Low-Complexity Architecture for PAPR Reduction in OFDM Systems with Near-Optimal Performance

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Abstract: ***To reduce the PAPR in OFDM systems selected mapping schemes (SLM) are widely used due its distortion less nature. However a major drawback of traditional SLM technique is high computational complexity to select a low PAPR signal it requires a bank of inverse fast Fourier (IFFT) operations. This paper proposes a novel architecture for PAPR reduction in OFDMs with low computational complexity. In this proposed method, frequency domain cyclic shifting, complex conjugate, sub-carrier reversal operations are performed to increase the PAPR reduction performance in OFDM systems whereas in traditional SLM scheme only frequency domain phase rotation can be performed to generate the candidate signals. Furthermore, to reduce the multiple IFFT problems, all of the frequency domain equivalent operations are converted into time-domain equivalents. It is shown that the sub carrier partitioning and re-assembling processes are important in realizing low complexity time domain equivalent operations. Moreover, it is shown theoretically and numerically that the computational complexity of the proposed scheme is significantly lower than the traditional SLM method and the PAPR reduction performance is within 0.001 dB of that SLM.***

Index Terms—Orthogonal frequency division multiplexing (OFDM), peak-to-average power ratio (PAPR), selected mapping (SLM).

1. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is one of the most promising techniques for achieving high-rate data transmissions due to its high spectral efficiency and inherent robustness toward multi-path channels. However, OFDM systems suffer a high peak to-average power ratio (PAPR) of the transmitted signals, which causes significant in-band distortion and out-of-band radiation when the signals are passed through a nonlinear power amplifier. The literature contains various proposals for PAPR reduction, including tone reservation (TR) companding, tone injection (TI), active constellation extension (ACE), interleaving, partial transmit sequence (PTS) and selected mapping (SLM) Of these techniques, SLM is the most commonly applied due to its distortion less nature. However, the computational complexity of the traditional SLM scheme is extremely high since it requires a bank of inverse fast Fourier transform (IFFT) operations. To address the problem multiple IFFTs, several low complexity SLM architectures have been proposed in which the frequency-domain phase rotation operations are converted into time-domain equivalent operations. For example, the IFFT operations are replaced by conversion vectors obtained by taking the IFFT of the phase rotation vectors. Similarly, the conversion vectors are specified in the form of perfect sequences. However, the practical usefulness of these time-domain approaches as an alternative to the traditional SLM method depends on the time-domain equivalent operations having a low computational complexity. As a result, only a limited selection of sequences can be applied. Furthermore, since the adopted sequences are not randomly generated, but are somewhat correlated due to the low-complexity requirement, the frequency-domain equivalent phase rotations are not truly random, and thus a substantial degradation of the PAPR reduction performance occurs. It is worth noting that a low-complexity interleaving-based PAPR reduction scheme has been proposed, where three frequency-domain operations, namely, frequency-domain cyclic shifting, complex conjugate, and sub-carrier reversal, are adopted to scramble the sub-carriers thus to increase the PAPR diversity of various candidate signals. To circumvent the multiple-IFFT problem, the frequency-domain operations are converted into time-domain equivalents. It is demonstrated that this scheme outperforms traditional interleaving-based PAPR reduction scheme. However, there is significant performance loss relative to that of the traditional SLM scheme. Therefore, scrambling the sub-carriers is inadequate to reach the optimal PAPR reduction performance. This paper proposes a novel architecture for reducing PAPR in OFDM systems with a lower computational complexity than the traditional SLM scheme whilst maintaining an equivalent

2. LITERATURE SURVEY

Jung-Chieh This letter considers selection of the optimal peak reduction tone (PRT) set for the tone reservation (TR) scheme to reduce the peak-to-average power ratio (PAPR) of an orthogonal frequency division multiplexing (OFDM) signal. In the TR scheme, PAPR reduction performance achieved by a randomly generated PRT set is superior to that by a consecutive PRT set and that achieved by an interleaved tone set. However, the optimal PRT set requires an exhaustive search of all combinations of possible PRT sets, which is known to be a nondeterministic polynomial-time (NP)-hard and cannot be solved for the practical number of tones. Inspired by the efficiency of the cross-entropy (CE) method for finding near-optimal solutions in huge search spaces, this letter proposes the application of the CE method to search the optimal PRT set. Computer simulation results show that the proposed CE method obtains near-optimal PRT sets and provides better PAPR performance.

