REDUCING PAPR BY ELIMINATING SIDE INFORMATION FOR SFBC MIMO-OFDM SYSTEMS

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Abstract: A novel peak-to-average power ratio (PAPR) reduction scheme designed as extended selected mapping (eSLM) for space-frequency block coding (SFBC) multi-input multi-output orthogonal frequency division multiplexing systems. In the eSLM method, it creates extension matrices comprising amplitude extensions and phase rotations to get the selected signal index with no side information and to reduce the PAPR, respectively. To minimize the computational complexity obtained by the IDFT operation in generating the candidate signals, a low-complexity eSLM scheme (LC-eSLM) is generated by establishing equivalent candidate signals in the time domain. Particularly, the extension matrices in both designs preserve the orthogonality of the SFBC code, promoting low-complexity decoding. The simulation results illustrate that the proposed eSLM scheme not only out performs existing blind SLM-based schemes. Compared with the valuable ordinary SLM scheme, the eSLM scheme involves a lowePr computational complexity with a work loss of less than 0.3 dB and causes no side information. The computational complexity of the LC-eSLM scheme is around 40%–50% lower than that of the eSLM scheme with only a minimal degradation in the PAPR reduction performance.

Keywords: Peak-to-average power ratio, SFBC, Multi-input Multi-output, blind detection, eSLM, OFDM

I. INTRODUCTION

Communication plays important role in our life. OFDM is very attractive digital modulation technique used in mobile communication for high speed data transmission. OFDM is a multicarrier wireless communication system used for transmitting large amount of data [1]. It is a multicarrier digital modulation scheme whose range is four times greater than that of single carrier modulation scheme. To meet out the requirement of higher data rates and high spectral efficiency a new digital multicarrier modulation scheme is developed in communication field which is suitable for both wired and wireless environment. New promising digital multicarrier modulation scheme to fulfill these requirements in the telecommunication field is Orthogonal Frequency Division Multiplexing (OFDM) [2]. The technology was first conceived in the 1960s and 1970s. A major shortcoming associated with OFDM is its high peak to average power ratio due to number of sub carriers which is responsible for signal degradation during transmission or seriously limit the efficiency of power amplifier. Due to efficient usage of bandwidth it is considered to be the modulation technique for next generation 4G networks [3]. The main objective of OFDM is to divide high data rate bit stream in to number of lower parallel bit stream which are used to modulate different number of subcarriers by using fourier transform. These subcarriers overlapped with each other in frequency domain form, thus to increasing the transmission rate, In OFDM a guard band is inserted to eliminate the effect of inter symbol interference (ISI). OFDM system has number of advantages like immunity to inter-symbol interference, high spectral efficiency, robustness in frequency selective fading channels[4].

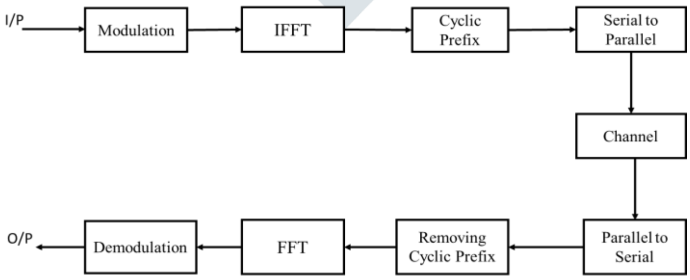

Figure 1.1 OFDM Block Diagram.

OFDM transmission systems suffer a high peak-to-average power ratio (PAPR), which results in severe in-band distortion in the nonlinear region of the power amplifier. The spectrum efficiency and energy efficiency are improved when the MIMO technology is introduced to the OFDM-based systems. However, it becomes more challenging to reduce PAPR in MIMO-OFDM systems especially when more carriers are aggregated in the LTE-A or 5G systems [5].

