Performance Analysis and PAPR Reduction of Coded OFDM (with RS-CC and Turbo coding) System using Modified SLM, PTS and DHT Pre-coding in Fading Environments

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Abstract— In wireless communication, parallel transmission of symbols using multi carriers is used to achieve high efficiency in terms of throughput and better transmission quality. Orthogonal Frequency Division Multiplexing (OFDM) is one of the techniques for parallel transmission of information. In multipath environment the performance of orthogonal frequency division multiplexing degrades which can be improved by introducing some kind of channel coding. Coded OFDM (COFDM) is the new candidate for application such as Digital audio Broadcast (DAB) and Digital Video Broadcast (DVB-T) due to its better performance in fading environments. However high peak to average power ratio (PAPR) is a major demerit of OFDM system, it leads to increased complexity and reduced efficiency of RF amplifier circuit. In this paper Modified Selective mapping (SLM), Partial Transmit sequence (PTS) and Discrete Hartley Transform (DHT) precoding schemes are proposed for PAPR reduction, where SLM, PTS and DHT precoding schemes are used in conjunction with post clipping and filtering processes. However clipping can degrade the BER performance but the degradation in performance can be compensated by using OFDM with channel coding; here we have used Reed Solomon (RS) codes along with convolution codes (CC) used as serial concatenation and TURBO codes as parallel concatenation code for channel coding purpose. The BER performances are simulated for Additive white Gaussian (AWGN), Rayleigh, Rician and Nakagami (m=3) channels and Complementary cumulative distribution functions (CCDF) curves are simulated for modified as well as ordinary SLM, PTS and DHT precoding techniques. The COFDM system implemented here is as per IEEE 802.11a.

Key-Words:- RS-CC codes, Turbo codes, Rayleigh, Rician, Nakagami-*m*, DHT pre-coding, PAPR, PTS, SLM, Clipping- filtering.

1 Introduction

Orthogonal frequency division multiplexing (OFDM) is a obvious choice for high speed transmission since quite a time. OFDM is a special case of multicarrier transmission, where a single data stream is transmitted over a number of lower rate subcarriers [1], [2]. OFDM is most preferred candidate for high speed communication in multipath environment due to its robustness to Inter symbol interference (ISI). It avoids ISI problem by sending many low speed transmissions simultaneously with addition of cyclic prefix. OFDM is presently used in a number of commercial wired and wireless applications such as digital audio broadcast (DAB), digital video Broadcast (DVB), and wireless LAN, MAN and DSL applications.

The two major drawbacks which hamper the performance of OFDM systems are poor bit error rate (BER) in fading environments and high Peak to average power ratio (PAPR).

The transmit signals in an OFDM system might produce high peak values in the time domain since many subcarrier components are added via an IFFT operation. Therefore, OFDM systems are known to have a high PAPR compared with single-carrier systems. When high PAPR OFDM signal pass through a nonlinear device such high power amplifiers (HPA), it causes the out-of-band radiation that affects signals in adjacent bands, and in-band distortions that result in rotation, attenuation and offset on the received signal. So a large back-off in input power is required to force the operation in linear region of HPA. Such HPA with large dynamic range are quite expensive and increase overall system cost. By reducing PAPR we reduce the overall cost as well as complexity of various components in the OFDM system.

The effect of fading on BER of OFDM system can be compensated by using channel coding which results in to a coded-OFDM system.

the remaining 48 sub-carriers provide separate wireless paths for sending the information in parallel fashion. The resulting sub-carrier frequency spacing is 0.3125 *MHz* (for a 20 *MHz* bandwidth with 64 possible frequency slots). The basic parameters for OFDM systems as per IEEE 802.11a standard are given in Table I.

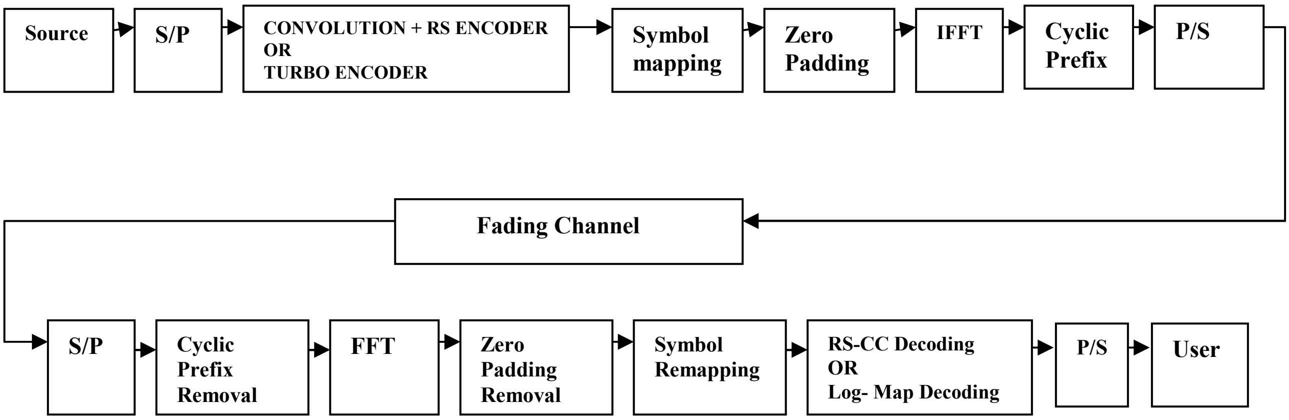


Fig 1. Coded -OFDM System

In this paper OFDM system with serial concatenated (RS- CC) and parallel concatenated code (Turbo) is implemented and BER performance is evaluated for Rayleigh, Rician, Nakagami-*3* and AWGN channels. PAPR for conventional SLM, PTS and DHT precoded OFDM system is calculated and then compared with scheme proposed here in this paper for PAPR reduction. All the parameters used for simulations comply with *IEEE 802.11 a* standard. Paper is organized as follows:

In section 2 *IEEE 802.11 a* standard is discussed, section 3 deals with implementation of COFDM system with serial (RS-CC) and parallel concatenated (Turbo) code. In section 4 fading environments description is given, section 5 discusses conventional PAPR reduction techniques. In section 6 proposed schemes for PAPR reduction is detailed, in section 7 simulation results are discussed and paper is concluded in section 8.

2 IEEE 802.11 *a* Specifications

IEEE 802.11[3-4] is a set of standards carrying out wireless local area network (WLAN) computer communication in the 2.4, 3.6 and 5 GHz frequency bands. This standard specifies an OFDM physical (PHY) layer that splits an information signal across 52 separate sub-carriers. 4 are pilot sub-carriers and

Table 1. OFDM Time Base Parameters in IEEE802.11a

Parameter	Value
FFT size (*nFFT*)	64
Number of subcarriers (*nDSC*)	52
FFT sampling frequency	20 *MHz*
Subcarrier spacing	312.5 *KHz*
Subcarrier index	{-26 to -1, +1 to +26}
Data symbol duration, T_d	3.2 μs
Cyclic prefix duration, T_{cp}	0.8 μs
Total symbol duration, T_s	4 μs
Modulation schemes	BPSK,QPSK,16 & 64-QAM

3 Implementing Coded-OFDM based system

The COFDM system is given in Fig.1 [4], [5]. The Turbo codes and RS codes with convolution codes are used for channel coding. The symbol mapping schemes used are BPSK, QPSK, 16-QAM and 64-QAM. The IFFT/FFT length used is 64. The zero padding is done for confirming the IFFT/FFT size and cyclic prefix is 25% of the IFFT/FFT size, thus making the total size of OFDM frame to 80 symbols. The cyclic prefix compensates the problem caused due to delay spread and maintains continuity of the signal which ensures orthogonal reception of received signal subcarriers.