Haibo Li In this letter, an improved tone reservation method, which is based on the least squares approximation, is proposed to reduce the peak-to-average-power ratio (PAPR) in orthogonal frequency division multiplexing (OFDM) systems. Compared with the clipping control method, the proposed scheme can generate the optimal peak-canceling signals with fast convergence. Simulation results show that the proposed scheme only needs two iterations at most to achieve almost the same PAPR reduction as that of the clipping control method.

Tao Jiang Provide the design criteria of the nonlinear companding transforms for reduction in peak-toaverage power ratio (PAPR) of multi-carrier modulation (MCM) signals, which can enable the original MCM signals to be transformed into the desirable distribution. As examples, some novel nonlinear companding transforms have been proposed to transform the amplitude or power of the original MCM signals into uniform distribution, which can effectively reduce the PAPR for different modulation formats and subcarrier sizes without any complexity increase and bandwidth expansion. It has been shown by computer simulations that the proposed schemes can significantly improve the performance of MCM systems including bit-error-rate and PAPR reduction.

Jung-Chieh Chen This letter considers the use of the tone injection (TI) scheme to reduce the peak-to-average power ratio (PAPR) of an orthogonal frequency division multiplexing (OFDM) signal. TI is a distortion less technique that can reduce PAPR significantly without data rate loss and does not require the extra side information. However, the optimal TI scheme requires an exhaustive search over all combinations of possible permutations of the expanded constellation, which is a potential problem for practical applications. To reduce the computational complexity while still improving PAPR statistics, this letter first formulates the PAPR reduction with TI scheme as a particular combinatorial optimization problem. Next, it proposes the application of the cross-entropy (CE) method to solve the problem. Computer simulation results show that the proposed CE method obtains the desired PAPR reduction with low computational complexity.

Brian Scott Korngold The high peak-to-average power ratio (PAR) in Orthogonal Frequency Division Multiplexing (OFDM) modulation systems can significantly reduce power efficiency and performance. Methods exist which alter or introduce new signal constellations to combat large signal peaks. We present a new PAR-reduction method that dynamically extends outer constellation points in active (data-carrying) channels, within margin-preserving constraints, in order to minimize the peak magnitude. This scheme simultaneously decreases the bit error rate slightly while substantially reducing the peak magnitude of an OFDM transmit block.

Furthermore, there is no loss in data rate and unlike other methods, no side information is required. PAR reduction for an approximated analog signal is considered, and about a 4.6 dB reduction at a 10 5 symbol-clip probability is obtained for 256-channel QPSK OFDM. The results show great promise for use in commercial systems.

Kitaek Bae For PAR reduction in OFDM systems, the clipping based Active Constellation Extension (ACE) technique is simple and attractive for practical implementation. However, we observe it cannot achieve the minimum PAR when the target clipping level is set below an initially unknown optimum value. To overcome this low clipping ratio problem, we propose a novel ACE algorithm with adaptive clipping control. Simulation results demonstrate that our proposed algorithm can reach the minimum PAR for severely low clipping ratios. In addition, we present the tradeoff between PAR and the loss in Eb/No over an AWGN channel in terms of the clipping ratio.

3. PAPR REDUCTION SCHEME IN FREQUENCY DOMAIN

This section describes the implementation of the proposed PAPR reduction scheme in the frequency domain. In the traditional SLM scheme, the candidate signals are generated using frequency-domain phase rotation only. By contrast, in the scheme proposed in the present study, the PAPR diversity of the candidate signals is increased by performing additional frequency-domain cyclic shifting, complex conjugate, and subcarrier reversal operations.

Consider an OFDM system with N sub-carriers. Let the sub-carriers be partitioned into S sub-carrier sets T_s, s = 0, 1. . . S − 1, where each set contains $\frac{N}{s}$ sub-carriers, in which $\frac{N}{s}$ is assumed to be a positive integer. In general, three methods exist for partitioning the sub-carriers in OFDM systems, namely the localized partitioning method (LPM), the distributed partitioning method (DPM), and the hybrid partitioning method (HPM).

