit Diversity

When there are more transmit antennas than receive antenna, this is called Transmit diversity. The simplest scenario uses two transmit and one receive antenna (MISO, 2x1).

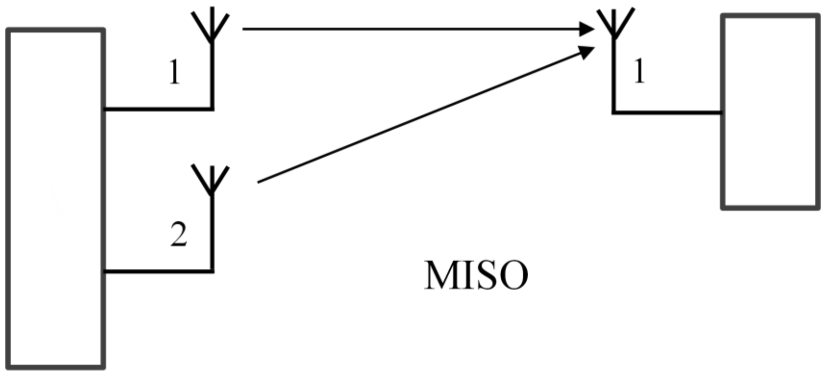


Fig. 6.4 MISO antenna configuration

In this case, the same data is transmitted redundantly over two antennas. To generate a redundant signal, space-time codes are used. Alamouti developed the first space-time codes for two transmit antennas with one receive antenna. This method can be extended to 2x2 antennas also.

6.2 Space-Time Block coding

The technique proposed by Alamouti [8] is a simple transmit diversity scheme which improves the signal quality at the receiver on one side of the link by simple processing across two transmit antennas on the opposite side. The obtained diversity order is equal to applying maximal-ratio receiver combining (MRRC) with two antennas at the receiver. The scheme may easily be generalized to two transmit antennas and M receive antennas to provide a diversity order of $2M$. This is done without any feedback from the receiver to the transmitter and with small computation complexity. The scheme requires no bandwidth expansion, as redundancy is applied in space across multiple antennas, not in time or frequency. The new transmit diversity scheme as shown in Fig.5.5,can improve the error performance, data rate, or capacity of wireless communications systems. The decreased sensitivity to fading may allow the use of higher level modulation schemes to increase the effective data rate.

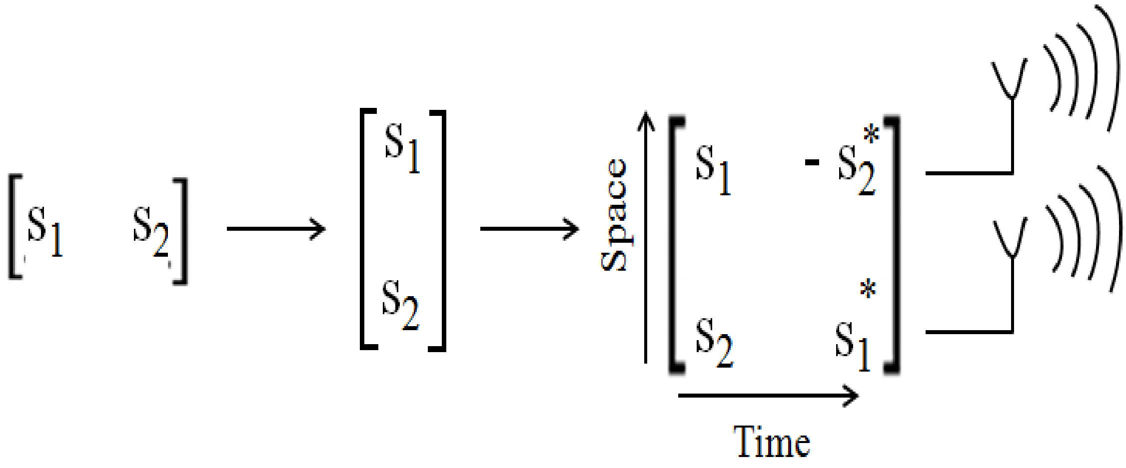


Fig. 6.5 Alamouti Space-Time coding scheme

Two-Branch Transmit Diversity with One Receiver

Fig.6.6 shows the baseband representation of the new two-branch transmit diversity scheme. The scheme uses two transmit antennas and one receive antenna and may be defined by the following three functions:

- the encoding and transmission sequence of information symbols at the transmitter;
- the combining scheme at the receiver;
- the decision rule for maximum likelihood detection.

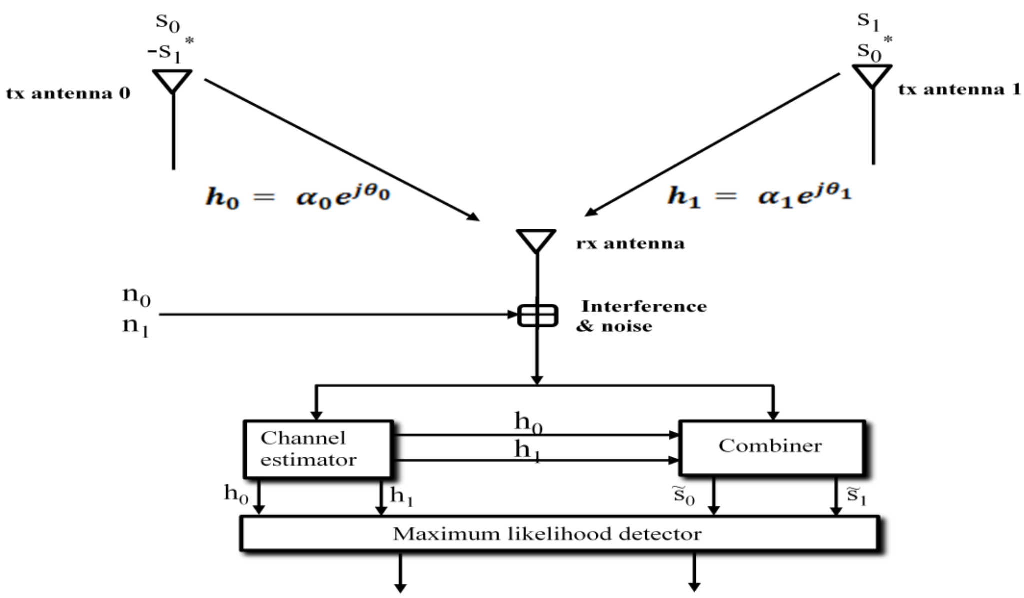


Fig. 6.6 Two-branch transmit diversity scheme with one receiver

The Encoding and Transmission Sequence:

At a given symbol period, two signals are simultaneously transmitted from the two antennas. The signal transmitted from antenna zero is denoted by S_1 and from antenna one by S_2. During the next symbol period signal ($- S_{2*}$) is transmitted from antenna zero, and signal (S_{1*}) is transmitted from antenna one where * is the complex conjugate operation. This sequence is shown in Fig.6.7. The encoding is done in space and time (space–time coding). The encoding, however, may also be done in space and frequency. Instead of two adjacent symbol periods, two adjacent carriers may be used (space–frequency coding).

	antenna 0	**antenna 1**
time *t*	s_0	s_1
time *t + T*	$-s_1^*$	s_0^*

Fig.6.7 The Encoding and Transmission sequence for the 2 branch Transmit diversity scheme.

The channel at time t may be modeled by a complex multiplicative distortion $\boldsymbol{h_0(t)}$ for transmit antenna zero and $\boldsymbol{h_1(t)}$ for transmit antenna one. Assuming that fading is constant across two consecutive symbols, we can write

$$h_0(t) = h_0(t+T) = h_0 = \propto_0 e^{j\theta_0}$$

$$h_1(t) = h_1(t+T) = h_1 = \propto_1 e^{j\theta_1} \qquad (6.1)$$

Where T is the symbol duration. The received signals can then be expressed as

$$r_0 = \quad r(t) = \quad h_0 s_0 + h_1 s_1 + n_0$$

$$r_1 = r(t+T) = -h_0 s_1^* + h_1 s_0^* + n_1 \qquad (6.2)$$

Where r_0 and r_1 are the received signals at time t and $t+T$ and n_0 and n_1 are complex random variables representing receiver noise and interference.