1.1 Peak-To-Average Power Ratio in OFDM System

It is defined as the giant variation or ratio between the average signal power and the maximum or minimum signal power. Theoretically, massive peaks in OFDM system may be expressed as Peak-to Average Power Ratio (PAPR) and it is typically outlined as

$$PAPR = \frac{P_{Peak}}{P_{Average}} = \frac{Max[|X_n|^2]}{E|X_n|^2}$$

Where P peak represents peak output power, P average means average output power [E]. Denotes the mean value, represents the transmitted OFDM signals that are obtained by taking IFFT operation on modulated input symbols. Mathematical, is expressed as

$$X_n = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k W_N^{nk}$$

For an OFDM system with sub-carriers, the peak power of received signals is N times the average power when phase values are the same.

The PAPR of baseband signal [2] will reach its theoretical maximum at PAPR (db) =10logN. Another commonly used parameter is the Crest Factor (CF), which is defined as the ratio between maximum amplitude of OFDM signal x (t) and root-mean-square (RMS) of the waveform [6].

The literature contains various PAPR reduction methods for overcoming this problem, including clipping, selected mapping (SLM), companding and partial transmit sequences (PTS). Among these methods, SLM is one of the most commonly used since it reduces the PAPR without distorting the MIMO-OFDM signals [5].

The high Peak-to-Average Power Ratio (PAPR) or Peak-to-Average Ratio (PAR) or Crest Factor of the Orthogonal Frequency Division Multiplexing (OFDM) systems can be reduced by using various PAPR reduction techniques namely: -

1.2 Multiple Signal Representation Techniques

a. Partial Transmit Sequence (PTS): -

Partial transmit sequences (PTS) is one amongst the foremost necessary strategies that cut back PAPR within the OFDM system. And it is conferred in 2 main steps. First, by dividing the initial OFDM signal into variety of sub-blocks. Secondly, adding the part turned sub-blocks to develop variety of candidate signals to select the one with smallest PAPR for transmission. There is another way which can also be used to express PTS method by multiplying the original OFDM signal with a number of phase sequences [6].

b. Selective Mapping (SLM): -

Selective mapping (SLM) could be a promising PAPR reduction technique of OFDM system. The basic idea of SLM is to produce U alternative transmit sequences from the same data source and then to select the transmit signal exhibiting the lowest PAPR. The idea stems from the fact that as the PAPR is determined by the sequence of the transmit data vectors; Xm multiplying the data vectors by some random phase will change the PAPR properties after the IFFT.

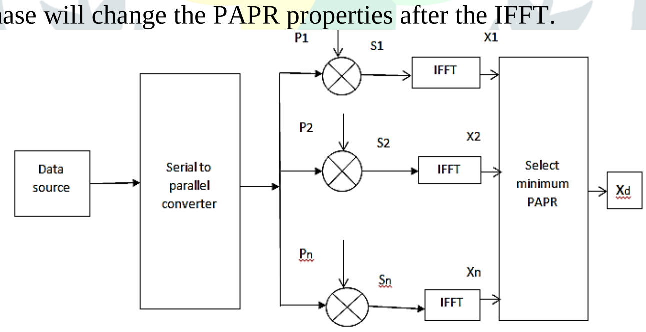

Figure 1.2: SLM block Diagram.

Selected mapping technique has to transmit the data to receiver, with the chosen signal, as side information. If there is any error in the received information, then it is difficult for the receiver to recover the information from the transmitted selected signal [1].

II. System Model

In this paper, we propose to address the PAPR problem by employing the extended SLM (eSLM) method for MIMO OFDM systems with SFBC coding which both avoids the need for SI and preserves the orthogonality of the original Alamouti SFBC code. The candidate signals are constructed by taking the Hadamard product of one of V extension matrices and the Alamouti-encoded blocks. The extension matrices comprise phase rotation and amplitude extension components, where the former components aim to reduce the PAPR, while the latter components facilitate the SI detection blindly. Compared with bSLM and pSLM schemes in which the phase rotation sequences for all the antennas are either the same or have a constant phase offset, the extension matrices proposed in the present study have a greater degree of freedom, and thus provide a greater potential to reduce the PAPR. The sufficient and necessary conditions for the extension matrices to satisfy the orthogonality of the SFBC coding scheme are derived [5].

In addition, a blind detection scheme is proposed for identifying the indices of the selected candidate signals. The simulation results show that the proposed eSLM scheme achieves a lower PAPR than existing blind schemes, such as bSLM and pSLM.