In LPM, each sub-carrier set consists of a number of adjacent and consecutive sub-carriers, i.e., T_s, = {i + s · $\frac{N}{s}$ |i = 0, 1, . . . , $\frac{N}{s}$ − 1}, s = 0, 1, . . . , S − 1. Meanwhile, in DPM, each sub-carrier set consists of multiple interleaved sub-carriers of equal spacing, i.e., T_s, = {s + i · S|i = 0, 1, . . . , $\frac{N}{s}$ −1}, s = 0, 1, . . . , S−1. Finally, in HPM, the subcarriers are partitioned into U localized sub-carrier sets, which are then further partitioned into V distributed sets. In one extreme, if we set U = 1, HPM becomes DPM. In the other extreme, if we set V = 1, HPM becomes LPM. Therefore, both LPM and DPM can be regarded as special cases of HPM. The total number of sub-carrier sets is therefore equal to S = U ·V, with the set indexes being denoted as s = u·V +v, where u = 0, 1, . . . , U − 1 and v = 0, 1, . . . , V − 1. In other words, the sub-carrier indexes of a given set are given by T_s, = {$\frac{N}{U}$ · ⌊ $\frac{s}{V}$ ⌋ + (s)V + i · V |i = 0, 1, . . . , $\frac{N}{s}$ − 1}, where ⌊·⌋ and (·)V denote the floor function and the modulo V operation, respectively.

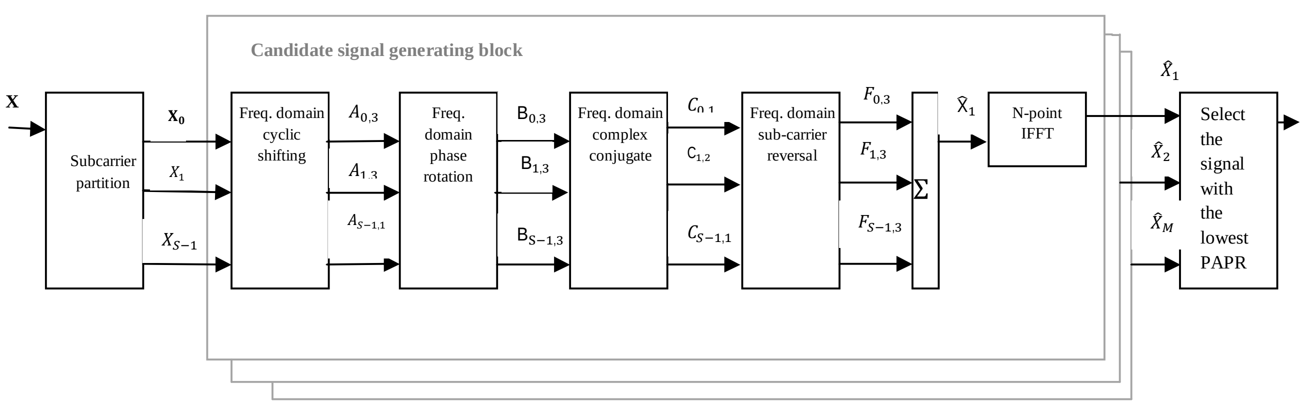

Fig.1. System architecture of proposed scheme in frequency domain

4. Time-Domain Signal Properties of OFDM Systems

In this section, the complexity of the frequency-domain architecture described above is reduced by means of four time-domain equivalent properties. In addition, a time-domain repetition property is introduced in order to further reduce the computational complexity. Note that all of the operations described in this section (both frequency-domain and time-domain) are performed on the sub-carriers of the same set.

5. Proposed Method

Each candidate signal requires the use of an N-point IFFT operation, and thus the scheme has a high computational complexity. To resolve this problem, this section utilizes the time-domain equivalent operations to construct low-complexity architecture for PAPR reduction.

