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An Efficient Scheme for PAPR Reduction of OFDM Based on Selected Mapping Without Side Information

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ABSTRACT

Orthogonal frequency division multiplexing (OFDM) has become a promising method for many wireless communication applications. However, one main drawback of OFDM systems is the high peak-to-average power ratio (PAPR). Selected mapping (SLM) is a well-known technique to decrease the problem of high PAPR in OFDM systems. In this method, transmitter is obliged to send some bits named side information (SI) for each data block. Such side information causes bandwidth efficiency to get decreased; in addition, incorrect detection of SI in receiver side make the whole data block be lost. In this paper, we propose a technique using linear feedback shift register (LFSR) in which side information bits are not explicitly sent. By considering an OFDM system through the use of 16-QAM modulation as an illustrative example, it is shown that the proposed technique from the view point of bit error rate (BER), probability of detection failure (P_{df}) and PAPR reduction performs very well.

KEYWORDS

Orthogonal Frequency Division Multiplexing (OFDM), Peak-to-Average Power Ratio (PAPR), Selected Mapping, Side Information.

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1- INTRODUCTION

Orthogonal frequency division multiplexing (OFDM), a form of the multicarrier technology, is one of the most popular access methods for many wireless digital communications. It can adopt high data rate links [1]–[4] over harsh multipath fading. However, the time-domain OFDM signal presents a large peak-to-average power ratio (PAPR) at the transmitter's output which is one of the challenging issues in this system. A high PAPR takes out the power amplifier from linear region and reduces the average power of the amplifier output at the transmitter side [5]-[7]. In addition, it causes nonlinear distortion and imposes alteration in the signal constellation. To decrease this PAPR, plenty of methods have been suggested [8]-[14], one of the most well-known of which is Selected Mapping method (SLM) which is able to achieve a better PAPR without distortion.

In SLM, the transmitter generates multiple signals which all represent alike information through multiplying the input data by predetermined phase sequences [15]. Afterwards, the signal with the lowest PAPR is chosen for transmission. In this method, the selected signal index, called side information (SI), must be sent by transmitters to allow for the recovery of the original data block at the receiver side. The transmitted SI bits cause bandwidth efficiency to be decreased; besides, when SI is incorrectly detected by receiver, all received frames be lost. Hence, use of channel coding is needed, which results in system complication and sacrifices data rate. However, for eliminating the explicit transmission of side information some methods have been introduced [16]-[19]. In [19], channel estimation and PAPR reduction are combined in which channel estimation is performed by selecting the place of pilot tone employed inside each data block, and then power distinction between data symbols and the pilot tone is used in order to recover SI in the receiver side. Recently, some schemes have been proposed in [20-21], using permutation and *m*-sequence respectively to eliminate the pilot tones and displaces them with data symbols inside the data block.

In this paper, we propose a new SLM scheme in which linear feedback shift register (LFSR) is used and enables receiver to recover data without explicit SI bits. We abandon the pilot tones and replace them with symbols existing inside the data block. The basic idea in the proposed technique is to fit the side information into transmitted symbols and some special locations in the transmitted data block are expanded, i.e. some transmitted symbols are extended. At the receiver side, it is tried to discover the places of the expanded symbols to recover the SI index. Because in our scheme, discovering the locations of expanded symbols corresponding to the SI index.

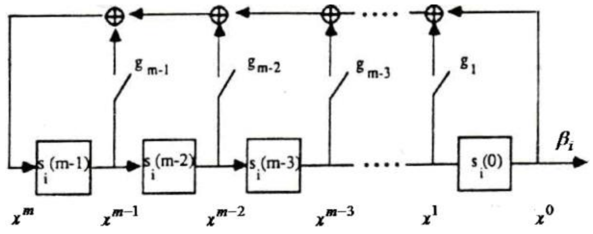

Figure 1: Shift register circuit

This paper is organized as follows. Distribution of PAPR in OFDM systems is presented in the next section. In Section 3, the LFSR circuit and maximum length binary sequence (*m*-sequence) are explained. In Section 4, we describe the proposed scheme in detail. In Section 5, we consider an example to evaluate the performance of the proposed method. Eventually, the paper is concluded in Section 6.