Furthermore, while the PAPR reduction performance of eSLM is slightly inferior to that of the oSLM scheme, it requires no SI and has a lower computational complexity.

Although the proposed eSLM scheme successfully lowers the PAPR without the need of SI, the proposed eSLM scheme and other existing SLM-based methods require to perform inverse discrete Fourier transformation (IDFT) on each of the candidate signals, which leads to high computational complexity especially with large number of subcarriers.

To reduce the computational complexity incurred by IDFT, a low-complexity eSLM scheme (LC-eSLM) is also proposed in which equivalent candidate signals are generated in the time domain. Notably, we show that the candidate signals in the LC-eSLM scheme are simply a special case of those in the eSLM scheme with repetitive extension factors. Through computer simulations, we show that the LC-eSLM scheme has a significantly lower complexity than the bSLM and pSLM schemes, while achieving a similar PAPR reduction performance [5]. The architecture of the eSLM-based transmitter for SFBC MIMO-OFDM systems with two transmit antennas is shown in Fig. 2.1. The frequency domain data symbols can be expressed as **X** = [X (0), X (1),..., X (N − 1)], where X(n) is independently and identically distributed (i.i.d.) with mean E[X(n)] = 0 and variance Eˆs = E[|X(n)|2]. In the SFBC MIMO-OFDM system with two transmit antennas, the data symbols are encoded pair wisely with Alamouti space frequency encoder as

$$[X1(2i)\ X1(2i+1)\ X2(2i)\ X2(2i+1)\ ] = [X(2i)\ X^*(2i+1)\ X(2i+1)\ -X^*(2i)\ ]$$

for 0 ≤ i ≤ N/2 − 1, where Xp(n) is the Alamouti encoded symbol modulated by n-th subcarrier in the p-th transmitting antenna. The block of SFBC data symbols conveyed by the p-th antenna is defined as **X**p = [Xp(0), Xp(1), . . . , Xp(N−1)], for p = 1, 2. Because SFBC code is an orthogonal block code, the two rows in are orthogonal. To reduce PAPR, the proposed eSLM scheme passes **X**1 and **X**2 through V extension matrices **E**(0), **E**(1), . . . ,**E**(V−1) ∈ C2×N. The extension matrices enable the PAPR to be effectively reduced without the need to transmit additional side information. In processing the datasymbols, the Alamouti encoded blocks **X**1 and **X**2 are respectively multiplied with the first and second rows of each E(v) in an element-wise manner as given

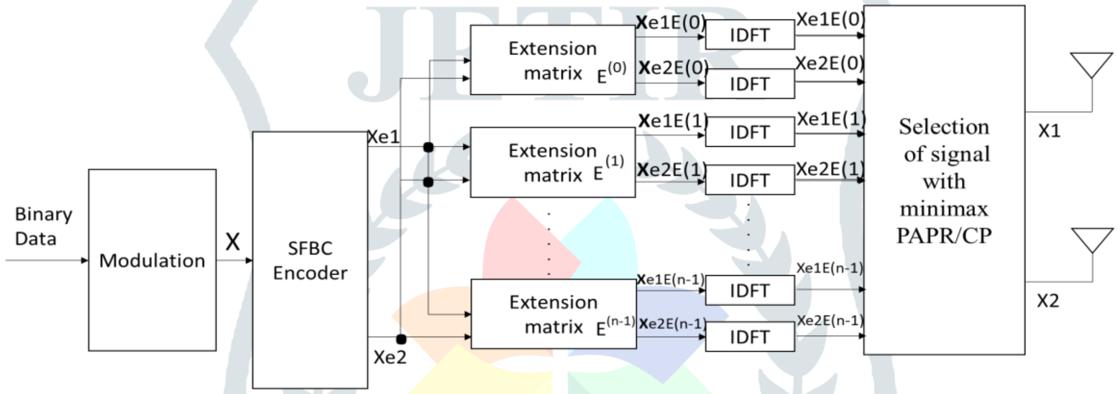

Figure 2.1: -Block Diagram of SFBC MIMO-OFDM System employing eSLM.

