PERFORMANCE ANALYSIS OF SPACE TIME BLOCK CODES WITH CHANNEL CODING

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CERTIFICATE

I hereby certify that the work which is being presented in this thesis entitled, **"PERFORMANCE ANALYSIS OF SPACE TIME BLOCK CODES WITH CHANNEL CODING,"** in partial fulfilment of requirements for the award of degree of Master of Engineering in Electronics and Communication from Thapar University, Patiala, is an authentic record of my own work carried under the supervision of Dr. Amit Kumar Kohli.

The matter presented in this thesis has not been submitted in any other University or Institute for the award of Master of Engineering.

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ABSTRACT

Wireless systems are rapidly developing to provide high speed multimedia data transfer services that were traditionally offered by wired networks. To support these services, channels with large capacities are required. Theoretical investigations in the past few years have shown that very high capacities can be obtained by employing multiple antenna elements at both the transmitter and the receiver end of a wireless system. In multiple input and multiple output (MIMO) system of wireless communication, Space time block coding (STBC) is one of the attractive technique for high bit-rate and high capacity transmission. The advantage of space-time codes lies in the fact that diversity advantage is obtained by shifting the complexity of adding additional antennas to the transmitter while allowing the complexity of receiver to be reasonably low. Methods for providing transmit diversity in communication over fading channels also been concerned. The advantage of space-time codes lies in the fact that diversity advantage is obtained by shifting the complexity of adding additional antennas to the transmitter while allowing the complexity of receiver to be reasonably low.

Broadband wireless communication must cope with critical performance limiting challenges that include time-selective and frequency-selective channels, as well as power and bandwidth constraints. It has been observed that, STBC technique is not enough to achieve the code rate up to the Shannon limit. In this thesis channel coding is used along with Space time block coding (STBC).

Recently, Convolutional codes (CC) have attracted much attention as the good error correcting codes achieving the near Shannon limit performance. In this thesis the concatenation scheme of STBC codes and CC (STBC-CC) is proposed and it has been shown that the STBC-CC can achieve the good error rate performance and coding gain. Referring to it as the STBC-CC, we evaluate the bit error rate (BER) of the STBC-CC with multiple transmit antennas in Rayleigh fading environments.

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Chapter 1

Introduction

General

In the recent year, the word PORTABLE is the most common word in today’s life. Everyone needs everything should be portable, though wired communication networks can provide the connectivity and better performance but portability is main issue with it, that is why wireless communication is getting much more emphasis and attention than that of its counterpart. But it has its own fundamental issue to overcome i.e. time varying multipath fading, it is this phenomenon which makes wireless transmission a challenge when compared to fiber, coaxial cable, line-of-sight microwave or even satellite transmission. Severe attenuation in a multipath wireless environment makes it extremely difficult for the receiver to determine the transmitted signal [4]. In additive white Gaussian noise (AWGN), using typical modulation and coding schemes, reducing the effective bit error rate (BER) from 10^{-2} to 10^{-3}may require only 1 or 2 dB higher signal to noise ratio (SNR). Achieving the same in multipath fading environment, however, may require up to 10 dB improvement in SNR [2].

Signal is reached to receiver from transmitter from various paths due to reflection, diffraction etc. in wireless communication. Each path has a different attenuation, time delay and phase shift, the signal from different paths add constructively sometimes or destructive sometimes, resulting in fluctuation signal strength. This phenomenon is known as multipath fading. Channel fading cause’s the performance degradation and renders reliable high data rate transmission a challenging problem in 4G wireless communication [14].

Theoretically, the most effective technique to mitigate multi path fading in a wireless channel is transmitter power control. If channel conditions as experienced by the receiver on one side of the link are known at the transmitter on the other side the transmitter can predistort the signal in order to overcome the effect of the channel at the receiver. There are two fundamental problems with this approach. The major problem is the required transmitter dynamic range. For the transmitter to overcome a certain level of fading, it must increase its power by that same level, which in most cases is not practical because of radiation power limitations and the size and cost of the amplifiers. The second problem is that the transmitter does not have any knowledge of the channel experienced by the receiver except in systems where the uplink (remote to base) and downlink (base to remote) transmissions are carried over the same frequency. Increasing the quality or reducing the effective error rate in a multipath fading channel is extremely difficult.

In most scattering environments, antenna diversity is a practical, effective and, hence, a widely applied technique for reducing the effect of multipath fading. It is principally used to combat fading. The basic idea is that if different copies of the same signal are available then there is a high probability that at least one of them is of a good quality [11]. Of course choosing the best copy and rejecting all others is not the optimal solution. This brings us to the problem of choosing the best way to combine all of them.

The popular forms of diversity are:

- Temporal diversity: Replicas of the information bearing signal are transmitted in different time slots, where the separation between the time slots is greater than the coherence of the channel [10].
- Frequency diversity: In this case, replicas of the information bearing signal are transmitted in different frequency bands, where the separation between the frequencies is greater than the coherence bandwidth of the channel.
- Antenna (spatial) diversity: It has been observed that antennas with a spacing of more than half a wavelength leads to spatially uncorrelated channels. The transmission of the replicas of the information bearing signal over these uncorrelated spatial channels leads to spatial diversity [16].

First category uses, implicit or explicit feedback of information from the receiver to the transmitter to configure the transmitter. In time division duplex systems, the same antenna weights are used for reception and transmission, so feedback is implicit in the appeal to channel symmetry. These weights are chosen during reception to maximize the SNR, and during transmission to weight the amplitudes of the transmitted signals.

Explicit feedback includes switched diversity with feedback based on the feedback of the channel response. However, in particular, vehicle movements or interference causes a mismatch between the state of channel perceived by transmitter and that perceived by receiver. Second category is with feed forward without feedback technique. This technique uses linear processing at the transmitter to spread the information across the antennas. At the receiver, information is obtained by either linear processing or maximum likelihood decoding 5 techniques. Feed forward information, is required to estimate the channel from the transmitter to receiver. These estimates are used to compensate for the channel response at the receiver. This scheme was first proposed by Wittneben.

The third category does not require feedback or feed forward information. Instead, it uses multiple transmit antenna combined with channel coding to provide diversity. Example of this approach is to combine phase sweeping transmitter diversity of with channel coding. Another scheme is to encode information by a channel code and transmit the code symbols using different antennas in orthogonal manner. This can be done either by frequency multiplexing, time multiplexing, or by using orthogonal spreading sequences for different antenna. The disadvantage of these schemes over previous two schemes is the loss in bandwidth efficiency due to the use of channel code. Using appropriate coding, it is possible to relax the orthogonality requirement need in these schemes and obtain the diversity as well as coding advantage offer without sacrificing bandwidth. This is possible when the whole system is viewed as a MIMO system and suitable codes are used [2].

In transmit diversity multiple antennas are used to achieve space diversity and to achieve time diversity the transmission is repeated n times. Scope of diversity is expanded by using multiple antennas at both the receiver and transmitter [3]. A major conclusion of these works is that the capacity of multi-antenna systems far exceeds that of a single antenna system. In particular, the capacity grows at least linearly with the number of transmit antennas as long as number of received antenna is greater than or equal to number of transmit antennas. Space-time Codes (STC) were first introduced by Tarokh from AT&T research labs in 1998 as a novel means of providing transmit diversity for the multiple antenna fading channel. There are two main types of STC [2]:

Space-time Trellis Codes (STTCs): These are a type of space–time code used in multiple-antenna wireless communications. This scheme transmits multiple, redundant copies of a trellis (or convolution) code distributed over time and a number of antennas (space). These multiple, 'diverse' copies of the data are used by the receiver to attempt to reconstruct the actual transmitted data. For an STC to be used, there must necessarily be multiple transmit antennas, but only a single receive antennas is required; nevertheless multiple receive antennas are often used since the performance of the system is improved by so doing.

In contrast to space–time block codes (STBCs), they are able to provide both coding gain and diversity gain and have a better bit-error rate performance. However, being based on trellis codes, they are more complex than STBCs to encode and decode; they rely on a Viterbi decoder at the receiver where STBCs need only linear processing. STTCs were discovered by Vahid Tarokh et al. in 1998.

Codes Space-time Block (STBC): Space-time block codes operate on a block of input symbols, producing a matrix output whose columns represent time and rows represents antennas. It contrast to single antenna block codes for the AWGN channel, space-time block codes do not generally provide coding gain, unless concatenated with an outer code. Their main feature [1], [2] is the provision of full diversity with a very simple decoding scheme.

It is found that in the sever condition, the gain obtain by using STBC along is unable to overcome the multipath effect. In order to overcome this one has to use channel coding technique along with STBC. The main objective of this thesis work is to compare performance of STBC with that of STBC with channel coding.

1.1 Over view of channel coding

Coding has the usefulness that it allows us to increase the rate at which information may be transmitted over a channel while maintaining a fixed error rate. Alternatively, coding allows us to reduce the information bit error rate while maintaining a fixed transmission rate. More generally, coding allows us, in principle (up to the Shannon limit) to design a communication system in which both information bit rate and error rate are independently and arbitratarily specified but subject to a constraint on bandwidth [14]. The price we pay, for seeking to reach closer to Shannon limit, is increased hardware complexity both at the transmitter where encoding is done and at the receiver where decoding is affected. In principle, with ingenious enough coding and unlimited complexity we would be able to reach the Shannon limit. That is, we would be able to transmit at channel capacity and with an error rate which may be made as small as desired. One measure of the efficiency of a code is precisely the extent to which it allows us to approach the Shannon limit. In this thesis Convolutional coder is used for the channel coding purpose

1.2 Convolutional coding

Convolution encoding with Viterbi decoding is a FEC technique that is particularly suited to a channel in which the transmitted signal is corrupted mainly by additive white Gaussian noise (AWGN). Convolution codes are usually described using two parameters: the code rate and the constraint length. The code rate, K/n, is expressed as a ratio of the number of bits into the convolution encoder (k) to the number of channel symbols output by the convolution encoder (n) in given encoder cycle. A Convolutional code introduces redundant bits into the data stream through the use of linear shift register. The information bits are input into shift register and the output encoded bits are obtained by modulo-2 addition of the input information bits and the contents of the shift register. Further we will discuss in details in chapter 2

1.3 Objective of thesis

- Performance analysis of Space Time Block Code along with Channel coding

1.4 Organisation of thesis

In Chapter 2 Discussed about channel coding and BPSK modulation technique

In chapter 3 Discussed diversity technique, Rayleigh channel, AWGN channel, Doppler shift

In chapter 4 Discussed STBC (Space Time Block Code),STBC encoder/decoder and STBC matrix notation

In chapter 5 Simulation results and discussion

In chapter 6 Conclusion and future scope

Chapter 2

Channel coding

Introduction

The purpose of forward correction code (FEC) is to improve the capacity of channel by adding some carefully designed redundant information to the data being transmitted through the channel [15]. The process of adding this redundant information is known as channel coding. Convolution coding and block coding are the two major forms of channel coding. Convolution codes operate on serial data or few bits block. Convolution encoding with Viterbi decoding is a FEC technique that is particularly suited to a channel in which the transmitted signal is corrupted mainly by additive white Gaussian noise (AWGN). Convolution codes are usually described using two parameters: the code rate and the constraint length[14]. The code rate, K/n, is expressed as a ratio of the number of bits into the convolution encoder (k) to the number of channel symbols output by the convolution encoder (n) in given encoder cycle.

2.1 Proposed system, STBC using channel coding:

Why use channel coding?

- The channel is not ideal.
- Propagation can introduces errors to the signal caused by path loss and path fading
- So channel coding is introduced to overcome these problems.

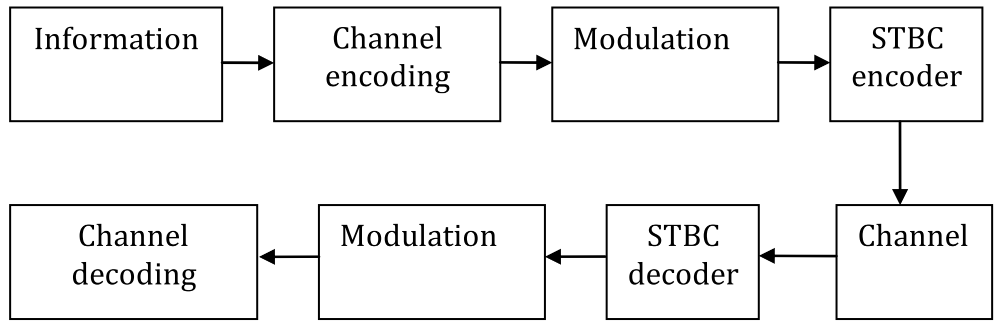


Fig. 2.1: Proposed system STBC using channel coding

2.2 Convolution encoder representation

A convolutional code introduces redundant bits into the data stream through the use of linear shift register. The information bits are input into shift register and the output encoded bits are obtained by modulo-2 addition of the input information bits and the contents of the shift register [14]. Further we will discuss in details. To describe a convolution code, one needs to characterize the encoding function G(m), so that given an input sequence m, one can readily compute the output sequence U. Several methods are used for representing a Convolutional encoder, the most popular being the connection pictorial, connection vectors and polynomials, the state diagram, the tree diagram.

The encoder can be represented in several different but equivalent ways representation [14]. They are .

- Generator/connection Representation
- State Diagram Representation
- Tree Diagram Representation
- Trellis Diagram Representation

2.2.1 Connection representation

We shall use the convolution encoder, shown in fig. 2.2, as a model for discussing convolution encoders. The figure illustrates a (2, 1) convolution encoder with constraints length $k = 3$. There are $n = 2$ modulo-2 adders; for every binary digit that enters the encoder, two code digits are outputs. Hence $code\ rate = 1/2$. At each input bit time, a bit is shifted into the left most stage and bits in the register are shifted one position to the right. Next, output switch samples the output of each modulo-2 adder (i.e., first the upper adder then the lower adder), thus forming the code symbol pair making up the branch word associated with the bit just inputted. The sampling is repeated for each inputted bit. The choice of connections between the adders and the stages of the register gives rise to the characteristics of the code. Any change of the choice of connections result in a different code. The connections are of course, not chosen or changed arbitrarily. The problems of choosing connections to yield good distance properties are complicated and have not been solved in general.

Unlike a block code that has a fixed word length n, a convolution code has no particular size. However, convolution codes are often forced into a block structure by periodic truncation. This requires a number of zero bits to be appended to the end of input data sequence, for the purpose of clearing of flushing the encoding shift register of the data bits. Since the added zeros carry no information, the effective code rate falls below k/n. To keep the code rate close to k/n, the truncation period is generally made as long as practical.

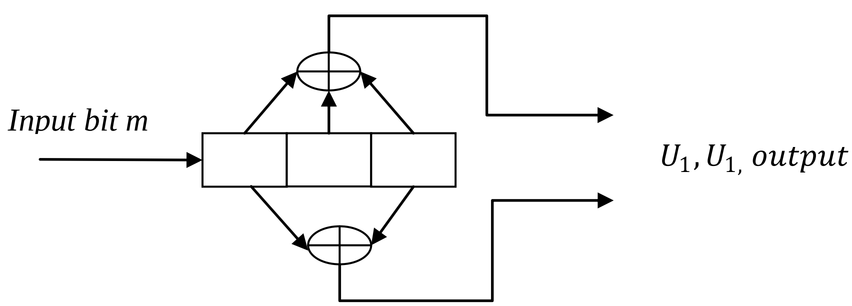


Fig. 2.2: Convolutional encoder [14]

2.2.2 A Polynomial representation

Sometimes, the encoder connections are characterized by generator polynomials. We can represent a convolution encoder bit a set of n generator polynomials, one each of the n modulo-2 adders. Each polynomial is of degree $(k - 1)$ or less and describes the connection of the encoding shift register to that modulo-2 adder, much the same way that a connection vector does. The coefficient of each term in the $(k-1)$ degree polynomial is either 0 or 1 depending on whether a connection exists or does not exist between shift register and modulo-2 adder in question [13], [14].

$$g_1 = 1\ 1\ 1$$
$$g_2 = 1\ 0\ 1$$

$$g_1(X) = 1 + X + X^2 \tag{2.1}$$

$$g_2(X) = 1 + X^2 \tag{2.2}$$

The output sequence is found as follows:

$U(X) = m(X)\, g_1(X)$ interlaced with $m(X)\, g_2(X)$. Message vector $m = 1\ 0\ 1$ as a polynomial that is, $m(X) = 1 + X^2$

$$m(X)g_1(X) = (1 + X^2)\,(1 + X + X^2) = 1 + X + X^3 + X^4$$

$$m(X)\, g_2\,(X) = (1 + X^2)\ \ (1 + X^2 = 1 + X^4$$

$$m(X)g_1(X) = 1 + X + 0\,X^2 + X^3 + X^4$$

$$m(X)g_2\,(X) = 1 + 0\,X + 0\,X^2 + 0\,X^3 + X^4$$

$$U(X) = (1{,}1) + (1{,}0)X + (0{,}0)X^2 + (1{,}0)X^3 + (1{,}1)X^4$$

$$U = 1\,1 \quad 1\,0 \quad 0\,0 \quad 1\,0 \quad 1\,1$$

In this example we started with another point of view- namely, that the convolution encoder can be treated as a set of cyclic code shift register. We represented the encoder with polynomial generators as used for describing code.