3.1 Reed Solomon codes

Reed Solomon codes are Maximum Distance Separable (MDS)codes, which means they achieve the maximum possible, minimum distance (d_{min}) for the forward error correction codes (FEC) with the specified parameters (*n, k*) [4], [5], [6], [7]. Reed-Solomon codes are "almost perfect" in the sense that the redundancy added by the encoder is at a minimum for any level of error correction. RS codes allow the correction of *erasures*, its block decoders are fast and powerful, which can operate at rates above 120Mb/s. These codes perform better when it comes to burst errors. The (*n, k*) RS encoder and decoder used here are designed as per parameters given in Table 2 below:

Table 2. RS-Encoder/ Decoder Parameters

Parameter	value	Range
m	Number of bits/ Symbol	3-16
n	Symbols/ Code word	$3\text{-}2^{m-1}$
k	Symbols/ Data word	$k<n$ (*n*-k should be even)
t	Number of errors	$(n\text{-}k)/2$

3.2 Convolution codes

Convolution code is a run-length type code where ***k*** input data bits leads to ***n*** bits of output codeword bits, *k & n* are very small (usually k=1-3, n=2-6), Input depends not only on current set of ***k*** input bits, but also on past input bits [3], [6]. The number of bits which affect current output code is called constraint length and denoted by *K*.

Where K= code memory + k

The specification used for generating convolution codes in this paper is as follows.

3.2.1 For 1/2 convolution code

1. Constraint length:7
2. Feedback connections: $(171)_8$ and $(133)_8$

3.2.2 For 2/3 convolution codes

1. Constraint length (5,4)
2. Feedback connections: $(23,35,0{:}0,5,13)_8$

RS codes together with Convolution codes forms a type of serial concatenated codes.

3.3 Turbo codes

Recursive systematic convolution (RSC) encoder is obtained from the nonrecursive nonsystematic (conventional) convolution encoder by feeding back one of its encoded outputs to its input [7], [8]. Turbo codes encoder (Fig. 2) use the parallel concatenated encoding scheme.

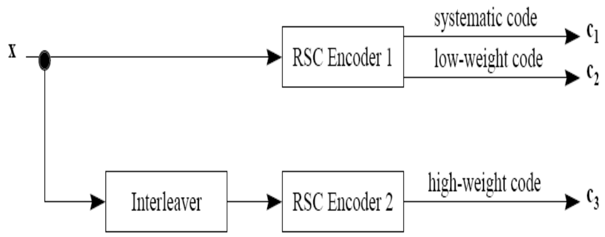


Fig. 2 Basic 1/3 Turbo Encoder

The input sequence **x** produces a low-weight recursive convolutional code sequence $\mathbf{c_2}$ for RSC Encoder 1. To avoid having RSC Encoder 2 produce another low-weight recursive output sequence, the interleaver permutes the input sequence **x** to obtain a different sequence that hopefully produces a high-weight recursive convolutional code sequence $\mathbf{c_3}$.

3.3.1 Log-MAP decoding

The turbo code decoder given in Fig. 3 is based on the serial concatenated decoding scheme [9], [10], [11]. MAP (maximum *a posteriori*) algorithm achieves soft decision decoding on a bit-by-bit basis by making two passes of a decoding trellis, as opposed to one in the case of Viterbi algorithm. One pass is made in the forward direction and one in the backward direction. The input to the turbo decoder is a sequence of received code values, R_k.

The turbo decoder consists of two component decoder DEC_1 to and DEC_2 for respective encoder *RSC Encoder1* and *2*. Each of these decoders is a MAP decoder. DEC_1 takes the received sequence systematic values y_k^s and parity values ${y_{k_1}}^p$ as its input from *Encoder1*.The output of DEC_1 is a sequence of soft estimates EXT_1 (extrinsic data) of the transmitted data d_k. This information is interleaved and then passed to the second decoder DEC_2. *DEC2* takes systematic received values y_k^s (interleaved) and ${y_{k_2}}^p$ from *Encoder2* along with the interleaved form of EXT_1 as its inputs. DEC_2 outputs constitute soft estimates EXT_2 of the transmitted data sequence d_k. This extrinsic data, formed without the aid of parity bits from the first code, is feedback to DEC_1 after deinterleaving. This procedure is repeated in an iterative manner. After several iterations DEC_2 outputs a value $\Lambda(d_k)$ likelihood representation of the estimate of d_k. Threshold operation results in to hard decision estimates of d_k. If $m=0....(M-1)$ represents the states of each decoder (DEC_1 or DEC_2) where the total numbers of states are $M = 2^{k-1}$ and K is the constraint length of the component codes from *Encoder1* and *2*. Let $\alpha_k(m)$ and $\beta_k(m)$ are the state probabilities that the component code is in state m at time instant k, in the forward and backward directions respectively.

Now the trellis is traversed in the reverse direction and the backward pass calculation of state probability β_k is done by multiplying the probability of arriving in the previous state $\beta_{k+1}(m)$ by the probability of the current state transition $\gamma_{k+1}(m',m)$, given the current received values $R_{k+1} = \{y_{k+1}^s, y_{k+1}^p\}$. This is expressed as follows

$$\beta_k(m) = \frac{\sum_{m'}\sum_{i=0}^{1}\gamma_i((y_{k+1}^s, y_{k+1}^p), m', m)\beta_{k+1}(m')}{\sum_{m}\sum_{m'}\sum_{i=0}^{1}\gamma_i((y_k^s, y_k^p), m', m)\alpha_{k-1}(m')} \quad (2)$$

Where β is the backward state probability. The transition probability for each branch between nodes

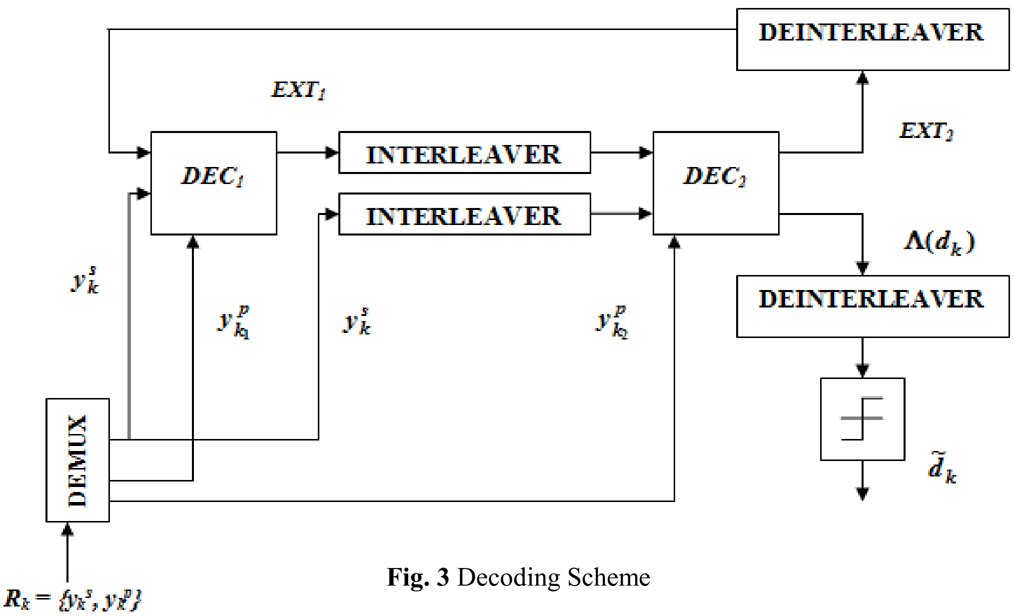


Fig. 3 Decoding Scheme

is given by the equation

$$\gamma_i(y_k^s, y_k^p), m', m) = p((y_k^s, y_k^p)|d_k = i, m', m)) \\ q(d_k = i|m', m)\,\pi(m'|m) \quad (3)$$