The Combining Scheme:

The combiner shown in Fig.6.6 builds the following two combined signals that are sent to the maximum likelihood detector :

$$\tilde{s}_0 = h_0^* r_0 + h_1 r_1^*$$

$$\tilde{s}_1 = h_1^* r_0 - h_0 r_1^* \qquad (6.3)$$

Substituting (10) and (11) into (12), we get

$$\tilde{s}_0 = (\alpha_0^2 + \alpha_1^2) s_0 + h_0^* n_0 + h_1 n_1^*$$

$$\tilde{s}_1 = (\alpha_0^2 + \alpha_1^2) s_1 - h_0 n_1^* + h_1^* n_0 \qquad (6.4)$$

The Maximum Likelihood Decision Rule:

These combined signals are then sent to the maximum likelihood detector. The resulting combined signals in (6.4) are equivalent to that obtained from two-branch MRRC. Therefore, the resulting diversity order from the new two-branch transmit diversity scheme with one receiver is equal to that of two-branch MRRC.

Two-Branch Transmit Diversity with M Receiver

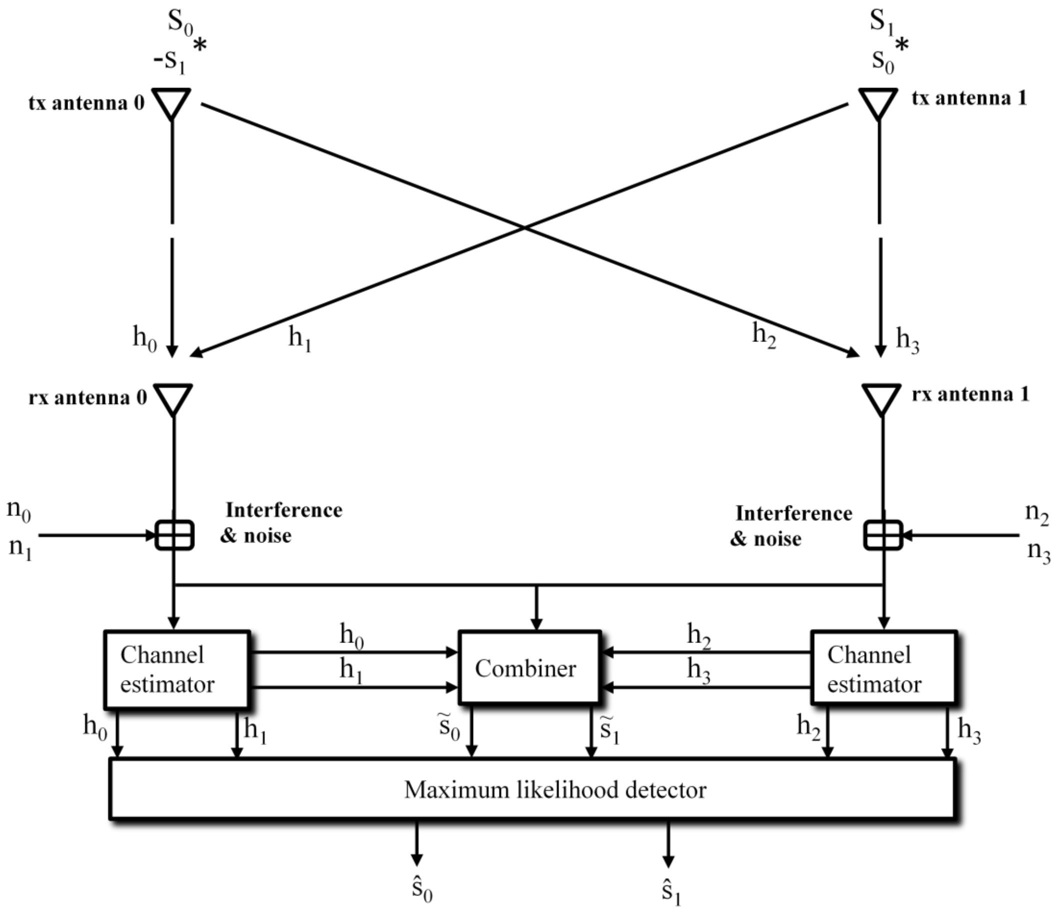


Fig.6.8 The new two-branch diversity scheme with two receivers

	rx antenna 0	**rx antenna 1**
tx antenna 0	h_0	h_2
tx antenna 1	h_1	h_3

Fig.6.9 Definition of channels between the transmit and receive antennas

	rx antenna 0	rx antenna 1
time t	r_0	r_2
time $t+T$	r_1	r_3

Fig.6.10 Notation for the received signals at the two receive antennas

Fig.6.8 shows the baseband representation of the new scheme with two transmit and two receive antennas. The encoding and transmission sequence of the information symbols for this configuration is identical to the case of a single receiver, shown in Fig.6.7. Fig.6.9 defines the channels between the transmit and receive antennas, and Fig.6.10 defines the notation for the received signal at the two receive antennas, where

$$r_0 = h_0 s_0 + h_1 s_1 + n_0$$
$$r_1 = -h_0 s_1^* + h_1 s_0^* + n_1$$
$$r_2 = h_2 s_0 + h_3 s_1 + n_2$$
$$r_3 = -h_2 s_1^* + h_3 s_0^* + n_3 \qquad (6.5)$$

$n_{0,}\ n_{1,}\ n_2\ and\ n_3$ are complex random variables representing receiver thermal noise and interference. The combiner in Fig.6.8 builds the following two signals that are sent to the maximum likelihood detector:

$$\tilde{s}_0 = h_0^* r_0 + h_1 r_1^* + h_2^* r_2 + h_3 r_3^*$$
$$\tilde{s}_1 = h_1^* r_0 - h_0 r_1^* + h_3^* r_2 - h_2 r_3^* \qquad (6.6)$$

Substituting the appropriate equations we have

$$\tilde{s}_0 = (\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2) s_0 + h_0^* n_0 + h_1 n_1^* + h_2^* n_2 + h_3 n_3^*$$
$$\tilde{s}_1 = (\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2) s_1 - h_0 n_1^* + h_1^* n_0 - h_2 n_3^* + h_3^* n_2 \qquad (6.7)$$

These combined signals are then sent to the maximum likelihood decoder which for signal s_0, uses the following decision criteria :

Choose s_i iff

$$(\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2 - 1)|s_i|^2 + d^2(\tilde{s}_0, s_i)$$
$$\leq (\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2 - 1)|s_k|^2 + d^2(\tilde{s}_0, s_k) \qquad (6.8)$$

Choose s_i iff

$$d^2(\tilde{s}_0, s_i) \leq d^2(\tilde{s}_0, s_k), \qquad \forall i \neq k. \tag{6.9}$$

Similarly, for s_1, using the decision rule is to choose signal s_i iff

$$(\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2 - 1)|s_i|^2 + d^2(\tilde{s}_1, s_i)$$
$$\leq (\alpha_0^2 + \alpha_1^2 + \alpha_2^2 + \alpha_3^2 - 1)|s_k|^2 + d^2(\tilde{s}_1, s_k) \tag{6.10}$$

choose s_i iff

$$d^2(\tilde{s}_1, s_i) \leq d^2(\tilde{s}_1, s_k), \qquad \forall \boldsymbol{i} \neq k. \tag{6.11}$$

The combined signals in (6.7) are equivalent to that of four-branch MRRC. Therefore, the resulting diversity order from the new two-branch transmit diversity scheme with two receivers is equal to that of the four-branch MRRC scheme. It is interesting to note that the combines signals from the two receive antennas are the simple addition of the combined signals from each receive antenna, i.e., the combining scheme is identical to the case with a single receive antenna. Hence, using two transmit and M receive antennas, a combiner can be used for each receive antenna and then simply add the combined signals from all the receive antennas to obtain the same diversity order as 2M-branch MRRC. In other words, using two antennas at the transmitter, the scheme doubles the diversity order of systems with one transmit and multiple receive antennas.