2.2.3 **A State representation and the state diagram**

A convolution encoder belongs to a class of devices known as finite state machine, which is the general name given to machine that have a memory of past signals. The adjective finite refers to the fact that there are only a finite number of unique states that the machine can encounter [16]. What is meant by the state of a finite state machine? In the most general sense, the state consists of the smallest amount of information that, together with a current input to the machine, can predict the output of the machine. The state provides some knowledge of the past signalling events and the restricted set of possible output in the future. A future state is restricted by past state. For a state 1/n convolution encoder, the state is represented by the contents of the rightmost $k-1$ stages. Knowledge of the state together with knowledge of the next input is necessary and sufficient to determine the next output.

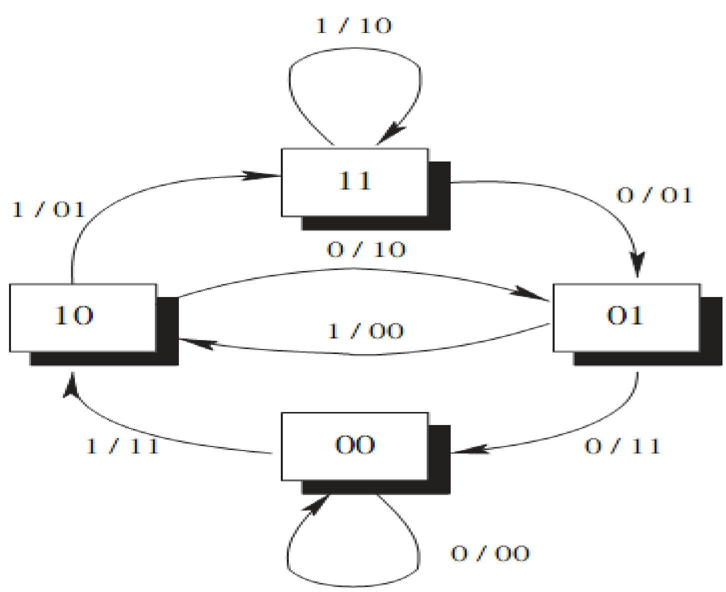


Fig. 2.3: State Representation and State Diagram [14]

Let the state of the encoder at time t_i be defined as $X_i = m_{i-1}, m_{i-2} \dots , m_i$. The *ith* codeword branch U_i is completely determined by state X_i and the present input bit m_i; thus the state X_i represents the past history of the encoder in determining the encoder output. The encoder state is said to be Markov, in the sense that the probabilities $p(X_{i+1}|X_i, X_{i-1 \dots\dots X_0})$ of being in state X_{i+1}, given all previous states, depends only on the most recent state X_i; that is the ,the probability is equal to $p(X_{i+1}|X_i)$. One way to represent simple encoder is with a state diagram; such a representation for the encoder in fig. 2.2 is shown in fig. 2.3. The states, shown in the boxes of the diagram, represent the possible contents of the right most $k-1$ stages of the register, and the path between the states represent the output branch words resulting from such state transition. The state of the register are designated $a = 00, b = 10,\ c = 01,\ d = 11$; the diagram shown in fig. 2.3 illustrates all the state transitions that are possible for the encoder in fig. 2.2. There are only two transitions emanating from each state, corresponding to the two possible input bits. Next to each path between states is written the output branch word associated with the state transition. In drawing the path, we use the convention that a solid lines denotes a path associated with the input bit, zero and a dashed line denotes a path associated with an input bit one. Notice that it is not possible in single transition to move from a given state to any arbitrary state. As a consequence of shifting in one bit at a time, there are only two possible state transitions that the register can make at each bit time. For example, the present encoder state is 00, the only possibilities for the state at the next shift are 00.

Example of convolution encoding:

For the encoder shown in fig. 2.2, show the state changes and the resulting output codeword sequence U for the message sequence $m = 11011$, followed by $k - 1 = 2$ zeros the flush the register. Assume that the initial contents of the register are all zeros [14].

Solution:

Table 2.1: Convolution encoding

Input bit m	Register contents	State at time t_i	State at time t_{i+1}	U_1	U_2
-	000	00	00	-	-
1	100	00	10	1	1
1	110	10	11	0	1
0	011	11	01	0	1
1	101	01	10	0	0
1	110	10	11	0	1
0	011	11	01	0	1
0	001	01	00	1	1

Output sequence: U = 11 01 01 00 01 01 11

2.2.4 Tree diagram

Although the state diagram completely characterizes the encoder, one cannot easily use it for tracking the encoder transitions as a function of time since the diagram cannot represent time history. The tree diagram adds the dimensions of time to the state of diagram. The tree diagram for the Convolutional encoder shown in fig. 2.2 is illustrated in fig. 2.4 [14]. At each successive input bit time the encoding procedure can be described by traversing the diagram from left to right, each tree branch describing an output branch word. The branching rule for finding a codeword sequence is as follows. If the input bit is zero, its associated branch word is found by moving to the next right most branches in the upward direction. If input bit is one, its branch word is found by moving to the next rightmost branch in the downward direction. Assuming that the initial contents of the encoder are all zeros. The diagram shows that if input bit is zero, the output branch is 00 and if the first input bit is one, the output branch is 11 Similarly, if the first input bit is one and the second input bit is a zero, the second output branch word is 10 or the first input bit is a one and second input bit is one, second output branch is 01. Following procedure we see that the input sequence 11011 traces the heavy line drawn on the tree diagram in fig 2.4. This path corresponds to the output code word sequence 1101010001.

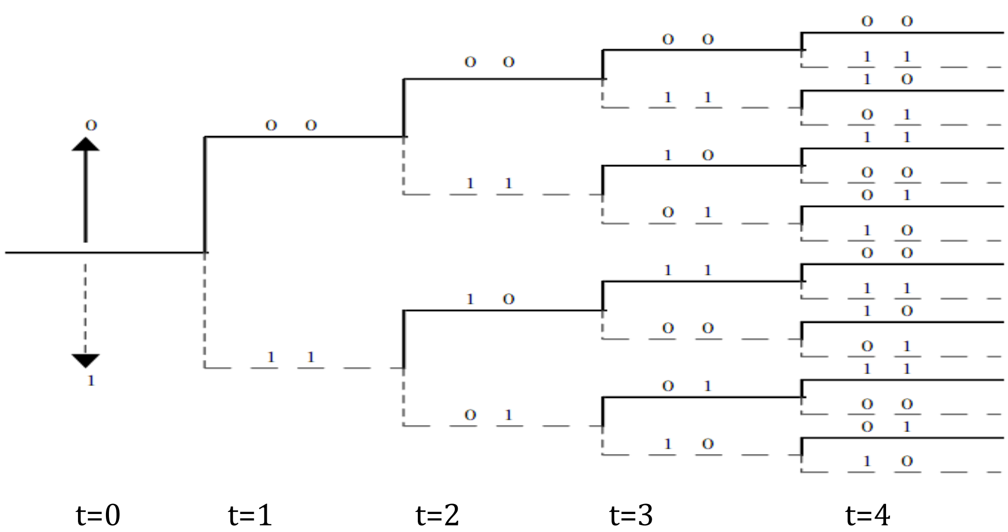


Fig. 2.4: Tree diagram [16]

2.2.5 Trellis diagram

Observation of the fig. 2.4 tree diagram shows that for this example, the structure repeats itself at time t_4, after third branching (in general the tree structure repeats after k branching, where k is the constraint length). We label each node in the tree of fig. 2.4 to correspond to the four possible states in the shift register, as follows: $a = 00$, $b = 10$, $c = 01$ and $d = 11$. The first branching of the tree structure, at time t_1, produces a pair of nodes labelled a and b. At each successive branching the number of nodes double. The second branching at time t_2, results in four nodes labelled a, b, c, and d. After the third branching, there are total of eight nodes: two are labelled a, two are labelled b, two are labelled c, and two are labelled d. We can see that all branches emanating from two nodes of the same state generate identical branch word sequences. From this point on, the upper and lower halves of the tree are identical. The reason for this should be obvious from examination of encoder in fig. 2.2. As the fourth input bit enters the encoder on left, the first input bit is ejected on the right and no longer influences the output branch words. Consequently, the input sequences 1 0 0 x y .... and 0 0 0 x y ..., where the left most bit is the earliest bit, generates the same branch words after the $(k = 3)rd$ branching. This means that any two nodes having the same state label at the same time t_i can be merged, since all succeeding path will be indistinguishable [14],[15],[16]. If we do this to tree structure of fig. 2.2, we obtain another diagram, called the trellis diagram. The trellis diagram, by exploiting the repetitive structure, provides a more manageable encoder description than does the tree diagram. The trellis diagram for the Convolutional encoder of fig. 2.2 is shown in fig. 2.5.

In drawing the trellis diagram, we use the same convention that we introduced with the state diagram, a solid line denotes the output generated by an input bit zero, and a dashed line denotes the output generated by an input bit one [8]. The nodes of the trellis characterize the encoder states; the first row nodes correspond to the state a= 00, the second and subsequent rows correspond to the states b= 10, c= 01 and d= 11. At each unit of time, the trellis requires 2^{k-1} nodes to represent the 2^{k-1} possible encoder states. The trellis in our example assumes a fixed periodic structure after trellis depth 3 is reached (at time t_4). In general case, the fixed structure prevails after depth k is reached. At this point and there after each of the state can be entered from either of two preceding states. Also each of the states can transition to one of the two states. Of the two outgoing branches, one corresponds to an input bit zero and the other corresponds to an input bit one. On fig. 2.5, the output branch words corresponding to the state transitions appear as labels on the trellis branches.

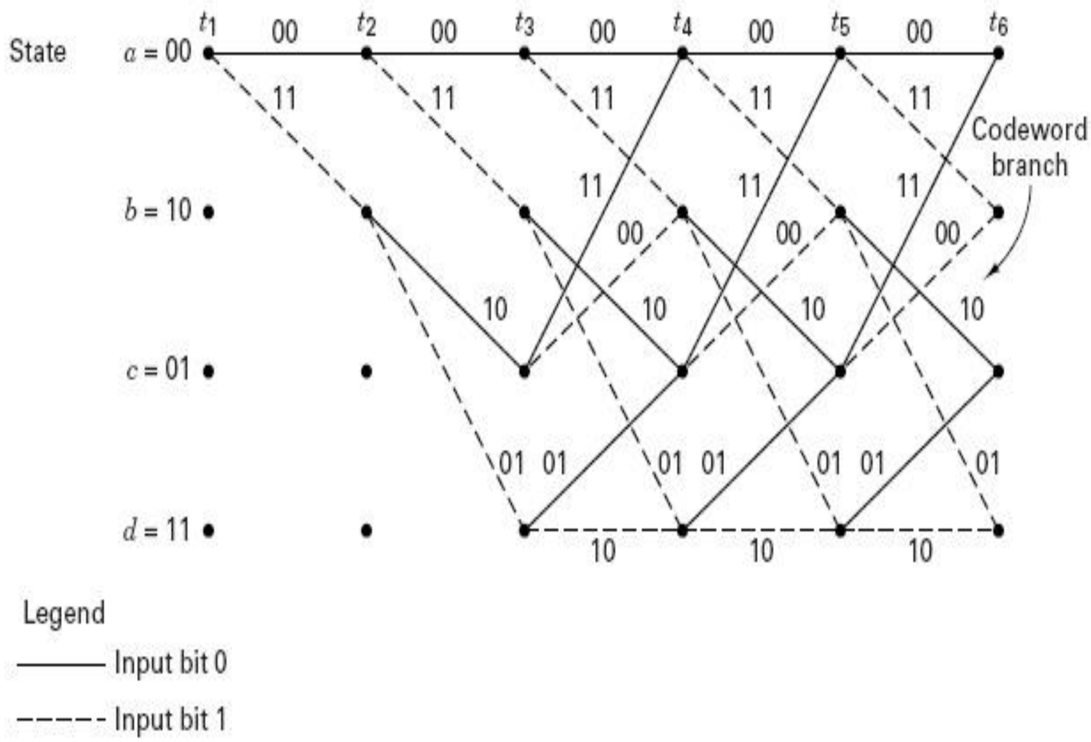


Fig. 2.5: Trellis diagram [16]

2.3 Viterbi Convolutional decoding algorithm:

Viterbi decoder can be represented in three parts

1. Find out the Hamming distance of each path.
2. Add the distance of each path.
3. Compare and select, the shortest path.

The Viterbi decoding algorithm was discovered and analyzed by Viterbi in 1967. The Viterbi algorithm essentially perform maximum likelihood decoding; however, it reduces the computational load by taking advantage of the special structure in the code trellis.

The advantage of Viterbi decoding, compared with brute force decoding, is that the complexity of a Viterbi decoder is not a function of the number of symbols in the codeword sequence. The algorithm involves calculating a measure of similarity or distance, between the received signal, at time t_i, and all the trellis paths entering each state at time t_i. The Viterbi algorithm removes from consideration those trellis path that could not possibly be candidates for the maximum likelihood choice [12] [18].

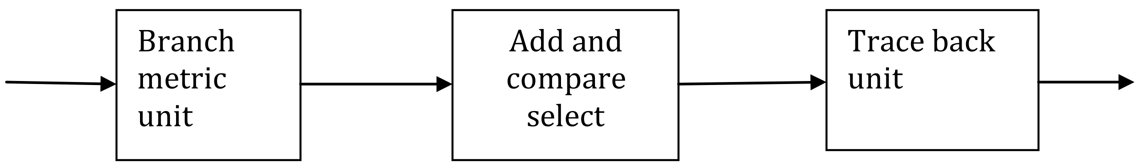


Fig. 2.6: Viterbi decoder

When two paths enter the same state, the one having the best metric is chosen; this path is called the surviving path. This selection of surviving path is performed for all the states. The decoder continues in this way to advance deeper into the trellis, making decision by eliminating the least likely path. The Viterbi algorithm is in fact maximum likelihood. Note that the goal of selecting the optimum path can be expressed equivalently, as choosing the codeword with the maximum likelihood metric or as choosing the codeword with the minimum distance metric [18].

2.3.1 An example of Viterbi Convolutional decoding

A similar Trellis can be used to represent the decoder. The basic idea behind the decoding procedure can best be understood by examining the fig. 2.5 for decoder trellis it is convenient at each time interval, to label each branch with the hamming distance between the received code symbols and the branch word corresponding to the same branch from the encoder trellis. A message sequence m, the corresponding codeword sequence U, and the noise corrupted received sequence $Z =$ 11 01 01 10 01 … …. .The branch words seen on the encoder trellis branches characterize the encoder in fig. 2.2 and are known a priori to both the encoder and the decoder. These encoder branch words are the code symbols that would be expected to come from the encoder output as a result of each of the state transition, the labels on the decoder trellis branches are accumulated by the encoder on the fly. That is as the code symbols are received. In order to label the decoder branches at (departing) time t_1 with the appropriate Hamming distance metric we look at the fig. 2.5 encoder trellis [17]. Here we see the state 00→ 00 transition yields an output branch word of 00 but we received 11. There for, on the decoder trellis we label the state 00→ 00 transition with Hamming distance between them, namely 2. Looking at the encoder trellis again we see that a state 00→ 10 transition yields an output branch of 11, which corresponds exactly with the code symbols we received at time t_1. There for, on the decoder trellis, we label the state 00→ 10 transition with hamming distance of 0. In summary, the metric entered on a decoder trellis branch represents the difference (distance) between what was received and what should have been received had the branch word associated with that branch been transmitted. In effect these metric describe a correlation like measure between a received branch word and each of the candidate branch words fig. 2.6. In receive sequence, black circle shows, it is wrong data.