The soft estimate is represented as a log likelihood ratio *(LLR)*, $\Lambda(d_k)$, is calculated as follows:

$$\Lambda(d_k) = \frac{\ln\left[\sum_{m}\sum_{m'}\gamma_1((y_k^s, y_k^p), m', m)\alpha_{k-1}(m')\beta_k(m)\right]}{\sum_{m}\sum_{m'}\gamma_0((y_k^s, y_k^p), m', m)\alpha_{k-1}(m')\beta_k(m)} \quad (4)$$

$\Lambda(d_k)$ represents the probability that the current data bit is a *'0'* (if $\Lambda(d_k)$ is negative) or a *'1'* (if $\Lambda(d_k)$ is positive). After a number of iterations, de-interleaved value of $\Lambda(d_k)$ from DEC_2 is converted

Forward state probability α_k is calculated by multiplying the state probability at the previous node $\alpha_{k-1}(m')$ by the branch transition probability $\gamma_{k-1}(m',m)$, given the received code pair $R_k = \{y_k^s, y_k^p\}$, $\alpha_k(m)$ is calculated as:

$$\alpha_k(m) = \frac{\sum_{m'}\sum_{i=0}^{1}\gamma_i((y_k^s, y_k^p), m', m)\alpha_{k-1}(m')}{\sum_{m}\sum_{m'}\sum_{i=0}^{1}\gamma_i((y_k^s, y_k^p), m', m)\alpha_{k-1}(m')} \quad (1)$$

Where m is the current state, m' is the previous state and i is the data bit *('0' or '1')* corresponding to each branch exiting a node.

to a hard decision estimate, $\tilde{d}_k$, of the transmitted data bit.

3.3.2 The Max Log-MAP Algorithm

The logarithmic domain compresses the dynamic range of all the values [10], [11]. Let us take the logarithm of $\gamma_i((y_k^s, y_k^p), m', m)$. Considering probabilities for AWGN channel as:

$$p(y_k^p | d_k = i, m', m) = \left(1/\sqrt{\pi} N_0\right) \exp\left(-1/N_0 [y_k^p - x_k^p(i, m', m)]^2\right)$$

$$p(y_k^s | d_k = i) = \left(1/\sqrt{\pi} N_0\right) \exp\left(-1/N_0 [y_k^s - x_k^s(i)]^2\right) \quad (5)$$

Where x_k^s and x_k^p are the transmitted systematic and parity bits respectively corresponding to the current branch at the time instant k, y_k^s and y_k^s are the received systematic and parity values, N_0 is the noise power spectral density. We can now write the logarithm of *Eq. 3* for *q (...) = 1* as:

$$\ln[\gamma_i((y_k^s, y_k^p), m', m] = \begin{pmatrix} 2y_k^s x_k^s(i)/N_0 \\ +(2y_k^p x_k^p(i)/N_0 \\ +\ln \Pr(m|m') + K) \end{pmatrix} \quad (6)$$

Constant K cancels out in the calculation of $\alpha_k(s_k)$ and $\beta_k(s_k)$. Logarithm of the α is given as:

$$\ln\alpha_k(m) = \ln\left[\sum_{m'} \sum_{i=0}^{1} \exp(\ln\gamma_i[(y_k^s, y_k^p), m', m] + \ln\alpha_{k-1}(m')\right] - \ln\left[\sum_m \sum_{m'} \sum_{i=0}^{1} \exp(\ln\gamma_i[(y_k^s, y_k^p), m', m] + \ln\alpha_{k-1}(m')\right] \quad (7)$$

However, we can approximate the sum of a series of log terms by considering only the maximum log value:

$$\ln[\exp(\delta_1) + \ldots + \exp(\delta_n)] \approx \max_{i \in (1..n)} \delta_i \quad (8)$$

Approximation makes it sub-optimal and its BER performance is therefore poorer then that of the MAP algorithm. A solution to the problem is called the Log MAP algorithm.

3.3.3 The Log-MAP Algorithm

Robertson *et al*. used the Jacobian logarithm to improve the approximation:

$$\begin{aligned} &\ln[\exp(\delta_1) + \exp(\delta_2)] \\ &= \max(\delta_1, \delta_2) + \ln(1 + \exp(-|\delta_2 - \delta_1|)) \\ &= \max(\delta_1, \delta_2) + f_c(|\delta_2 - \delta_1|) \end{aligned} \quad (9)$$

Where $f_c(.)$ is a correction function. In fact, the expression $\ln[\exp(\delta_1) + .. + \exp(\delta_n)]$ can be computed exactly using the Jacobian logarithm, by finding the maximum term and applying the correction recursively through a sequence of δ values. The recursion is initialized with eq.9. We assume $\ln[\exp(\delta_1) + .. + \exp(\delta_n)]$ is known, hence:

$$\ln[\exp(\delta_1) + .. + \exp(\delta_n)] = \ln(\Delta + \exp(\delta_n))$$

$$\textit{Where } \Delta = \exp(\delta_1) + \ldots + \exp(\delta_{n-1}) = \exp(\delta) \quad (10)$$

$$= \max(\ln\Delta, \delta_n) + f_c(|\ln\Delta - \delta_n|) = \max(\delta, \delta_n) + f_c(|\delta - \delta_n|)$$

When implementing the Log-MAP algorithm, all maximizations over two values are augmented by the correction function $f_c(.)$ as defined by Eq. 9. By correcting the Max Log-MAP algorithm in this way, the precision of the MAP algorithm has been preserved. The incorporation of the correction function increases the complexity slightly relative to the Max Log-MAP algorithm. However, since $f_c(.)$ depends only on $|\delta_2 - \delta|_1$, this effect can be minimized by storing the correction values in a simple look-up table.

4 Fading Environments

4.1 Rayleigh Fading

Let the mobile antenna receives a large number (say N) of replicas of same signal. The transmitted signal at frequency ω_c reaches the receiver via number of paths. The amplitude and phase of the i^{th} path are a_i, and φ_i. If there is no direct path or line-of sight (LOS) component, the received signal $s(t)$ can be expressed as

$$s(t) = \sum_{i=1}^{N} a_i \cos(\omega_c t + \varphi_i) \quad (11)$$

If there is a relative motion between transmitter and receiver the Doppler shift has to be considered. If ω_{d_i} represents the shift in i^{th} component, the received signal can be expressed as

$$s(t) = \sum_{i=1}^{N} a_i \cos(\omega_c t + \omega_{d_i} t + \varphi_i) \quad (12)$$

The phase φ_i is assumed to be uniformly distributed over [0,2π]. If N is large, the in-phase and quadrature components of received signal becomes zero mean Gaussian with standard deviation σ. The probability density function (PDF) of the received signal envelope can be given as: [7], [11], [12]

$$f(r) = \frac{r}{\sigma^2}\left\{\frac{-r^2}{2\sigma^2}\right\} \qquad r \geq 0 \tag{13}$$

4.2 Rician Fading

Rician fading is similar to Rayleigh fading except for the fact that there exists a strong LOS component along with reflected waves [7]. In the presence of such a path, if the k_d is the strength of the LOS component and ω_d is the Doppler shift along the LOS path, the transmitted signal given in Eq. 12 can be written as

$$s(t) = \sum_{i=1}^{N-1} a_i \cos(\omega_c t + \omega_{d_i} t + \varphi_i) + k_d \cos(\omega_c t + \omega_d t) \tag{14}$$

The envelope in this case has a Rician density function given by

$$f(r) = \frac{r}{\sigma^2}\left\{-\frac{r^2 + k_d{}^2}{2\sigma^2}\right\} I_0\left\{\frac{rk_d}{\sigma^2}\right\}, \qquad r \geq 0 \tag{15}$$

Where $I_0(.)$ is the 0th order modified Bessel function of the first kind.