6.3 Channel Estimation & Detection

Channel estimation is done to obtain Channel State Information (CSI). CSI refers to known channel properties of a communication link. This information describes how a signal propagates from the transmitter to the receiver and represents the combined effect of scattering, fading, and power decay with path length. The efficacy of detection process at the receiver depends on the accuracy of CSI. In channel estimation, the transmitted sequence is assumed to be known at the receiver. In a pilot-based scheme, a known sequence is transmitted and this is used to estimate the channel. In a MIMO channel, the receiver receives a superposition of the transmitted signals and must separate the constituent signals based on channel knowledge. Maximum likelihood (ML) decoding, which amounts to exhaustive comparisons of the received signal to all possible transmitted signals. This is computationally prohibitive for higher order constellations such as 64-QAM. Lower complexity sub-optimal receivers include the zero-forcing receiver (ZF) or the minimum mean square error (MMSE) receiver, the design principles of which are similar to equalization principles for SISO links with Inter-Symbol Interference (ISI). An attractive alternative to ZF and MMSE receivers is the vertical BLAST (V-BLAST) algorithm, which is essentially a successive cancellation technique [20].

6.4 Spatial Multiplexing and BLAST Architectures

The basic premise of spatial multiplexing is to send M_t independent symbols per symbol period using the dimensions of space and time. To obtain full diversity order, an encoded bit stream must be transmitted over all M_t transmit antennas. This can be done through a serial encoding, illustrated in Fig.6.11.

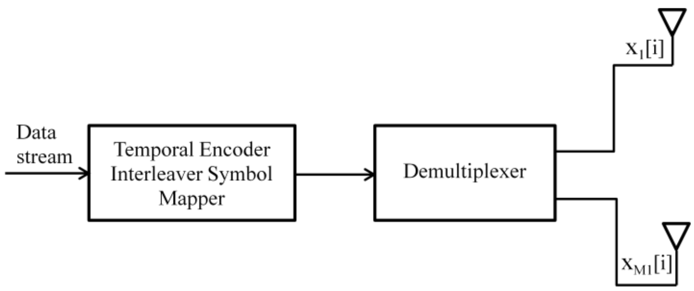


Fig.6.11 Spatial multiplexing with serial encoding

With serial encoding, the bit stream is temporally encoded over the channel block length T to form the codeword $[x_1, \ldots, x_T]$. The codeword is interleaved and mapped to a constellation point, then demultiplexed onto the different antennas. The first M_t symbols are transmitted from the M_t antennas over the first symbol time, and this process continues until the entire codeword has been transmitted. We denote the symbol sent over the *k*th antenna at time *i* as $x_k[i]$. If a codeword is sufficiently long, it is transmitted over all M_t transmit antennas and received by all M_r receive antennas, resulting in full diversity gain. However, the codeword length T required to achieve this full diversity is $M_t\, M_r$, and decoding complexity makes serial encoding impractical.

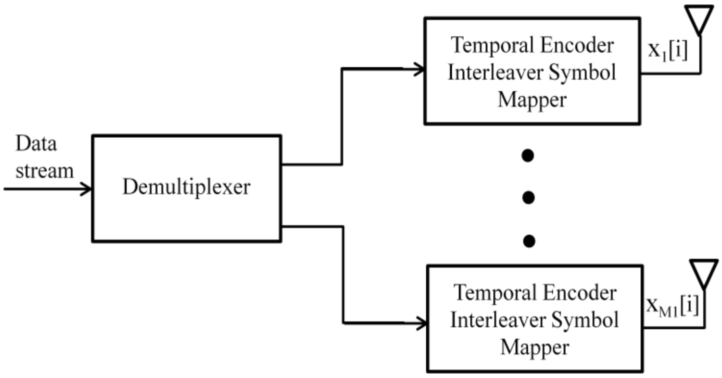


Fig.6.12 Spatial multiplexing with parallel encoding : V-BLAST

A simpler method to achieve spatial multiplexing, pioneered at Bell Laboratories as one of Bell Labs Layered Space Time (BLAST) architectures for MIMO channels, is parallel encoding, illustrated in Fig.6.12. With parallel encoding the data stream is demultiplexed into M_t independent streams. Each of the resulting substreams is passed through an SISO temporal encoder with block length T, interleaved, mapped to a signal constellation point, and transmitted over its corresponding transmit antenna. Specifically, the k th SISO encoder generates the codeword { $x_k\,[i]$, $i = 1,2,\ldots,T$}, which is transmitted sequentially over the kth antenna. This process can be considered to be the encoding of the serial data into a vertical vector and hence is also referred to as *vertical encoding* or V-BLAST. Vertical encoding can achieve at most a diversity order of M_r, since each coded symbol is transmitted from one antenna and received by M_r antennas. This system has a simple encoding complexity that is linear in the number of antennas.

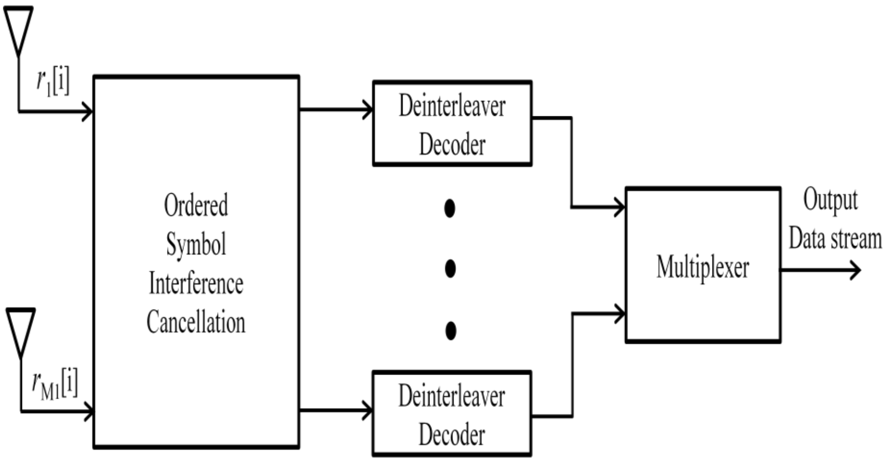


Fig.6.13 V-BLAST receiver with linear complexity

However, optimal decoding still requires joint detection of the code words from each of the transmit antennas, since all transmitted symbols are received by all the receive antennas. The receiver complexity can be significantly reduced through the use of *symbol interference cancellation,* as shown in Fig.6.13. This cancellation, which exploits the synchronicity of the symbols transmitted from each antenna, works as follows. First the M_t transmitted symbols are ordered in terms of their received SNR. An estimate of the received symbol with the highest SNR is made while treating all other symbols as noise. This estimated symbol is subtracted out, and the symbol with the next highest SNR estimated while treating the remaining symbols as noise. This process repeats until all M_t transmitted symbols have been estimated. After cancelling out interfering symbols, the coded sub stream associated with each transmit antenna can be individually decoded, resulting in a receiver complexity that is linear in the number of transmit antennas.

The simplicity of parallel encoding and the diversity benefits of serial encoding can be obtained by using a creative combination of the two techniques called *diagonal encoding* or D-BLAST, illustrate in Fig.6.14.

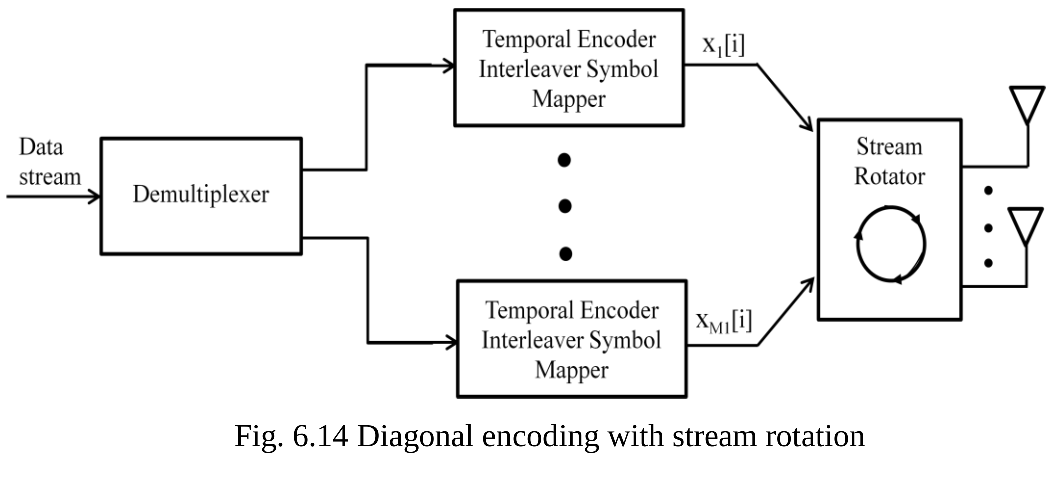


Fig. 6.14 Diagonal encoding with stream rotation

In D-BLAST, the data stream is first parallel encoded. However, rather than transmitting each codeword with one antenna, the codeword symbols are rotated across antennas, so that a codeword is transmitted by all M_t antennas. The operation of the stream rotation is shown in Fig.6.15.