Table 2.2: Transmitted/ Received sequence

Input seq.. , m	1	1	0	1	1
Transmitted codeword, U	11	01	01	00	01
Received seq.. , Z	11	01	01	10	01

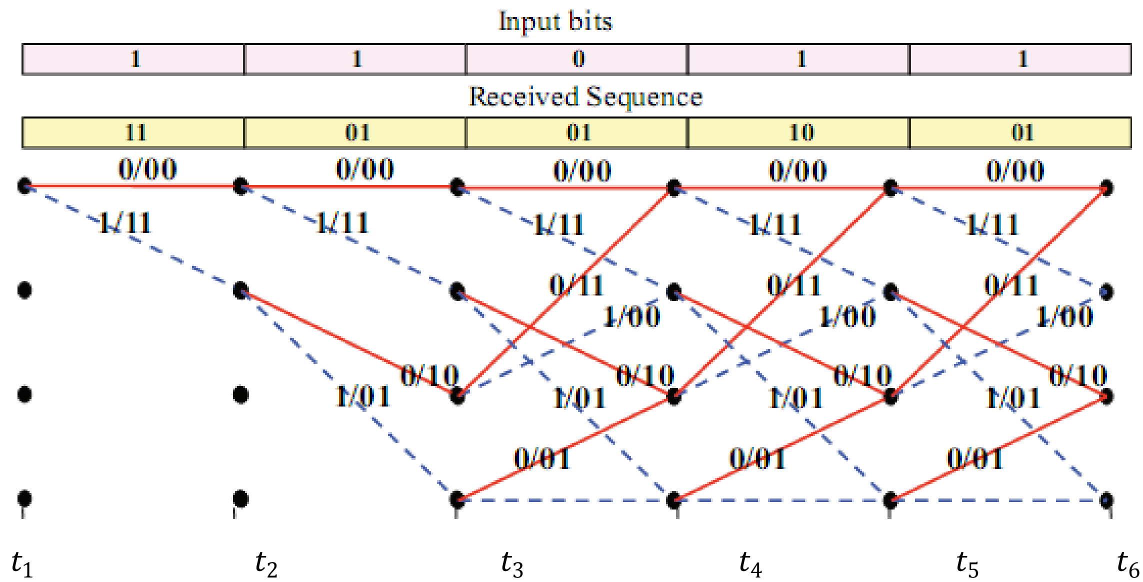


Fig. 2.7: Viterbi decoder1

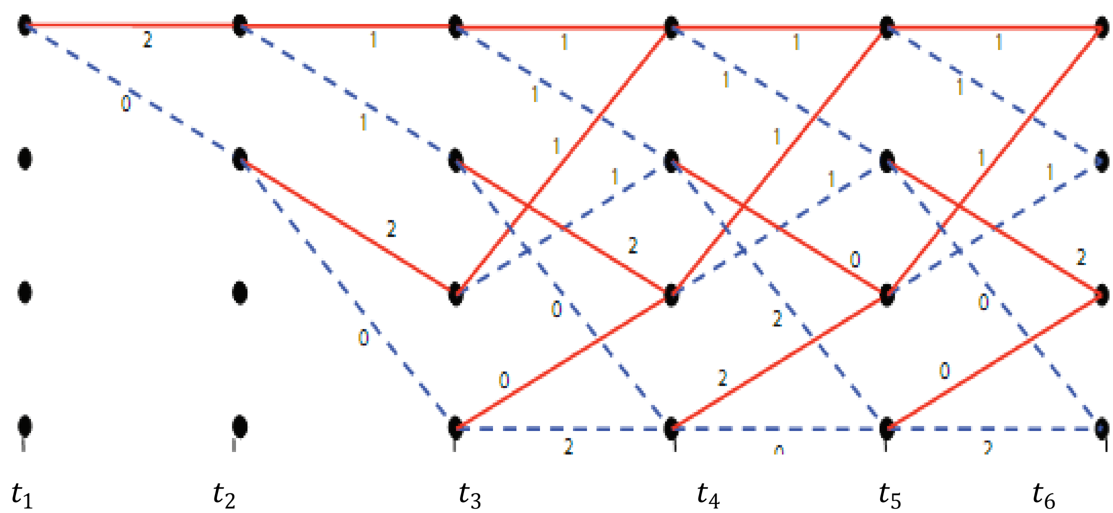


Fig. 2.8: Viterbi decoder2

The basis of Viterbi decoding is the following observation: if any two path in the trellis merge to a single state, one of them can always be eliminated in the search for an optimum path, for example fig. 2.6 shows two path merging at time t_5 to state 00, let us define the cumulative Hamming path metric of a given path at time t_i as the sum of the branch hamming distance metric along that path up to time t_i. In fig. 2.6 the upper path has metric 4; the lower has metric 1. The upper path cannot be a portion of the optimum path because the lower path which enter the same state, has a lower metric [14], [16], [17].

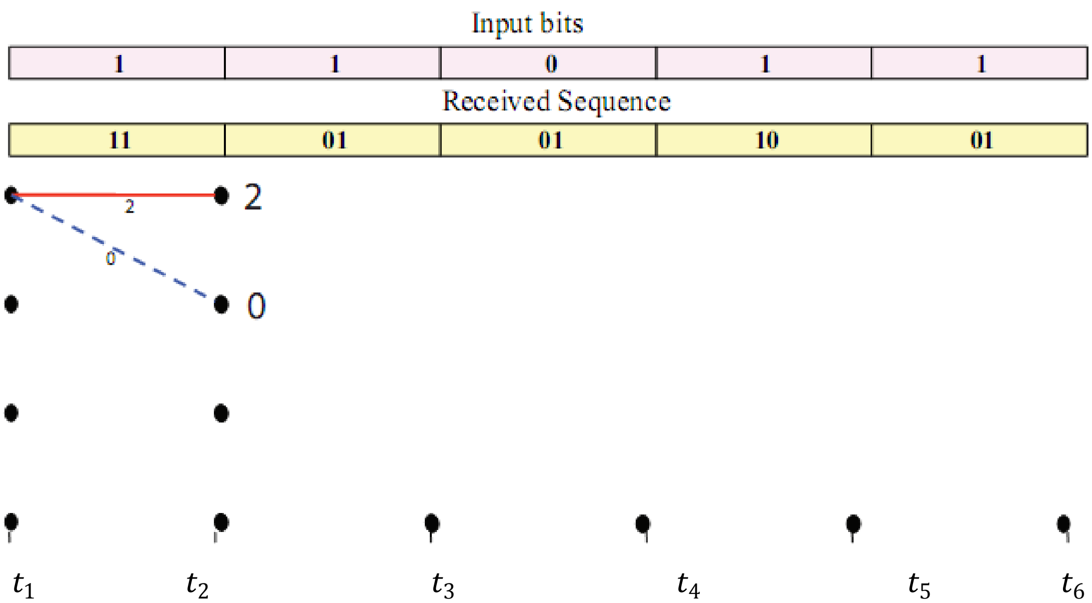


Fig. 2.9: Viterbi decoder3

Assume that the input data sequence m, codeword U, and received sequence Z are as shown in fig. 2.5. Assume that that the decoder knows the correct initial state of the trellis (this assumption is not necessary in practice, but simplifies the explanation.) at time t_1 the received code symbols are 11. From state 00 the only possible transition are to state 00 or state 10 ,state 00→ 00 transition has branch metric 2, in similar fashion find out the Hamming distance of each path. At each succeeding step in the decoding process, there will always be two possible paths entering each state, and one of each pair can be eliminated [14]. The decoder continues in this way to advance deeper into the trellis and to make decisions on the input data bits by eliminating all paths but one.

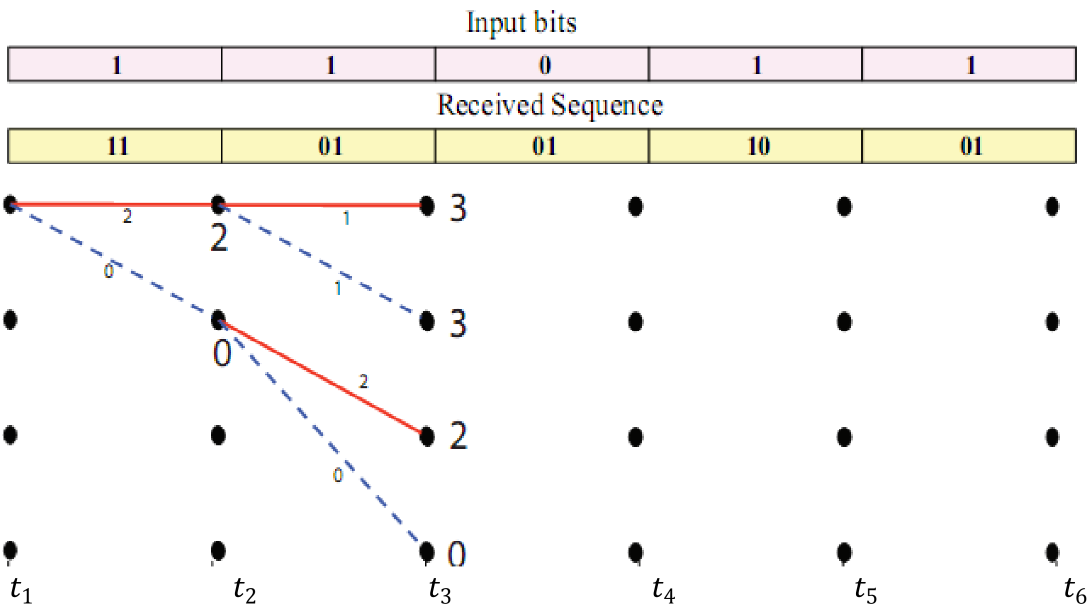


Fig. 2.10: Viterbi decoder

This process continue, after that we will find out the path between t_3 to t_4, t_4 to t_5 and t_5 to t_6

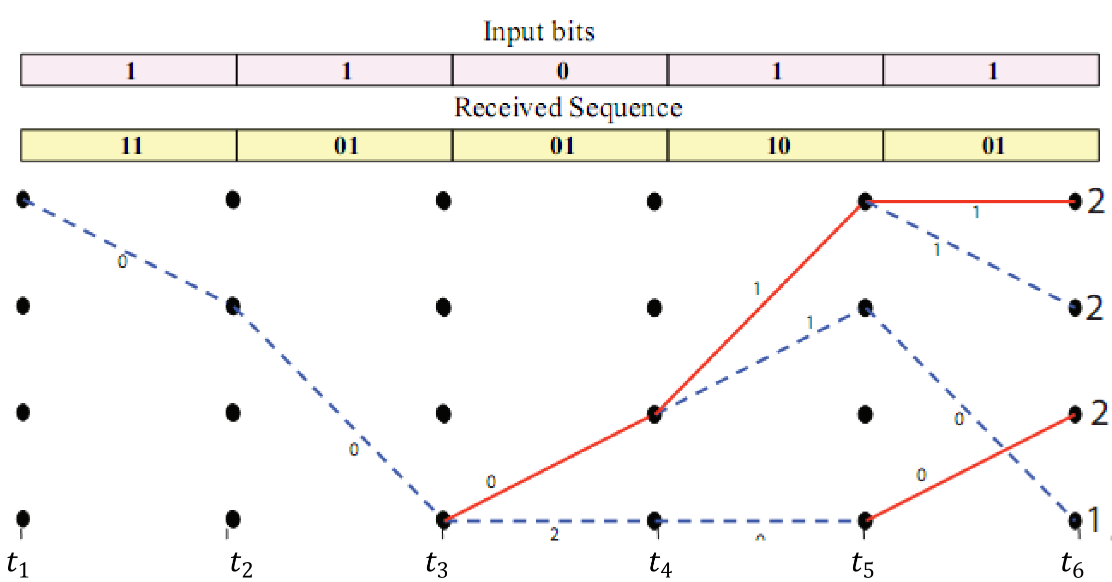


Fig. 2.11: Viterbi decoder5

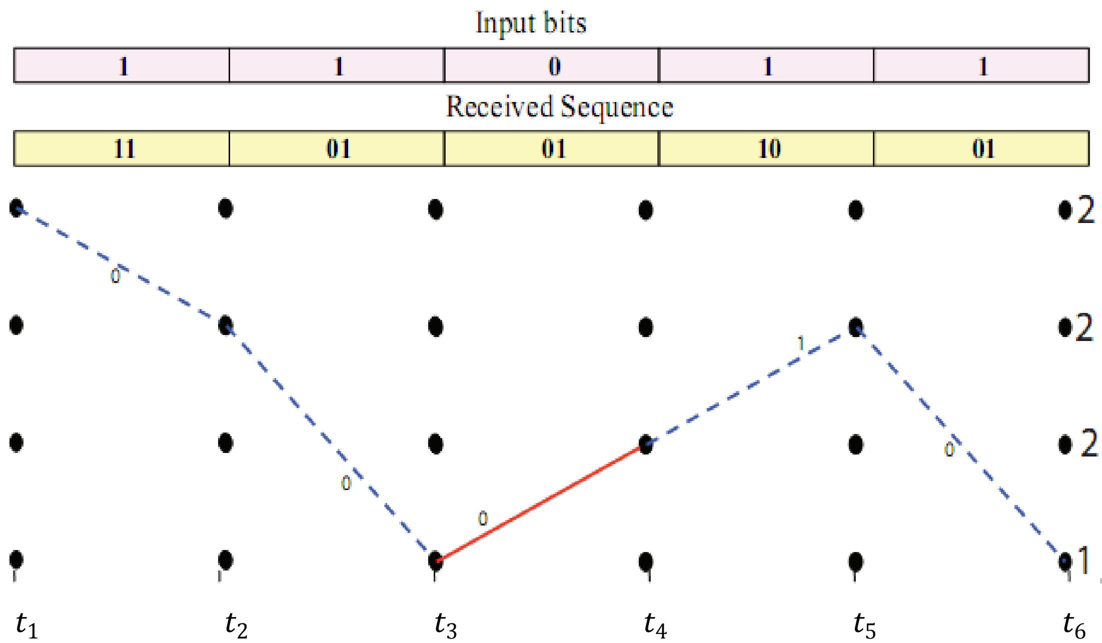


Fig. 2.12: Viterbi decoder6

It is the final step, where we receive the correct information. Actual codeword, in between t_4 and t_5 is 00, but because of noise, data is corrupted and we are receiving 10, by using the Viterbi decoder we can remove this error.

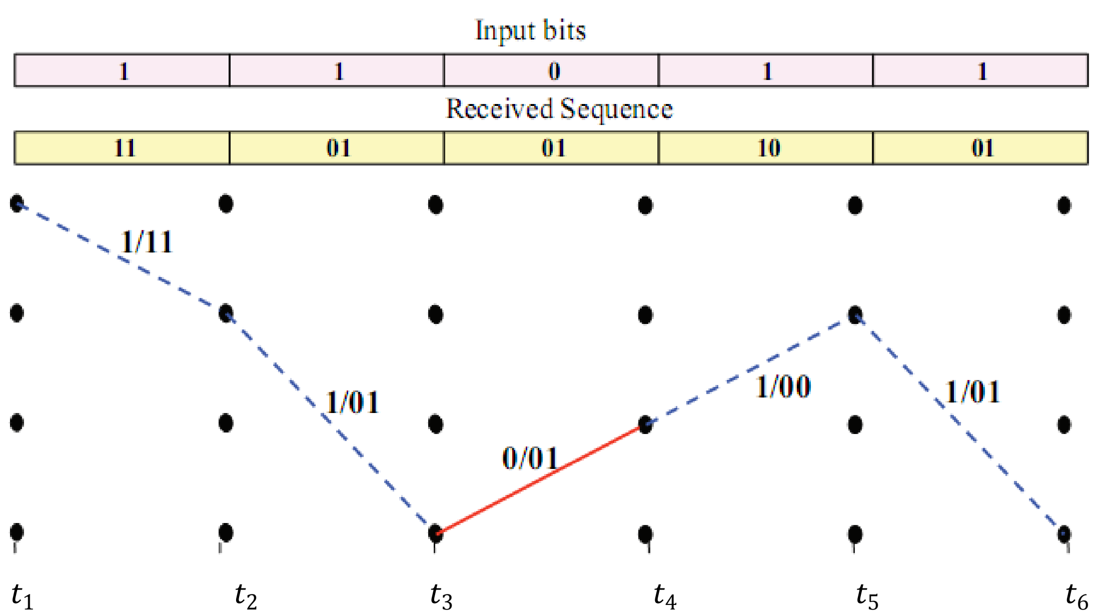


Fig. 2.13: Viterbi decoder7

2.3.2 Binary Phase Shift Keying (BPSK):

The phase of the carrier is varied to represent binary ones and zeros which is used to transmit data via changing and modulating of carrier wave is called Phase Shift keying and if the phase shift uses two phases differing by 180 degree to represent binary digits, the modulation is called BPSK [15]. The principle equation of BPSK as following:

$$s(t) = \begin{cases} A\cos(2\pi f_c t) & \text{Binary 1} \\ A\cos(2\pi f_c t + \pi) & \text{Binary 0} \\ A\cos(2\pi f_c t) & \text{Binary 1} \\ A\cos(2\pi f_c t) & \text{Binary 0} \end{cases}$$

Where

A = constant

f_c = the carrier frequency

t = the bit duration

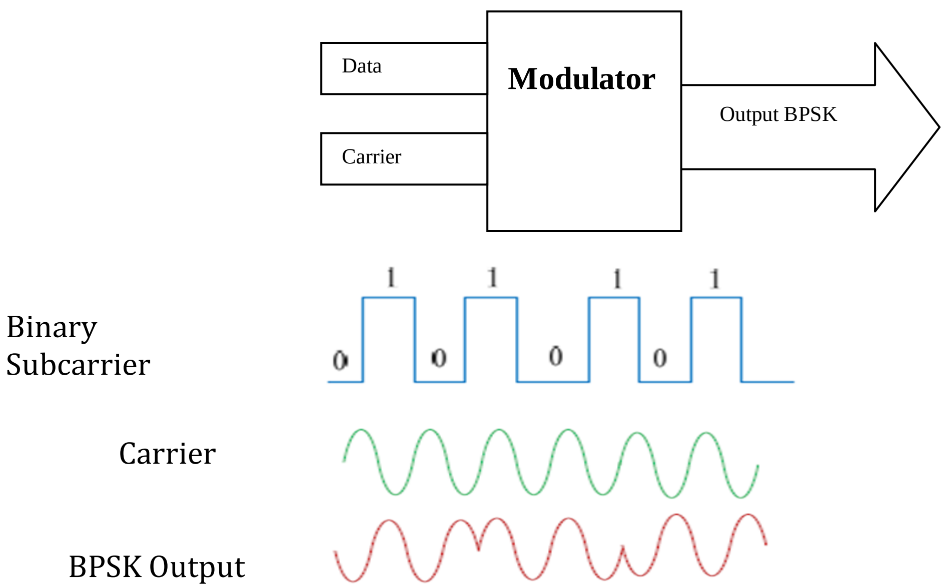


Fig. 2.14: Block diagram of BPSK

Chapter 3
SPACE TIME CODING

Introduction

An effective and practical way to approaching the capacity of multiple-input multiple output (MIMO) wireless channels is to employ space-time (ST) coding. Space-time coding is a coding technique designed for use with multiple transmit antennas. Coding is performed in both spatial and temporal domains to introduce correlation between signals transmitted from various antennas at various time periods. The spatial-temporal correlation is used to exploit the MIMO channel fading and minimize transmission errors at the receiver. Space-time coding can achieve transmit diversity and power gain over spatially uncoded systems without sacrificing the bandwidth [2].