Below figure presents a block diagram of the proposed architecture, in which the frequency-domain data vector **X** is partitioned into **S** data vectors $\mathbf{X_s}$ of size N×1, s = 0, 1, . . . , S−1, in accordance with. Note that in implementing the proposed architecture, the HPM sub-carrier partitioning method is adopted in order to maximize the PAPR diversity. As shown in Fig. 3, having partitioned the sub-carriers, an IFFT operation is performed on $\mathbf{X_s}$ to obtain the corresponding time-domain data vector x s of size N × 1. It is noted that although the proposed scheme still requires S N-point IFFT operations, the computational complexity of the proposed architecture is much lower than that of the traditional SLM scheme since S is much smaller than the number of candidate signals M. Furthermore, the computational complexity of each N-point IFFT is significantly decreased in the proposed scheme since most of the elements of $\mathbf{X_s}$ are zero.

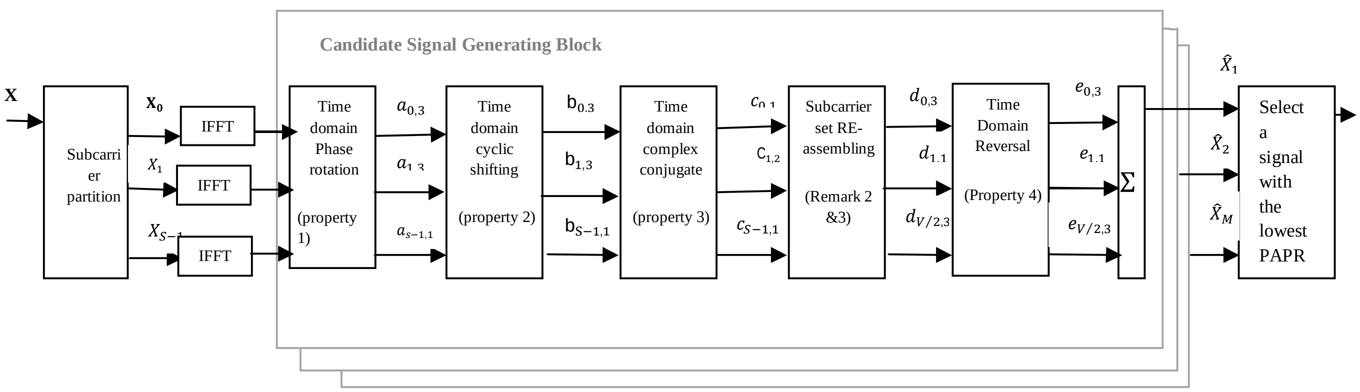

Fig.2. System architecture of proposed scheme in time domain

Following the IFFT operations, the time-domain data vectors$\mathbf{x_s}$, s = 0, 1,…, S − 1, are processed by **M** CSGBs in order to generate **M** candidate signals. It is noted that the summation of $\mathbf{x_s}$ generates the original time-domain transmitted signal. In each of the M CSGBs, each $\mathbf{X_s}$ is first processed by the time-domain phase rotation block. The resulting output signal is denoted as $a_{s,m}$, where the *n* th element is given by

$$a_{s,m}[n] = x_s[n] \cdot exp\left\{\frac{j2\pi n \cdot l_{s,m}}{N}\right\} \quad n = 0,1,\ldots,N-1 \qquad (1)$$

In which $x_s[n]$ is the *n* th element of $\mathbf{X_s}$ and $l_{s,m}$ is the number of frequency-domain cyclic shifts of Γs for the *m* th candidate signal. the selection of the cyclic shift value $l_{s,m}$ for a given m is not arbitrary, but is based on a joint consideration over various *s*. The second block of the CSGB performs a time-domain cyclic shifting operation, i.e., the time-domain equivalent of the frequency-domain phase rotation operation. Therefore, the *n* th element of the output signal $b_{s,m}$ has the form

$$b_{s,m}[n] = a_s[(n - w_{s,m})N], \quad n = 0,1,\ldots,N-1 \qquad (2)$$

Where $w_{s,m}$ denotes the number of cyclic shifts of the s th sub-carrier set for the m th CSGB. It will be recalled that the frequency-domain phase rotation operation is not arbitrary. In practice, phase rotation is not completely random, and thus the PAPR reduction performance is slightly degraded. However, this drawback is minor compared to the substantial reduction achieved in the computational complexity of the proposed scheme.