$$X_p^{(v)} = E^{(v)}[p]\ \cdot X_p\ , p = 1,2$$

Where E(v)[p] denotes the p-th row of E(v) and ◦ is the elementby- element multiplication. The extension matrices generates V different pairs of extended blocks (X(v) 1 ,X(v) 2), for 0 ≤ v ≤ V −1. These two extended blocks (X(v) 1 ,X(v) 2) are then passed through the inverse discrete Fourier transform (IDFT) operation to generate the time-domain signals (x(v) 1 , x(v) 2), where the kth time-domain sample at the pth antenna is written as:

$$X_p^{(v)}(k) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} X_p^{(v)}(n) e^{j2\pi\, kn/N}\ , p = 1,2$$

Because the signals of the two antennas are transmitted simultaneously, the overall PAPR of SFBC MIMO-OFDM should be considered jointly over the two antennas. As a result, the candidate pair (x(ˆv) 1 , x(ˆv) 2) is selected for transmission in accordance with

$$\hat{v} = arg\ arg \quad \left(PAPR\left(X_1^{(v)}\right), PAPR\left(X_2^{(v)}\right)\right)$$

Where the PAPR value of the transmitted candidate signal of each antenna x(v) p is given as

$$PAPR\left(X_p^{(v)}\right) = \frac{\left|X_p^{(v)}(k)\right|^2}{E\left[\left|X_p^{(v)}(k)\right|^2\right]}\ , p = 1,2$$

Notably, as V increases, the PAPR of the eSLM-based MIMO OFDM signal is reduced at a greater extent [5].

2.1 Extension Matrices

The extension matrices described above have two fundamental advantages in addition to reducing the PAPR. First, the pairwise orthogonality between the original encoded signals X1 and X2 is preserved following multiplication with the extension matrix (i.e., X(v) 1 and X(v)2). By assuming that the channel coefficients are almost identical for two adjacent subcarriers, the SFBC blocks can be decoded using simple low-complexity linear operations. Second, the extension matrix enables the receiver to indicate which candidate is chosen without the need of SI, and hence the spectrum efficiency is improved [5]. Each extension matrix consists of N/2 basic extension units U(v) i for 0 ≤ i ≤ N/2 − 1, expressed as

$$E^{(v)} = \left[U_0^{(v)}, U_1^{(v)}, U_2^{(v)}, \cdots\cdots\cdots, U_{\frac{N}{2-1}}^{(v)} \right], 0 \le v \le V-1$$

Obviously, the value of {U(v) i } is crucial to the bit error rate, PAPR reduction performance and SI detection performance. Specifically, both the amplitude extension and the phase rotation are incorporated into the U(v) i as

$$U_i^{(v)} = \left[U_{i,1}^{(v)} e^{j\theta_{i,1}^{(v)}} \; U_{i,2}^{(v)} e^{j\theta_{i,2}^{(v)}} \; U_{i,3}^{(v)} e^{j\theta_{i,3}^{(v)}} \; U_{i,4}^{(v)} e^{j\theta_{i,4}^{(v)}} \right], 0 \le i \le \frac{N}{2-1}$$

where U(v) i,j∈ R is equal to or greater than one, i.e., U(v) i,j ≥ 1, and θ(v) i,j is ranged between [0, 2π].

Finally, the required average power at each transmitter antenna is amplified since the extension factor U ≥ 1. The energy increase of the eSLM scheme resulting from the parameters U, M, and K is expressed as

$$\Delta_{(dB)} = 10 \cdot \left[\frac{(U^2+1)K + 2(M-K)}{2M} \right]$$

That is, the symbol energy Eˆs of the transmitter should be normalized to fulfill the average power constraint Es as given by

$$\hat{E}_s = \frac{2M}{(U^2+1)K + 2(M-K)} \cdot E_s$$

III. Low-Complexity eSLM Scheme (LC- eSLM)

Although the proposed eSLM scheme effectively reduces the PAPR without the need for side information (i.e., as in the conventional SLM method), it requires 2V IDFT operations of size N to generate the candidate signals. To simplify the computational complexity, this section therefore proposes an alternative low-complexity eSLM (LC-eSLM) scheme. Specifically, the proposed scheme generates equivalent candidate signals in the time domain.