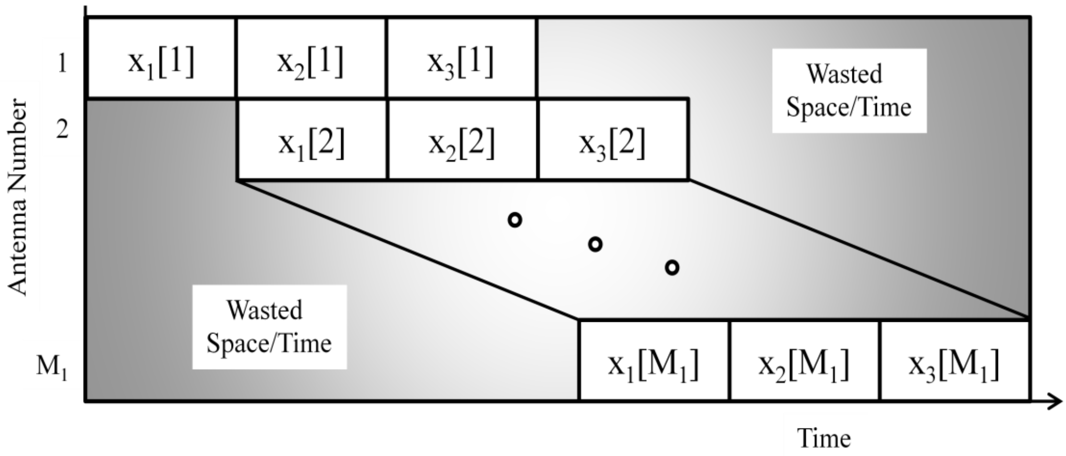


Fig.6.15 Stream rotation

Suppose the *i*th encoder generates the codeword $x_i = [x_i[1], \ldots x_i[T]]$. The stream rotator transmits each symbol on a different antenna, so $x_i[1]$ is sent on antenna 1 over symbol time *i*, $x_i[2]$ is sent on antenna 2 over symbol time *i* +1, and so forth. If the code block length T exceeds M_t then the rotation begins again on antenna 1. As a result, the codeword is spread across all spatial dimensions. Transmission schemes based on D-BLAST can achieve full $M_t M_r$ diversity gain if the temporal coding with stream rotation is capacity-achieving. Moreover, the D-BLAST system can achieve maximum capacity with outage if the wasted space-time dimensions along the diagonals are neglected. Receiver complexity is also linear in the number of transmit antennas, since the receiver decodes each diagonal code independently [20].

Chapter 7

Simulation Results and Discussions

7.1 Simulation setup

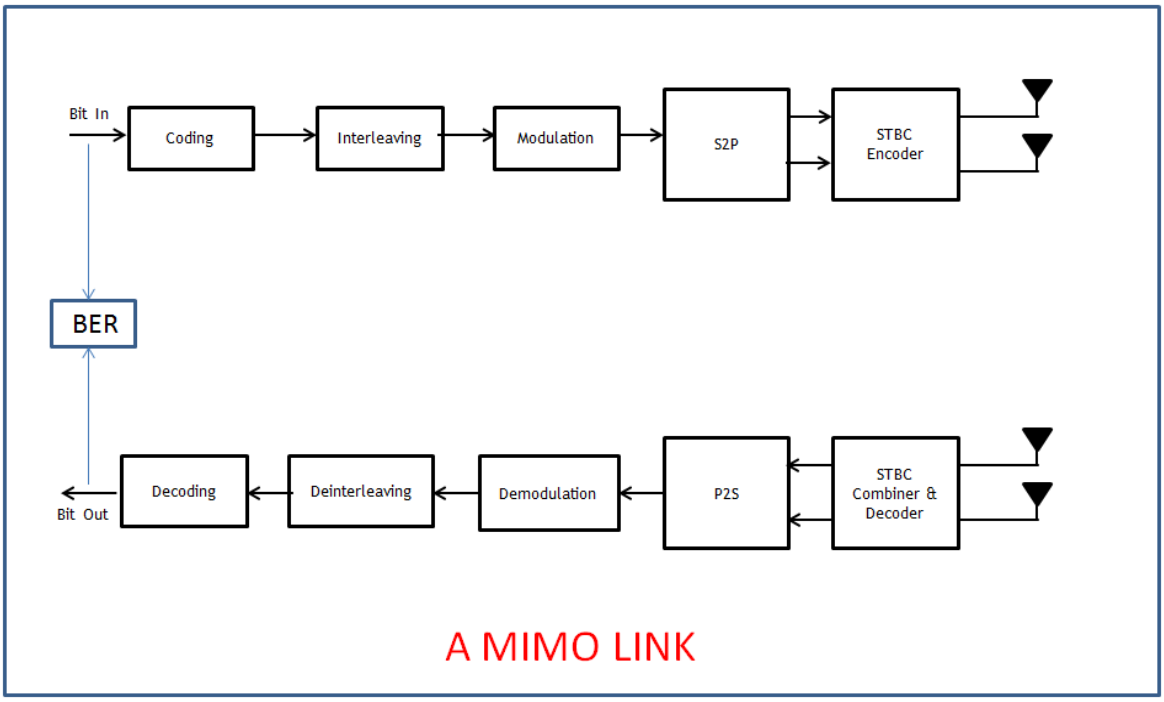


Fig.7.1 MIMO 2x2 STBC Communication link

In our work, we consider, 2x2 MIMO link with STBC as sketched in Fig.7.1. As mentioned earlier, deep fades are equivalent to burst errors. These can be dispersed into random errors using an interleaver. Hence, we use HC and CC as the coding techniques, each concatenated with a random interleaver. The number of bits corrupted in a single block entirely depends on the occurrence of fading pattern and interleaving does not guarantee avoidance of multiple errors in a single block but is presumed to minimize which in turn should improve the error performance. To investigate the effect of adding redundancy, we have considered different code rates for both codes. In HC, the code parameters are (7,4), (15,11),(31,26),(63,57) and in CC the code rates are : 1/2,3/4, 5/6, 9/10. The performance of various code rates are also compared with uncoded scheme.Higher code rates of CC are obtained by considering puncturing technique. The type of interleaving adopted here is random interleaver with seed 123. The error performance also depends on the type of modulation scheme implemented. We have chosen coherent BPSK and QPSK. The two parameters by which these two schemes differ are bandwidth efficiency and the distance between the symbols on the signal space (constellation). As the order of the modulation increases, the signal space becomes narrower. Thus higher order modulation schemes are more prone to fading. However, there should be a tradeoff between the bandwidth efficiency and error performance. This analysis will help in identifying the suitable modulation scheme for a particular application.

We study the performance in the presence of Rayleigh and Rician fading distributions that represent multiple reflectors and scatterers and, in the case of Rician, includes dominant line of sight components. The number of paths considered in both cases are three with the following parameters: path delay (s) [0, 8e-6, 16e-6] , path gain (dB) [0, -10, -20], Doppler shift '30 Hz', sampling rate of the channel '1e6 Hz'. The K-factor used in case of Rician fading is [4, 7, 10]. The product of path delay with signal bandwidth determines the type of fading, i.e., whether flat or frequency-selective. In our case, this product is greater than 0.1 making the fading to be frequency-selective. The channel is taken to be an Additive White Gaussian Noise (AWGN) channel and vary Eb/No values in the range 0 to 12 dB. At the receiving end, the process involved are channel estimation, coherently combining signals from multiple antennas, followed by demodulation, random de-interleaving (seed '123'), and then soft decoding. Simulation of a complete 2x2 MIMO - STBC wireless communication link is done in MATLAB. The various schemes of a MIMO link considered for error performance analysis are MIMO-HC (CC)-BPSK-RAYLEIGH,MIMO-HC (CC)-QPSK-RAYLEIGH, MIMO-HC(CC)-BPSK-RICIAN, MIMO-HC(CC)-QPSK-RICIAN.