The third block in Fig. 2 performs the time-domain complex conjugate operation, i.e., the equivalent operation of the frequency-domain complex conjugate process. Since the system arbitrarily chooses whether or not to perform the complex conjugate operation, the nth element of the output signal $C_{s,m}$ has the form

$$C_{s,m}[n] = b_{s,m}[n] \quad \text{or} \quad b^{*}_{s,m}[(-n)N] \qquad (3)$$

The sub-carrier sets must be properly re-assembled before performing the time domain signal reversal operation. In particular, when using the HPM partitioning method, the sub-carrier sets must be reassembled in such a way that the partition is equivalent to that obtained using DPM (Remark 3).

$$\bar{\boldsymbol{c}}_{s,m} = \sum_{u=0}^{U-1} \boldsymbol{C}_{(u\cdot V+\bar{s}),\boldsymbol{m}}, \qquad \bar{s} = 0,1,\ldots,V-1 \qquad (4)$$

In the second step, sub-carrier sets $\bar{\Gamma}_{\bar{s}}$ and $\bar{\Gamma}_{S-\bar{s}}$,$\bar{s} = 1,2,\ldots.\frac{v}{2}-1$ are combined to form a single sub-carrier set $\breve{\Gamma}_q, q = 1,2,\ldots.\frac{v}{2}-1$, while leaving $\breve{\Gamma}_{\bar{s}=0}$ and $\breve{\Gamma}_{\bar{s}=v/2}$ unchanged, i.e.,

$$d_{q,m} = \begin{cases} \bar{c}_{0,m}, & q = 0 \\ \bar{c}_{\frac{v}{2},m}, & q = \frac{v}{2} \\ \bar{c}_{q,m} + \bar{c}_{V-q,m}, & q = 1,2,\ldots,\frac{v}{2}-1 \end{cases} \qquad (5)$$

Substituting (4) into (5) yields

$$d_{q,m} = \begin{cases} \sum_{u=0}^{U-1} c_{u\cdot V,m}, & q = 0 \\ \sum_{u=0}^{U-1} c_{u\cdot V+\frac{v}{2},m}, & q = \frac{v}{2} \\ \sum_{u=0}^{U-1} c_{u\cdot V+q,m} + \sum_{u=0}^{U-1} c_{u\cdot V+(V-q),m}, & q = 1,2,\ldots,\frac{v}{2}-1 \end{cases} \qquad (6)$$

It is worth noting that all operations before re-assembling are performed by HPM. Thus, the resulting signals are not equivalent to those obtained by DPM from the beginning. It is noted that the Sub-carrier Set Re-assembling block of the proposed time-domain architecture uses the time-domain repetition property in order to reduce the computational complexity. Thus, as shown in Fig. 4, in the first step, it is necessary only to generate the first N/V elements of $\bar{c}_{\bar{s},m}, \bar{s} = 0,1,\ldots,V-1$. In other words, the remaining N − N/V elements of c ¯0,m and ¯ cV/2,m can be obtained directly using Property 5. Similarly, for ¯ cs,m ¯ , s ¯ = 1, 2, . . . , V/2 − 1, V/2 + 1, . . . , V − 1, only the first N/2 elements are obtained from the first N/V elements using the time-domain repetition property and are sent to the second stage. In the second step, the first N/2 elements of ¯c q,m + ¯cV −q,m, q = 1, 2, . . . , V 2 − 1 are obtained and the remaining N/2 elements are then generated using the timedomain repetition property once again. It is worth noting that applying the time-domain repetition property does not involve any complex multiplications when V = 2, 4 since βs,i ¯ ∈ {±1, ±j}.

Figure 3 presents an illustrative example of the reassembling operation for the case of eight time-domain signals, c s,m, s = 0, 1, . . . , 7, corresponding to the eight frequency domain sub-carrier sets shown in Fig. 1(c). The sub-carrier sets are firstly re-assembled such that the partition is converted from a HPM form to a DPM form, i.e., $\bar{T}_{\bar{s}=0} = T_0 \cup T_4 = \{0, 4, 8, 12\}$, $\bar{T}_{\bar{s}=1} = T_1 \cup T_5 = \{1, 5, 9, 13\}$, $\bar{T}_{\bar{s}=2} = T_2 \cup T_6 = \{2, 6, 10, 14\}$ and $\bar{T}_{\bar{s}=3} = T_3 \cup T_7 = \{3, 7, 11, 15\}$, resulting in V = 4 DPM sub-carrier sets. Sub-carrier sets $\bar{T}_1$ and $\bar{T}_3$ are then combined to form a single sub-carrier set $\breve{T}_1$. In addition, we have $\breve{T}_0 = \bar{T}_0$ and $\breve{T}_2 = \bar{T}_2$. Therefore, the output of the re-assembling operation comprises three signals, namely $d_{q,m}$, q = 0, 1 and 2, as given in (19).