2- DISTRIBUTION OF PAPR IN OFDM SYSTEMS

The distribution of PAPR bears stochastic characteristics in an OFDM system. The complementary cumulative distribution function (CCDF) of PAPR helps to measure the probability that the PAPR of a certain data block exceeds the given threshold. The PAPR distribution of OFDM signals can greatly simplify the system design process. If the number of subcarriers *N* is small, the CCDF of PAPR is given below taken from [22].

$$P(PAPR > z) = 1 - P(PAPR \le z) = 1 - (1 - e^{-z})^N \tag{1}$$

In the OFDM system with a very large *N*, an empirical approximation expression of the CCDF of PAPR can be written as [23]:

$$P(PAPR > z) = 1 - (1 - e^{-z})^{2.8N} \tag{2}$$

Reference [24] derives a simple approximation of the CCDF of PAPR by using the extreme value theory as:

$$P(PAPR > z) \approx 1 - \exp\left\{ -Ne^{-z} \sqrt{\frac{\pi}{3}\log N} \right\} \tag{3}$$

However, in an OFDM system using M-Quadrature Amplitude Modulation (MQAM), when all the subcarriers have the same phase, the maximum of PAPR happens. Therefore, the expression of the upper bound of PAPR, as shown in [25], can be written as:

$$\frac{3N}{M-1} \le PAPR_{\max} \le \frac{3N(\sqrt{M}-1)^2}{M-1} \tag{4}$$

3- MAXIMUM LENGTH BINARY SEQUENCE

One of the most important pseudo-random sequences is the maximum length binary sequence (*m*-sequence) generated by linear feedback shift register (LFSR) circuit. Fig.1 shows the LFSR circuit consisting of *m* memory elements (s_i) each containing a 0 or 1. The first value considered for $S=(s(m-1),s(m-2),...,s(0))$ is said to be the seed. Since each of the memory elements, s_i contains 0 or 1, there exist 2^m possible seeds for LFSR. When all-zero seed happens, the all-zero sequence appear in the output. By deducting the all-zero state, the number of maximum possible seeds is 2^m-1 [26]-[28]. However, for the goal of producing random sequences with a so long period it is required to use the well-chosen feedback in the LFSR circuit in Figure 1. Any LFSR circuit is associated with a generator polynomial determining situation of feedbacks. Accordingly, the well-chosen feedback depends on selecting a suitable generator polynomial. In this paper, we specify the generator polynomial by $g(x)$:

$$g(x)=g_m x^m+g_{m-1}x^{m-1}+...+g_1 x+g_0 \tag{5}$$

where g_0=g_m=1 the other coefficients g_i's are 1 when feedbacks in Figure 1 are connected and 0 otherwise. Furthermore, all algebraic operations are performed in modulo-2 and the output sequence is composed of 0 and 1 values. In design of LFSR circuits, the number of memory elements is equal to the generator polynomial degree. A sequence generated by an LFSR with period 2^m-1 is called a maximal length sequence, or *m*-sequence. To create an *m*-sequence it is required to employ a primitive polynomial of degree *m*. (reference [29] defines the primitive polynomial). At the *i*th clock pulse, the LFSR state is a vector denoted as $s_i=(s_i(m-1),s_i(m-2),...,s_i(0))$, the output sequence is $\beta_i=s_i(0)$. It is significant to say that in LFSR there are *N* cyclic shifts for a sequence with period *N* and the output sequence (β) is an infinite vector with period *N*. By employing the primitive polynomial, the period of output sequence is equal to 2^m-1 containing 2^{m-1} 1's and $2^{m-1}-1$0's. Based on the above subjects, in the LFSR circuit if $g(x)$ is a primitive polynomial of degree *m*, the output sequence is an *m*-sequence and relation (6) is concluded.

$$\text{The number of possible seeds } = \text{Period of } m\text{-sequence} = 2^m-1 \tag{6}$$

4- THE PROPOSED SLM TECHNIQUE

In this part, we use marking in the shape of $G^{(\Omega)}$ to express a vector *G* is composed of Ω scalar quantities.