3.1 Multipath Propagation

Multipath is the result of reflection of wireless signals by objects in the environment between the transmitter and receiver. The objects can be anything present on the signal travelling path, i.e. buildings, trees, vehicles, hills or even human beings. Thus, multipath scenario includes random number of received signal from the same transmission source; depending on the location of transmitter and receiver, a direct transmission path referred to as the Line Of Sight (LOS) path may be present or may not be present. When LOS component is present (or when one of the components is much stronger than others), then the environment is modelled as Ricean channel, and when no LOS signal is present, the environment is described as Rayleigh channel. Multi-paths arrive at the receiver with random phase offsets, because each reflected wave follows a different path from transmitter to reach the receiver. The reflected waves interfere with direct LOS wave, which causes a severe degradation of network performance. The resultant is random signal fades as the reflections destructively (or constructively) superimpose one another, which effectively cancels part of signal energy for a brief period of time. The severity of fading will depend on delay spread of the reflected signal, as embodied by their relative phases and their relative power [19].

A common approach to represent the multipath channel is channel impulse response which gives us the delay spread of the channel. Delay spread is the time spread between the arrival of the first and last multipath signal seen by receiver. In a digital system, delay spread can lead to ISI. In fig. 4.1, delay spread amounts to τ_{max} is shown. It is noted that delay spread is always measured with respect to the first arriving component. Let's assume a system transmitting in the time intervals T_{sym}. The longest path with respect to the earliest path arrives at the receiver with a delay of $\tau_{max;}$ in other words, the last path arrives τ_{max} seconds after the first path arrives. This means that a received symbol can theoretically be influenced by previous symbols, which is termed as ISI.

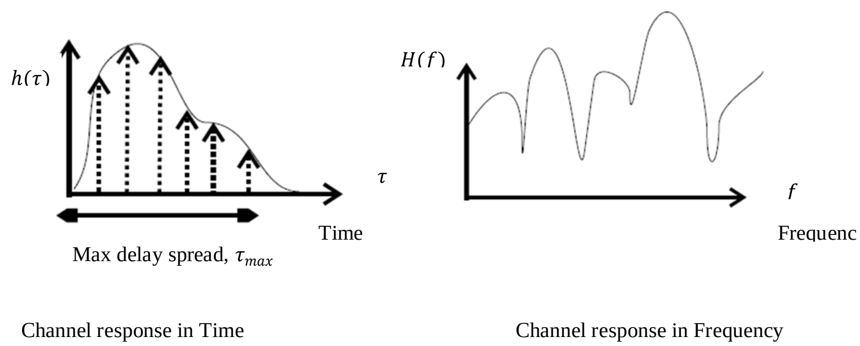


Fig. 3.1: Channel Impulse Responses and Corresponding Frequency Response [13]

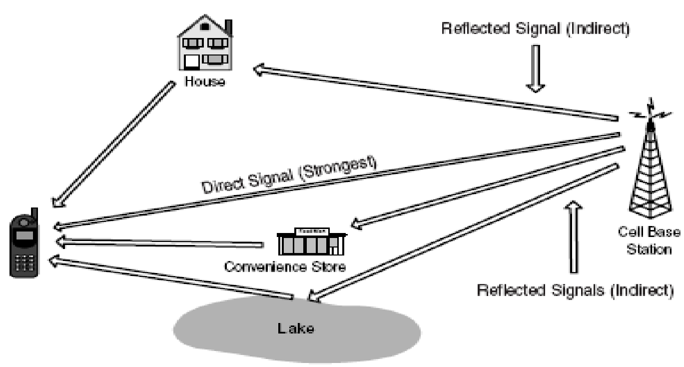


Fig. 3.2: Multipath fading [13]

With high data rate, T_{sym} can be very small; thus the number of symbols that are affected by ISI can be in multiple of tens or more. Combating the influence of such large ISI at the receiver is very challenging and sometimes may become unattainable at very severe channel conditions.

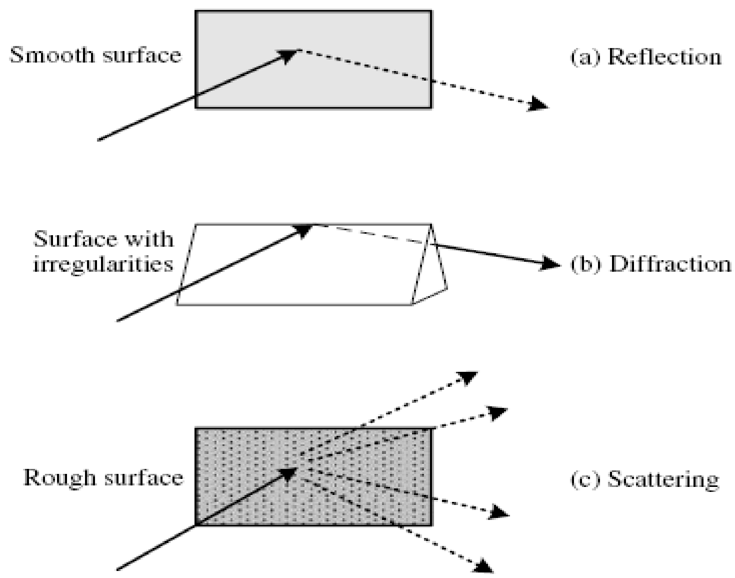


Fig. 3.3: The three major radio propagation mechanisms [13]

3.2 Fading channel features

- N discrete fading paths, each with its own delay and average power gain. A channel for which N = 1 is called a frequency-flat fading channel. A channel for which N > 1 is experienced as a frequency-selective fading channel by a signal of sufficiently wide bandwidth.
- A Rayleigh or Rician model for each path.

3.2.1 Doppler shift

Due to Doppler Effect, if a transmitter is moving away from a receiver, the frequency of the received signal is lower than the one sent out from the transmitter; otherwise, the frequency is increased. In wireless communications, there are many factors that can cause relative movement between a transmitter and a receiver. It can be the movement of a mobile such as a cell phone; it can be the movement of some background objectives, which causes the change of path length between the transmitter and the receiver. The lengths of signal path are often different, which correspond to different movement speeds of transmitter signals, and in turn different frequency shifts on the signal paths. As a result, a frequency spread is caused in the signal spectrum [15].

Corresponding to Doppler spectrum spread, there is a concept called coherence time, which is related to the reciprocal of the maximum Doppler shit. Coherence time is used to measure a time interval, in which a smaller amount of fading is occurred. Specifically, if the baseband signal varies faster than the coherence time, the distortion from Doppler spread fading is negligible. Such a situation is called slow fading. Otherwise, if the baseband signal varies more slowly than the coherence time, the distortion from Doppler spread fading may be significant. This situation is called fast fading [13].

Let us consider mobile user at position x, moving with velocity V, with signal arriving at angleθ. With a nominal carrier frequency of f_0 the received signal is

$$r(t) = Ae^{j(2\pi f_0 t - \beta x cos\theta)} \tag{3.1}$$

with phase (where $\beta = 2\pi/\lambda = \ wave\ number$)

$$\emptyset(t) = 2\pi f_0 t - \beta V t cos\theta \tag{3.2}$$

and resulting Doppler frequency

$$f_D = \left(\frac{V}{C}\lambda\right) cos\theta \tag{3.3}$$

another way: from physics, for a radial (approaching) velocity V_r

$$f_D = f_0 \frac{1}{\frac{c}{V_r} - 1} = \frac{f_0 V_r}{c - V_r} \approx \frac{f_0 V_r}{c} \tag{3.4}$$

but $V_r = V cos\theta$ and $c = f_0 \lambda$ so

$$f_D = \frac{V}{\lambda}\ cos\theta \tag{3.5}$$

often, V and θ will vary with time, so f_D will vary corresponding, causing Doppler spread. Often this is modelled statically. In general, the Doppler spread characterizes the rate of channel variations.

3.2.1.1 Amplitude Variation Due To Motion

Now consider the case where two incoming waves arrives at the mobile at angle of 0 and θ. Assume that they are of equal amplitude. Then, from equation 3.1, the received signal is

$$r(t) = Ae^{j2\pi f_0 t\left(e^{-j\alpha} + e^{-j\alpha\ cos\theta}\right)} \qquad (3.6)$$

where $\alpha = \beta x$ This can be expressed as

$$r(t) = Ae^{j2\pi f_0 t\left(e^{-j\alpha\ (U+D)} + e^{-j\alpha\ (U-D)}\right)} \qquad (3.7)$$

$$U = (1 + cos\theta)/2, D = (1 - cos\theta)/2$$

factoring, we have

$$r(t) = A\ \ e^{j2\pi f_0 t}\ e^{-j\alpha U}\left(e^{-j\alpha\ D} + e^{-j\alpha D}\ \right) \qquad (3.8)$$

notice that

$$e^{\ -j\alpha D} + e^{-j\alpha D} = 2\cos\left[\frac{\alpha(1 - cos\theta)}{2}\right] \qquad (3.9\ )$$

so

$$r(t) = Ae^{j2\pi f_0 t}\ e^{-j\beta x\ \left(\frac{1 + cos\theta}{2}\right)\left(2cos\ \left[\frac{\beta x\ (1 - cos\theta\ )}{2}\right]\right)} \qquad (3.10)$$

thus the phase of signal due to Doppler is

$$\emptyset(t) = \frac{1}{2}\beta x(1 + cos\theta) \qquad (3.11)$$

and the Doppler frequency is

$$f_D = \frac{1}{2\pi}\frac{1}{2}\frac{2\pi}{\lambda}V(1 + cos\theta) = \frac{V}{2\lambda}(1 + cos\theta) \qquad (3.12)$$

but the signal amplitude fluctuates due to multipath according to

$$\Theta_{sw} = \frac{\beta x}{2}(1 - cos\theta) \tag{3.13}$$

$$f_{sw} = \frac{V}{2\lambda}(1 - cos\theta) \tag{3.14}$$

3.2.2 Shadow fading or Shadowing

Shadow fading is another troublesome effect of wireless channel. Wireless signals are obstructed by large obstacles, like huge buildings, high hills, etc. These large objects cause reflections off their surface and attenuation of signals passing through them, resulting in shadowing, or shadow fading. These shadows can result in large areas with high path loss, causing problems with communications. The amount of shadowing depends on the size of the object, the structure of the material, and the frequency of the RF (Radio Frequency) signal. Large attenuations by huge obstacles can result in deep fading behind them. Under this condition, most of the received signal energy comes from reflected and diffracted paths of the original signal, because LOS is absent due to large object between the transmitter and the receiver

3.2.3 Rayleigh fading

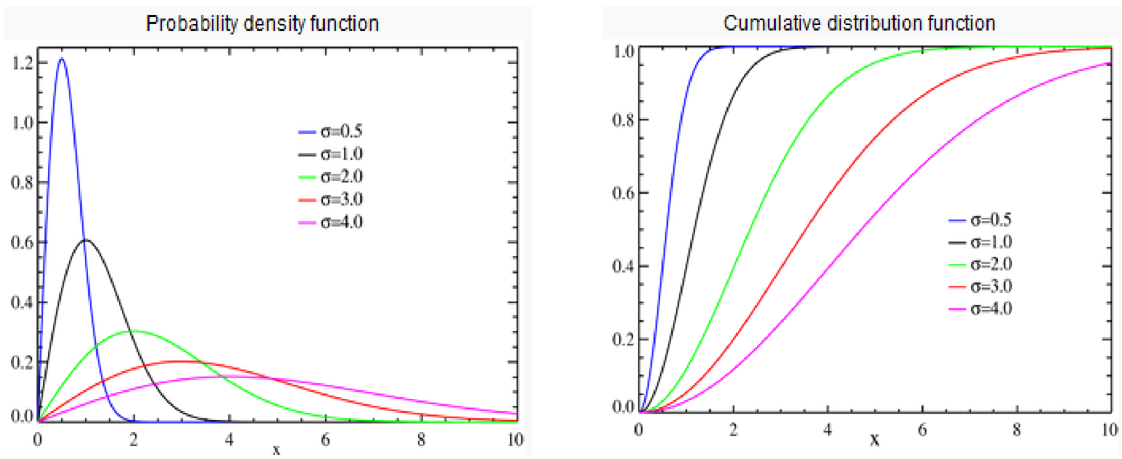
Rayleigh fading is a statistical model for the effect of a propagation environment on a radio signal, such as that used by wireless devices. Rayleigh fading models assume that the magnitude of a signal that has passed through such a transmission medium (also called a communications channel) will vary randomly, or fade, according to a Rayleigh distribution the radial component of the sum of two uncorrelated Gaussian random variables. Rayleigh fading is viewed as a reasonable model for tropospheric and ionospheric signal propagation as well as the effect of heavily built-up urban environments on radio signals. Rayleigh fading is most applicable when there is no dominant propagation along a line of sight between the transmitter and receiver [15].

3.2.3.1 Rayleigh distribution

A Rayleigh random variable has a one sided PDF. The form of the PDF and CDF are given (for any $\sigma > 0$) by

$$f_x(x) = \left(\frac{x}{\sigma^2}\right) \exp\left(-\frac{x^2}{2\sigma^2}\right) u(x) \tag{3.15}$$

$$F_x(x) = 1 - exp\left(-\frac{x^2}{2\sigma^2}\right) \tag{3.16}$$



(a) (b)

Fig. 3.4: (a) PDF (b) CDF of Rayleigh random variable [13]

3.2.4 Additive white Gaussian noise (AWGN)

It is a channel model in which the only impairment to communication is a linear addition of wideband or white noise with a constant spectral density (expressed as watts per hertz of bandwidth) and a Gaussian distribution of amplitude. The model does not account for fading, frequency selectivity, interference, nonlinearity or dispersion. However, it produces simple and tractable mathematical models which are useful for gaining insight into the underlying behavior of a system before these other phenomena are considered. Wideband Gaussian noise comes from many natural sources, such as the thermal vibrations of atoms in conductors (referred to as thermal noise or Johnson-Nyquist noise), shot noise, black body radiation from the earth and other warm objects, and from celestial sources such as the Sun.

The AWGN channel is a good model for many satellite and deep space communication links. It is not a good model for most terrestrial links because of multipath, terrain blocking, interference, etc. However, for terrestrial path modeling, AWGN is commonly used to simulate background noise of the channel under study, in addition to multipath, terrain blocking, interference, ground clutter and self interference that modern radio systems encounter in terrestrial operation.

Basic characteristic of the AWGN channels are

- The noise is additive, i.e., the received signal equals the transmit signal plus some noise, where the noise is statistically independent of the signal.
- The noise is white, i.e., the power spectral density is flat, so the autocorrelation of the noise in time domain is zero for any non-zero time offset.
- The noise samples have a Gaussian distribution.

Additive white Gaussian noise (AWGN) is the commonly used to transmit signal while signals travel from the channel and simulate background noise of channel.

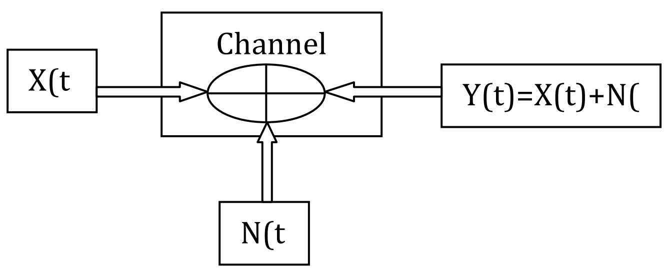


Fig. 3.5: Received Signal through an AWGN channel

The mathematical expression in received signal $Y(t) = X(t) + N(t)$ is shown in fig. 4.2 that passed through the AWGN channel where s(t) is transmitted signal and n(t) is background noise [24]. Where $X(t)$ is the input waveform, regarded as a real random process, and $N(t)$ is a real white Gaussian noise process with single-sided noise power density N_0 which is independent of $X(t)$. Moreover, the input $X(t)$ is assumed to be both power-limited and band-limited. The average input power of the input waveform $X(t)$ is limited to some constant P. The channel band B is a positive-frequency interval with bandwidth W Hz. The channel is said to be baseband if $B = [0, W]$, and passband otherwise. The (positive-frequency) support of the Fourier transform of any sample function $X(t)$ of the input process $X(t)$ is limited to B. The signal-to-noise ratio SNR of this channel model is then

$$SNR = P/N_0W \tag{3.17}$$

where N_0W is the total noise power in the band B. The parameter N_0 is defined by convention to make this relationship true; i.e. N_0 is the noise power per positive-frequency Hz. Therefore the double-sided power spectral density of $N(t)$must be $S_{nn}(f) = \frac{N_0}{2}$, at least over the bands $\pm B$.