In other words, after applying data modulation and SFBC encoding, each of the codewords X1 and X2 is partitioned by 2M disjoint sub-vectors Xp,0, Xp,1, · · ·, Xp,2M−1 (p = 1, 2), where the n-th element of Xi, m is expressed as

$$[X_{p,m}]_n = \{[X_p]_n, \; n = g \cdot 2M + m, \quad \forall_g = 0,1,\cdots\cdots\cdots, \frac{N}{2M} - 1 \; 0, \; otherwise$$

for p = 1, 2 and m = 0, 1, . . ., 2M − 1. In each sub-vector, the G = N/2M nonzero entries are equally spaced by 2M elements. The space-frequency encoded symbol vectors can thus be written as

$$X_p = \sum_{m=0}^{2M-1} X_{i,m}, \qquad p = 1,2$$

To generate time-domain candidate signals, each sub-vector is passed through IDFT prior to phase rotation. The corresponding time-domain signal of each sub-vector is therefore given by

$$X_{p,m} = IDGT(X_{p,m}), \qquad p = 1,2, m = 0,1,\cdots\cdots\cdots 2M-1.$$

The time-domain signals corresponding to each transmit antenna are then passed through a Candidate Signal Generating Block (CSGB) to generate V candidate signals. As shown in Fig. 4, the candidate signals are obtained by linearly combining the time-domain signals with various amplitude extensions and phase rotations. Specifically, the v-th candidate signal of the p-th antenna, x(v) p, is obtained as

$$X_p^{(v)} = \sum_{m=0}^{2M-1} A_{p,m}^{(v)} e^{j\phi_{p,m}^{(v)}} x_{p,m}$$

where A(v) p, m and (v) p,m is respectively the m-th amplitude extension and phase rotation corresponding to the v-th candidate signal at the p-th antenna.

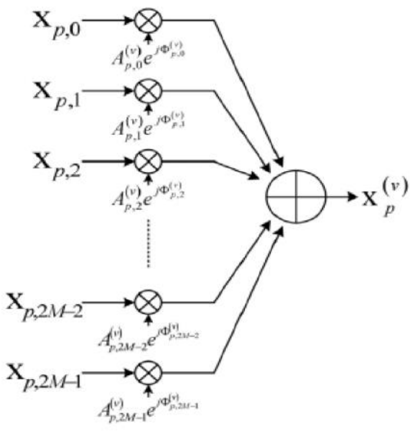

Fig. 3.2. Architecture of candidate signal generating block (CSGB) in the LC-eSLM scheme.

As described in the following proposition, the candidate signals produced in the LC-eSLM scheme can be regarded as a special case of those generated by the proposed eSLM scheme.