4.3 Nakagami-m Fading

The Nakagami distribution, also known as the "*m*-distribution," provides greater flexibility in matching experimental data [13], [14]. Nakagami distribution fits in experimental data better than Rayleigh or Rician distributions. The fading model for the received signal envelope, proposed by Nakagami, has the PDF given by

$$f(r) = \frac{2}{\Gamma(m)}\left(\frac{m}{\Omega}\right)^m r^{2m-1} e^{-mr^2/\Omega}, \ \Omega \text{ and } r \geq 0 \tag{16}$$

In Eq.16 the shape factor, $m \geq 1/2$, Ω is the second moment of r and $\Gamma(.)$ is the gamma function. The parameter Ω controls the spread of the distribution. In the range, $0.5 \leq m \leq 1$ the Nakagami distribution models high frequency channels where the fading is more severe than the Rayleigh fading. When m=1 the distribution become Rayleigh distribution. For $m \geq 1$ the effect of fading is less severe. There is no fading effect when $m=\infty$. For $1<m<2$, Nakagami distribution fits into Rician distribution. In our paper we used following method for generation of Nakagami-*m* distribution.

Let $x_i(t)$ and $y_i(t)$ are zero mean Gaussian processes. The variables x_i and y_i are random variables corresponding to these processes respectively. For fixed t, $E[x_i] = E[y_i] = 0$ and $E[x_i^2] = E[y_i^2]$. Let $r_0^2 = x_0^2$, (or equivalently $r_0^2 = y_0^2$) and $r_i^2 = x_i^2 + y_i^2$, where i is any positive integer and r_i is Rayleigh distributed. We note that r_0 is semi-positive Gaussian distributed. It can be shown that the Nakagami fading model with parameter m of the envelope r is defined as [15]

$$r = \frac{1}{m}\sqrt{\sum_{i=1}^{m} x_i{}^2 + y_i{}^2} \tag{17}$$

5 PAPR and its reduction Techniques

An OFDM signal consists of a number of independently modulated sub carriers, which can give a large peak-to-average power ratio (PAPR) when added up coherently. When N signals are added with the same phase, they produce a peak power that is N times the average power [16], [17], [18],[19]. PAPR for a signal $s(t)$ transmitted in time interval τ is defined as:

$$PAPR\{s(t),\tau\} = \frac{\max[s(t)]^2}{E\{[s(t)]^2\}} \tag{18}$$

Where: $\max[s(t)]^2$ is the peak signal power for $t \in \tau$

$E\{[s(t)]^2\}$ is the average signal power

Transmitted signal with N- sub carriers is represented by:

$$s(t) = \sum_{n=-\infty}^{\infty}\left[\sum_{k=0}^{N-1} d_{n,k}\phi_k(t - nT)\right] \tag{19}$$

Where

$$\phi_k(t) = \begin{cases} e^{j2\pi f_k T}, t \in (0,T) \\ 0, otherwise \end{cases}$$

$$f_k = f_0 + \frac{k}{T}, k = 0,1,2....N-1$$

$d_{n,k}$ = symbol transmitted during n^{th} timing interval using k^{th} subcarrier (assuming 1)

T = Symbol duration, N = Number of OFDM sub-carriers and $f_k = k^{\text{th}}$ sub-carrier frequency, with f_0 being the lowest.

Peak value of the signal

$$\max[s(t)]^2 = \max[s(t).s(t)^*] = N^2 \quad (20)$$

Similarly the mean square value (average signal power) is

$$E\{[s(t)]^2\} = E[(s(t).s(t)^*] = N \quad (21)$$

So when all the subcarriers are equally modulated, and all the subcarriers align in phase and the peak value hits the maximum, the PAPR will be

$$PAPR = N \quad (22)$$

Some of the widely used PAPR reductions methods are Partial transmit sequence (PTS), Selective Mapping (SLM) and Discrete Hartley Transform (DHT) pre-coding.

Selected Mapping scheme is shown in Fig.4. The main purpose of this technique is to generate a set of data blocks at the transmitter end which represent the original information and then to choose the most favorable block among them for transmission [19], [20], [21], [22]. The frequency domain constellation points are multiplied with a set of randomly generated phase vectors which is available at both the transmitter and the receiver. Then, the block with the least PAPR is sent to the IFFT to create the time-domain signal to be transmitted. The index of the phase matrix which gives the minimum PAPR signal is also sent as side channel information (CSI). At the receiver, the received time domain signal is multiplied with the conjugate phase vector (using theCSI) and sent through the FFT to retrieve the original signal.

Every data block is multiplied by U dissimilar phase sequences, each of length N, $P^{(u)} = [p_{u,0,} p_{u,1,} ....p_{u,N-1}]^T$, where $u = 1,2...U$, and $P_{u,v} = e^{j\phi_{u,v}}$ and $\phi_{u,v} \in [0,2\pi)$ for $v = 0,1....N-1$.

This resultant signal is

$$X_n^u = X_n p_{u,n}$$

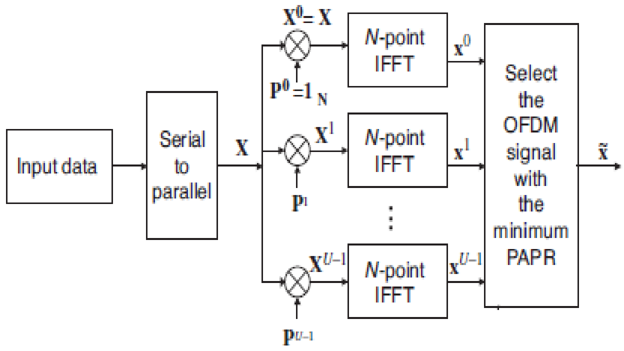


Fig. 4 OFDM system with SLM technique

After taking its IFFT the sequence generated is $x^u = \left[x^u[0], x^u[1].....x^u[N-1]\right]^T$ among which $\tilde{x} = x^{\tilde{u}}$ with lowest PAPR is selected for transmission.

$$\tilde{u} = \arg\min_{u=1,2...U}\left(\max\left|x^{u[n]}\right|\right)$$

The Transmitted OFDM signal is given by

$$x(t) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x_{u,n} \exp(j2\pi n\Delta f t) \quad (23)$$

Here $\Delta f = \frac{1}{NT}$; *NT= Duration of Data block*

Partial Transmit Sequence is one of the most sought technique for PAPR reduction in OFDM (Fig. 5). Here the input frequency domain data block is first partitioned into disjoint sub-blocks. Then each of the sub-blocks are then padded with zeros appropriately and weighted by complex phase factors.