3.2.4.1 Gaussian distribution

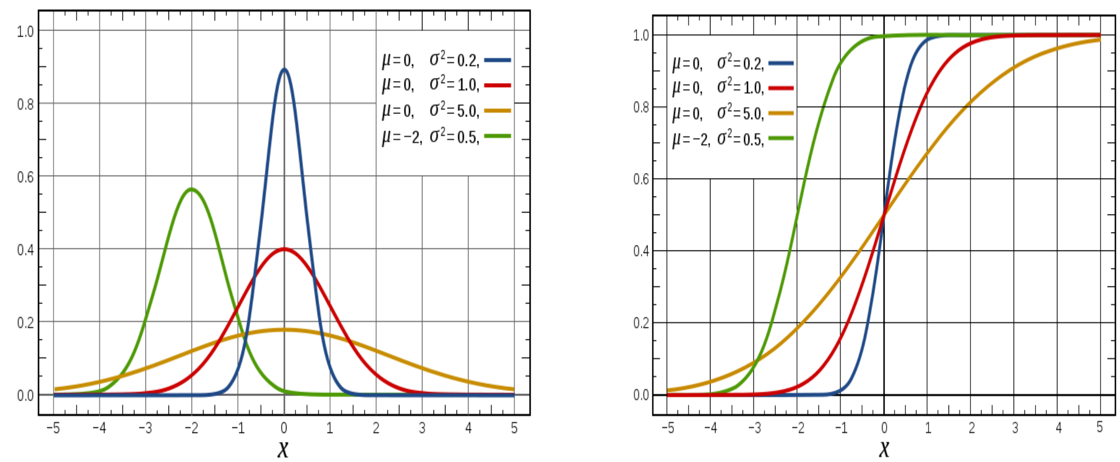
A Gaussian random variable is one whose probability density function can be written in the general form

$$f_x(x) = \frac{1}{\sqrt{2\pi\sigma^2}} \frac{e^{-(x-m)^2}}{2\sigma^2}, -\infty < x < \infty \tag{3.18}$$

The PDF of the Gaussian random variable has two parameters, m and σ, which have the interpretation of the mean and standard deviation respectively. The parameter σ^2 is referred to as the variance. In general, the Gaussian PDF is centred about the point $x = m$ and has a width that is proportional to σ. The CDF is required whenever we want to find the probability that a Gaussian random variable lies above or below some threshold or in some interval. The CDF of a Gaussian random variable is written as.

$$F_x(x) = \int_{-\infty}^{x} (\frac{1}{\sqrt{2\sigma^2\pi}})\, e^{-(y-m)^2/2\sigma^2} dy \tag{3.19}$$

it can be shown that it is impossible to express this integral in closed form. The fallowing fig. 4.3 (a) and (b) shows the PDF and CDF of a Gaussian random variable.



(a) (b)

Fig. 3.6: (a) PDF (b) CDF of Gaussian distribution [15]

Because Gaussian random variables are so commonly used in such a wide variety of applications, it is standard practice to introduce a shorthand notation to describe a Gaussian random variable $X \sim N(m, \sigma^2)$. This is read X is distributed normally (or Gaussian) with mean m, and variance σ^2.

3.3 Diversity Techniques

One of the most powerful techniques to mitigate the effects of fading is to use diversity-combining of independently fading signal paths. Diversity-combining uses the fact that independent signal paths have a low probability of experiencing deep fades simultaneously. Thus, the idea behind diversity is to send the same data over independent fading paths. These independent paths are combined in some way such that the fading of the resultant signal is reduced. For example, consider a system with two antennas at either the transmitter or receiver that experience independent fading. If the antennas are spaced sufficiently far apart, it is unlikely that they both experience deep fades at the same time. By selecting the antenna with the strongest signal, called selection combining, we obtain a much better signal than if we just had one antenna.

Diversity techniques that mitigate the effect of multipath fading are called microdiversity, and that is the focus of this chapter. Diversity to mitigate the effects of shadowing from buildings and objects is called macrodiversity. Macrodiversity is generally implemented by combining signals received by several base stations or access points. This requires coordination among the different base stations or access points. Such coordination is implemented as part of the networking protocols in infrastructure-based wireless networks.

3.3.1 Time diversity

Time diversity can be achieved by transmitting identical messages in different time slots, which results in uncorrelated fading signals at the receiver. It is used in digital communication systems to combat that the transmissions channel may suffer from error bursts due to time-varying channel conditions. The error bursts may be caused by fading in combination with a moving receiver, transmitter or obstacle, or by intermittent electromagnetic interference, for example from crosstalk in a cable, or co-channel interference from radio transmitters. In telecommunication, a burst error or error burst is a contiguous sequence of symbols, received over a data transmission channel, such that the first and last symbols are in error and there exists no contiguous subsequence of m correctly received symbols within the error burst [2]. In mobile communications, error control coding is combined with interleaving to achieve time diversity. In this case, the replicas of the transmitted signals are usually provided to the receiver in the form of redundancy in the time domain introduced by error control coding. The time separation between the replicas of the transmitted signals is provided by time interleaving to obtain independent fades at the input of the decoder. Since time interleaving results in decoding delays, this technique is usually effective for fast fading environments where the coherence time of the channel is small. For slow fading channels, a large interleaver can lead to a significant delay which is intolerable for delay sensitive applications such as voice transmission. This constraint rules out time diversity for some mobile radio systems. For example, when a mobile radio station is stationary, time diversity cannot help to reduce fades. One of the drawbacks of the scheme is that due to the redundancy introduced in the time domain, there is a loss in bandwidth efficiency.

3.3.2 Space Diversity

Space diversity has been a popular technique in wireless microwave communications**.** Space diversity is also called antenna diversity. It is typically implemented using multiple antennas or antenna arrays arranged together in space for transmission and/or reception. The multiple antennas are separated physically by a proper distance so that the individual signals are uncorrelated [13]. The separation requirements vary with antenna height, propagation environment and frequency. Typically a separation of a few wavelengths is enough to obtain uncorrelated signals.

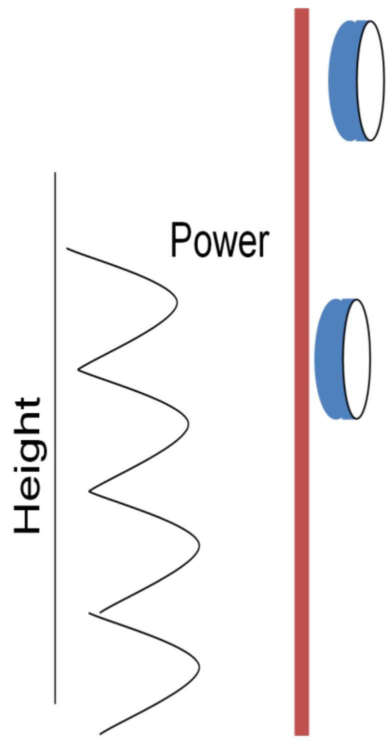


Fig. 3.7 Space diversity

In space diversity, the replicas of the transmitted signals are usually provided to the receiver in the form of redundancy in the space domain. Unlike time and frequency diversity, space diversity does not induce any loss in bandwidth efficiency. This property is very attractive for future high data rate wireless communications. Depending on whether multiple antennas are used for transmission or reception, we can classify space diversity into two categories: receive diversity and transmit diversity In receive diversity, multiple antennas are used at the receiver site to pick up independent copies of the transmit signals. The replicas of the transmitted signals are properly combined to increase the overall received SNR and mitigate multipath fading. In transmit diversity, multiple antennas are deployed at the transmitter site. Messages are processed at the transmitter and then spread across multiple antennas [2],[3],[5]. The details of transmit diversity is discussed in next chapter.

3.4 Diversity combining methods

Diversity combining is the technique applied to combine the multiple received signals of a diversity reception device into a single improved signal [11].

3.4.1 Switched combining

In a switched combining diversity system as shown in fig. 3.3 (a) and (b). The receiver scans all the diversity branches and selects a particular branch with the SNR above a certain predetermined threshold. This signal is selected as the output, until its SNR drops below the threshold. When this happens, the receiver starts scanning again and switches to another branch[2],[13]. This scheme is also scanning diversity.

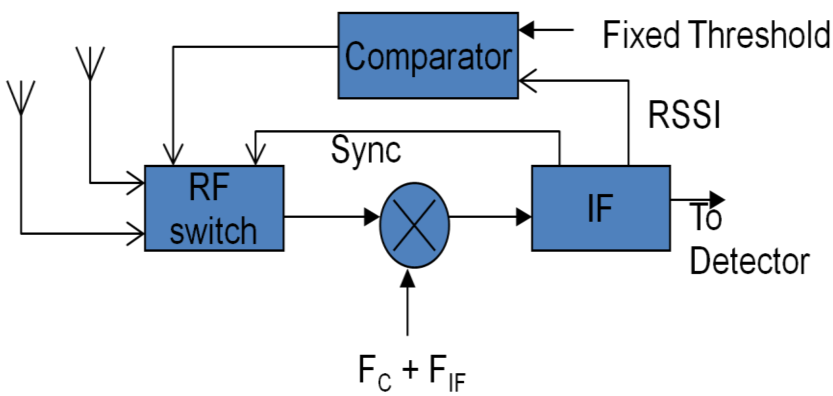


Fig 3.8 (a): Scaning diversity

Compared to selection diversity, switched diversity is inferior since it does not continually pick up the best instantaneous signal. However, it is simpler to implement as it does not require simultaneous and continuous monitoring of all the diversity branches. For both the selection and switched diversity schemes, the output signal is equal to only one of all the diversity branches. In addition, they do not require any knowledge of channel state information. Therefore, these two schemes can be used in conjunction with coherent as well as noncoherent modulations.

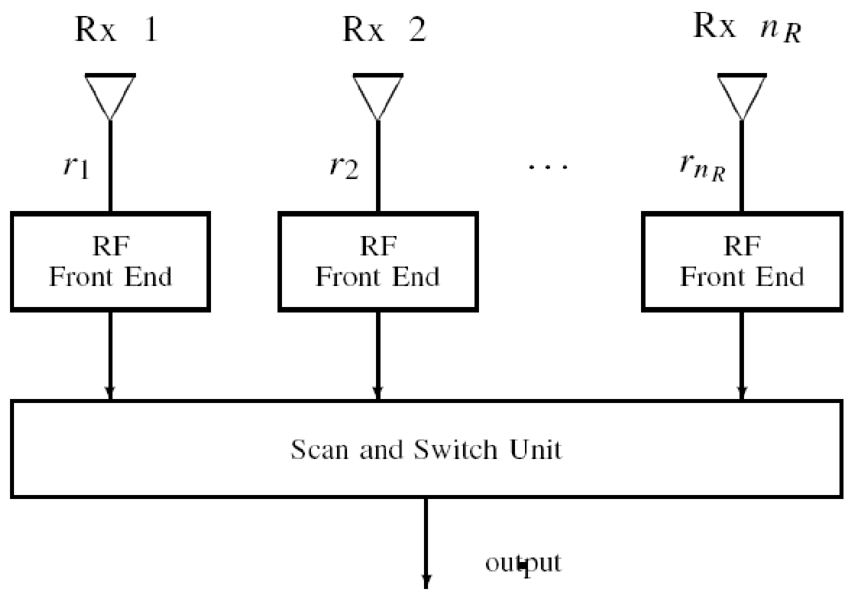


Fig. 3.8 (b): Scaning diversity [13]

3.4.2 Maximal ratio combining

Maximum ratio combining is a linear combining method.

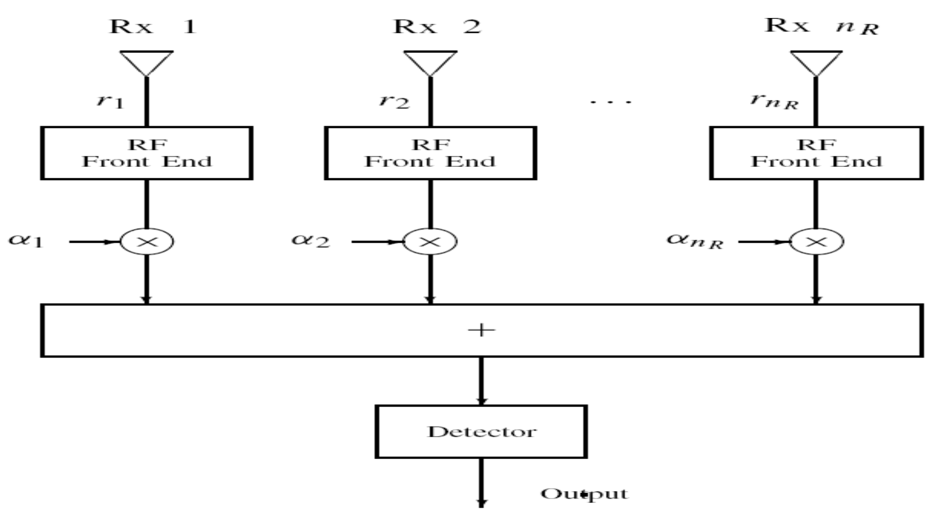


Fig. 3.9: Maximal ratio combining [13]

In a general linear combining process, various signal inputs are individually weighted and added together to get an output signal.

Chapter 4
SPACE TIME BLOCK CODE

Introduction

STBC basically it forms combination of two words space diversity and time diversity [2]. In modern wireless communication systems, data are sent from one or more transmit antennas to one or more receive antennas. In this paper, we consider one model for sending data over fading channels using multiple transmit antennas and one or more receive antennas. The motivation for using multiple transmit antennas is that multiple copies of data sent from different physical positions maybe corrupted in different ways during transmission, which may better allow the receiver to recover the intended data. Other words, through appropriate signal processing at the receiver, the independent paths provided by the multiple transmit antennas can be considered as one channel that is more reliable than any of the single independent paths. This idea is known as space diversity. The model we consider also sends multiple copies of data at different time steps, again increasing the likelihood that the receiver can recover the intended data [2],[3]. This idea is known as time diversity.

4.1 Rate of STBC

The rate of STBC is defined as ratio between the number of symbol the encoder takes as the input(k) and the total time consumed in transmitted the entire symbol(T) [3]. It is given by

$$R = \frac{k}{T}$$

An STBC is usually represented by matrix each row represents a time slot and each column represents one antenna's transmissions over time.

STBC operates on a block of data-stream which is to be transmitted and produces a matrix whose rows and columns represent antennas and time, respectively [3]. The usual and simplest representation of STBC is shown below

The maximum code rate of an O-STBC for two transmit antennas is one, which speaks for the well-known Alamouti STBC If you increase the transmit antenna and input symbol, then you reduce the code rate [5], [7]. Two examples of O-STBC for four and eight transmit antennas with rate 3/4 and 1/2 respectively

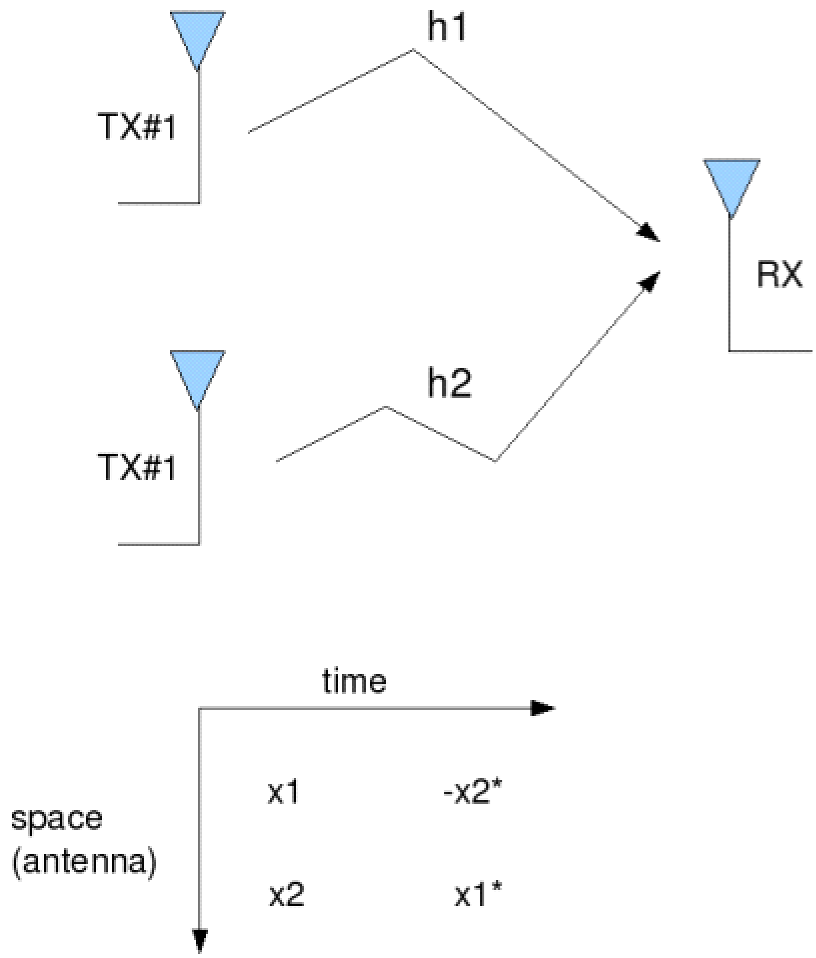
$$c_4 = \begin{bmatrix} c_1 & 0 & c_2 & -c_3 \\ 0 & c_1 & c_3^* & c_2^* \\ -c_2^* & -c_3 & c_1^* & 0 \\ c_3^* & -c_2 & 0 & c_1^* \end{bmatrix}$$

$$c_8 = \begin{bmatrix} c_1 & c_1 & c_2^* & -c_2^* & c_3^* & -c_3^* & c_2^* & -c_4^* \\ c_1^* & -c_1^* & c_2 & c_2 & c_3 & c_3 & c_4 & c_4 \\ -c_2^* & c_2^* & c_1 & c_1 & c_4 & -c_4 & -c_3 & c_1 \\ -c_2 & -c_2 & c_1^* & -c_1^* & c_4^* & c_4^* & -c_3^* & -c_3^* \\ -c_3^* & c_3^* & -c_4 & c_4 & c_1 & c_1 & c_2 & -c_2 \\ -c_3 & -c_3 & -c_4^* & -c_4^* & c_1^* & -c_1^* & c_2^* & c_2^* \\ -c_4^* & c_4^* & c_3 & -c_3 & -c_2 & -c_2 & c_1 & c_1 \\ -c_4 & -c_4 & c_3^* & c_3^* & -c_2^* & -c_2^* & c_1^* & -c_1^* \end{bmatrix}$$

