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# Performance Comparison of Decoding Algorithms used in MIMO-OFDM Systems while Combating Channel Impairments

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**ABSTRACT : At wideband frequencies the channel impairments are more severe since the coherent bandwidth of the transmitted signal is smaller than that of the channel. As the characteristic of the channel is found to be time varying, the mitigation methods are too fragile against such impairments like fading, attenuation, noise, etc. So it is mandatory that the methods to be used in broadband systems to combat these problems should be robust against fast fading channels which may sink deeply in noise due to attenuation. In this paper a comparative study of two decoding algorithms that are found to be effective in a deep fading, noisy channel is furnished. The method aims at spatial diversity exploitation together with sub band multiplexing and channel coding to mitigate the impairments. In Multi Input Multi Output-Orthogonal Frequency Division Multiplexing (MIMO-OFDM) a broadband channel can be partitioned into narrowband channels, so that the frequency selective fading may change into flat fading which can be easily mitigated by OFDM technique. Then coding the data by Forward Error Correction (FEC) codes the error due to noise can be minimized. Zero forcing and Minimum mean square error estimate algorithms are used for decoding of MIMO-OFDM and their comparative performance is discussed here.**

**Keywords:** Orthogonal frequency divisionmultiplexing (OFDM),Convolutional code, Multi Input Multi Output (MIMO), Zero forcing, Minimum Mean Square Error (MMSE) estimate.

## I. INTRODUCTION

Future generation communication systems can only be operated in higher frequency bands since they have to incorporate more features and facilities to fulfill the rising demands of the users. As the frequency is very high the time period of a pulse is so short that they experience deep fading and other impairments. Paradoxically this channel condition may be exploited to improve signal quality also. This can be accomplished by assuming the channel as independently and identically distributed (i.i.d). Then multiple antennas can be used to exploit this channel behavior. The spectral efficiency of the system is also quite high because all the transmitting antennas are using the same band of frequencies. This type of multiple antenna system is known as Multiple Input Multiple Output (MIMO) [1] system.

Since Orthogonal Frequency Division Multiplexing splits a wideband carrier into narrow band sub carriers this in turn changes a frequency-selective fading into a set of parallel-flat frequency fading channels. Now the time period of the sub carrier is longer than the coherence time of the channel. This renders multi-channel equalization particularly simple [1]. So MIMO-OFDM [2] can be used as a multiplexing and modulating technique to ensure good quality of signal reception. In such a system with N subcarriers the individual data streams are first passed through OFDM modulators which perform an IFFT on blocks of length N followed by a parallel-to-serial conversion. A DFT processor can be used to generate the orthogonal subcarriers. A cyclic prefix (CP) is added to the transmitting signal which is a guard interval, and it serves to eliminate interference between symbols (ISI). Then this data is send through the multiple antennas.

There are two types of errors that degrade the performance of any communication system. They are burst errors and random errors. Interleavers are used to spread the burst error into random, and this makes the type of error only as random. In this method an interleaver of appreciable length is used. To combat the other type of error Convolutional codes [3] are used here since they are the forward error correction codes and found to be more robust against random errors. The main decoding strategy for Convolutional codes is based on Viterbi algorithm. The Zero forcing and Minimum Mean Square Error Estimate are the two simple decoding techniques in MIMO systems. The performance comparison of these two techniques in deep fading and high noisy channel environments is carried out.

## II. MIMO-OFDM

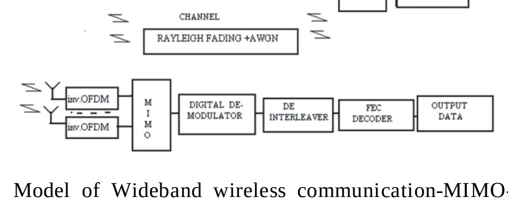


Fig. 1. Model of Wideband wireless communication-MIMO-OFDM.

Multiple antennas at the transmitter and receiver are mainly used for achieving appreciable 'spectral diversity gain' as well as an increase in data rate. The zero forcing [4] and MMSE algorithm [5] is used in this design since they are very simple and ensures an increase in the data rate while minimizing the error due to noise. In broadband channels the multipath characteristic of the environment causes the MIMO channel to be frequency selective. OFDM can transform such a frequency selective channel into a set of parallel frequency-flat channel, and this reduces the effect of fading due to multipath delay spread. The guard period introduced by adding cyclic prefix takes care of inter symbol interference (ISI). The combination of the two powerful techniques, MIMO and OFDM [6], is very attractive, and has a most promising broadband wireless access scheme. Channel estimation is carried out by the pilot symbols embedded in the OFDM signal. The computational complexity of estimation increases rapidly by increasing the number of antennas at the transmitter and receiver, and the performance substantially degrades with the estimation error.