7.2 Results & Discussions

In Figure 7.2, we plot the path gain (dB) vs the information bit for Rayleigh distribution. It shows the four fading envelopes for each of the four independent channels in our 2x2 MIMO link. Figure 7.3 is same as Fig.7.2 but for Rician distribution.

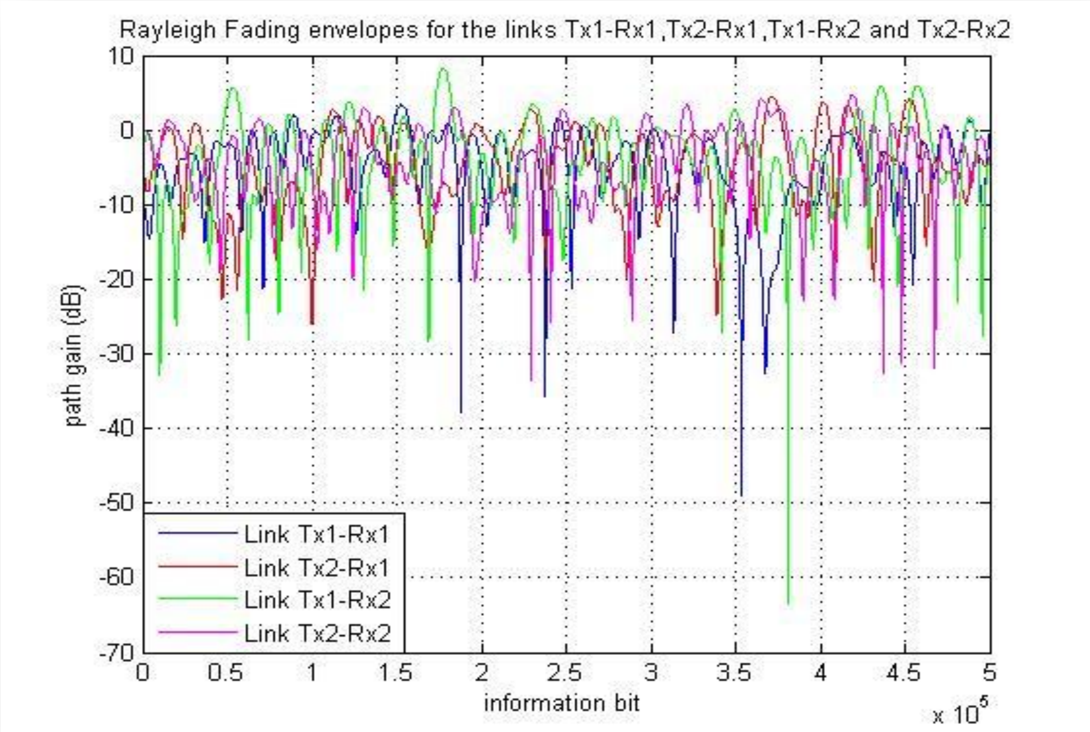


Fig.7.2 MIMO 2x2 Rayleigh channel Fading Envelopes

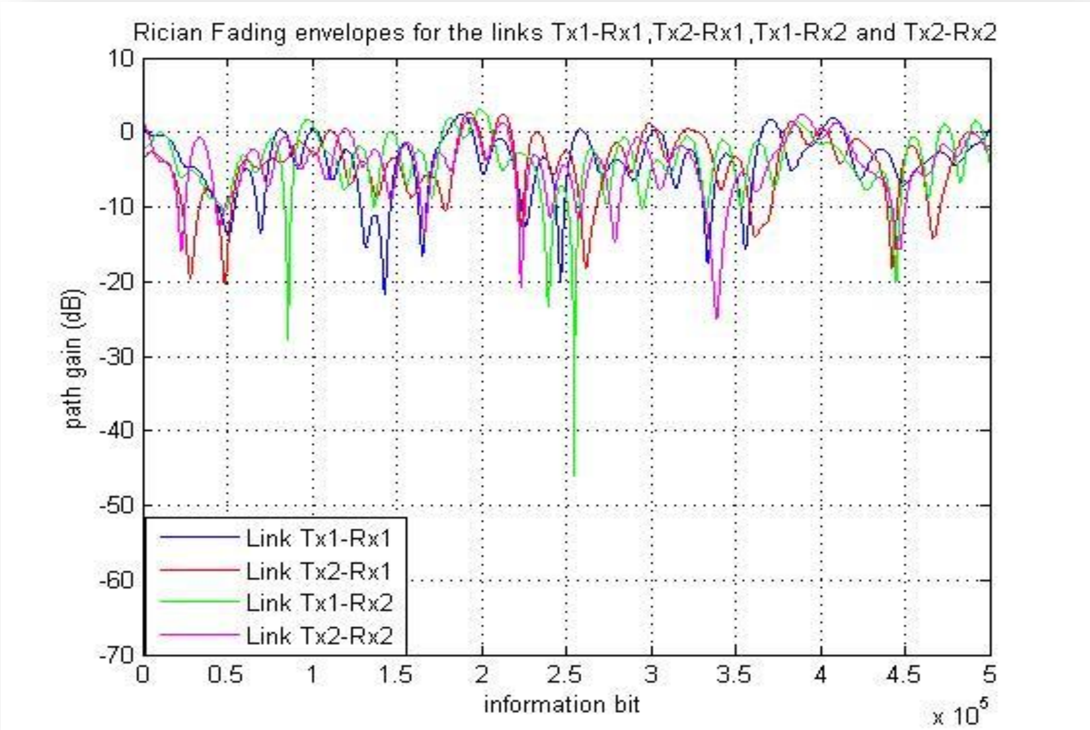


Fig.7.3 MIMO 2x2 Rician channel Fading Envelopes

It is interesting to note that in both the plots on many occasions all the four channels fade together. But, they deep fade together rarely demonstrating diversity. This means complete data loss is rare and channel codes may help improve performance. In the case of Rayleigh, the deep fades happen more often but last for short durations owing to the absence of line of sight components. In the case of Rician, the deep fades are rare but last for longer durations. This implies in equivalent terms the burst errors are wider in Rician compared to Rayleigh. This width should be a small fraction of the packet length in order to avoid multiple errors. Hence, we have chosen a packet length of 100000 bits.

Fig.7.4 shows the Error performance vs Eb/No for various coding schemes, modulation schemes under MIMO Rayleigh and Rician fading environments. From Fig.7.4 we observe that HC and CC have the following common behavior:

1. The BER behavior with respect to code rate and Eb/No are very similar for Rayleigh and Rician fading distributions, in both BPSK and QPSK modulation schemes.
2. BPSK performs better than QPSK with respect to both code rate and Eb/No.

The difference in the behavior of MIMO-HC and MIMO-CC is that, the dependence on code rate in CC is more pronounced compared to HC.

1. In HC, compared to uncoded BER, for smaller redundancy the BER worsens. This is likely to be due to increased temporal exposure to fading environment leading to increased multiple errors within a block. However, as the redundancy is increased, the BER does improve though modestly.
2. In CC, it appears that uncoded transmission performs better than coded transmission even up to the code rate of 0.5. We believe this could be due to puncturing technique employed to achieve higher code rates.