A- Design of transmitter in the proposed SLM scheme

Suppose an OFDM system with *N* subcarriers which *N* complex data symbols d_n are transmitted concurrently on subcarriers; in addition, a data block is represented in the form of $D^{(N)}=d_n$. In SLM, with the aim of generating a sum of various OFDM signals, the original data is multiplied by *L* predetermined phase vectors

$Q_l^{(N)}=q_{l,n}=[q_{l,0},q_{l,1},...,q_{l,N-1}]$, $l\in\{0,1,...,L-1\}$

where Q_l is composed of *N* complex numbers $q_{l,n}$ and $\left|q_{l,n}\right|=1$, where $|\cdot|$ denotes the absolute value operator. Therefore, *L* various vectors $D_l^{(N)}=d_{l,n}$ are produced with $d_{l,n}=q_{l,n}.d_n$. In time domain, corresponding to the OFDM signal $I_l(t)$ is considered as:

$$I_l(t)=\frac{1}{\sqrt{N}}\sum_{n=0}^{N-1}d_{l,n}.e^{j2\pi f_n}f(t-\tau)$$
$$f(t)=\begin{cases}1 & 0<t\le\tau\\ 0 & t\le 0 \quad t>\tau\end{cases} \tag{7}$$

wherein τ is time duration of the OFDM signal, f_n is the *n*th subcarrier frequency and *f*(*t*) specifies each interval. Then, among the produced signals $I_l(t)$, the one with the lowest PAPR is selected for sending. In other words, the SLM vector that produces the OFDM waveform with minimum PAPR is selected for transmission. Fig.2 depicts a block diagram of the SLM method. In order to discover the original data block D at receiver, the number of $\lfloor \log_2^L \rfloor$ SI bits must be sent by the transmitter. The SI bits specify the special phase vector having minimized PAPR among the *L* phase sequences.

In this paper, we suggest a scheme based on the LFSR circuit and *m*-sequence in which some special locations in the transmitted data block *D* are expanded. Determining the expanded symbols is equivalent to discovering the SI index. In our scheme, the phase vectors $Q_l^{(N)}=q_{l,n}=[q_{l,0},q_{l,1},...,q_{l,N-1}]$, $l\in\{0,1,...,L-1\}$ are such that, for each $Q_l^{(N)}$, the absolute values of some elements are set to a constant value *C* >1, while the absolute values of the other elements are set to 1. Note that the places whose absolute values are *C* >1 are not identical in the phase vectors Q_l. That is, no two separate vectors Q_l can be united with identical *C* places. These locations, in general, are locations of 1's in *m*-sequence and are represented by γ_l. For example, if a supposed phase vector Q_l is associated with $\gamma_l=\{3,6,7,9\}$, it indicates that for each subvector of phase vector Q_l only the elements existing in locations *n*=3,6,7 and 9 have the absolute value equal to $C>1$. In other words, γ_l shows the SI index. Hence, for a given phase vector Q_l there is

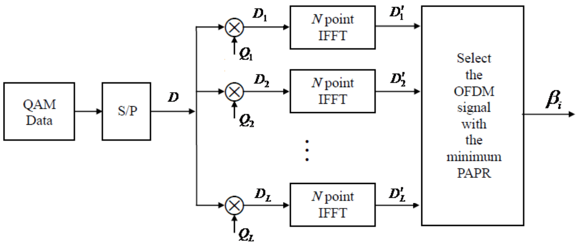

Figure 2: Block diagram of SLM technique

one and only one associated γ_l and vice-versa. The phases of elements $q_{l,n}$, as in conventional SLM, are set to any random values. In the proposed scheme, for creating the set of phase vectors we act as follows.