A MIMO system takes the advantage of the spatial diversity obtained by spatially separated antennas in a dense multipath scattering environment. As shown in Fig.1, in the transmitter, the source bit stream is first coded by channel codes meant for error correction and then the coded bits are interleaved to spread the bits randomly and then mapped into a constellation by the digital modulator, (*e.g.*, PSK, QPSK, QAM etc). Now the modulated and coded data is send to the MIMO-OFDM transmitter. Here, each of the parallel output symbol streams undergoes orthogonal frequency division multiplexing and being transmitted through multiple antennas simultaneously.

The received symbol stream from IF/RF components over the receiver antennas are first OFDM demodulated in which; the refined frequency pilots from all the receiver antennas are used for channel estimation. The estimated channel matrix aids the decoding of transmitted bits in correct manner. The inverse- OFDM performed symbols are then digital-demodulated and decoded after de-interleaving. At last, the decoded source bit streams are taken to the data sink. Bit error rate can be calculated by comparing the original data with the estimated output.

In Zero Forcing as well as in MMSE- MIMO, the information symbols are mapped in parallel streams and each stream is transmitted from a separate transmitter element. If there are a sufficiently large number of scatters in the channel, and the antenna spacing is reasonably large, then the fading across paths from each transmitter to receiver antenna will be independent. The channel delay spread is assumed to be small enough to cause flat fading on the signals transmitted from various antennas. In the following figure (Fig. 2) a MIMO-OFDM system is shown. Here two transmitter and two receiver antennas are used. In MIMO the number of receiver antennas may be equal to or higher than the number of transmit antennas. The information carrying signals are multiplexed among all the transmitting antennas and transmit simultaneously. While, at the receiver end the signals are received by all receivers, with different channel responses (different level of fading due to multipath delay spread). The relation between the transmitted signal and the received signal is given as:

**Rayleigh faded Channel**

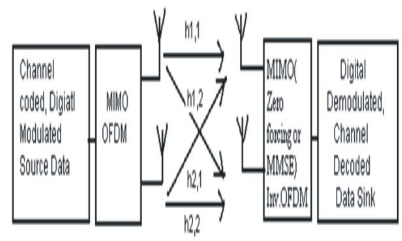


Fig. 2. MIMO Transmitter-Receiver model (Multi Channel-Rayleigh faded).

$$Y = Hx + n \qquad \text{... (1)}$$

where, $y$ is a $N \times 1$ vector whose elements represent the received signal at each receiver antenna element ($N$), $H$ is an $M \times N$ channel matrix representing MIMO Quasi static flat Rayleigh fading that is estimated through pilot symbols in the OFDM, while $x$ is the $M \times 1$ vector of transmit symbols ($M$) and n is the $N \times 1$ noise vector of independent AWGN samples. In the receiver, either zero forcing (ZF) decoding or the minimum mean square error (MMSE) [4], [5] criterion is used. To decode the transmitted symbols of the first layer, the receiver initially estimates the channel matrix H using pilot symbols. It then considers symbols from a particular transmit antenna as desired signal and suppresses signals from other antennas using array processing. That is, while decoding the symbols from transmit antenna $j$, the signals from all other antennas are suppressed, and, the receiver multiplies vector $y(1)$ with $W_j$ [4] (an $N \times 1$ weight vector) given by

$$W_j = H^{+}_{j}$$

here $H^{+}_{j}$ is the $j^{th}$ row of pseudo inverse of matrix $H$. The MIMO-OFDM offers both spatial and frequency diversity in broadband communications. In this paper, it is already stated, the system is composed of 2 transmitting antennas and 2 receiving antennas (Fig. 2). The frequency selective fading channel between each pair of transmitter antennas and receiver antennas has $L$ (here 4) independent delay paths and the same power delay profile. The MIMO channel is constant over each OFDM block period [7]. The channel impulse response hij from transmit antenna $i$ and receive antenna $j$ can be modeled as:

$$h_{ij} = \sum_{l=0}^{L-1} \alpha_{i,j}(l)\delta(t - t_1) \qquad \text{... (2)}$$

where $t_1$ is the delay of the $l^{th}$ path, and $\alpha_{i,j}(l)$ is the complex amplitude of the $l^{th}$ path between transmit antenna $i$ and receive antenna $j$. The time delay $\tau_l$ and variance $\delta_l^2$ are assumed to be the same for each transmit-receive channel link. The power of the $L$ paths are normalized, such that

$$\sum_{l=0}^{L-1} \delta^2{}_l = 1 \qquad \text{... (3)}$$

The channel frequency response is given by:

$$H_{i,j}(j) = \sum_{l=0}^{L-1} \alpha_{i,j}(l) e^{-j2\pi f \tau 1} \qquad \text{... (4)}$$