The sub-carrier set re-assembling operation is followed by the time-domain signal reversal process (see Fig. 2). As with the complex conjugate operation, the system arbitrarily chooses whether or not to perform the reversal operation. The nth element of the resulting signal $e_{q,m}$, q = 0, 1. . . $\frac{V}{2}$, therefore has the form

$$e_{q,m}[n] = d_{q,m}[n] or\ d_{q,m}[(-n)_N], q = 0,1, \dots, \frac{V}{2}. \qquad (7)$$

Finally, the m th candidate signal is obtained by adding all the $e_{q,m}$ of the m th CSGB, to give

$$\hat{X}_m = \sum_{q=0}^{V/2} e_{q,m} \qquad (8)$$

Having generated M candidate signals, the signal with the lowest PAPR is selected for transmission. It should be noted that the proposed scheme requires various operations at the transmitter, but the related parameters can be stored at both the transmitter and receiver with code book. Therefore, the number of side information bits depends only onthe number of candidate signals. If M candidate signals are generated, the scheme requires only log_2[M] bits to transmit side information. In addition, the side information is assumed to be transmitted through the control channel, where channel coding is adopted to protect the side information from being erroneously detected.

Subcarrier index k	0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
Subcarrier set index s	4	1	6	3	0	5	6	3	0	5	2	7	4	1	2	7
Combined Subcarrier set index $\bar{s}$	4	1	6	3	0	5	6	3	0	5	2	7	4	1	2	7
Index q in (19)	0	1	2	1	0	1	2	1	0	1	2	1	0	1	2	1
Subcarrier index k	X[12]	X[5]	$X^*[10]$	$X^*[3]$	X[0]	X[9]	$X^*[14]$	$X^*[7]$	X[4]	X[13]	X[2]	X[11]	X[8]	X[1]	X[6]	X[15]

Fig.3. Illustrative example of Sub-carrier Set Re-assembling operation (N = 16, U = 2, V = 4).

5. Results

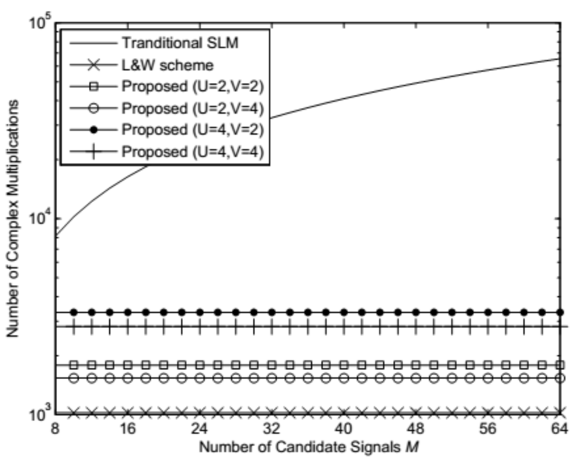

Fig.4. Number of complex multiplications as function of number of candidate signals M (N = 256).

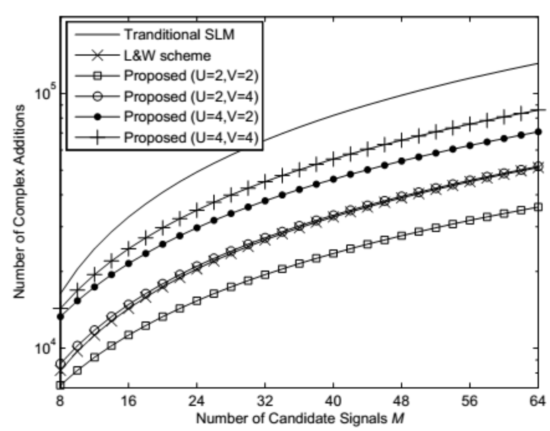

Fig.5. Number of complex additions as function of number of candidate signals M (N = 256).