Step1: The phase vectors $Q_l^{(N)}$ are divided into δ subvectors of length U :

$$U = N/\delta = 2^m - 1 \tag{8}$$

Step2: For a given subvector of length $U = 2^m - 1$, the absolute values of elements are set to 0 or 1 based on the *m*-sequence. That is, for a supposed primitive polynomial of degree *m* (and different seeds) all of the possible *m*-sequences are produced each of which has the length of $2^m - 1$. It is noteworthy that in the proposed scheme, the number of phase vectors is equal to the number of possible *m*-sequences. From (6) it is concluded that in our scheme the number of phase vectors is equal to the length of each subvector, this in turn equals $2^m - 1$. From this, we can have

$$L = U = 2^m - 1 \tag{9}$$

In Figure 3, we illustrate this subject via an example. In this example, an OFDM system with *N*=14 subcarriers is considered. Each phase vector is divided into $\delta = 2$ subvectors of length $U = 2^m - 1 = 2^3 - 1 = 7$. Corresponding to *m*-sequence of length *U* are produced by employing the primitive polynomial of degree *m*=3 ($g(x) = x^3 + x^2 + 1$). According to (9) the number of phase vectors is equal to *L*=7.

Step3: In each subvector obtained from step 2, the ones and zeros are mapped to C >1 and 1, respectively. For a supposed subvector, the absolute values of the elements which are set to a constant value C >1 form a set γ_l. In fact, γ_l shows the side information that is not clearly sent because it is fit into the vector Q_l. For example, in figure 4 the ones and zeros obtained from step3 are mapped to C >1 and 1, respectively and two sub vectors of length $U = 7$ for an OFDM system with *N*=14 subcarriers are generated. In figure 4, we show that the absolute values of the elements because the phases of the elements, like conventional SLM, are chosen randomly.

Note that the number of the primitive polynomial of degree *m* is $\frac{1}{m}\varphi(2^m - 1)$, where φ is the Euler's function.

	sub vector 1							sub vector 2						
l=1	0	0	1	1	1	0	1	0	0	1	1	1	0	1
l=2	0	1	0	0	1	1	1	0	1	0	0	1	1	1
l=3	0	1	1	1	0	1	0	0	1	1	1	0	1	0
l=4	1	0	0	1	1	1	0	1	0	0	1	1	1	0
l=5	1	0	1	0	0	1	1	1	0	1	0	0	1	1
l=6	1	1	0	1	0	0	1	1	1	0	1	0	0	1
l=7	1	1	1	0	1	0	0	1	1	1	0	1	0	0

Figure 3: List of *L*=7 phase vectors for an OFDM system with *N*=14 subcarriers before using the mapping (step2). Each phase vector Q_l is divided into $\delta = 2$ subvectors of length $U = 2^m - 1 = 2^3 - 1 = 7$. Corresponding to *m*-sequences of length *U* are produced by employing the generator primitive polynomial of degree *m*=3 and $g(x) = x^3 + x^2 + 1$. Subvector 2 is a duplicate of subvector 1.

	sub vector 1							sub vector 2						
l=1	1	1	C	C	C	1	C	1	1	C	C	C	1	C
l=2	1	C	1	1	C	C	C	1	C	1	1	C	C	C
l=3	1	C	C	C	1	C	1	1	C	C	C	1	C	1
l=4	C	1	1	C	C	C	1	C	1	1	C	C	C	1
l=5	C	1	C	1	1	C	C	C	1	C	1	1	C	C
l=6	C	C	1	C	1	1	C	C	C	1	C	1	1	C
l=7	C	C	C	1	C	1	1	C	C	C	1	C	1	1

Figure 4: List of *L*=7 phase vectors for an OFDM system with *N*=14 subcarriers after using the mapping (step3). Only the absolute values of the elements have been shown because the phases of the elements are chosen randomly.

The Euler's function $\varphi(\lambda)$ for integer λ is defined as the number of positive integers that are not greater than λ and are co-prime to λ (two numbers are co-prime when they share no factor other than 1). Therefore, in our technique, for detecting the phase vectors at the receiver side, it is necessary that both transmitter and receiver arrive at agreement about the value of "*m*" and primitive polynomial *g*(*x*) of degree *m*. We explain the detection algorithm in detail in Section *B*.