In order to improve the BER performance of the proposed two transmitter and two receiver antenna systems, first zero forcing is used as the decoding strategy and then MMSE equalizer is used for decoding in order to compare the performance against the same multipath delay spread and channel noise. Both the methods are proved to be effective, but in the deep fading channel some other mitigating method are to be included for good result. Further the error in the received bits can be minimized by clubbing error correction codes and interleaver.

**Zero forcing and MMSE equalizer decoders**

Let us now try to understand the math for extracting the two symbols which interfered with each other. In the first time slot, the received signal on the first receive antenna is,

$$y_1 = h_{1,1}\, x_1 + h_{1,2}\, x_2 + n_1 = [h_{1,1} \;\; h_{1,2}] \begin{bmatrix} x1 \\ x2 \end{bmatrix} + n_1 \qquad \text{...(5)}$$

The received signal on the second receive antenna is

$$y_1 = h_{2,1}\, x_1 + h_{2,2}\, x_2 + n_2 = [h_{2,1} \;\; h_{2,2}] \begin{bmatrix} x1 \\ x2 \end{bmatrix} + n_2 \qquad \text{... (6)}$$

where $y_1$ and $y_2$ are the received symbol on the first and second antenna respectively, $h_{1,1}$ is the channel impulse response from the first transmit antenna to first receiver antenna, $h_{1,2}$ is the channel impulse response from second transmit antenna to first receiver antenna, $h_{2,1}$ is the channel impulse response from second transmit antenna to first receive antenna, $h_{2,2}$ is the channel impulse response from second transmit antenna to second receive antenna, and $x_1$ and $x_2$ are the transmitted symbols and $n_1$, $n_2$ are the noise on first and second receive antennas. We assume that the receiver knows $h_{1,1}$, $h_{1,2}$, $h_{2,1}$ and $h_{2,2}$ through channel estimation. The receiver also knows $y_1$ and $y_2$. The unknown are $x_1$and $x_2$. From the knowledge of $y$

*i.e.* $y = Hx + n$

One can solve for $x$, the only processing that have to be carried out is to find a matrix $W$ which satisfies the condition $WH = I$ while $I$ is the identity matrix. The zero forcing estimation for finding out.

$$W = (H^H H)^{-1} H^H$$

And for the Minimum Mean Square Error (MMSE) linear detector for meeting this constraint is given by,

$$W = (H^H H + N_0 I)^{-1} H^H$$

where $N_0$ is the noise variance and $I$ is the identity matrix. This matrix is also known as the pseudo inverse for a general $m \times n$ matrix.

## III. OFDM SYSTEM

The block diagram of a typical OFDM [6], [7] transmitter is shown in Fig. 3. The data source is assumed to be a sequence of discrete digits: $\{x_0, x_1, x_2, \ldots,\}$

where, $x_i$ = 1 or 0, i = 0, 1, 2, 3, ... and

$P(x_i = 1) = P(x_i = 0) = 1/2$

The output of the Convolutional encoder [9] is fed into a block interleaver that spreads the coded bits randomly. Then the bits are digital modulated (PSK, QPSK, QAM etc) [12] and send for OFDM technique which is a serial-to parallel converter first, that partitions the input data arriving at the rate $R$ into $N$ parallel information symbols each at a reduced data rate of $R/N$ [10]. The number of bits in each of the $N$ output sequences of a serial-to-parallel converter is determined by the constellation of the signal mapper ie Digital modulator.

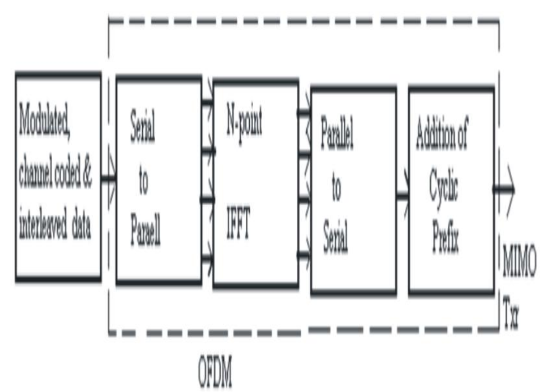


Fig. 3. Model of OFDM-MIMO Transmitter.