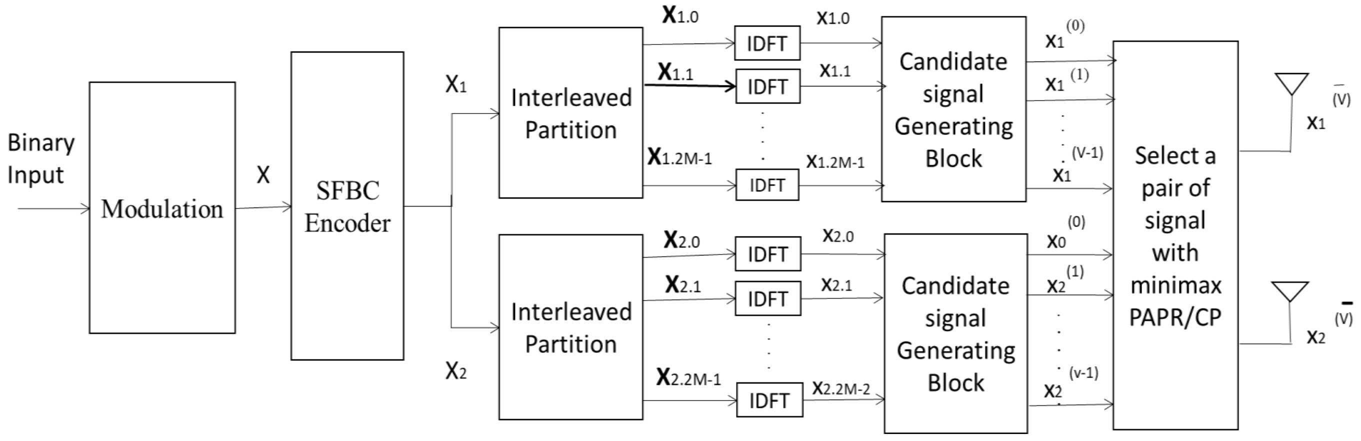

Figure3.1 Block diagram of the proposed LC-eSLM scheme in a SFBC MIMO-OFDM system.

IV. Computational Complexity of ESLM and LC-ESLM Schemes

Analyzes the computational complexities of the eSLM and LC-eSLM schemes by evaluating the required number of complex additions and complex multiplications in each case. Since the Alamouti encoder simply requires complex conjugations, and the corresponding complexity is relatively small and constant irrespective of the number of candidate signals, the SFBC complexity is ignored in comparing the costs of the two schemes. In the eSLM scheme, the Hadamard production of the data blocks and extension blocks requires 4 · N/2 complex multiplications.

Furthermore, each IDFT operation on the oversampled time-domain signals involves (LN)log2(LN) complex additions and (LN/2) log2(LN) complex multiplications for an oversampling factor of L. Hence, the numbers of complex multiplications and complex additions required to construct V pairs of candidate signals in the eSLM scheme are given respectively as

$$MUL_{eSLM} = 2VN + VLN(LN)$$
$$ADD_{eSLM} = 2VLN(LN)$$

If the phase rotations are chosen from a finite set {0, π2, π, 3π 2}, rotating the signal phase simply alters the sign of the signal or exchanges the real and imaginary parts. As a result, it is necessary only to consider the complexity of the amplitude extensions when generating the frequency-domain candidate signals in the eSLM scheme. Only K basic data units are multiplied with amplitude extension factors (1, U) or (U, 1) within each group.

Two complex multiplications are required to generate each 2 × 2 sub-block which is to be extended. Hence, the number of complex multiplications required to generate each frequency-domain candidate signal is equal to 2 ·KN/2M reduces to

$$MUL_{eSLM} = v\left(2K\cdot\frac{N}{2M} + 2\cdot\frac{LN}{2}(LN)\right)$$
$$= \frac{VKN}{M} + VLN(LN)$$

For the LC-eSLM scheme, the space-frequency encoded symbols are interleaved with an equal spacing of 2M. Based on the Cooley-Tukey algorithm [7], each IDFT operation involves LN + (LN/4M) log2(LN/2M) complex multiplications and (LN/2M) log2(LN/2M) complex additions for oversampling by a factor L.

In addition, each candidate signal constructed involves a total of 2MLN complex multiplications and (2M − 1) LN complex additions. Hence, the total numbers of complex multiplications and complex additions required in the LC-eSLM scheme to construct V pairs of candidate signals for both antennas are equal

$$MUL_{LC-eSLM} = 4M\left(\frac{LN}{4M}\left(\frac{LN}{2M}\right) + LN\right) + 4VMLN$$
$$= LN\left(\frac{LN}{2M}\right) + 4MLN(V+1)$$
$$ADD_{LC-eSLM} = 4M\left(\frac{LN}{2M}\left(\frac{LN}{2M}\right)\right) + 2V(2M-1)LN$$
$$= 2LN\left(\frac{LN}{2M}\right) + (4M-2)VLN$$

If the phase rotations are chosen from a finite set {0, π2, π, 3π 2}, it is necessary only to consider the complexity associated with amplitude extension when generating the time-domain candidate signals.

Furthermore, since only K basic data units in each group require amplitude multiplication, only K out of the 2M time-domain signals xi,0, xi,1, · · ·, xi,2M−1 are extended by a factor of U at each antenna.

$$MUL_{LC-eSLM} = 4M\left(\frac{LN}{4M}\left(\frac{LN}{2M}\right) + LN\right) + 2VKLN$$
$$= LN\left(\frac{LN}{2M}\right) + 2LN(2M + KV)$$

Comparing equations, the LC-eSLM scheme requires less number of multiplications with increasing number of candidate signals V.