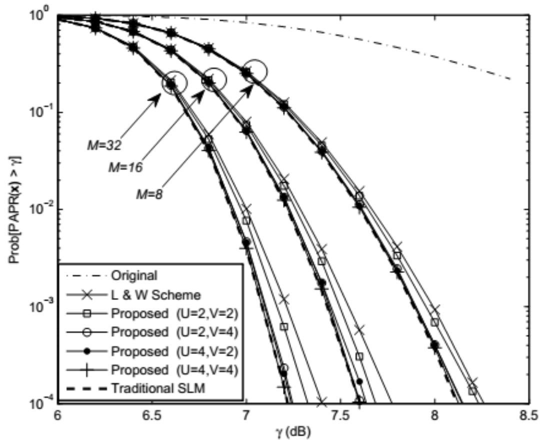

Fig.6. PAPR reduction performance of various schemes (16-QAM, N = 256).

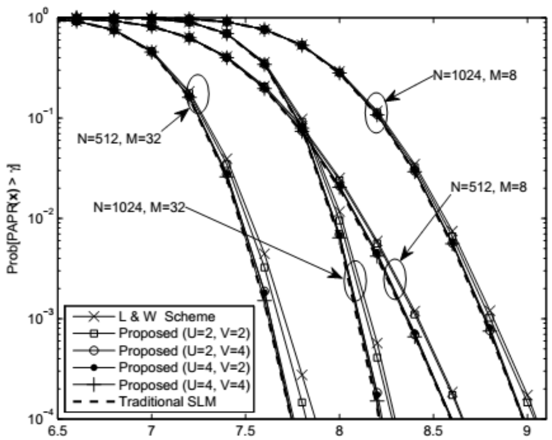

Fig.8. PAPR reduction performance of various schemes (16-QAM, N = 512, 1024).

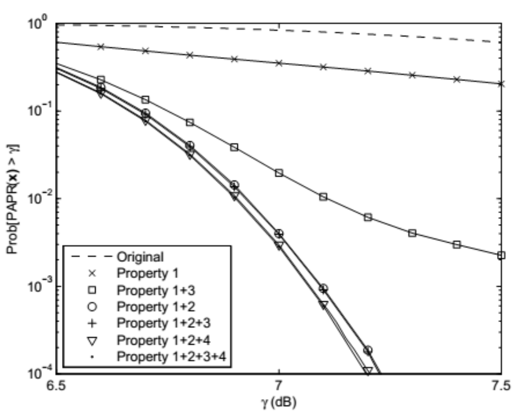

Fig.7. PAPR reduction performance of various combinations of frequency domain operations (16-QAM, M = 32, N = 256, U = 4, V = 4).

6. Conclusion

This paper has presented a new low-complexity architecture for PAPR reduction in OFDM systems. Compared to the traditional SLM scheme, in which the candidate signals are generated using frequency-domain phase rotation only, the architecture proposed in this study additionally uses frequencydomain cyclic shifting, complex conjugate and sub-carrier reversal operations to maximize the PAPR diversity of the candidate signals. In order to avoid the multiple-IFFT problem inherent in the traditional SLM method, the proposed scheme converts all four frequency-domain operations into time-domain equivalent operations. It has been shown that the computational complexity of the proposed approach can be minimized through an appropriate partitioning and reassembling of the sub-carriers in the OFDM system. In addition, the theoretical analysis results have shown that the number of complex multiplications and complex additions required in the proposed scheme for (U, V) = (4,4) are 8.59% and 68.75%, respectively, of those required in the traditional SLM scheme. Furthermore, the simulation results have shown that the performance loss of the proposed scheme relative to that of the traditional SLM scheme is less than 0.001 dB for 16-QAM, M = 32, N = 256, U = 4, V = 4, and Pr(PAPR(x) > γ) = 10−4. In other words, the proposed scheme closely approximates the PAPR reduction performance of the traditional SLM method, but with a significantly reduced computational complexity.

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