For example, when a BPSK [14] modulation is used, each output sequence carries one bit of information. For QPSK [14] and 16-QAM, each symbol carries 2 and 4 bits of information respectively. The output of the digital modulator can be thought of as a discrete complex signal which during any arbitrary OFDM symbol duration is a vector of N complex numbers given by:

$$C = [c_0, c_1, \ldots., c_{N-1}] \qquad \text{... (7)}$$

This discrete signal representing the data is fed to an $N$-point IFFT [11] block to obtain a transformed discrete signal given by:

$$D = [d_0, d_1, \ldots, d_{N-1}] \qquad \text{... (8)}$$

The relationship between the input and the output discrete signals of IFFT block is given by the well known Inverse Discrete Fourier Transform (DFT) [6].

That is

$$Dn = \sum_{m=0}^{N-1} C_m e^{j2\pi mn/N}; n = 0,1,...., N-1 \quad ... (9)$$

where it is assumed that the input sequence $C$ is periodic with period equal to $N$. Then parallel-to-serial conversion takes place and all the sub carriers are added together to a composite signal form. In order to combat multipath delay spread a portion (20% in this work) of this composite signal is added as prefix, so that symbol time ($t_s$) extends from ts to $t_s + t_c$.

The transmitted analog signal after digital to analog converter is, $x(t)$ and is given by

$$x(t) = \sum_{m=0}^{N-1} Ame^{j2\pi fmt} \quad ... (10)$$

where $f_m = m/(N\,T_b)$ and $t = nT_b$ is the original duration of bit and fm is the frequency of the $m^{th}$ carrier. The complex continuous-time modulating signal given in (10) can be represented as [6]:

$$x(t) = m_I(t) + j\, m_Q(t) \quad ... (11)$$

where $m_I(t) = Re\{x(t)\}$ ... [12(*a*)]

is the in phase component of the transmitted signal and

$$m_Q(t) = Im\{x(t)\} \quad ... [12(b)]$$

is the Quadrature component of the transmitted signal. This analog signal should be up converted and amplified before transmission. So it is modulated with a strong carrier $fc$, which is represented as

$$X(t) = m_I(t) \cos 2\pi fct + m_Q(t) \sin 2\pi fct \quad ... (13)$$

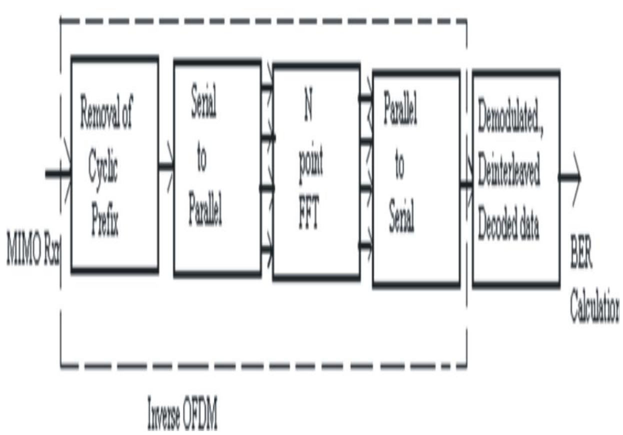


Fig. 4. OFDM Receiver model.

The transmitted signal while going through the deep fading channel undergo severe impairments and being received, $\{y(t)\}$ at the receiver antennas with much degradation it amplitude, phase etc.

Thus the objective at the receiver is to recover $m_I(t)$ and $m_Q(t)$ from $y(t)$. This can be accomplished using the in phase and quadrature-phase detector. The MIMO decoders and pilot estimators in OFDM are used to reduce the signal loss to certain extent. The OFDM demodulated baseband signals $m_I(t)$ and $m_Q(t)$ are sampled and fed to the *A*/*D* converter which produces a vector of complex numbers $[d_0, d_1, ..., d_{N-1}]$. Then the Cyclic prefix is removed. The complex numbers $[c_0, c_1, ..., c_{N-1}]$ can then be recovered using FFT operation and is subjected for de-mapping, de-interleaving and decoding, to recover the transmitted information from this OFDM demodulated output.

The following table gives an idea about the system parameters [12] used for the simulation of this method.