After making all of the phase vectors $Q_l^{(N)}$, the proposed scheme performs like the conventional SLM. That is, through an element-wise multiplication of the data block $D^{(N)}$ by each phase vector, the number of *L* vectors $D_l^{(N)} = d_{l,n}$ with $d_{l,n} = q_{l,n}.d_n$ and also the *L* corresponding $I_l(t)$ are produced. Eventually, the vector having the lowest PAPR is selected for sending. All over this paper, this special vector is associated with the vectors $D_V^{(N)}$ and $Q_V^{(N)}$ where

$$Q_V^{(N)} = q_{V,n} \ , \quad D_V^{(N)} = d_{V,n} \ , \qquad n \in \{1,2,..., N\} \tag{10}$$

In the proposed method, by multiplying the data block $D^{(N)}$ with the phase vectors $Q_l^{(N)}$, the average energy for each transmitted symbol raises because of $|d_{l,n}| = |q_{l,n}|.|d_n|$, with $|q_{l,n}|$ equal to C>1 or 1, leading to $E\left[|d_{l,n}|^2\right] > E\left[|d_n|^2\right]$, where E[.] is the expectation operator.

B- Design of receiver in the proposed SLM scheme

In this paper, we investigate our scheme on a Flat Rayleigh fading channel. The frequency-domain of each transmitted symbol $d_{V,n}$ after passing the Flat Rayleigh fading channel becomes:

$$y_{V,n} = h_n.d_{V,n} + n_n \quad , \quad n \in \{1,2,..., N\} \tag{11}$$

where h_n, a real sample, indicates the fading affecting the nth subcarrier and n_n, a complex sample, exhibits complex Gaussian noise with zero mean and variance σ^2. The receiver, according to the agreement between receiver and transmitter, is aware of the m value and the corresponding primitive polynomial $g(x)$ of degree m. That is, the receiver side knows that for each U-sample subvector $(U = 2^m - 1)$ inside the transmitted data $(Y_V^{(N)})$, some symbols have been expanded by a factor C, (the number of 2^{m-1} symbols), and other symbols have not been expanded (the number of $2^{m-1}-1$ symbols). We assume that at the receiver side the fading samples are perfectly known, that is, perfect channel state information (CSI) is considered. By finding the locations of the expanded symbols, the receiver can recover the SI index. Therefore, at the receiver side, it is tried to discover the places of the expanded symbols correctly to recover the side information. To exhibit the detection algorithm in the receiver side, it is necessary to explain some properties of the m-sequence [30]:

Property 1 ***(The Window Property):***

By sliding a window of width m along an m-sequence, the number of $2^m - 1$ m-fold windows are extracted, each of which is seen exactly once.

For example, "000100110101111" is an m-sequence produced by employing m=4, $f(x) = x^4 + x + 1$ and seed=(0001). Slide a window of width 4 along this sequence and suppose this sequence is cyclically written. In this example, the contents of window #1 is 0001, the contents of the window #2 is 0010,..., and the contents of the window #5 is 0011 and so on.

Property 2 (Indicator position Property):

In an m-sequence, a string of successive 0's of length $m-1$ happens exactly once. Hereafter, we denote the string of successive 0's of length $m-1$ by an "*indicator window*". If position of the *indicator window* in m-sequence is distinguished, positions of other binaries will become specified.

Lemma 1 ***(Z-locations):***

From property 2 it is concluded that:

If position of the *indicator window* in m-sequence becomes distinguished, the locations of ones (2^{m-1} 1's) will become specified.

Detection algorithm

In the proposed scheme, a SI detection by receiver is made of the following stages (see Figure 5).

a) A window of width $m-1$ is slid along the received frame and $2^m - 1$ successive $m-1$ fold windows ($w_i^{(m-1)}$ $i \in \{1,2,..., 2^m - 1\}$) are extracted. It is tried to find the window to which the *indicator window* is attributed.

b) By supposing each of the windows ($w_i^{(m-1)}$) as *indicator window*, according to *Lemma 1*, locations of 1's for each window are specified. For each window $w_i^{(m-1)}$ the received samples existing in locations of 1's are specified by $R_{i,r}$, $r \in \{1,2,..., \delta*(2^{m-1})\}$. Similarly, for each window $w_i^{(m-1)}$ the fading samples existing in locations of 1's are specified by $F_{i,r}$, $r \in \{1,2,..., \delta*(2^{m-1})\}$.