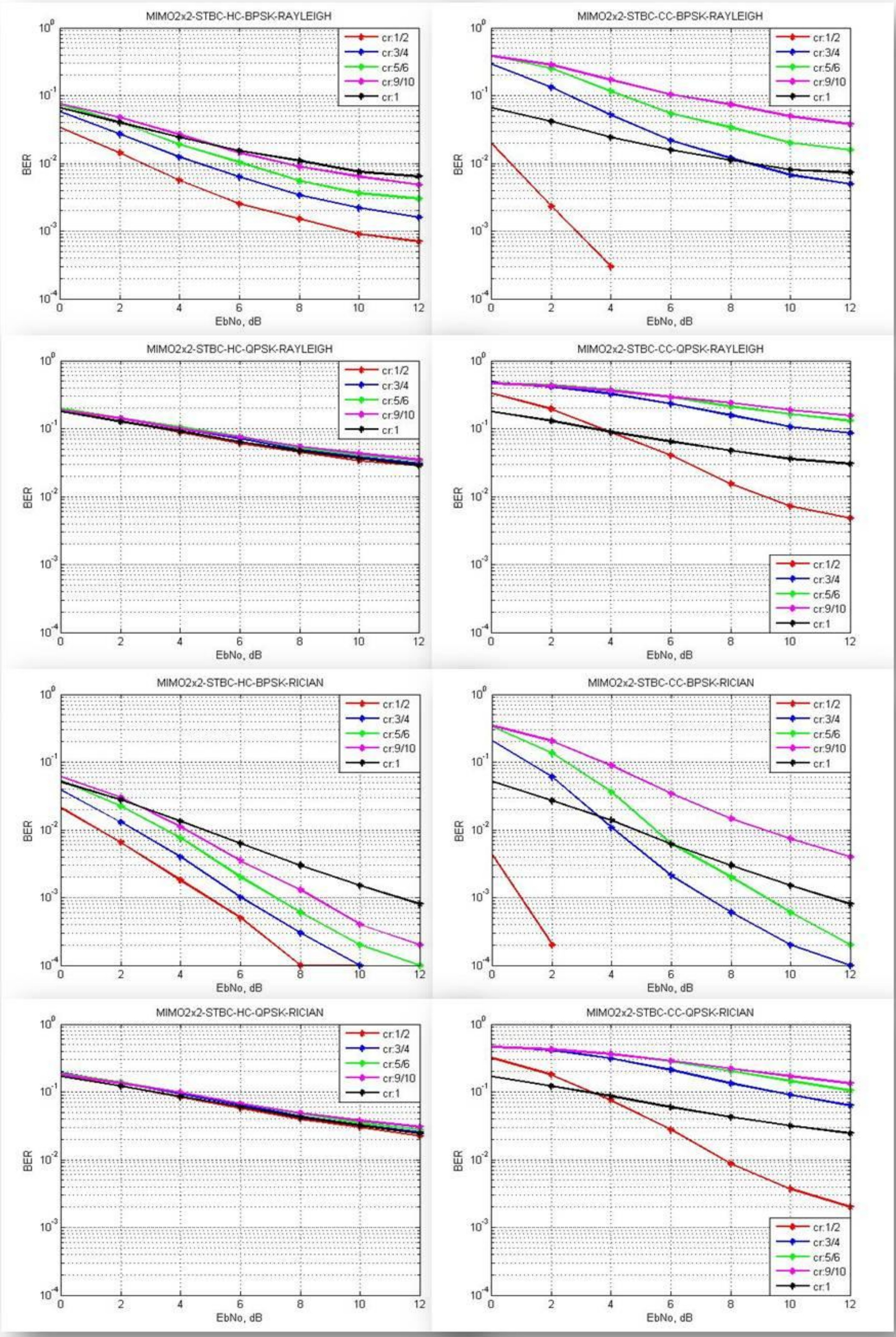


Fig.7.4 Eb/No vs BER for Coding schemes – Hamming, Convolutional, Modulation schemes – BPSK, QPSK and Fading channels – Rayleigh, Rician

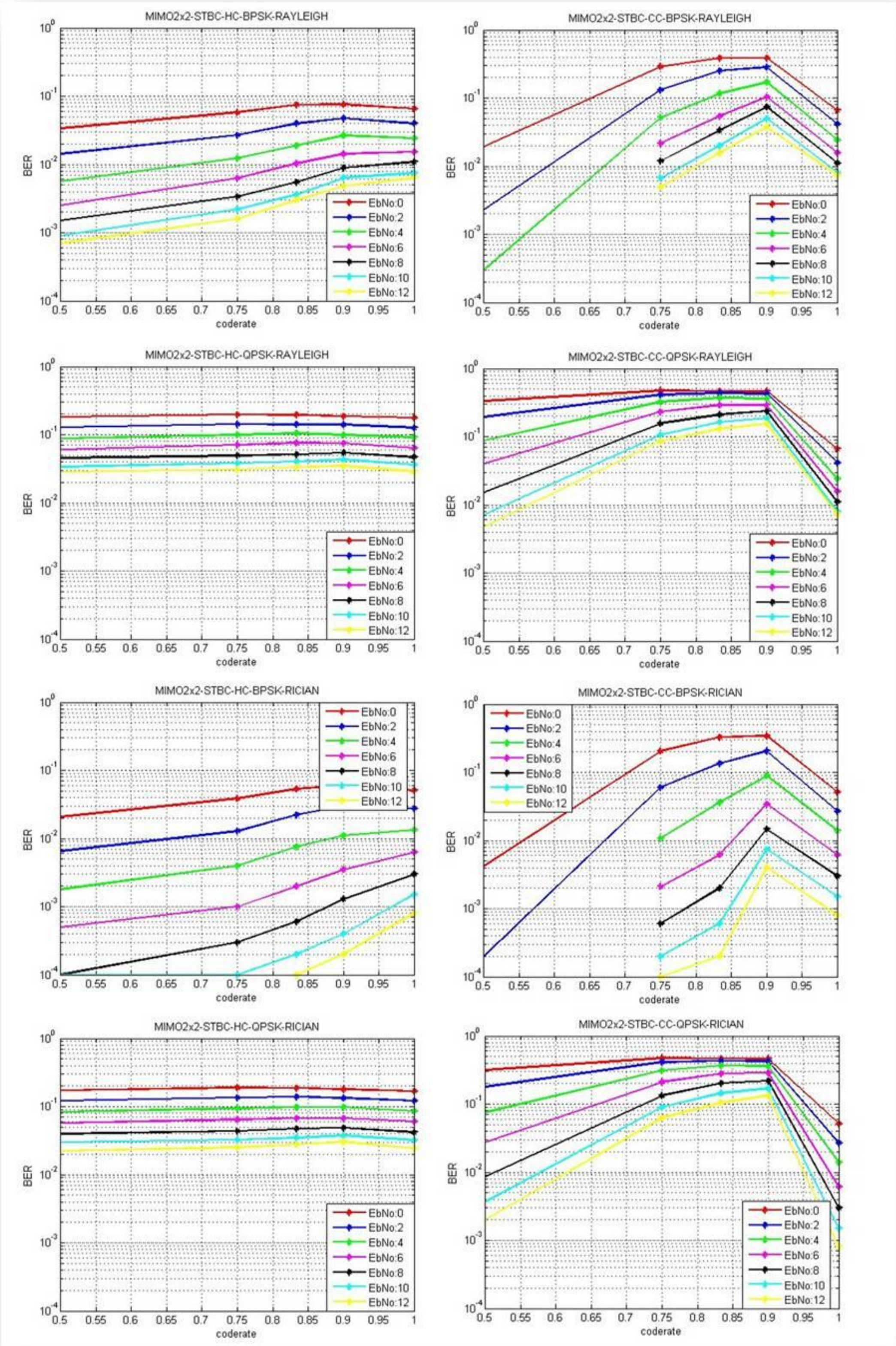


Fig.7.5 Code rate vs BER for Coding schemes – Hamming, Convolutional, Modulation schemes – BPSK, QPSK and Fading channels – Rayleigh, Rician

Fig.7.5 shows the error performance vs code rate for various coding schemes, modulation schemes under MIMO Rayleigh and Rician fading environments. From figure, we observe that there are three kinds of behavior.

1. With HC-QPSK, for both fading distributions, there is no significant improvement with code rate. Here the curves are flat.
2. With CC-BPSK, for fading distributions, the curves have ‘Humpy’ behavior with code rate. There is a threshold value for the code rate, below which the coding improves the BER performance compared to the uncoded one. This threshold depends on Eb/No.
3. With HC-BPSK, the curve monotonically falls with decreasing code rate.

Chapter 8

CONCLUSION

The importance of error control coding techniques for MIMO fading channel has been studied. Wireless multipath fading channel, various coding schemes and MIMO have also been studied. From literature survey, it is observed that, by considering various coding schemes, under different fading environment, significant enhancement in bit error performance and coding gain can be achieved. Hence we set our goal to analyze the performance of various coding schemes under MIMO fading environment. The performance parameters, we have considered are bit error rate, code rate and coding gain. We have carried out simulations of a complete wireless communication MIMO link by considering Hamming and Convolutional codes, BPSK and QPSK modulation schemes under Rayleigh and Rician fading environment. We have presented the BER vs E_b/N_o and BER vs coderate plots for the above mentioned schemes.From the simulation results, it is observed that, coding does improve the error performance, but E_b/N_o has dominant role than the code rate in error recovery. With frequency selective fading, the number of and delay in multiple paths increases, this leads to Inter Symbol Interference (ISI). The MIMO-STBC-OFDM technology is best suited for mitigating ISI in frequency selective fading channel. In our future work we will analyze the error performance of various coding techniques and modulation schemes in the presence of Rayleigh and Rician frequency selective fading channels by employing MIMO-STBC-OFDM technique.