**Table 1: System parameters.**

| | |
|---|---|
| Number of FFT points | 64 |
| Number of sub-carriers | 52 |
| Number of data sub-carriers | 48 |
| Number of pilot sub-carriers | 4 |
| Cyclic prefix | 16 chips |
| Modulation scheme | BPSK ,QPSK |
| Coding | ½ Convolutional, constraint length 9 |
| No. of Transmitting and Receiving Antenna | 2, 2 |
| MIMO Detection | Zero Forcing, MMSE algorithm |
| Interleaver size | 10 × 10 |

## 1V. CONVOLUTIONAL CODES

Convolutional codes [3] [8] are widely used as forward error correction codes in practical communication systems for random error correction. It is defined by three parameters: *n, k,* and *K*. It process input data *k* bits at a time and produces an output of *n* bits for each incoming k bits. Since the convolutional codes have memory, which is characterized by the constraint factor *K*, the current *n* bit output of an (*n, k, K*) code depends not only on the value of current block of *k* input bits but also on the previous $K-1$ blocks of *k* input bits. The main decoding strategy for convolutional codes is based on the widely used Viterbi algorithm. As a result of the wide acceptance of convolutional codes, there have been several approaches to modify and extend this basic coding scheme.

The Convolutional encoder [9] used in this paper is shown in Fig. 5. The information bits are fed in small groups of *k*-bits at a time to a shift register. The output encoded bits are obtained by modulo-2 addition (XOR operation) of the input information bits and the contents of the shift registers which are a few previous information bits. Here $k = 1$ and $n = 2$ and $K = 9$ The code rate of (2, 1, 9) Convolutional code of 753,561 standard is ½.

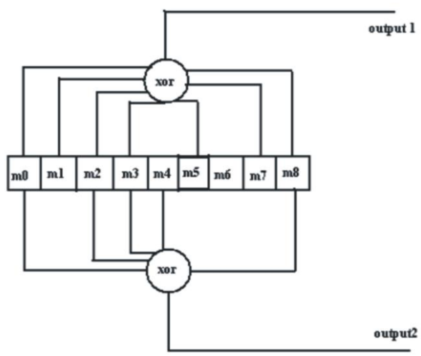


Fig. 5. (753, 561) Convolutional encoder.

The shift register of the encoder is initialized to all-zero-state before encoding operation starts. As each (input) bit enters into the shift register, it is convoluted with the previous one; by the way it is shown in the Fig. 5. The encoded, output bits from encoder are transferred to the next block for interleaving.

For the decoder design, trellis diagram representation is mandatory. A trellis is a tree like structure that is illustrating the state transitions of the encoder for a particular input. *i.e.*, it shows the present state, next state, and probable output for an input bit of 1 or 0. For 753 and 561, ½ rate Convolutional code the trellis diagram has 256 states ($2^{9-1}$ since the constraint length is 9), so the received code word is analyzed with this trellis diagram and the maximum likelihood probability of the transmitted binary information is found out. Viterbi algorithm is based on this maximum likelihood analysis and the same technique is followed in this work too.

**Hard-Decision Viterbi Algorithm**

There are mainly two types of convolutional decoding. One is hard decision and the other is soft decision Viterbi coding.

For a convolutional code, the input sequence $x$ is "convoluted" with the previous $K - 1$ bits (according to the encoder polynomial) to the encoded sequence $c$. Sequence $c$ is further processed and transmitted across a noisy channel and the received sequence $r$ is obtained. The Viterbi algorithm computes a maximum likelihood (ML) estimate on the estimated code sequence $y$ from the received sequence $r$ such that it maximizes the probability $p(r|y)$ that sequence $r$ is received conditioned on the estimated code sequence $y$. Sequence $y$ must be one of the allowable code sequences and cannot be any arbitrary sequence.

For a rate $r$ convolutional code, the encoder inputs $k$ bits in parallel and outputs $n$ bits in parallel at each time step. The input sequence is denoted as

$$x = [x_0(1), x_0(2), ..., x_1(k), x_{L+m-1}(1), ..., x_{L+m-1}(k)]$$

and the coded sequence is denoted as

$$c = [c_0(1), c_0(2), ..., c_1(n), c_{L+m-1}(1), ..., c_{L+m-1}(n)]$$

where $L$ denotes the length of input information sequence and $m$ (which is nothing but $K - 1$)denotes the maximum length of the shift registers. Additional m zero bits are required at the tail of the information sequence to take the convolutional encoder back to the all-zero state. It is required that the encoder start and end at the all-zero state. The subscript denotes the time index while the superscript denotes the bit within a particular input k-bit or output n bit block. The received and estimated sequences r and y can be described similarly as