c) The average energy of symbols existing in the locations of 1's for each window $w_i^{(m-1)}$ is given by:

$$\eta_i = \frac{E[|R_{i,r}|^2] - \sigma^2}{E[(F_{i,r})^2]} \quad , \quad r \in \{1,2,..., \delta*(2^{m-1})\} \tag{12}$$

where $E[|R_{i,r}|^2] \approx \frac{1}{\delta*(2^{m-1})} \sum_{r=1}^{\delta*(2^{m-1})} |R_{i,r}|^2$

and $E[(F_{i,r})^2] \approx \frac{1}{\delta*(2^{m-1})} \sum_{r=1}^{\delta*(2^{m-1})} (F_{i,r})^2$

Hence, by computing (12) for each window $(w_i^{(m-1)})$ we can determine the one with the highest η_i. This window is guessed as the *indicator window*. Also, According to the Property 2, locations of other binaries will become specified.

d) By mapping the ones and zeros to C >1 and 1, respectively, we can guess the phase vector.

Such evaluations may cause erroneous detection of the index "V"; therefore, in this case the receiver cannot discover the SI index correctly. However, the system designer must be aware that for a constant m and a larger amount of *C*, a larger increase of the average energy can be achieved. This increase of energy can cause the error performance to decrease.

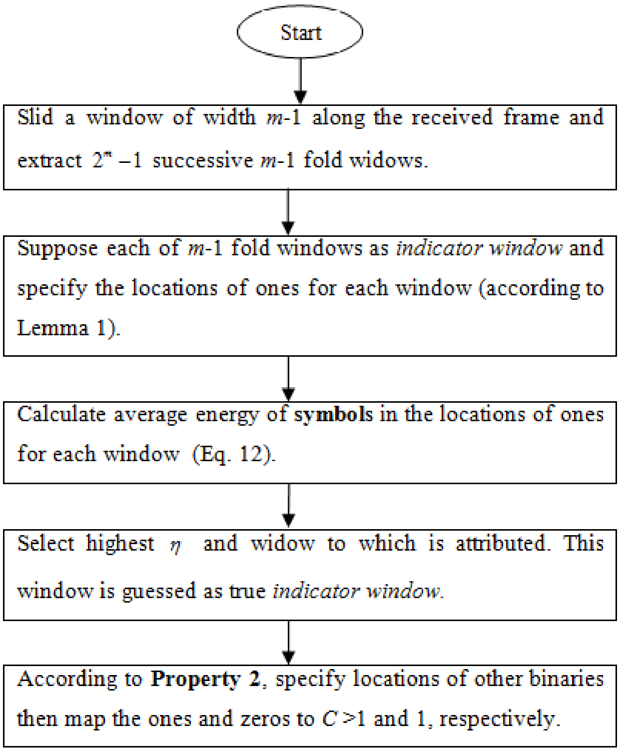

Figure 5: Algorithm of SI detection in receiver.

TABLE I
LFSR PARAMETERS USED IN THE EXAMPLE

m	$g(x)$	Length of subvectors $(U=2^m-1)$	Number of subvectors (δ)	$N=U*\delta$
4	x^4+x^3+1	$U=2^4-1=15$	4	60
			8	120
			17	255
			34	510
			68	1020

5- AN EXAMPLE

In this section, to further explain the proposed technique we assume an OFDM system with *N* subcarriers and 16-QAM modulation. In this example, *m*=4 and corresponding primitive polynomial $g(x)=x^4+x^3+1$ have been employed to generate the *m*-sequences. Hence, the length of each subvector is equal to $U=2^m-1=15$. Table I shows the LFSR parameters used in the example for an OFDM system with *N*=60,120,255,510 and 1020 subcarriers. The phases of $q_{l,n}$ with equal probabilities are set to either 0 or π. We suppose the flat Rayleigh Fading channel for transmission and evaluate our scheme in terms of the probability of the SI detection failure (P_{df}), PAPR reduction and bit error rate (BER).