4.2 Alamouti STBC

The scheme is as follows

- Consider that we have a transmission sequence, for example s_1, s_2……s_n.
- In normal transmission, we will be sending s_1, in the first time slot, s_2 in the second time slot, s_3 and so on.
- However, ALAMOUTI suggested that we group the symbols into groups of two. In the first time slot, send s_1 and s_2 from the first and second antenna and in second time slot, send and from the first and second antenna respectively. In the third time slot, send s_3 and s_4 from the first and second antenna and in fourth time slot, send and from the first and second antenna respectively and so on. [10]
- Notice that though we are grouping two symbols, we still need two time slots to send two symbols. Hence, there is no change in the data rate.
- This forms the simple explanation of the transmission scheme with Alamouti Space Time Block coding.



$$X_1 = S_1,\ X_2 = S_2,\ X_2^* = -S_2^*,\ X_1^* = S_1^*$$

Fig. 4.1: 2-Transmit, 1-Receive Alamouti STBC coding [10]

4.3 Alamouti uses two transmit antennas and one receive antenna simultaneously

Its most attractive features are given by

- Diversity order of two.
- Only one receive antenna.
- Low complexity detection principle.
- No feedback requirements.

A diversity order of two is also obtainable by the ideal receive diversity system (with $Nr = 2$) that has been presented as a reference in the previous chapter. However, the Alamouti scheme allows to shift the multiplicity of RF-chains from the receiver to the transmitter, making it an attractive method for enhancing the downlink to small mobile units with a low power consumption profile. The detection algorithm complexity is low for both schemes, optimum maximum-likelihood detection performance is attainable with simple linear processing [3], [8]. To sum up, the Alamouti STBC transmission is capable of delivering the same diversity order as the MRRC system while utilizing only one receive antenna. The performances of the two schemes only differ by a constant penalty of 3dB in average SNR for the Alamouti scheme. The reason for this is simply because the total transmit power is being split up on the two transmit antennas. At this point some general remarks on the application of space-time coding are indicated. Again comparing it with some receive diversity technique.

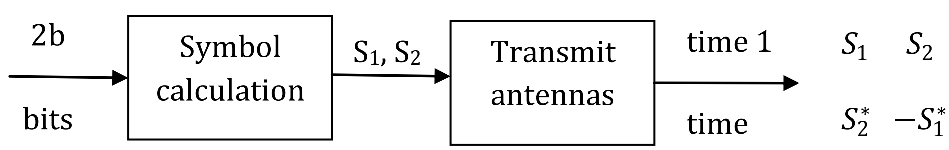


Fig. 4.2 Transmitter block diagram for Alamouti code

Table 4.1: Space time block code

	Antenna 1	Antenna 2
First clock cycle	S_1	S_2
Second clock cycle	S_2^*	$-S_1^*$

like MRRC two major differences arise. The first issue is the unavoidable decoding delay that is inherent to the space-time coding principle. Secondly, channel estimation issues are more complex to solve since the paths from multiple transmit antennas need to be estimated, thus either alternate pilot insertion or orthogonal pilot signals need to be employed. However, these drawbacks are moderate considering the Alamouti scheme that is analyzed in the following, since it only introduces the minimal decoding delay of one additional symbol period and utilizes only two transmit antennas.

4.4 Alamouti transmission scheme

The to-be transmitted bit sequence is mapped to symbols and the resulting symbol stream is subsequently grouped into blocks being two symbols long. These blocks are then successively transmitted via two transmit antennas during two symbol intervals in the following fashion [3]. Let s_2 and s_2 be the elements of one of the above mentioned symbol blocks. According to Table 4.1 an Alamouti STBC scheme is realized by transmitting s_1 via the first antenna and s_2 via the second antenna during the first symbol period. Whilst the next symbol cycle s_1^* is emitted from antenna one and $-s_2^*$ from antenna two respectively. It is assumed that the channel path gains from the two transmit antennas to the single receive antenna (denoted by h_1 and h_2) are not changing during those two symbol cycles. The receive signals r_1 and r_2 for the two considered time intervals are then obtained as:

$$r_1 = h_1 S_1 + h_2 S_2 + n_1$$

$$r_2 = h_1 S_2^* - h_2 S_1^* + n_2$$

with n_1 and n_2 being the noise

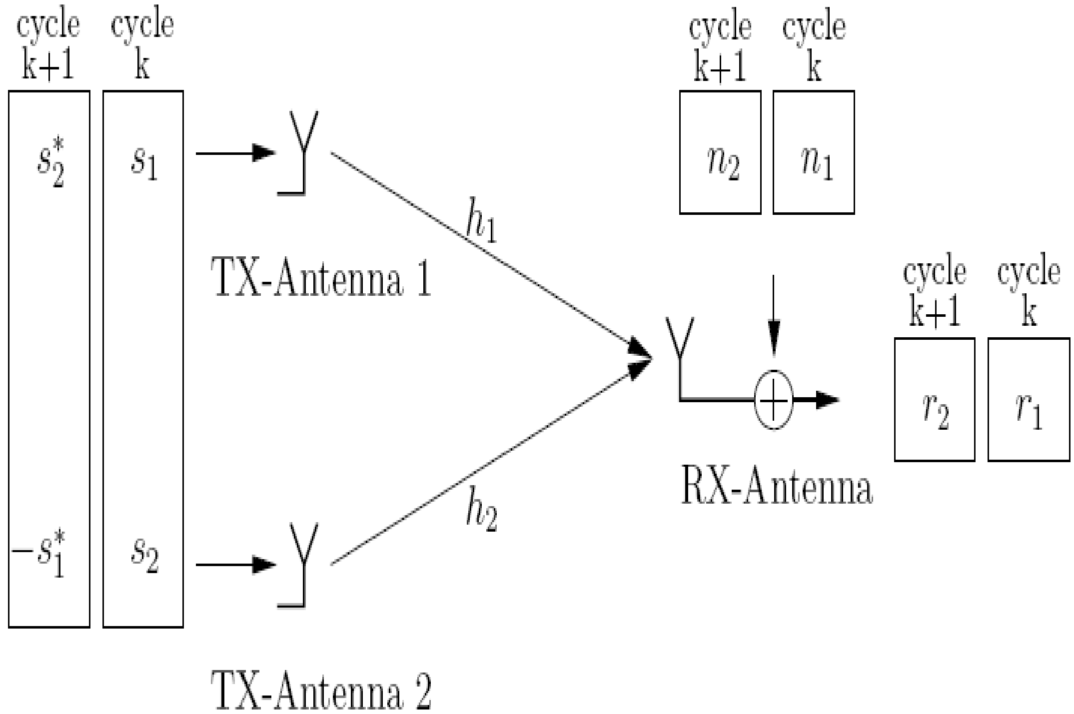


Fig. 4.3: Two-Branch Transmit Diversity with One Receiver [10]

Fig. 4.2 shows the baseband representation of the new two branch transmit diversity scheme [3]. 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.

4.4.1The 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_0) and from antenna one by (s_1). During the next symbol period signal (-s_1^*) is transmitted from antenna zero, and signal is transmitted from antenna one where is the complex conjugate operation. This sequence is shown in Table 4.1, the encoding is done in space and time (space–time coding) [3]. 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)

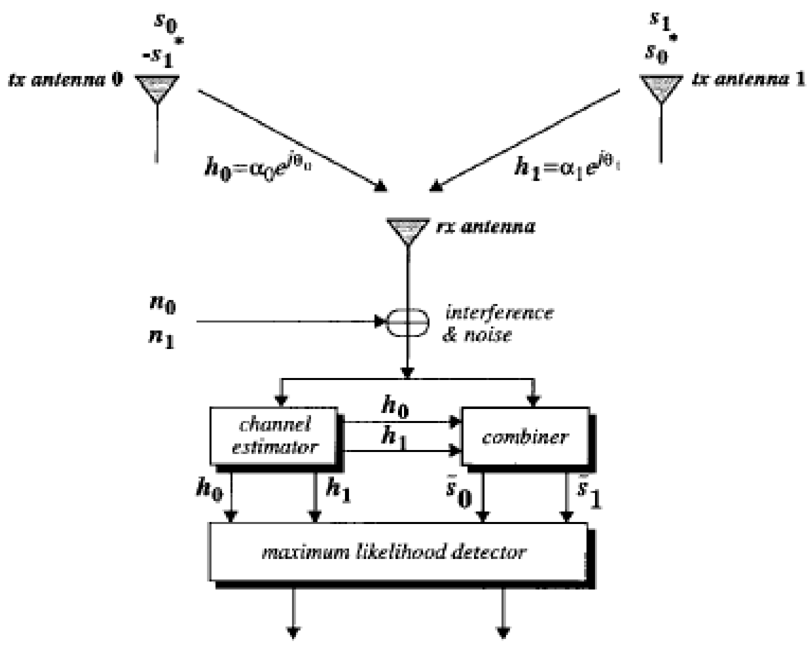


Fig. 4.4: Two transmit antenna and one receiver antenna [3]

The channel at time may be modeled by a complex multiplicative distortion $h_0(t)$ for transmit antenna zero (s_0) and $h_0(t)$ for transmit antenna one(s_1). Assuming that fading is constant across two consecutive symbols, we can write across two consecutive symbols, we can write.

$$h_0(t) = h_0\,(t+T) = h_0 = \alpha_0 e^{j\theta_0} \qquad (4.1)$$

$$h_1(t) = h_1(t+T) = h_1 = \alpha_1 e^{j\theta_1} \qquad (4.2)$$

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

$$r_0 = r(t) = h_0 s_0 + h_1 s_1 + n_0 \qquad (4.3)$$

$$r_1 = r(t+T) = -h_0 S_1^* + h_1 S_0^* + n_1 \qquad (4.4)$$

where r_0, r_1 is received signal and n_0, n_1 is noise

4.4.2 A STBC matrix notation

For more complex space-time block codes the previous way of description becomes impractical. In the following a different mathematical representation of the Alamouti scheme is given, outlining a straightforward and more general way to describe space-time block codes. Starting point is the formulation of the space-time code matrix S [2],[3]:

$$S = \begin{bmatrix} s_1 & s_2 \\ s_2^* & -s_1^* \end{bmatrix}$$

Thus the rows correspond to the symbol intervals and the columns identify the respective transmit antenna. The received signals, channel path gains and noise samples are stacked in 2 × 1 column vectors

$$r = \begin{bmatrix} r_1 \\ r_2 \end{bmatrix}, h = \begin{bmatrix} h_1 \\ h_2 \end{bmatrix}, n = \begin{bmatrix} n_1 \\ n_2 \end{bmatrix}$$

Above equation. we can also write in matrix form

$$\begin{bmatrix} h_0 & h_1 \\ h_1^* & -h_0^* \end{bmatrix} \begin{bmatrix} r_1 \\ r_2 \end{bmatrix} + \begin{bmatrix} n_0 \\ n_1^* \end{bmatrix}$$

4.5 The combining scheme

The combiner shown in fig. 4.3 builds the following two combined signals that are sent to the maximum likelihood detector [1], [11].

$$\bar{S}_0 = h_0^* r_0 + r_1^* h_1 \quad (4.5)$$

$$\bar{S}_1 = h_1^* r_0 - r_0^* h_1 \quad (4.6)$$

It is important to note that this combining scheme is different from the MRRC

$$\bar{S}_0 = (\alpha_0^2 + \alpha_1^2)s_0 + h_0^* n_0 + n_1^* h_1 \qquad (4.7)$$

$$\bar{S}_0 = (\alpha_0^2 + \alpha_1^2)s_1 - n_1^* h_0 + h_1^* n_0 \qquad (4.8)$$

4.5.1 The Maximum likelihood decision rule

These combined signals are then sent to the maximum likelihood detector which, for each of the signals and, uses the decision rules. The resulting combined signals are equivalent to that obtained from two-branch MRRC. The only difference is phase rotations on the noise components which do not degrade the effective SNR. Therefore, the resulting diversity order from the new two-branch transmit diversity scheme with one receiver is equal to that of two-branches MRRC.

The performance of the new scheme with two transmitters and a single receiver is 3dB poorer than two-branch MRRC. As the 3dB penalty is incurred because the simulations assume that each transmit antenna radiates half the energy in order to ensure the same total radiated power as with one transmit antenna [3]. If each transmit antenna in the new scheme was to radiate the same energy as the single transmit antenna for MRRC, however, the performance would be identical. In other words, if the BER was drawn against the average SNR per transmit antenna, then the performance curves for the new scheme would shift 3dB to the left and overlap with the MRRC curves. Nevertheless, even with the equal total radiated power assumption, the diversity gain for the new scheme with one receive antenna at a BER of 10^{-4} is about 15 dB [3].

- **Theorem*:* A complex orthogonal design of size (4× 4) does not exist**.

Proof: The proof is divided into six steps.

Step I: In this step, we provide necessary and sufficient conditions for a matrix of (4× 4) indeterminates to be a complex linear processing generalized orthogonal design. To this end, let L_c be a complex linear processing generalized orthogonal design of size n=4. Each entry of is a linear combination x_1, x_1^*, x_2, x_2^*. It follows that

$$LC = x_1 A_1 + x_1^* B_1 + x_2 A_2 + x_2^* B_2 + \cdots \ldots\ldots\ldots .. + x_4^* B_4 \qquad (4.5\text{ a})$$

Where $A_1 B_1$ ……. $A_4 B_4$ it follows 4× 4 $matrices$,since

$$Lc\, L_c^* = L_c^* Lc = (|x_1|^2 + |x_2|^2 + \cdots \ldots .\, |x_4|^2) I$$

We can conclude from the above

$$A_i A_i^* + B_i B_i^* = A_i^* A_i + B_i^* B_i = I, \quad i = 1,2,3,4 \qquad (4.5\, b)$$

$$A_i A_j^* + B_j B_i^* = A_i^* A_j + B_j^* B_i = 0, \qquad 1 \le i \ne j \le 4 \qquad (4.5\, c)$$

$$B_i A_j^* + B_j A_i^* = A_i B_j^* + A_j B_i^* = 0, 1 \le i \ne j \le 4 \qquad (4.5\, d)$$

$$A_i B_i^* = A_i^* B_i = 0\, ; \quad 1 \le i \le 4$$

Conversely, any set of 4× 4 complex matrices $A_1 B_1, \ldots, A_4 B_4$ satisfying the above equations defines a linear processing complex orthogonal design. Satisfying the above equations defines a linear processing complex orthogonal design [6].