$$r = [r_0(1), ..., r_0(n), \ldots, r_{L+m-1}(1), ..., r_{L+m-1}(n)]$$

and from this received bits the output bits are estimated as

$$y = (y_0(1), ..., y_0(n), ..., y_{L+m-1}(1), ..., y_{L+m-1}(n)).$$

The Viterbi algorithm utilizes the trellis diagram to compute the path metrics. Each state (node) in the trellis diagram is assigned a value, the partial path metric. The partial path metric is determined from state $s = 0$ at time $t = 0$ to a particular state $s = k$ at time $t = 0$. At each state, the "best" partial path metric is chosen from the paths terminated at that state. The "best" partial path metric may be the one with smaller metric.

The selected metric represents the survivor path and the remaining metrics represent the non survivor paths. The survivor paths are stored while the non survivor paths are discarded in the trellis diagram. The Viterbi algorithm selects the single survivor path left at the end of the process as the ML path. Trace-back of the ML path on the trellis diagram would then provide the ML decoded sequence.

The hard-decision Viterbi algorithm (HDVA) can be implemented as follows:

*Sk*, *t* is the state in the trellis diagram that corresponds to state *Sk* at time *t*. Every state in the trellis is assigned a value denoted *V*(*Sk*,*t*).

1. (*a*) Initialize time $t = 0$.

   (*b*) Initialize $V(S0,0) = 0$.

2. (*a*) Set time $t = t + 1$.

   (*b*) Compute the partial path metrics for all paths going to state *Sk* at time *t*.

Find the $t^{\text{th}}$ branch metric

$$M(r_t / y_t) = \sum_{j=1}^{n} M(r_t^{(j)} / y_t^{(j)})$$

This is calculated from the Hamming distance

$$\sum_{j=1}^{n} | r_t^{(j)} - y_t^{(j)} |$$

Second, compute the $t^{\text{th}}$ partial path

$$M^t(r / y) = \sum_{i=0}^{t} M(r_i / y_i)$$

This is calculated from $V(Sk, t-1) + M(rt \mid yt)$.

(*b*) Compute the partial path metrics for all paths going to state *Sk*

3. (*a*) Set $V(Sk, t)$ to the "best" partial path metric going to state *Sk* at time *t*. Conventionally, the "best" partial path metric is the partial path metric with the smallest value.

   (*b*) If there is a tie for the "best" partial path metric, then any one of the tied partial path metric may be chosen.

4. Store the "best" partial path metric and its associated survivor bit and state paths.

5. If $t < L + m - 1$, return to Step 2.

The result of the Viterbi algorithm is a unique trellis path that corresponds to the ML codeword.

As an example, the trellis diagram of the code shown in Fig. 5, has two hundred and sixty four states and each state has two incoming and two outgoing branches. At any depth of the trellis, each state can be reached through two paths from the previous stage and as per the VA, the path with lower accumulated path metric is chosen. In the process, the 'accumulated path metric' is updated by adding the metric of the incoming branch with the 'accumulated path metric' of the state from where the branch originated. No decision about a received codeword is taken from such operations and the decoding decision is deliberately delayed to reduce the possibility of erroneous decision. Typically, the delay in decision making is $m \times k$ code words where $L$ is an integer, *e.g.* 5 or 6. For the code in Fig. 5 the decision delay (trace back length) of $5 \times 9 = 45$ code words may be sufficient for most occasions.

This means, we decide about the first received codeword after receiving the 46th codeword. The decision strategy is simple. Upon receiving the 46th codeword and carrying out a trace back to the initial state, we compare the 'accumulated path metrics' of all the states (sixty-four in our example) and chose the state with minimum overall 'accumulated path metric' as the 'winning node' for the first codeword and declare this codeword as the most likely transmitted first codeword.

**Interleaver**

Most FEC's are robust against random errors but not burst errors. An interleaver spreads the coded information, bits-by-bits, so that the transmitted bits of information may not have any correlation. If burst error occurs in the channel or in the transmitter, it corrupts a long string of the transmitted information. But at the receiver before decoding this corrupted information it is once again interleaved (De-interleaver) and changes into random error. Any FEC which is meant for random error correction can correct such an error.

There are many types of interleavers [8]. One common interleaver is row/column interleaver. This interleaver reads m rows of length n of the information bits, and then it transmits the n columns of length m. At the receiving end the n columns of length m are used to reconstruct the m rows.