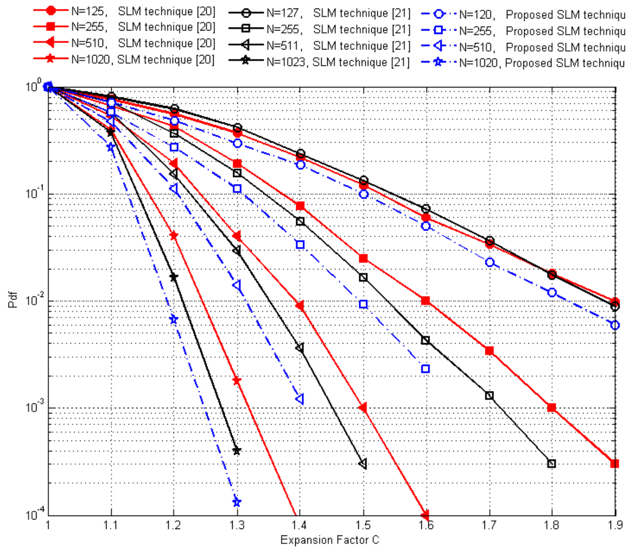

Figure 6. Probability of SI detection failure versus the expansion factor *C* for the proposed SLM technique compared to SLM techniques in [20-21].

C- Probability of SI detection failure (P_{df})

The probability of the SI detection failure, P_{df}, declares the probability that the receiver cannot discover the side information correctly; in other words, an entire received OFDM frame is lost. The results here are achieved for SNR=14dB and system parameters as in Table I. The probability of the SI detection failure P_{df} vs. expansion factor *C* for the proposed SLM technique and the SLM techniques introduced in [20] , [21] are shown in figure 6. It is obvious that the proposed SLM has improved the SI detection performance. From figure 6, it is clear that the values of *N* and *C* much influence the P_{df} performance. Our SI detection algorithm in a higher value of constant *C* causes more distinction between the expanded and non-expanded symbols. Therefore, the proposed scheme especially in a higher value of *C* acts better. Additionally, whenever *N* rises, zero locations of each frame are also increased. Hence, it provides further valid estimates of the average energies in (12). Consequently, in the proposed scheme, the more increase in *N*, the better result can be achieved for any value of the expansion factor *C*. Other modulations, such as QPSK or 64-QAM, can be applied in the proposed scheme for the analysis of P_{df} compared to the 16-QAM. For a given set of system parameters (such as the number of subcarriers (*N*), the agreed primitive polynomial *g*(*x*), etc.) if the lower-order signal constellation, for example QPSK, is used the value of P_{df} will be smaller. But by replacing

64-QAM or 256-QAM with 16-QAM this probability will be worsen. As a conclusion, if the probability of the SI detection failure is made small enough (e.g., by increasing the expansion factor *C* and/or the number of subcarriers), this probability in a case of QPSK will be smaller than the higher-order signal constellation.

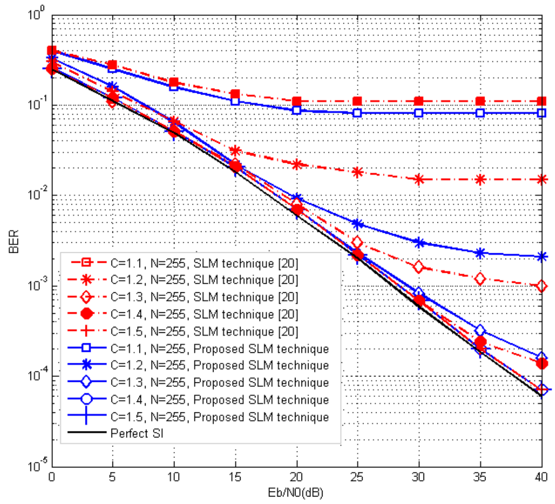

Figure 7a. Comparison between BER performance of the proposed SLM and the SLM technique introduced in [20].