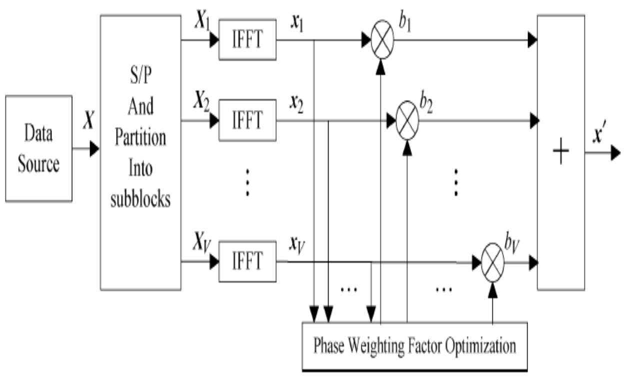


Fig. 5 Block diagram of PTS technique

The data vector $X = [x_{0,} x_1, x_2, ...x_{N-1}]^T$ is divided in V disjoint sets, $\{X_v, v = 1,2..V\}$, using same number of carrier for each group then V group sum:

$$X' = \sum_{v=1}^{V} X_v b_v \qquad (24)$$

Where $b_v = e^{j\phi v}$ are the phase factors. In time domain the x_v (IFFT of X_v) is called Partial transmit sequence [20-24]. The phase factor is chosen such that PAPR of x' is minimum.

$$\left[\widetilde{b}_1, \ldots\ldots\ldots \widetilde{b}_V\right] = \underset{[b_1 \ldots\ldots b_V]}{\arg\min} \left(\max_{n=0,1 \ldots. N-1} \left| \sum_{v=1}^{V} b_v x_v [n] \right| \right) \qquad (25)$$

Corresponding time domain signal with lowest PAPR is:

$$x' = \sum_{v=1}^{V} \widetilde{b}_v x_v$$

The PTS technique requires V IFFT operations for each data block and $\lfloor \log_2 W^V \rfloor$ bits of side information. The PAPR performance of the PTS technique is affected by not only the number of sub-blocks V and the number of the allowed phase factors W, but also by the sub-block partitioning. In fact, there are three different kinds of the sub-block partitioning schemes: adjacent, interleaved, and pseudo-random. Among these, the pseudo-random one has been known to provide the best performance.

PTS technique suffers from the complexity of searching for the optimum set of phase vector, especially when the number of sub-block increases

Discrete Hartley Transform (DHT) precoding can also be used as a mean to reduce PAPR in multi carrier system [25], [26], [27].
The N-point DHT of a real sequence is defined by

$$S_k = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} s_n H_k^{kn}, k = 0,1,2 \ldots N-1 \qquad (26)$$

Where $H_N^k = \sin(2\pi k/N) + \cos(2\pi k/N)$

If $s = [s_{N-1,} s_{N-2}, \ldots.. s_0]^T$ is the transform input vector and $S = [S_{N-1}, S_{N-2}, \ldots. S_0]^T$ be the transform output vector. The equation (26) can be written as

$$S = Hs \qquad (27)$$

Where

$$H = \frac{1}{\sqrt{N}} \begin{vmatrix} H_N^{(N-1)^2} & H_N^{(N-1)(N-2)} & \ldots. & 1 \\ H_N^{(N-1)/(N-2)} & H_N^{(N-2)^2} & \ldots.. & 1 \\ .. & . & . & . \\ 1 & 1 & 1 & 1 \end{vmatrix}$$

DHT and Inverse DHT transform matrices are same ***(H=H^-1^)***. Fig: 6 shows the DHT precoded COFDM system.

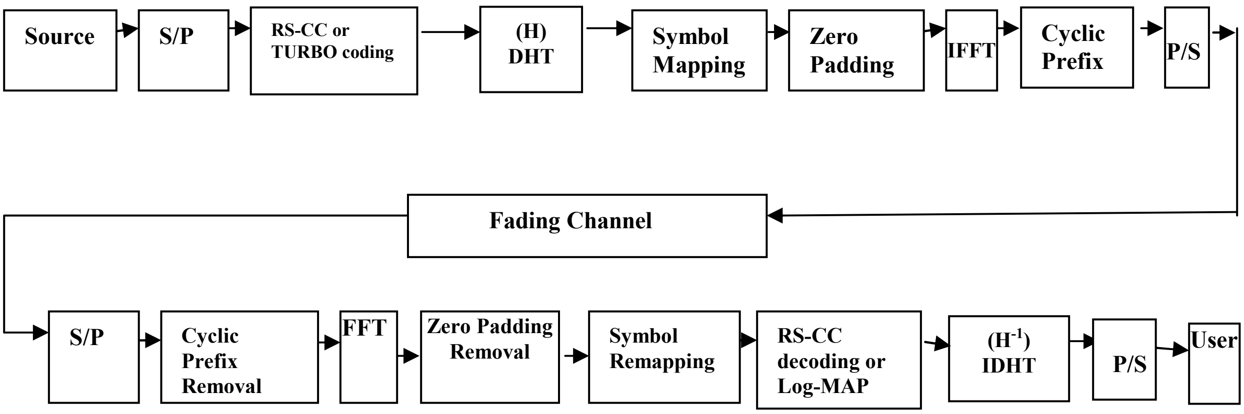


Fig. 6 DHT Precoded -OFDM System

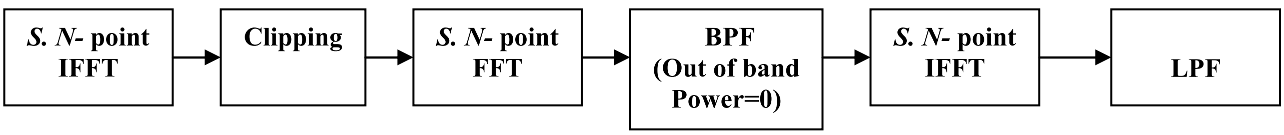


Fig. 7 clipping and filtering scheme used post PTS, SLM and DHT precoding

6 Proposed Schemes for PAPR reduction

In all the above three schemes if we use clipping method to further suppress high power values, the PAPR can be reduced further [18] [19]. A maximum peak amplitude K is chosen so that the OFDM signal does not exceed the limits of this region, symbols that exceed this maximum amplitude, will be clipped.

If K is the clipping level the clipped signal can be given as

$$S_c(t) = \begin{cases} Ke^{j\theta(t)}, |s(t)| > K \\ s(t), |s(t)| < K \end{cases} \tag{28}$$

Where $\theta(t)$ is the phase of $s(t)$.

The clipping ratio is given by

$$CR = \frac{A}{\sigma} \tag{29}$$

Where σ is the rms value of OFDM signal.

However Clipping may introduce in band as well as out of band distortion which may cause BER degradation .This can be avoided by using filtering. Since we are using Coded-OFDM system here the BER is much better then corresponding uncoded OFDM system with clipping and filtering.

The Modified PTS, SLM and DHT pre-coding scheme where clipping and filtering scheme with over sampling factor S is used post PTS, SLM and DHT precoded system is given in Fig. 7.

7 Simulation Parameters and Results

All the simulations are done in MATLAB 7.0 and simulation parameter are chosen as per IEEE 802.11a standard. For BPSK, QPSK, 16-QAM (15, 13) RS codes and for 64-QAM (60, 58) RS codes are used, which are the nearest values of (n, k) as per Table 2, convolutions codes used are of 1/ 2 and 2/3 code rate. Turbo code rate is 1/3 with generator matrices [1 1 1] and [1 0 1] with log-Map decoding, rest of the parameters are as per Table 1.

PTS technique is simulated for block size V= 2, 4, 8 and phase weights W = {1,-1}. For filtering a FIR BPF filter with eqiripple response is used, considering over sampling factor S > 4, BW= 1 MHz, stop Band attenuation ranging to 35-40 dB and frequency vectors varying from 1.3 to 2.5 MHz.

Performance analysis in terms of BER for various cases is as follows. Fig. 8, Fig. 9 and Fig. 10 shows BER comparison for various mapping with RS-1/2 CC and RS-2/3 CC codes in Rician , Rayleigh and Nakagami (m=3) fading environments.

Fig 11 (a) and (b) compares BER performance coded OFDM system in Rayleigh and Rician environments, where Turbo coding is used for channel coding. Fig. 12 shows a comparison of BER in AWGN, Rayleigh, Rician and Nakagami environments for 16-QAM scheme.