Data from Convolutional Encoder is entered in row wise into the Interleaver as shown below. Therefore the Convolutional coded information is interleaved, [9] and sends to the next block for digital modulation in the way shown in the matrix form representation of the $10 \times 10$ interleaver.

$$\begin{pmatrix} c_0 & c_1 & c_2 & c_3 & c_4 & c_5 & c_6 & c_7 & c_8 & c_9 \\ c_{10} \ldots\ldots & & & & & & & & & \\ \vdots & & & & & & & & & \\ \vdots & & & & & & & & & \\ c_{90} & c_{91} & .. & .. & .. & & & & c_{98} & c_{99} \end{pmatrix}$$

**10 ×10 Interleaver**

Data is entered in column wise as c0, c1,c2, c3,…c9 in the first row, $c10$, $c11$, $c12$, …, $c18$ in the second row and so on. It is taken from the Interleaver as $c0$, $c10$, $c20$, $c30$ …, $c11$, $c21$, $c31$, … $c89$, $c99$. So it is very clear that the coded information sequence is well spread. In the receiver, the MIMO-OFDM detected signals are (they are assumed to be corrupted at the channel) passed through the de-interleaver and the reverse process takes place.

$$\begin{pmatrix} c_0 & c_{10} & c_{20} \ldots\ldots c_{80} & & c_{90} \\ c_1 & .. & .. \; .. \; .. \; .. & & c_{91} \\ \vdots & & & & \\ \vdots & & & & c_{98} \\ c_9 & c_{19} & .. \; .. \; .. & c_{88} & c_{99} \end{pmatrix}$$

**10 ×10 De-Interleaver**

So the burst error which corrupts a long stream of information is interleaved and spread as random error. Now the type of error is turned into random and it is easily corrected by the Viterbi decoder.

**Modulation Scheme**

PSK or QPSK[13] modulations choice depends on various factors like the bit rate and sensitivity to errors. Coherent modulation is achieved by transmitting the IQ constellation data vectors with absolute phase angles. *i.e.* if BPSK is used then 0° or 180° and QPSK is used then $\pi/4$, $3\pi/4$, $5\pi/4$ and $7\pi/4$ would be transmitted. At the receiver, it would compare the received phase to the above given angles. Phase rotations and amplitude scaling, is, however, overcome by using the channel equalization before demodulation. Tracking of the channel requires continual updates in the channel equalization, thus regular pilot symbols must be inserted into the transmission system.

## V. RESULTS

The deep fading condition is modeled by using Rayleigh fading of delay length 10 tap and 20 tap respectively. AWGN is also added and OFDM of N = 64 is used. Cyclic prefix of 16 bit is added to combat ISI. Bit Error Rate is controlled using Convolutional code of constraint length 9. Mat lab 2007*b* is used for the simulation. As mentioned before, burst errors deteriorate the performance of any communications system. The occurrence of burst errors is due to either by impulsive noise or by deep frequency fades. The interleaver smartly splits the burst errors into random errors, so no other coding is necessary to combat burst errors (*e.g.*, RS Codes, RM Codes etc).

The zero forcing is the very simple detection technique of MIMO. The decoding is simply based on the knowledge of channel impulse response. It is obtained from the pilot signals used in the transmission. Here the number of transmitting and receiving antennas is two, so the channel response is not much complicated and also the transmission rate is increased by the factor two. OFDM takes care of frequency selective fading and also minimizes the symbol interference due to the inclusion of cyclic prefix.

Under the same channel conditions and coding, the same process is repeated for MMSE detection technique of MIMO. Here the noise effect is also estimated and that is used in the decoding process. Since same number of transmitting and receiving antennas and same number of bits are used for simulation the transmission rate is same as that of the previous method. But the bit error rate is much decreased in the case of MMSE estimation. The comparative performance analysis is made with all simulation results.

In the first graph Fig. 6, the QPSK modulated MIMO-OFDM with convolutional code simulation result is given. A Rayleigh fading of 10-tap delay and 20 tap delay are used and MIMO decoding is by using Zero forcing algorithm.

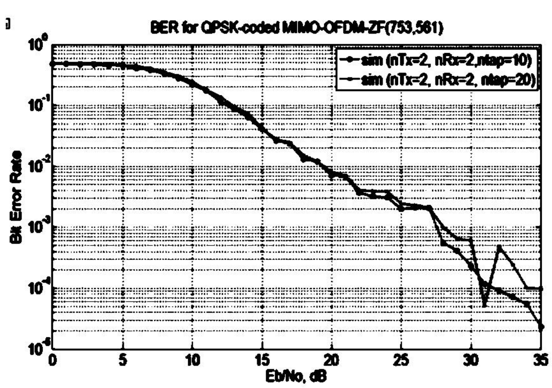


Fig. 6. BER Vs Eb/No, QPSK-MIMO-OFDM-ZF.