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Error Correcting Codes for MIMO Fading channel

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Abstract

Error control coding techniques are important to achieve reliable communication in the presence of fading that occurs in wireless communication channel. Multi-Input Multi-Output (MIMO) systems are considered where higher data rates are needed, as in the case of fourth generation cellular communication. This paper sets the context of the problem viz. MIMO communication in a fading channel, identifies the limitation and then provides a survey of error control coding techniques relevant for immunizing MIMO communication in such a situation. This sets the scope for a detailed study that we are currently undertaking.

Introduction

The present third generation wireless cellular communication technology employs single antenna both at the transmitter and at the receiver, which results in point-to-point communication. This configuration is known as Single Input Single Output (SISO). The reliable communication depends on the signal strength. If the signal path is in deep fade, there is no guarantee for reliable communication. Since the channel capacity depends on the signal-to-noise ratio and the number of antennas used both at the transmitter and at the receiver, the capacity provided by SISO is limited. If the transmitted signal, passes through multiple signal paths, each path fades independently, there is a possibility that atleast one path is strong, which ensures reliable communication. This technique is called “diversity” [17]. To achieve this, we need to use multiple antennas at the transmitter side and at the receiver side. This technology is called as Multiple Input Multiple Output (MIMO). Hence the multipath fading channels are considered as an advantage scenario in MIMO. This is the most preferred technique in fourth generation cellular communication, since higher data rates can be achieved. Since fading due to multipath propagation in wireless channel can be modeled as burst mode error [19], the error control channel coding techniques developed for wired channel can be efficiently implemented for wireless fading channel.

Fading Channel

The wireless channel is a multipath propagation channel. Multipath in the radio channel causes rapid fluctuation of signal amplitude, called small scale fading or simply fading. Fading is caused by destructive interference of two or more versions of the transmitted signal arriving at the receiver at slightly different times with different amplitudes and phases. Delayed signals are the result of reflections/scatterings from terrain features such as trees, hills, or mountains or objects such as people, vehicles or buildings. The received signal may vary widely in amplitude and phase over a short period of time or travel distance.

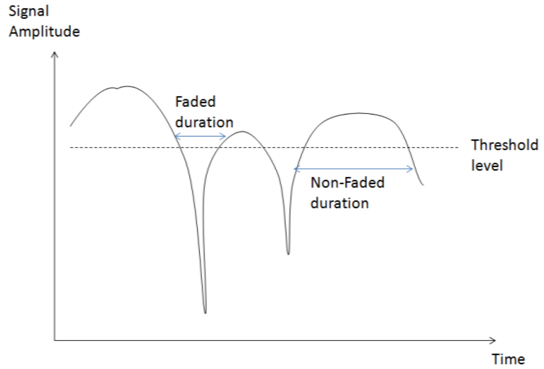

Fig 1: Fading as a function of signal amplitude and time.

Statistical characterization of the variation of the envelope of the received signal over time leads to fading distributions. Two most common fading distributions we encounter are Rayleigh Fading and Rician Fading. If all the multipath components have approximately the same amplitude, the envelope of the received signal is Rayleigh distributed. No dominant signal component, such as line of sight component must exist. When there is a single dominant, stationary signal component present, the fading envelope is Ricean. The Ricean distribution slowly degenerates to Rayleigh when the dominant component fades away [20].

Importance of Error detection and correction

Error detection and correction are techniques that enable reliable delivery of digital data over unreliable communication channels. Many communication channels are subject to channel noise, and thus errors may be introduced during transmission from the source to a receiver. Error detection techniques allow detecting such errors, while error correction enables reconstruction of the original data. The general idea for achieving error detection and correction is to add some redundancy (i.e., some extra data) to a message. The parity bits are added in a known fashion, which is to be known both to the encoder and the decoder. The Shannon theorem states that given a noisy channel with channel capacity C and information transmitted at a rate R, then there exists a coding technique, that allow the probability of error at the receiver to be made arbitrarily small. This means that, theoretically it is possible to transmit information at a rate below the capacity C, without error [16]. This rate depends on the noisiness of the channel.

Error-correcting codes are usually distinguished between convolutional codes and block codes: Convolutional codes are processed on a bit-by-bit basis and the Viterbi decoder allows optimal decoding. Block codes are processed on a block-by-block basis. Examples of block codes are Hamming codes, Reed-Solomon codes and Low-Density parity check code (LDPC). A code having a minimum distance ‘ d_{min} ‘ is capable of correcting all patterns of errors ‘ $t = [(d_{min} - 1)/2]$ ‘ or fewer errors in a code word, and is referred to as a random error correcting code.

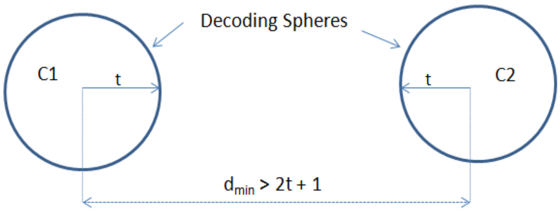

Fig 2 : Error Detection & Correction using Decoding Spheres

In block codes, to detect ‘t’ errors per block, the minimum distance of the block code should be $d_{min} \geq t+1$. i.e $t \leq d_{min} - 1$ errors can be detected. If the number of errors is equal to dmin, then one of the two codewords which form the pair for minimum distance might get changed into the other codeword. Hence the number of errors less than d_{min}, can be detected, because it will not make any codeword into another codeword. It takes d_{min} changes to transform from one codeword to another codeword. Similarly, to correct ‘t’ errors per block, we must have $d_{min} \geq 2t+1$. Considering the above figure, the codewords C1 and C2 are spatially separated by the minimum distance d_{min}. i.e, if the number of errors occurs in C2 is equal to d_{min}, then C2 is transformed to C1. If ‘t’ errors make the decoding sphere boundary around the codewords C1 and C2, and if the circumference of these two spheres touch each other, the $d_{min} = 2t$. In this case, if any ‘t’ error occurs, the erroneous codeword can be mapped to the codeword corresponding to the decoding sphere. If the erroneous codeword is exactly ‘t’ distance from C1 and ‘t’ distance from C2, the correct decision can not be made. To avoid this ambiguity, the condition for correcting 't' errors per block should be $d_{min} \geq 2t + 1$.When these two decoding spheres intersects, the decoding will be incorrect. In order to increase the error correcting capability of a block code, the minimum distance should be larger.

Bit interleaving is a well-known technique for dispersing the errors that occur in burst when the received signal level fades, and which are likely to exceed the correcting capability of a code. Before a message is transmitted, the entire bit stream is interleaved. By interleaving, the burst mode error is converted to dispersed error. Hence, the coding schemes designed to correct random or dispersed error can be used to correct burst mode error with interleaving. Note that the interleaving process does not involve adding redundancy. Concatenated coding schemes are used to provide even more protection against bit errors than is possible with a single coding scheme.

MIMO Communication

In wireless communication, multiple antennas can be utilized in order to enhance the bit rate, and signal-to-noise-plus-interference ratio of wireless systems. Increased capacity is achieved by introducing multiple spatial channels by diversity at both transmitter and receiver. Sensitivity to fading is reduced by the spatial diversity provided by space-time coding. In a MIMO system, given total transmit power can be divided among multiple spatial paths. For a given fixed bandwidth, there is always a fundamental tradeoff between bandwidth efficiency (high bit rates) and power efficiency (small error rates). Conventional single-antenna transmission techniques aiming at an optimal wireless system performance operate in the time domain and/or in the frequency domain. With MIMO, spatial domain is exploited by using space-time coding to overcome the detrimental effects of multipath fading.

A. Channel Capacity

For a Single-Input Single-Output antenna system, the capacity is given by $C = log(1+SNR)$. With MIMO, the capacity is $C \approx m.log(1+SNR)$, where m is the minimum number of antennas in the transmitter and receiver sides [17].