Fig.13 (a) Shows BER comparison of various mapping schemes with TURBO codes in Nakagami *(m=3)* environment and 13 (b) compares the performance of serial and parallel concatenation codes .

Fig. 14 (a) compares the PAPR of Plain COFDM, SLM and DHT precoded COFDM with Modified SLM and DHT precoded COFDM systems, Fig. 14 (b) compares performance of Plain COFDM, PTS (V =2, 4, 8) with modified PTS version and Fig. 14 (c) compares all of the above schemes for 16- QAM.

Fig. 15 (a) compares the PAPR of Plain COFDM, SLM, DHT precoded COFDM with Modified SLM and DHT precoded COFDM systems, Fig. 15 (b) compares performance of Plain COFDM, PTS (V =2, 4, 8) with modified PTS versions and Fig. 15 (c) compares all of the above schemes for 64- QAM.

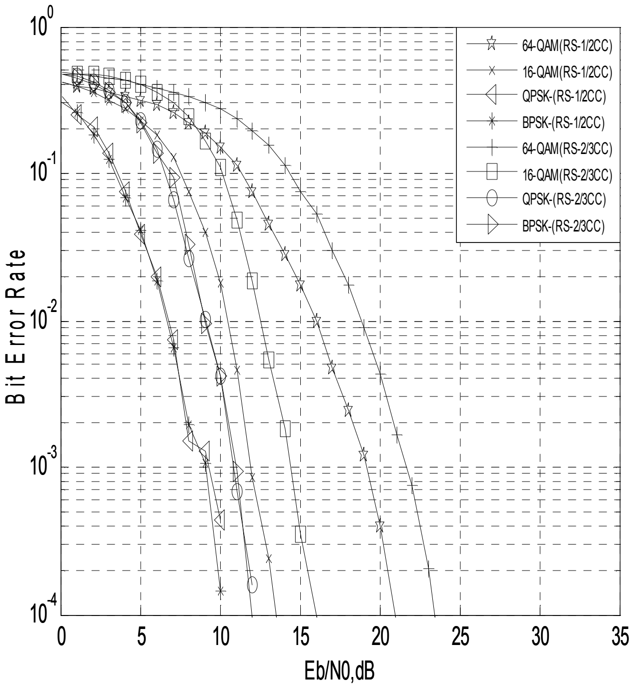


Fig .8. BER Comparison of COFDM System with RS codes and 1/2, 2/3 convolution codes in Rician Channel

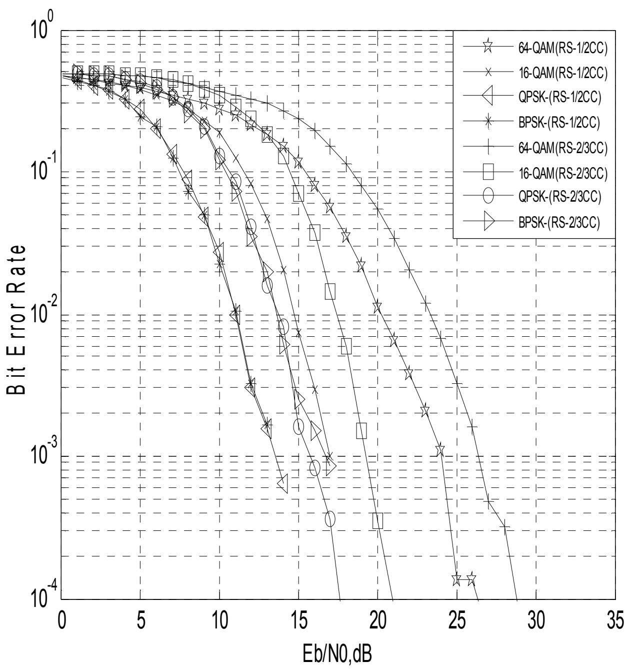


Fig .9 BER Comparison of COFDM System with RS codes and 1/2, 2/3 convolution codes in Rayleigh Channel

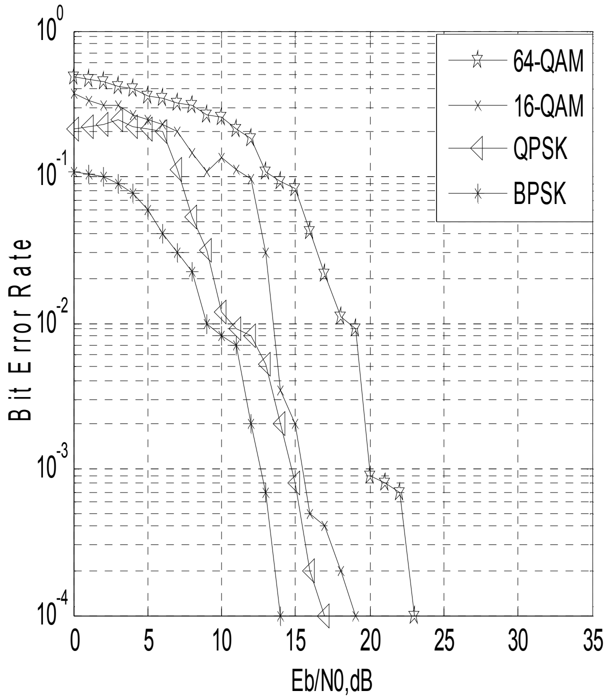


Fig. 11 (a) BER Comparison of COFDM using Turbo codes in Rayleigh fading

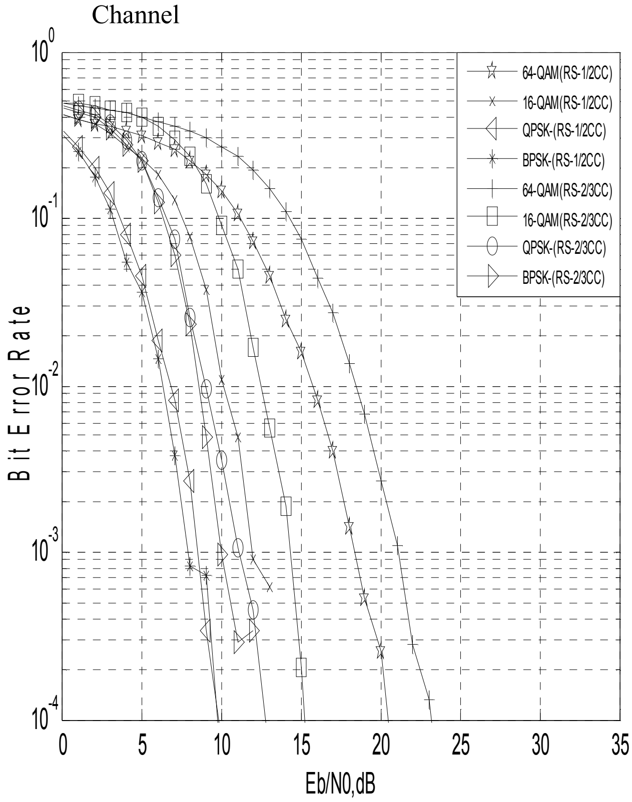


Fig. 10 BER Comparison of COFDM System with RS codes and 1/2, 2/3 convolution code in Nakagami-3 fading environment

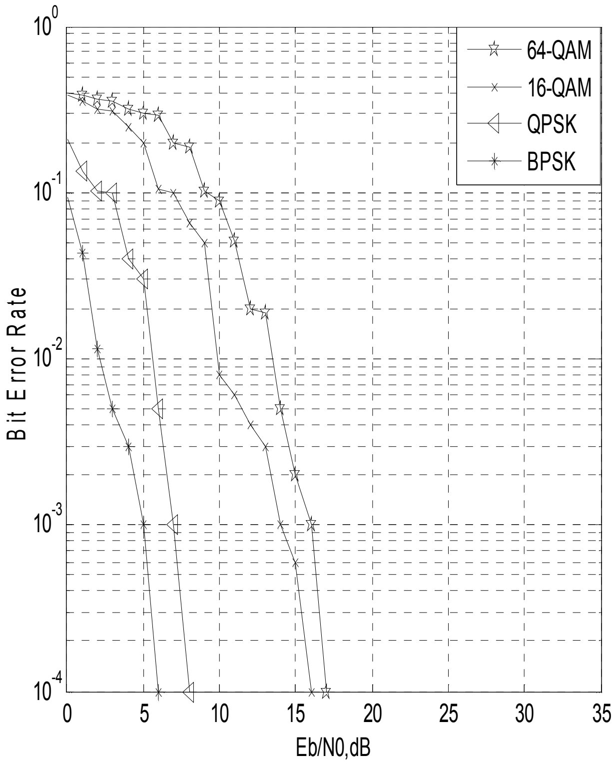


Fig. 11 (b) BER Comparison of COFDM using Turbo codes in Rician fading Environment.