V. Simulation Results

This section evaluates the performance of the eSLM and LC-eSLM schemes by means of Monte Carlo simulations In performing the simulations, the data symbols are modulated by QPSK or 16-QAM schemes with Gray mapping.

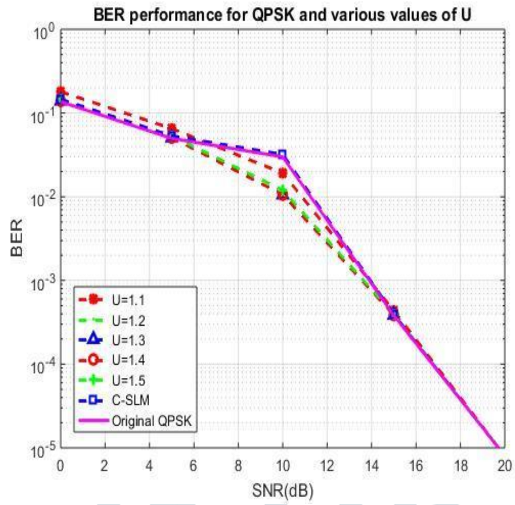

Figure 5.1: - BER Calculation Performance Graph.

Fig 4.1 shows the BER Performance of eSLM scheme as a function of SNR given the use of QPSK and 16-QAM modulation, respectively. In both the figures the number of sub-carriers is set as N= 512. Furthermore, for both modulation schemes, the BER performance is given both for various values of amplitude extension U and the case of perfect SI

Observing fig:4.1 It is seen that BER performance degrades severely with an increasing SNR given an amplitude extension of U=1.1. The degradation of BER is less pronounced as U increases since the SI detection performance improves. However, for large value of U, the average energy of each data symbol reduces after power normalization.

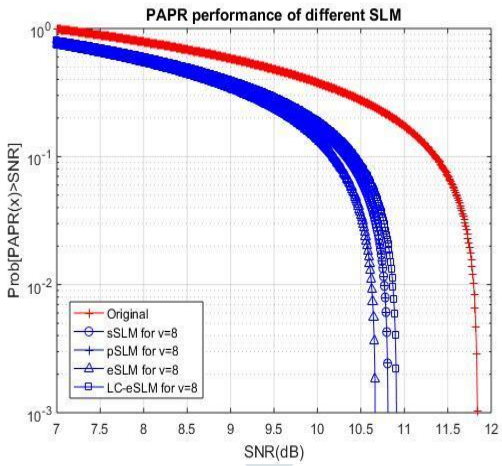

Figure 5.2: - PAPR Performance Graph

The above Fig 5.2 compares the CCDFs of the eSLM and LC-eSLM schemes with those of various SLM-based methods presented i.e.., sSLM, bSLM and pSLM, given the use of V=8 candidate signals. And the (M, K) parameters are set as (4, 1) for V=8.

It is seen that sSLM and pSLM schemes provide an equal PAPR reduction performance since they all apply equal phase rotations to the candidate signals at both antennas. In addition, the proposed eSLM scheme achieves a better PAPR reduction performance than the sSLM, pSLM, schemes as a result of its greater degree of freedom in generating the candidate signals. However, the LC-eSLM scheme has a slightly poorer performance than the sSLM, pSLM, schemes due to its use of repetitive phase rotation sequences for each group.

VI. CONCLUSION

This study has proposed two PAPR reduction schemes, namely eSLM and LC-eSLM, for SFBC MIMO-OFDM systems. Both schemes yield an effective reduction in the PAPR without the need for side information. Furthermore, both methods retain the orthogonality of the SFBC code, and therefore able data symbol demodulation to be performed at the receiver side using only low-complexity linear operations. Notably, the proposed LC-eSLM scheme provides a good tradeoff between the computational complexity and the PAPR reduction performance by generating candidate signals in the time domain. Consequently, it provides a particularly attractive solution for PAPR reduction in practical SFBC MIMO-OFDM systems.

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