Then in the second, Fig. 7, under same conditions MMSE equalized MIMO-OFDM is shown.

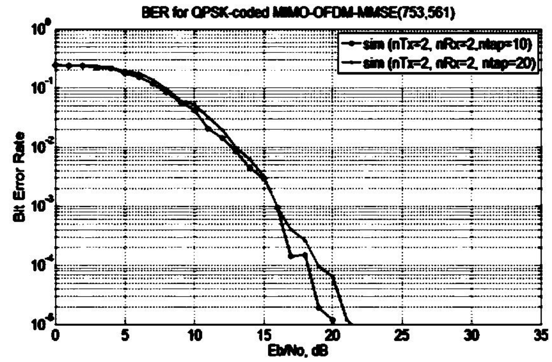


Fig. 7. BER Vs Eb/No, QPSK-MIMO-OFDM-MMSE.

From the above two graphs it is very clear that MMSE MIMO is superior to ZF MIMO in view of reduction of SNR to obtain a BER of $10^{-4}$. In severe fading channel (say 20-tap) in order to achieve a BER of $10^{-4}$ , 35dB is required in ZF-MIMO while it is less than 19dB if the decoding is according to MMSE.

Similarly in 10-tap delay channel for a SNR of 20dB it is easy to get a BER of $10^{-5}$ while in ZF-MIMO it is more than 35dB to be spent to achieve the same BER. If PSK is used then the requirement of SNR can further be reduced.

The following simulated results are representing

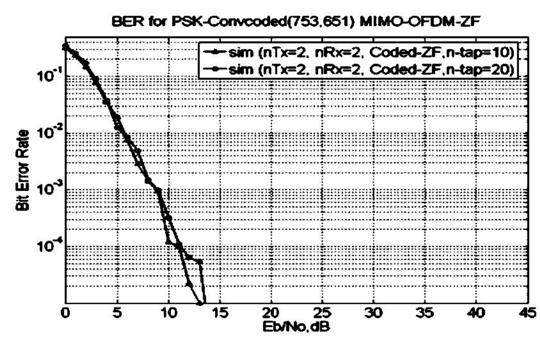


Fig. 8. BER Vs Eb/No, PSK-MIMO-OFDM-ZF.

PSK modulated convoluted codes under the same channel conditions.

For a delay of 10-tap if ZF is used as decoding strategy of MIMO, BER of $10^{-5}$ can be attained with just 14dB and if MMSE is used then at 13 dB it is possible to get same error rate. More or less is the case of 20-tap delay also.

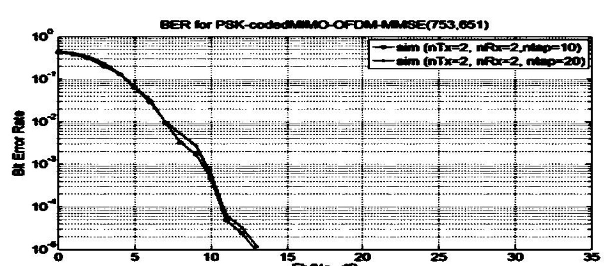


Fig. 9. BER Vs Eb/No, PSK-MIMO-OFDM-MMSE.

## VI. CONCLUSION

The performance of the two algorithms is quite well for MIMO-OFDM in deep fading condition. Pilot symbols in the OFDM signal gives a good estimate of the channel and this helps a better performance of both the algorithms used in the MIMO technique. Even though the zero forcing MIMO algorithm ensures double throughput, the data rate is unchanged because it is reduced to half while encoding by Convolutional code. So overall there is no reduction in data rate. The same code rate for MMSE but it enjoys more error correction capability for less Signal-to-Noise ratio. The simulation of the entire work is done on MATLAB 2007*b* platform. The zero forcing MIMO-OFDM with QPSK modulation and MMSE MIMO-OFDM for the same conditions are first analyzed. The number of bits in the transmitted stream is of $10^5$ and the performance is again repeated for digital modulation of PSK. Among these two modulation techniques QPSK combines two bits and reduces the transmission rate by two but the price to be paid is more SNR. On the other hand PSK is affected less by noise, so the SNR requirement is also less. But overall analysis shows, under deep fading conditions, the MMSE-MIMO-OFDM, Coded by Convolutional code and modulated by PSK is the better choice.

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