B. Higher Bit Rates with Spatial Multiplexing

Spatial multiplexing techniques simultaneously transmit independent information sequences, over multiple antennas. Spatial Multiplexing takes the high rate signal and breaks it down to lower rate streams. Using M transmit antennas, the overall bit rate compared to a single-antenna system is thus enhanced by a factor of M without requiring extra bandwidth or extra transmission power. Channel coding is often employed, in order to guarantee a certain error performance. Since the individual data streams are superimposed during transmission, they have to be separated at the receiver using an interference cancellation type of algorithm (typically in conjunction with multiple receive antennas). Spatial multiplexing scheme can be implemented with Bell-Labs Layered Space-Time Architecture (BLAST) [4].

C. Enhancing Error performance through Space-Time coding

Similar to channel coding, multiple antennas can also be used to improve the error rate of a system, by transmitting and/or receiving redundant signals representing the same information sequence. By means of two-dimensional coding in time and space, commonly referred to as space-time coding, the information sequence is spread out over multiple transmit antennas. At the receiver, an appropriate combining of the redundant signals has to be performed. Optionally, multiple receive antennas can be used, in order to further improve the error performance (receiver diversity). The

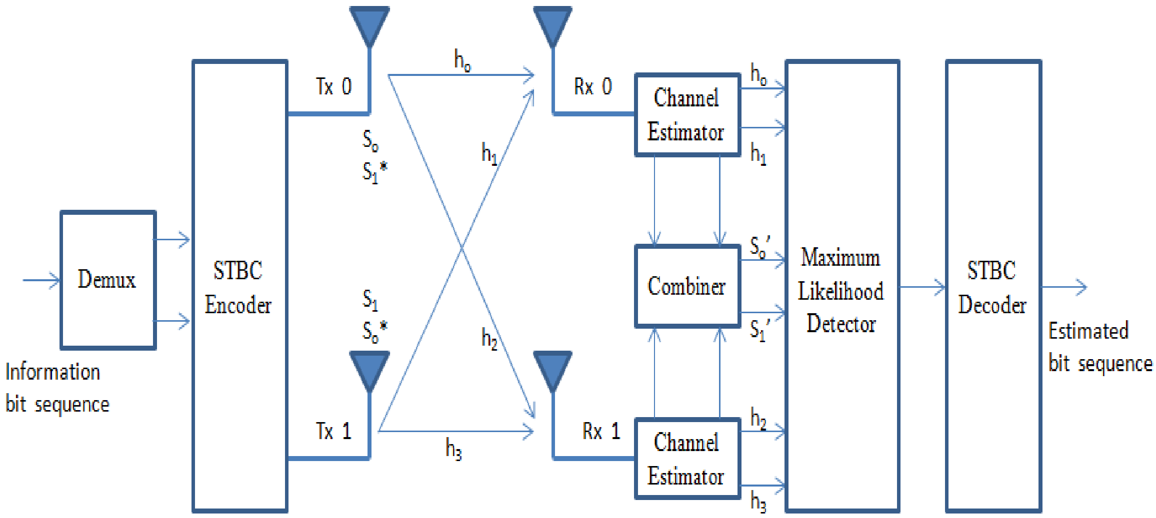

Fig 3 : 2x2 MIMO – STBC Channel

advantage over conventional channel coding is that redundancy can be accommodated in the spatial domain, rather than in time domain. Correspondingly, a diversity gain and a coding gain can be achieved without lowering the effective bit rate compared to single-antenna transmission. Well-known spatial diversity technique for systems with multiple transmit antennas is Alamouti's transmit diversity scheme[12].

Literature Survey

In [6], performance analysis of orthogonal space time block codes (OSTBC) concatenated with channel coding is done by considering quasi-static and block-fading Rayleigh as well as Rician fading. The channel codes considered are convolutional codes, turbo codes, trellis coded modulation (TCM), and multiple trellis coded modulation (MTCM). Union bounds evaluated for convolutional and turbo codes with BPSK modulation, and for 4-state, 8-PSK TCM and MTCM codes. The OSTBC for two-transmit antenna used here is the Alamouti scheme [12]. A random interleaver is placed between the channel code and OSTBC. Convolutional codes, with rate–1/2 is concatenated with Alamouti OSTBC with one receive antenna. With BPSK, the coding gain suffers about 1 dB (at BER= 10^{-5}) with a transmit correlation of ρ_t = 0.7. For a turbo coded Alamouti system, containing a rate- 1/3 code with four-state constituent recursive convolutional codes the degradation due to spatial correlation of $\rho_t = 0.7$ is about 0.8 dB at BER = 10^{-7} . With TCM, 2-Tx and 1-Rx antennas and Rayleigh block fading channel, the performance loss due to spatial correlation of $\rho_t = 0.7$ is about 1.2 dB at BER = 10^{-5}. Under Rician fading, with parameter K = 5dB, and two transmit and one receive antennas the loss due to a spatial correlation of $\rho_t = 0.7$ is around 1dB.

In [10], a study on turbo-space-time coded modulation (turbo – STCM) scheme utilizing parallel concatenated systematic space-time codes (STC) with multilevel modulation and multiple transmit/receive antennas is carried out. The turbo-STCM encoder consists of two systematic recursive space–time component codes and it features full rate. Simulation results are provided for 4 state 4-PSK over the block-fading channel. It is shown that at frame error rate (FER) = 10^{-2}, and for a two transmit-one receive antenna configuration, the performance with recursive codes has an advantage of about 1.3 dB over the configuration with non recursive codes. For the same FER and antenna configuration, turbo-STCM provides an advantage of 2.7 dB over conventional 4-state STCs and 0.6 dB over conventional 32-state space–time codes.

In [3], the performance of convolutional codes (CC), Reed-Solomon (RS) block code as well as concatenated coding schemes that are used to encode the data stream in wireless communications are investigated. It is shown that by concatenating two different codes we can improve the total bit-error rate (BER). RS codes are preferred in correcting burst errors while convolutional codes are good for correcting random errors that are caused due to a fading channel. The simulation results confirm the outperformance of the concatenated codes especially when RS is the outer code and CC is the inner code, compared to when CC as the outer code and RS as the inner code. Due to the fact, that RS codes are best suited for correcting burst errors, total BER of RS-CC has significant coding gain, and it increases as Eb/No increases.

In [1], a concatenated code that achieves full system diversity by appropriately selecting the outer convolutional code (CC) with an inner reduced-rank space-time block code (STBC) using Trellis diagram is proposed. The advantage of the lower rank STBC is that the number of RF chains can be reduced. The number of RF chains considered are 8. For simulation, considered BPSK transmission, and the channel is modeled as i.i.d quasi-static Rayleigh flat-fading channel. The coding gain is largest when the full rank STBC is employed, while it is smallest when no STBC is involved. From a practical point of view, the concatenation of rank four STBC and rate 1/2 CC offers the best trade off since it requires only half the number of RF chains with a small coding gain loss. It is clear that the coding gain increases as the rank of the STBC gets higher. The concatenation of rate 1/2 CC and Alamouti scheme offers the best trade off between coding gain and system complexity (i.e., number of RF chains).

In [5], upper bound on the error probability of low-density parity-check (LDPC) coded modulation schemes operating on Rayleigh and Rician MIMO fading channels are obtained. LDPC code is concatenated with the orthogonal space-time block code (OSTBC). Over single-input single output (SISO) fading channel, the SNR difference between the bit error probabilities and the corresponding upper bound is about 2 - 4 dB. The error probability decreases faster as the Rician factor ‘K’ increases from zero, five to twenty. This is because a Rician channel converges to an AWGN channel as K goes to infinity. The final SNR difference, as K goes up is thus expected to be about 1.5 dB. In the case of concatenated MIMO system with 4PSK and 8QAM modulation, the derived bounds are about 2.5 dB away from the simulation results. The difference decreases to 1.5 dB for the system with four transmit and four receive antennas, where the orthogonal space-time block code is adopted.

Conclusion

The importance of error control coding techniques for MIMO fading channel is reviewed. From literature survey, it is observed that, by considering various coding schemes, under different fading environment, significant enhancement in bit error performance and coding gain can be achieved. It is also observed that, various ways of deploying error correcting codes are not tried exhaustively. This gives us opportunity to explore various ways of deploying error correcting codes. Our aim is to analyse the performance of various coding schemes under MIMO fading environment. The performance parameters to be considered are bit error rate, effective data throughput and coding gain.

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