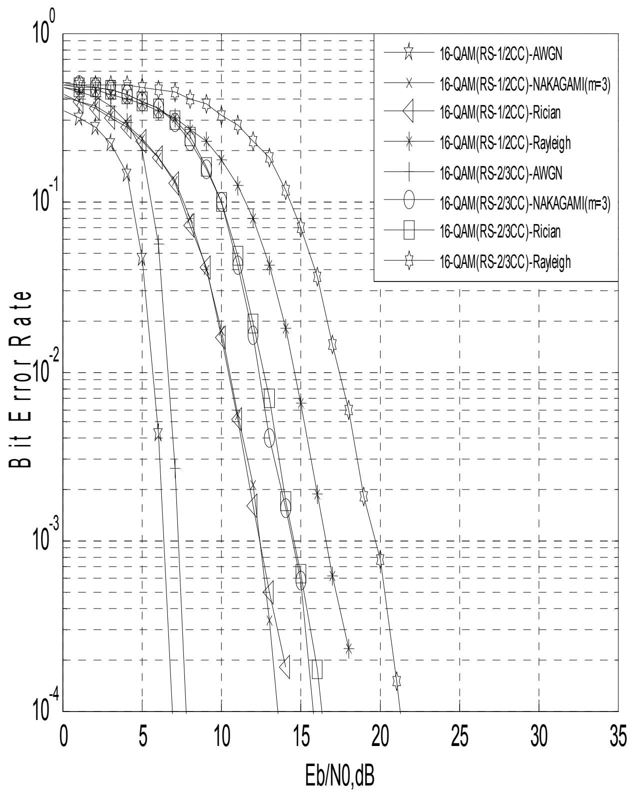
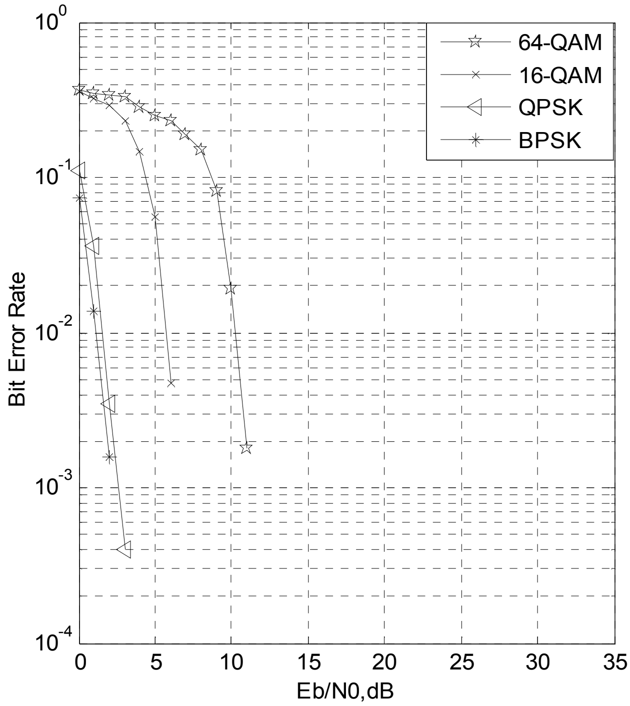
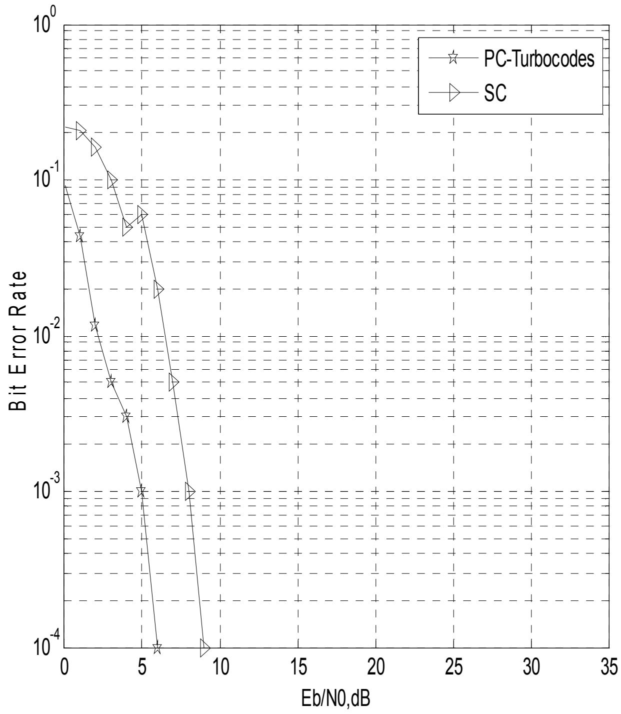


Fig. 12 BER Comparison of COFDM System with RS Codes and 1/2, 2/3 Convolution Code in AWGN, Rayleigh, Rician and Nakagami fading for 16-QAM mapping

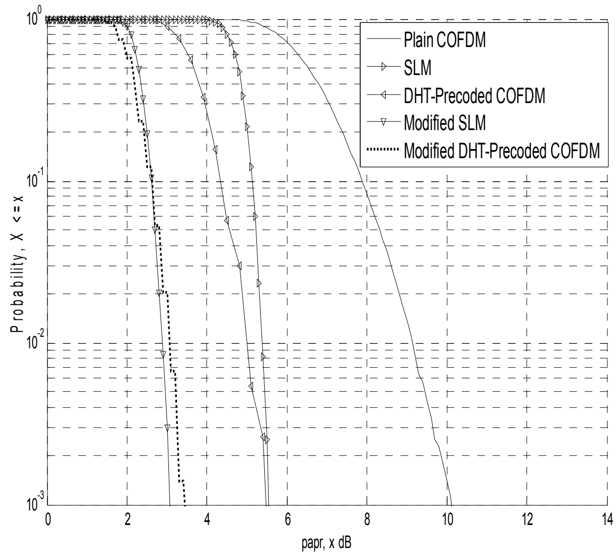


(a)

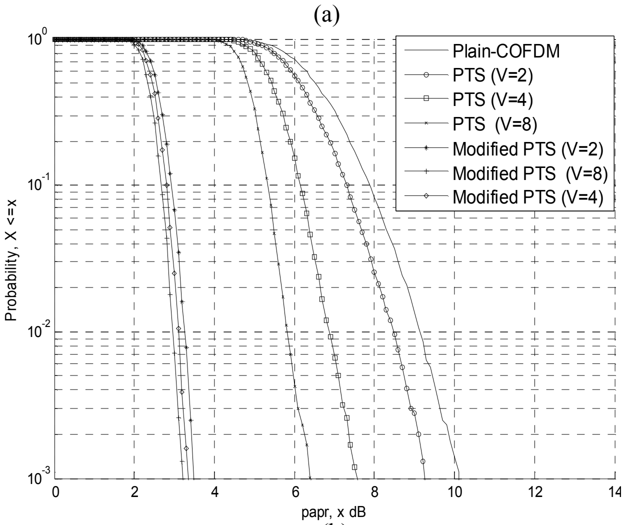


(b)