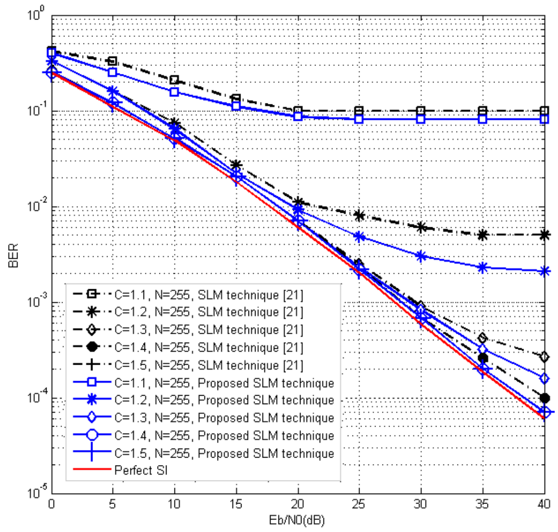

Figure 7b. Comparison between BER performance of the proposed SLM and the SLM technique introduced in [21].

D- BER performance evaluation

The simulation results here are obtained for system parameters as given in Table I and various values of constant *C*. Figures 7a and 7b show the BER performance of the proposed SLM technique compared to the SLM technique introduced in [20-21]. As seen from these figures, the proposed scheme has improved the BER performance. But the receivers in [20], [21] suffer from the differences between symbols energies and average symbol energy. The proposed receiver can overcome this problem because it estimates the symbol energy especially for high SNRs. The scenario of perfect SI is gained by employing the strong channel code absolutely devoted to the SI protection as a result of which, system complication and reduction of data rate will happen. As figures 7a and 7b show, the proposed scheme improves the BER performance especially for *C*=1.2, and 1.3 and it comes closer to the perfect SI performance for *C*=1.4, and 1.5.

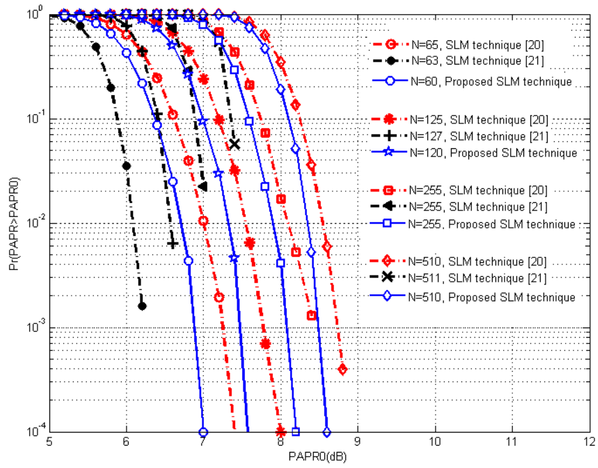

Figure 8. CCDF of the PAPR obtained with the proposed SLM method and the SLM techniques introduced in [20-21] with oversampling factor of 4.

E- Evaluation of PAPR reduction

The CCDF of PAPR has been shown in Figure 8. The simulation results of the proposed SLM technique are achieved for the 16-QAM modulation, *C* = 1.4, *N* = 60, 120, 255, and 510 subcarriers for system parameters according to Table I. These results have been achieved by application of oversampling=4 [31]. The CCDF of the proposed SLM technique compared to the SLM techniques introduced in [20-21] are shown in figure 8. It is obvious that the PAPR reduction performance of the proposed scheme, for all *N* values, is better than that of the SLM techniques introduced in [20], but not that of the other technique introduced in [21].

6- Conclusion

In this paper, we have proposed a new SLM scheme to reduce PAPR without transmitting the explicit SI bits. In the proposed method, we have used LFSR circuit which enables the phase coefficient vector to be random. Our examinations performed via considering an OFDM system by employing the 16-QAM modulation. As the results show, from the view point of bit error rate, the probability of the SI detection failure and PAPR reduction, the proposed method performs better than that of the SLM technique introduced in [20]. It could be concluded from the results that both the BER and Pdf performances of the proposed scheme compared to the SLM technique introduced in [21] have significant improvement. However, the technique introduced in [21] results in a better PAPR reduction performance than that of the proposed method; in other words, there is a trade-off between PAPR reduction and both the BER and Pdf performances.

7- ACKNOWLEDGMENT

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