Step II: In this step, we will prove that given a complex linear processing generalized orthogonal design L_c we could construct another complex linear processing generalized orthogonal design ϵ such that for any row, one of X_j and X_j^* does not occur in the entries of that row of ϵ . In other words for any $i = 1,2,3,4$

$$\varepsilon_{ik} = \sum_{j=1}^{4} a_{i,j,k}\, x_j + \sum_{j=1}^{4} b_{i,j,k}\, x_j^* \qquad (4.5\, e)$$

Where for any fixed i either $b_{i,j,k} = 0$ for all $k = 1,2,3,4$ or $a_{i,j,k} = 0\ for\ all\ k = 1,2,3,4$ Using equation (a), we first observe that

$$A_i = A_{i(I)} = A_{i(A_i^* A_i + B_i^* B_i)} = A_i A_i^* A_i \qquad (4.5 f)$$

Hence,

$$A_i^* A_i = A_i^* A_i A_i^* A_i = (A_i^* A_i)^2$$

Similarly $B_iB_i^*$ = $(B_iB_i^*)^2$ are idempotent for i = 1,2,3,4 since $A_iA_i^*$ + $B_iB_i^*$ = I, The matrices $A_iA_i^*$ and $B_iB_i^*$ represent projection onto perpendicular vector spacesW_i , and $W_i^{\perp}$ thus are digonalizable with all eigen value in set {0,1}. if Pi = $rank(wi)$and Pi = $rank(W_i^{\perp})$ = 4 − Pi then exactly p_i (respectively, q_i) of Eigen value of $A_iA_i^*$ (respectively $B_iB_i^*$) are 1. Next, using (above equation), we observe that for $i \neq j$

$$A_iA_i^* \ A_jA_j^* = -A_iB_j^* \, B_iA_j^* = A_jB_i^* \, B_iA_j^*$$

$$= -A_j \, B_i^* \, B_j \, A_i^* = A_j \, A_j^* \, A_i \, A_i^*$$

Thus the matrices $A_iA_i^*$ and $B_jB_j^*$ commute. Similarly, it follows that $\{A_iA_i^* \, , B_iB_i^*, i =$ $1,2,3,4 \}$ is a commuting family of diagonalizable matrices. Hence, these matrices are simultaneously diagonalizable. Since the Eigen values $A_iA_i^*$ of are in the set{0,1}, we conclude that there exists a unitary transformation U such that

$$UA_iA_i^*U^* = D_i^1$$

$$UB_iB_i^*U^* = D_i^2$$

whereD_i^1, $D_i^2, i = 1,2,3,4$ are diagonal matrices with diagonal entries in the set{0,1} . Moreover, because,

$$D_i^1 + D_i^2 = U(\ A_iA_i^* + B_iB_i^*)U^* = I \qquad (4.5g)$$

The (i,j)th entry of D_i^1 is zero (respectively, one) if and only if the (i,j)th entry of D_i^2 is one (respectively, zero). Since,

$$D_i^1U \, B_iU^* = UA_iA_i^*U^*UB_iU^* = 0 \qquad (4.5h)$$

implies that the nonzero entries of $U \ A_i \ U^*$ appear in those rows, where the corresponding diagonal element of D_i^2 is zero. Thus the nonzero entries of $U A_i U^*$ and $U B_i U^*$ occur in different rows.

Let

$$\varepsilon = \sum_{1}^{4}(UA_iU_i^*x_i + UB_iU^*x_i^*) \qquad (4.5i)$$

then it follows from the matrix equations given in Step I that is a complex linear processing generalized orthogonal design with the desired property [6].

***Step III**:* We can now assume without any loss of generality That $\mathcal{L}_c$is a complex linear processing generalized orthogonal design with the properties described in Step II $\mathcal{L}_c$. In this step, we apply Construction II to and study the properties of the associated real linear processing generalized orthogonal design. By interchanging x_i withx_i^* everywhere in the design if necessary, we can further assume that only x_1,x_2,x_3,x_4 occur in the first row of. We next apply Construction II to and construct a real orthogonal design of size in variables $x_1^2,\ x_1^2,\ x_2^1, \dots\dots x_4^2$. The matrix can be written as

$$0 = C_1x_1^1 + C_2x_1^2 + C_3x_2^1 + \cdots + C_8x_4^2 \qquad (4.5j)$$

Where $c_1,\ c_2,\ c_3\ and\ c_4$ are real matrices. Furthermore, assuming the property established in Step II, we can easily observe by direct computation that

$$C_2 = J_1C_1$$

$$C_4 = J_2C_3$$

$$C_6 = J_3C_5$$

$$C_8 = J_2C_3$$

Where,

$$J_i = F_i * \begin{pmatrix} 0 & 1 \\ -1 & 0 \end{pmatrix}$$

Where F_i is a diagonal matrix of size whose diagonal entries belong to the set. Moreover, the th entry of F_i, i=1,2,3,4 equals 1 . We let $\in_i = (\in_{i,1} \dots\dots \in_{i,4})$ denote the vector whose th component is the (j,j)th element of The F_ith element of is equal to 1 (respectively, -1) if X_i (respectively, x_i^*) occurs in row j.

$$0^{T0} = 0^T 0 = \left(\sum_{1}^{4}[(x_i^1)^2 + (x_i^2)^2]\right) I \qquad (4.5\,k)$$

we arrive at the following set of equations:

$$C_i x_i^T = I_S, i = 1,2, \dots \dots \dots 8 \qquad (4.5\,l)$$

$$C_i x_j^T = -C_j C_i^T, 1 \leq i < j \leq 8$$

Let $E_i = C_i\, C_1^T$, using above two equation,

$$E_1 = I_S$$

$$E_2 = J_1$$

$$E_4 = J_2 E_3$$

$$E_6 = J_3 E_3$$

$$E_8 = J_4 E_7$$

$$E_i^T = -E_i, i=2, \dots .8$$

$$E_i^T = E_i E_i^T = I_8, i=1 \dots ..8$$

$$E_j^T E_i = -E_i^T E_j, 1 \leq i \neq j \leq 8 \qquad (4.5m)$$

Step IV: We next prove that the matrices E_{2i-1}, E_{2i}, $i = 1, 2, 3, 4$ aanticommute with j_1 and j_2 but commute with j_j, $j \neq 1$ and $j \neq i$. First, we observe that by

$$E_{2i-1} J_1^T + J_1 E_{2i-1}^T = E_{2i-1} E_2^T + E_2 E_{2i-1}^T = 0 \qquad (4.5n)$$

$$E_{2i-1}\, J_i + E_{2i-1}^T J_i^T = E_{2i} + E_{2i}^T = 0 \qquad (4.50)$$

Since the matrices j_1, j_i and are antisymmetric, the above equations prove that E_{2i-1} anticommutes with j_1 and j_i. Furthermore, since $(j_j\, E_{2i-1})^T = E^T_{2j-1}\ j^T_j$, we conclude from above equationthat when j$\neq$ 1 and j$\neq$ i

$$E_{2i-1}\ J_i E^T_{2i-1} + E_{2i-1}\, E^T_{2j-1} J^T_i = E_{2j}\, E^T_{2i-1} + E_{2i-1}\, E^T_{2j} = 0 \qquad (4.5\,p)$$

$$E_{2j-1}\, E^T_{2i-1} + E_{2i-1}\, E^T_{2j-1} = 0 \qquad (4.5\,q)$$

Since j_j anticommutes with E_{2i-1} we arrive at,

$$E_{2j-1}\ J_j\, E_{2i-1} = E_{2j-1}\, E_{2i-1}\ J_j \qquad (4.5r)$$

Because E_{2j-1} is orthogonal, it is invertible and thus whenever j$\neq$ 1and j$\neq$ i. We have

$$J_j\ E_{2i-1} = E_{2i-1}\ J_j \qquad (4.5s)$$

The assertion forE_{2i} now easily follows since. $E_{2i} = j_i\ E_{2i-1}$

Step V: Recall that let $\in_i = (\in_{i,1} \ldots\ldots \in_{i,4})$ is the vector whose th component is the jth element of (j, j) . In this step, we prove that any two vectors$\in_i$ and$_j$ have Hamming distance exactly equal to two. To this end, since E_{2i-1} commutes withJ_j for when j$\neq$ 1, j$\neq$ i and anticommutes with J_1 and J_i, we can easily conclude from the non-singularity of that for E_{2i-1} and for . Thus the Hamming distance of any two distinct vectors $\in_i$and$\in_j$ is neither zero nor four. We first prove that the Hamming distance of any two distinct vectors $\in_i$ and $\in_j$ cannot be one. To this end, let us suppose that two distinct vectors $\in_i$and$\in_j$ have Hamming distance one and differ only in the ith position. Then in the th row of $\mathcal{L}_c$, we have either occurrences x_i of and x_j or occurrences of and x_j but not both. In any other row of $\mathcal{L}_c$, we have either occurrences of and or occurrences of and but not both. It is easy to see that the columns of $\mathcal{L}_c$ cannot be orthogonal to each other [6].

We next prove that the Hamming distance of any two distinct vectors $\in_i$ and $\in_j$ cannot be three. To this end, let us suppose that two distinct vectors $\in_i$ and $\in_j$ have Hamming distance three. Since for all $k = 1, 2, 3\ and\ 4$ we conclude that for all. We can now choose L$\neq i, j$ and observe that the vector $\in_l$ is distinct with both $\in_i$ and moreover, it coincides with both and in the first position. It follows using a simple counting argument that has Hamming distance one with either or. But we just proved that this is not possible. We conclude that any two distinct vectors and have Hamming distance exactly equal to two.

Step VI: In this step, we will arrive at a contradiction that concludes the proof.

Because any two distinct vectors $\in_i$ and $\in_j$ have Hamming distance exactly equal to two, the matrix H whose ith row is $\in_i$ a Hadamard matrix. It follows that any two distinct columns of H also have Hamming distance 2. Thus we can now assume without loss of generality that x_1, x_2, x_3, x_4 occur in row one and $x_{1,}^* \; x_2^*, x_{3,}^* \; x_4^*$ occur in row two of the first row of $\mathcal{L}_c$ is thus expressible as $x_1 v_1 +, x_2 v_2, + x_3 v_3 + x_4 v_4$ and the second row of is of the form for appropriate vectors. Because

$$L_c^* L_c = (|x_1|^2 + |x_2|^2 + \cdots \ldots . |x_4|^2) I \qquad (4.5t)$$

we observe that $v_{i,} \; w_i$ are vectors of unit length. Moreover, if j$\neq i$ the vectors v_i and v_j are orthogonal to each other. Since the first and second rows of $\mathcal{L}_c$ are orthogonal, we observe that w_3 is orthogonal to v_i, $i = 1,2,3,4$. This means that $\{ \mathrm{w}_{3,} \, v_1, \, v_3, \, v_4 \}$ contains a set of five orthonormal vectors in complex space of dimension. This contradiction proves result.

- **Advantage of Alamouti:**

  - The transmit diversity scheme can improve the error performance, data rate, or capacity of wireless communications systems [3].
  - This scheme is effective in all of the applications where system capacity is limited by multipath fading.

- **Disadvantage of Alamouti:**

Delay effect: With branch transmit diversity, if the transformed copies of the signals are transmitted at distinct intervals from all the antennas, the decoding delay is symbol periods. That is, for the two-branch diversity scheme, the delay is two symbol periods. For a multicarrier system, however, if the copies are sent at the same time and on different carrier frequencies, then the decoding delay is only one symbol period [2],[3].

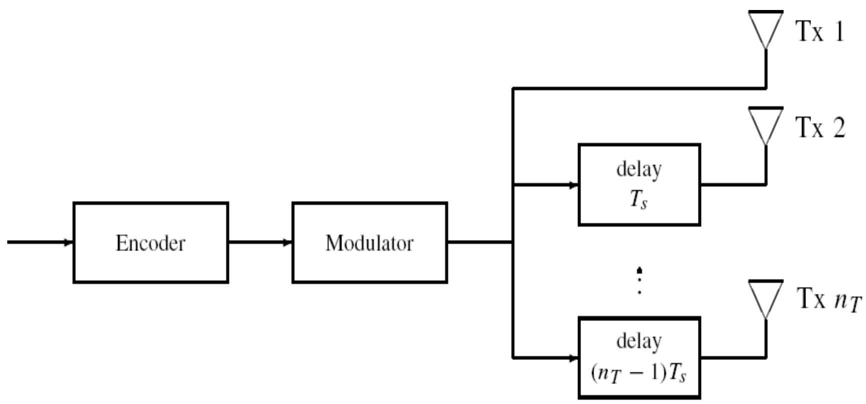


Fig. 4.5: Delay transmit diversity scheme

we can define the following two quantities to account for the decoding performance of a S.T.B.C

- Transmit Diversity Gain: The minimum rank D of the matrix a overall pairs of distinct code words. It accounts for the slope of the bit error rate (BER) or block error rate (BLER) curve of the STBC. A STBC is said to achieve full diversity if $D = N_T$, where N_T is number of transmitted antennas.
- Diversity Product*:* Diversity product accounts for the coding gain, which determines the left or right shift of the BER curve of the STBC.

Chapter 5
Simulation Result and Discussion

The transmit diversity technique, using two transmit diversity and one receive antenna the scheme provides the same diversity order as the MRRC with one transmit antenna and two receive antenna but it is generalized only for two transmit antenna and M receive antenna. And Alamouti scheme provide full rate, i.e code rate equal to one. STBC with Channel coding

- Convolutional encoding can be used to improve the performance of wireless system.
- Viterbi algorithm is an optimum decoder.
- Using convolutional coding, the information can be extracted without any error from noisy channel.
- When the SNR is lower than some value, there will be some error and the error rate increases as the SNR decreases. When SNR is lower than some certain value, the convolutional encoder cannot extract the information.

5.1 Comparison of results

The results for **S**ingle Input Single Output (SISO), Receiver diversity, space time block code (STBC) and space time block code with channel coding has been discussed here. SISO systems are the simplest antenna technology. With single antennas used, single frequencies are vulnerable to space limits and frequency fading. The term "SISO", as applied to wireless technology, refers to the antenna technology that uses a single antenna at both the transmitter side and the receiver side. SISO systems are sometimes troubled by multipath effects. Electromagnetic wavefronts are dispersed when they encounter signal-path obstructions like buildings, hills, tunnels, valleys and utility wires. In such cases, the scattered electromagnetic waves take many paths to reach the destination. That causes problems like cut-out (cliff effect), fading, and intermittent reception (picket fencing). In a digital communications system, it can cause a reduction in data speed and an increase in the number of errors.

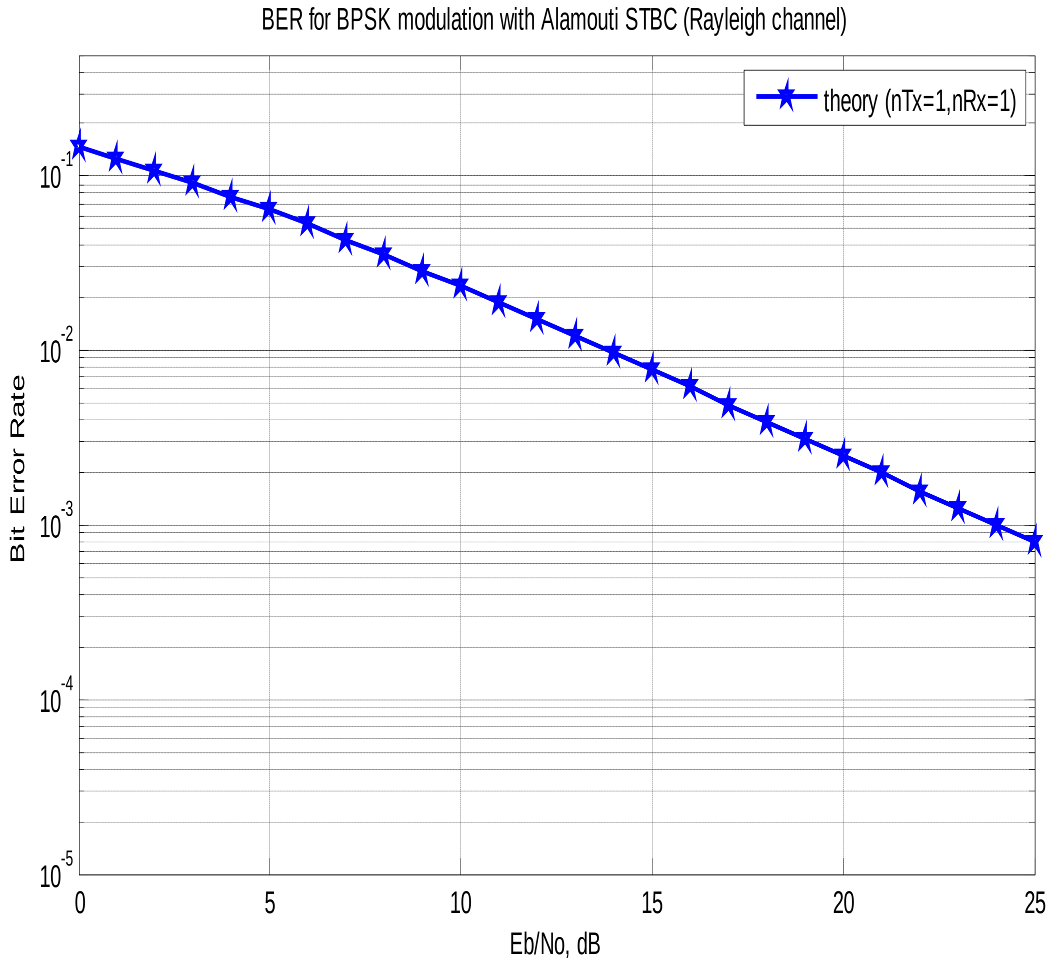


Fig. 5.1: BER performance of Single transmitter and single receiver

A form of smart antenna technology that uses a single antenna at the transmitter and a multiple antennas at the receiver on a wireless device to improve the transmission distance. From fig. 5.2, it has been observed that when we increase the antenna at the receiver side performance of the system is improved.