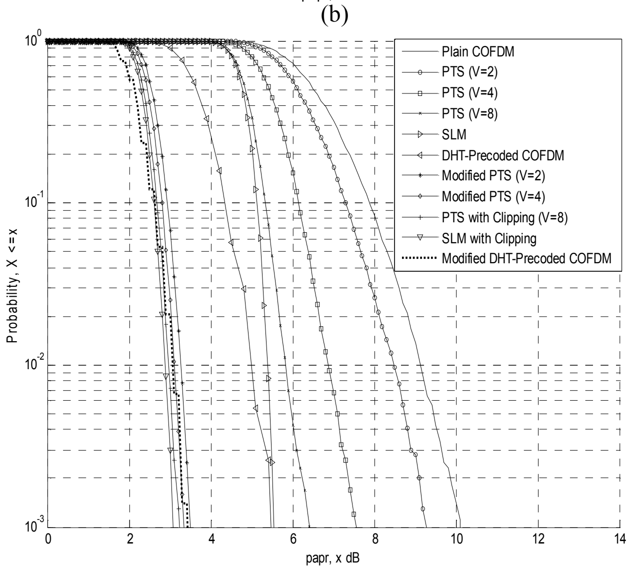
Fig. 13 (a) BER of COFDM using Turbo codes in Nakagami fading (b) Comparison of serial and parallel concatenation



(a)

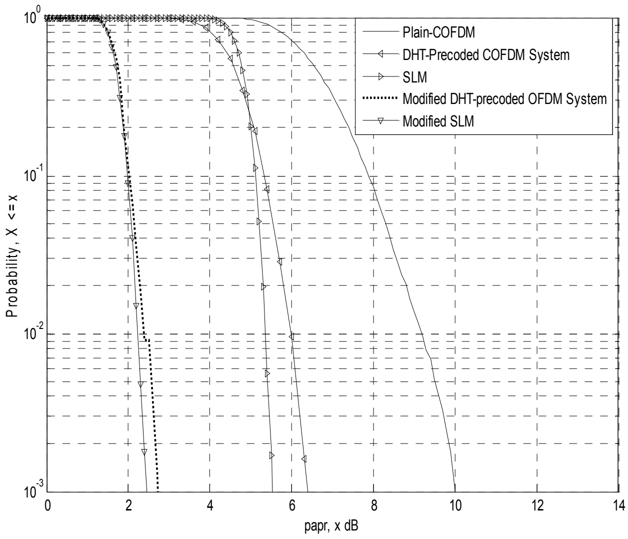


(b)

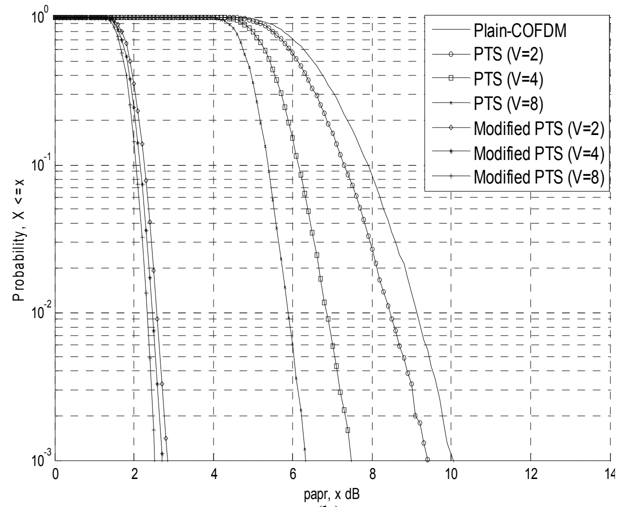


(c)

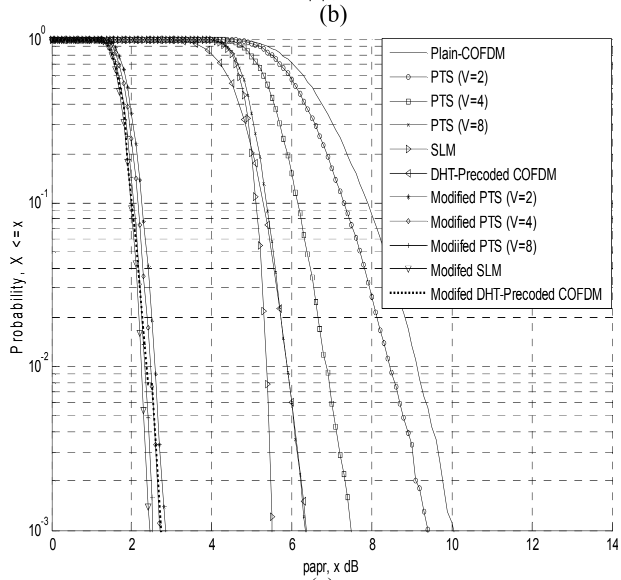
Fig. 14 PAPR Comparison of (a) DHT-precoded, SLM and their modified versions (b) PTS (V =2, 4, 8) and Modified PTS for COFDM (c) all schemes with 16-QAM



(a)



(b)



(c)

Fig. 15 PAPR Comparison of (a) DHT precoded, SLM and their modified versions (b) PTS (V =2, 4, 8) and Modified PTS for COFDM (c) all schemes with 64-QAM

8. Conclusion

The results of all above simulations for various codings, fading and mapping schemes can be summated as follows:

Based on fading

$BER_{AWGN} < BER_{Nakagami-m} < BER_{Rician} < BER_{Rayleigh}$

Based on Modulation scheme

$BER_{BPSK} < BER_{QPSK} < BER_{16-QAM} < BER_{64-QAM}$

Based on convolution coding

$BER_{1/2-conv} < BER_{2/3-conv}$

Based on concatenation

$BER_{PC-turbo} < BER_{SC}$.

Results show that BPSK scheme has best BER performance when mapping is concerned. In Serial concatenation RS-1/2 CC gives better performance tan RS-2/3 CC combination. In terms of concatenation Parallel concatenation using Turbo codes gives a better performance than their serial counterparts. Nakagami fading environments suits well for practical data and fading is most severe in Rayleigh fading.

In terms of PAPR analysis, the PAPR performance for post -clipped and filtered versions (modified) for PTS, SLM and DHT precoded system versions is better than unclipped versions. However this is achieved at cost of increase in complexity of circuit. Value of PAPR (in dB) for a given CDF value is given below:

Table 3. PAPR (in dB) Comparison For Various Schemes

SCHEME	16-QAM	64-QAM
Plain- COFDM	*7.9*	*8.6*
DHT Precoded	*5.2*	*5.6*
SLM	*5.1*	*5.3*
PTS (V=2)	*7.3*	*7.9*
PTS (V=4)	*6.2*	*6.5*
PTS (V=8)	*5.4*	*5.7*
Modified DHT	*2*	*2.2*
Modified SLM	*2*	*2.2*
Modified PTS (V=2)	*2.3*	*2.4*
Modified PTS (V=4)	*2.2*	*2.3*
Modified PTS (V=8)	*2*	*2.2*

We can see the modified scheme brings down PAPR drastically. With PTS, further increasing the block size *V*, PAPR reduces more. DHT precoded schemes also proved as a better candidate as a PAPR reduction technique.

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