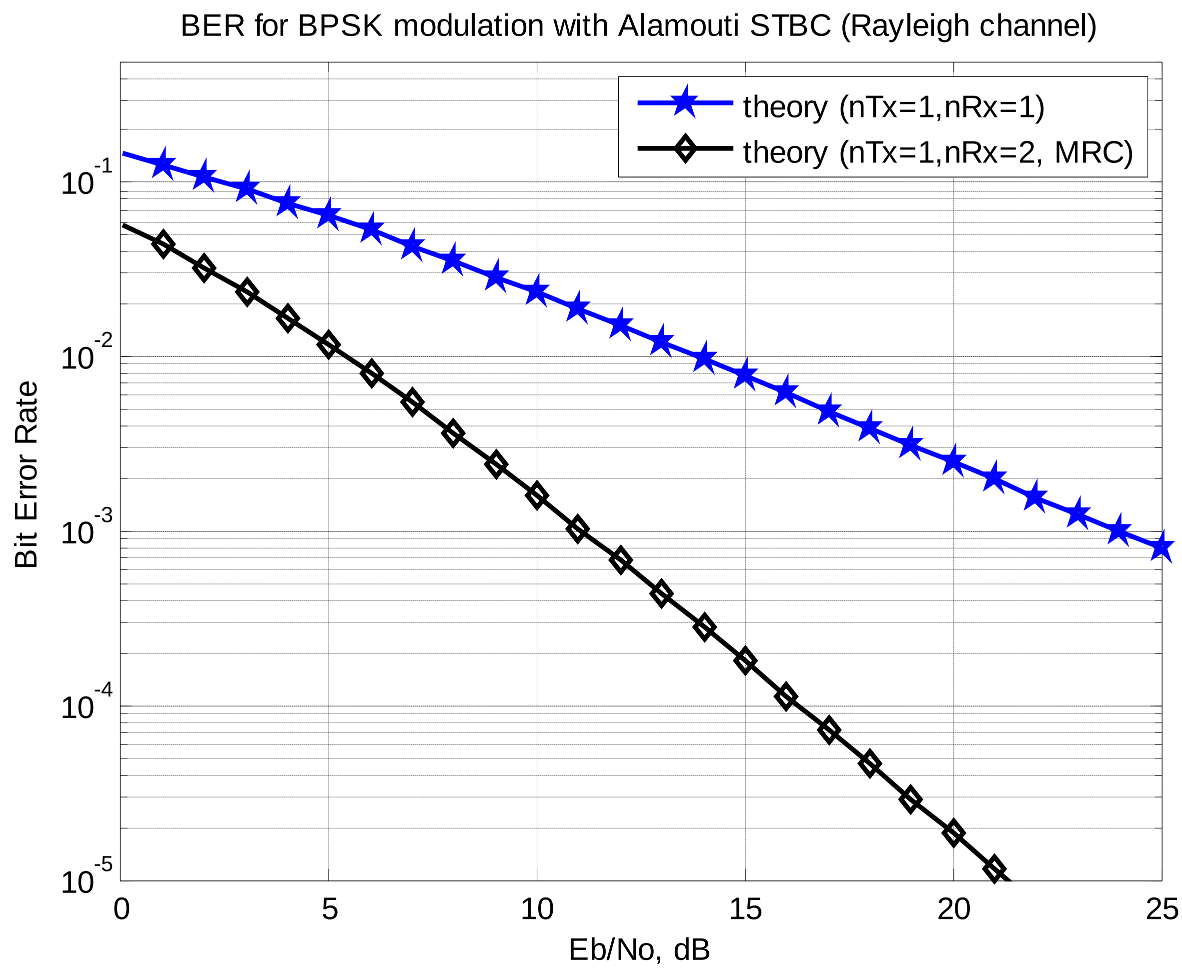


Fig. 5.2: BER performance of Non-diversity and Receiver diversity

Result of STBC (Space time block code) Alamouti code, when we compare transmit diversity and receiver diversity, performance of the system is same but 3dB losses in STBC (Space Time Block Code) Alamouti code.

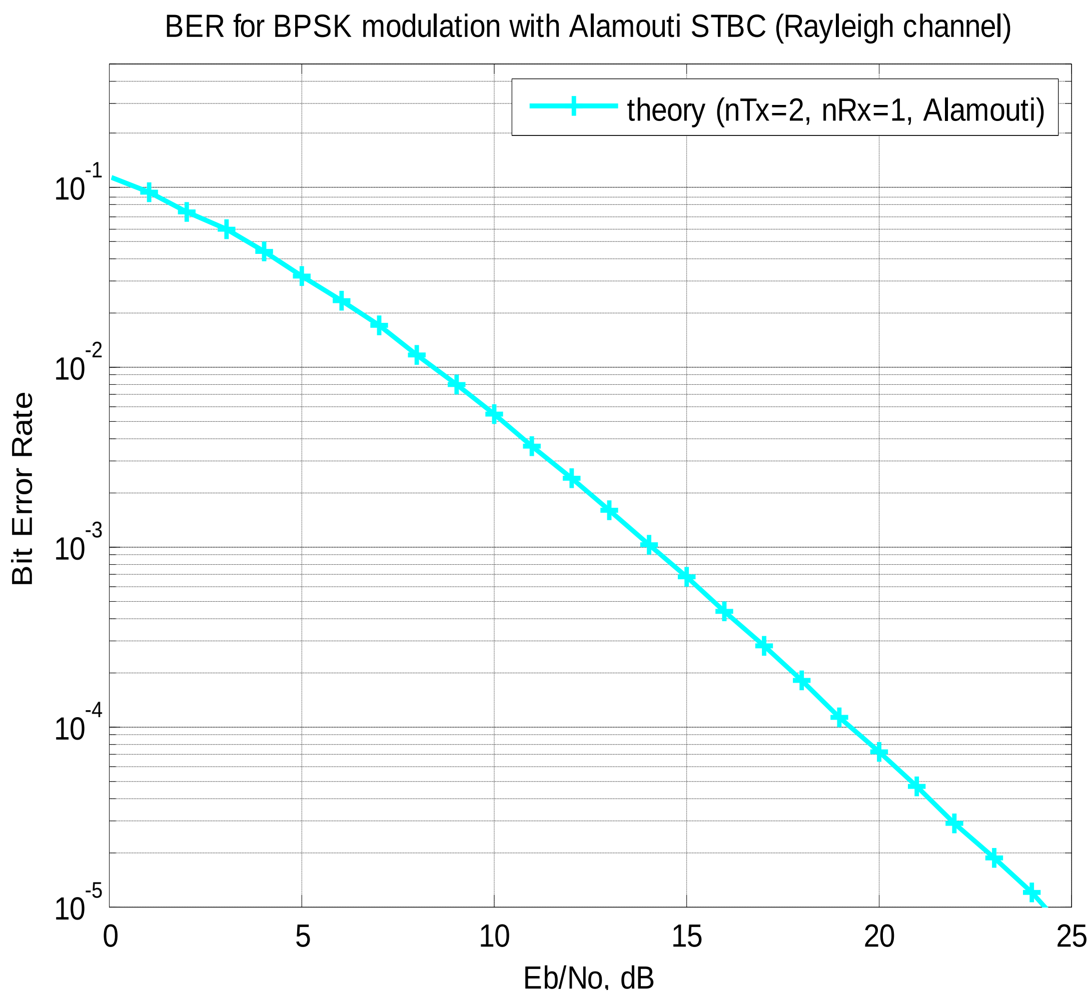


Fig. 5.3: BER performance of Space Time Block Code

In practice, channel is not ideal due to propagation delay error introduces in the system. After using the channel coding with STBC Alamouti code, we get the better performance in the wireless system as clearly shown in MATLAB simulated graph.

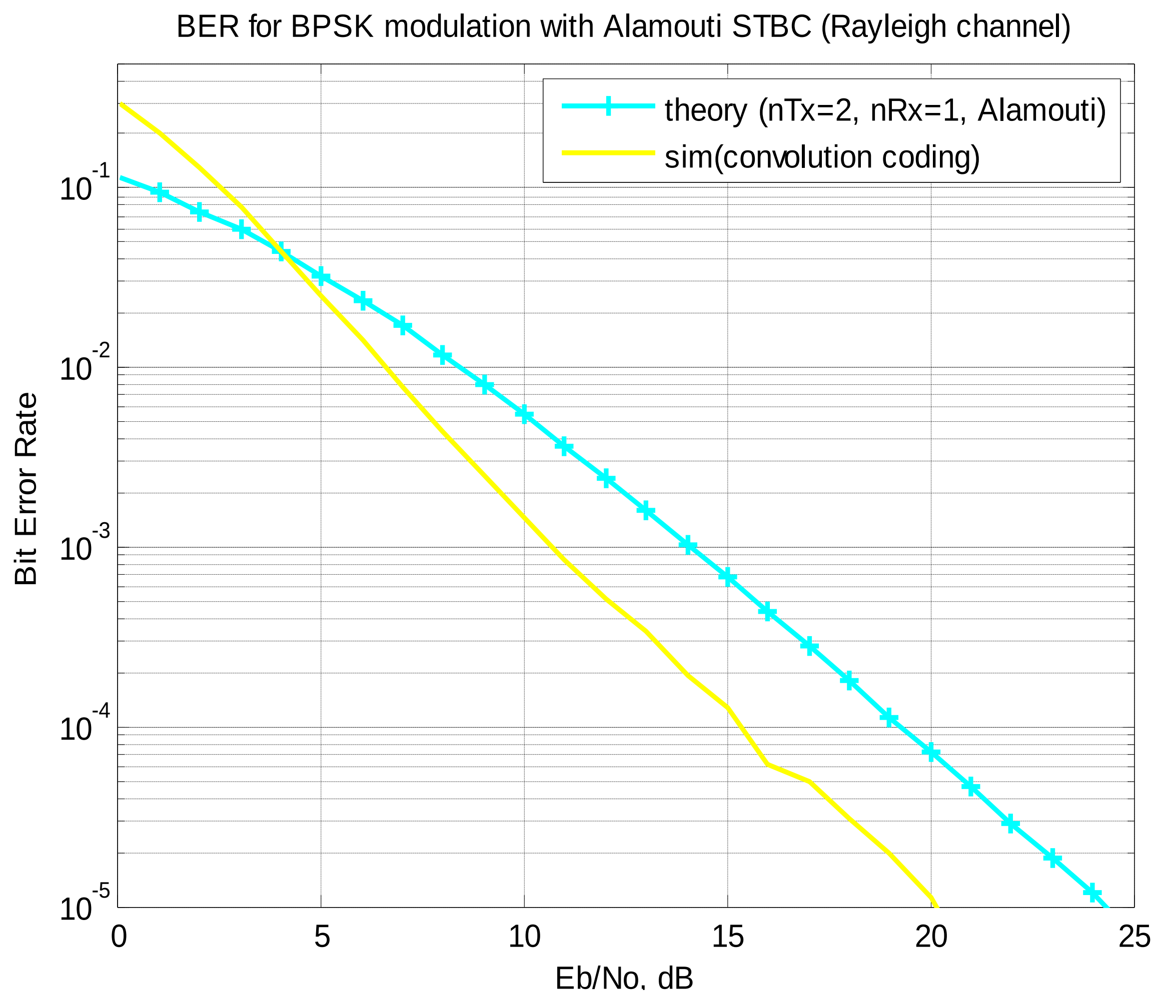


Fig. 5.4: BER performance of Space Time Block Code with channel coding

Overall performance of the system:

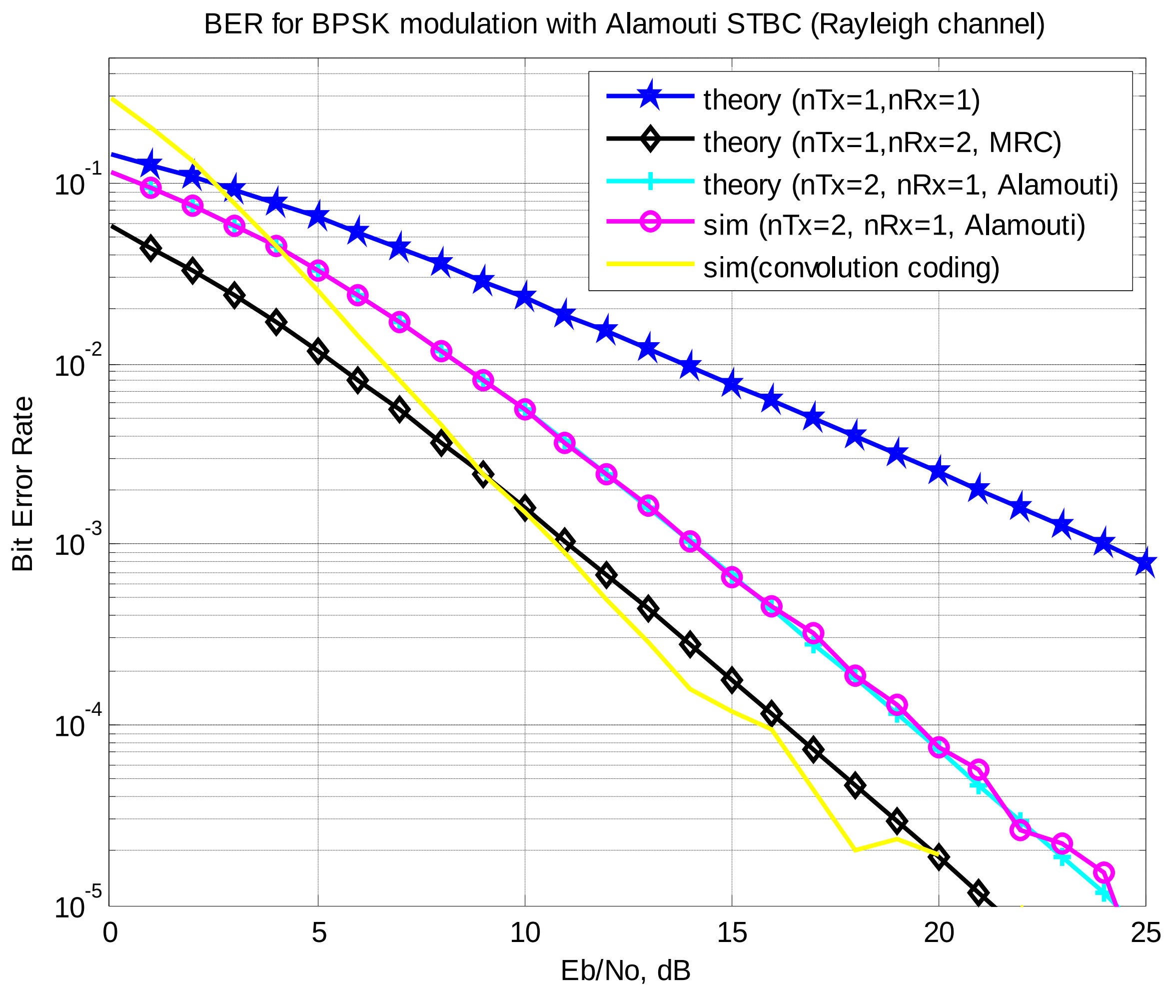


Fig. 5.5: BER comparisons of different diversity technique

Table 5.1: Comparisons of BER

SNR (dB)	BER for MRRC	BER for STBC	BER for STBC using CC
0	0.0581	0.1151	0.2952
1	0.0441	0.0939	0.2047
2	0.0328	0.0748	0.1292
3	0.0238	0.0582	0.0780
4	0.0169	0.0442	0.0450
5	0.0118	0.0329	0.0252
6	0.0081	0.0239	0.0142
7	0.0055	0.0170	0.0078
8	0.0037	0.0119	0.0045
9	0.0024	0.0082	0.0025
10	0.0016	0.0055	0.0015
11	0.0010	0.0037	0.0009
12	0.0007	0.0024	0.0005
13	0.0004	0.0016	0.0003
14	0.0003	0.0010	0.0002
15	0.0002	0.0007	0.0001
16	0.0001	0.0004	0.0001
17	0.0001	0.0003	0.0000
18	0.0000	0.0002	0.0000
19	0.0000	0.0001	0.0000
20	0.0000	0.0001	0.0000
21	0.0000	0.0000	0.0000
22	0.0000	0.0000	0.0000
23	0.0000	0.0000	0.0000
24	0.0000	0.0000	0.0000
25	0.0000	0.0000	0.0000
26	0.0000	0.0000	0.0000

From fig. 5.5 and table 5.1 it has been observed, at BER 10^{-3} STBC has 3dB penalty, as compared to MRCC. STBC with the Convolutional code has overcome this 3 dB penalty.

Chapter 6

Conclusion and Future Scope

Conclusion

Alamouti is the first person who gave the transmitter diversity technique, using two transmit antennas and one receive antenna. An obvious application of the scheme is to provide diversity improvement at all the remote units in a wireless system, using two transmit antennas at the base station instead of two receive antennas at all the remote terminals. The scheme does not require any feedback from the receiver to the transmitter and its computation complexity is a similar to MRRC. When compared with MRCC if the total radiated power is to remain the same, the transmit diversity scheme has 3dB disadvantage because of the simultaneous transmission of two distinct symbols from two antennas.

In digital communications, a channel code is a broadly used term mostly referring to the forward error correction Code. Channel coding is a viable method to reduce information rate through the channel and increase reliability. This goal is achieved by adding redundancy to the information symbol vector resulting in a longer coded vector of symbols that are distinguishable at the output of the channel.

In this thesis STBC with CC has been used. It has been seen that with this scheme, 3dB disadvantage of STBC is improved.

Future Scope

- Other modulation types, such as QAM, FSK, and DPSK, may be tested.
- Other channel coding types such as Reed Solomon (RS) code may be tested

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