Efficient PTS Scheme with Adaptive Selection Method for Dominant Samples in OFDM Systems

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***Abstract*—In OFDM systems, high peak-to-average power ratio (PAPR) of OFDM signals is one of the most important problem. As a solution of the PAPR problem in OFDM systems, partial transmit sequence (PTS) scheme is quite a suitable scheme due to its PAPR reduction performance and distortion characteristic. However, high computational complexity is serious problem in the PTS scheme. In this paper, in order to reduce the computational complexity of the previous PTS schemes, efficient PTS scheme is proposed. Although the proposed PTS scheme uses dominant time-domain samples similar to some previous low-complexity PTS schemes, they utilize more efficient selection method for dominant time-domain samples. The proposed PTS scheme lowers the computational complexity compared to the previous PTS schemes while achieving the optimal PAPR reduction performance.**

***Keywords*—Dominant time-domain samples, orthogonal frequency division multiplexing (OFDM), partial transmit sequence (PTS), peak-to-average power ratio (PAPR).**

I. Introduction

ORTHOGONAL frequency division multiplexing (OFDM) is a technique of encoding data on multiple sub-carriers in the wireless communication systems. OFDM has high spectral efficiency and easy adaptation to severe channel without complex time-domain equalization. A communication system using OFDM is robust against inter-symbol interference (ISI) and fading caused by multipath propagation, as well as narrow-band co-channel interference. In addition, the OFDM system is efficient for hardware implementation because it can use fast Fourier transform (FFT) instead of discrete Fourier transform (DFT). Besides, it has low sensitivity to time synchronization errors. Due to these various and valuable advantages described as above, OFDM has been adopted as one of the most popular modulation technique for wireless communications. OFDM has been used in the various applications such as digital audio broadcasting (DAB), digital video broadcasting (DVB), digital media broadcasting (DMB), wireless local area network (WLAN) IEEE 802.11 [1], long term evolution (LTE), and LTE Advanced 4G mobile phone standards. Also, OFDM is a candidate for 5G mobile phone standards.

This work was supported in part by the National Research Foundation of Korea through the Korea government (MSIP) under Grant NRF-2016R1A2B2012960, and in part by the Brain Korea 21 Plus Project in 2017.

However, OFDM has not only the advantages but also some disadvantages. High peak-to-average power ratio (PAPR) of OFDM signals is one of the most serious problems in OFDM systems. Due to the non-linear property of high power amplifier (HPA), HPA output of OFDM signals with high PAPR causes in-band distortion and out-of-band radiation, which result in degradation of communication quality such as bit error rate (BER).

In order to solve the PAPR problems in OFDM systems, various PAPR reduction schemes have been proposed [2]–[8]. Among these PAPR reduction schemes, SLM and PTS schemes can effectively reduce the PAPR of OFDM signals without causing signal distortion. SLM and PTS schemes require several inverse fast Fourier transform (IFFT) to generate candidate signals among which a candidate signal with the lowest PAPR is selected and then transmitted. In an OFDM system, since FFT and IFFT operations account for a substantial part of hardware complexity, IFFT operations to generate candidate signals in SLM or PTS schemes can lay an additional burden on the system. In general, PTS schemes require less IFFT operations to generate several candidate signals, compared with SLM schemes.

The conventional PTS scheme [8] has great PAPR reduction performance with simple idea that a signal with the lowest PAPR is selected for transmission among candidate signals generated by efficient methods. However, since the conventional PTS scheme requires a lot of computational complexity, various low-complexity PTS schemes have been proposed [9]–[19]. Recently, a new type of reduced-complexity PTS (RC-PTS) [17] was proposed to reduce the computational complexity by estimating the PAPRs of candidate OFDM signals based on only the selected dominant time-domain samples and find the signal with the lowest estimated PAPR. After that, two improved PTS schemes, low-complexity PTS (LC-PTS) [18] and the other low-complexity PTS (OLC-PTS) [19] have been proposed to further reduce the computational complexity and enhance the PAPR reduction performance of RC-PTS.

In this paper, a new low-complexity PTS scheme is proposed to further reduce computational complexity compared with PTS schemes by using a more efficient selection method for dominant samples compared with RC-PTS and LC-PTS. This PTS scheme uses advanced sub-signal rotation method, which has adaptive property for further reducing its computational complexity. The proposed PTS scheme has much lower computational complexity compared with RC-PTS and LC-PTS while achieving the optimal PAPR reduction performance, which is the same as that of the conventional PTS.

The rest of the paper is organized as follows. In Section II, for easy understanding of the proposed PTS schemes, some important preliminaries are reviewed. Then, the efficient PTS scheme with adaptive selection method for dominant time-domain samples is proposed in Section III. In Section IV, the computational complexity of the proposed PTS scheme is analyzed and the simulation results of the computational complexity and the PAPR reduction performance are provided. Finally, conclusions are given in Section V.

II. Preliminaries

A. OFDM and PAPR

In OFDM systems, a serial block of N modulated symbols is organized from serial original data bits and converted to a parallel block to generate the corresponding N frequency-domain OFDM symbols. A time-domain OFDM signal is generated by adding N OFDM symbols modulated onto the corresponding N orthogonal sub-carriers (sub-channels) with same bandwidth. The complex baseband OFDM signal x_t is obtained as

$$x_t = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k e^{j2\pi k \Delta f t}, \ 0 \le t \le NT \tag{1}$$

where $j = \sqrt{-1}$, Δf denotes the sub-carrier bandwidth, and NT denotes the period of the N OFDM symbols. Note that sub-carriers of the OFDM signal satisfy the condition $\Delta f = 1/NT$ for the orthogonal relationship.

The PAPR of the original OFDM signal is defined as

$$PAPR = \frac{\max\limits_{0 \le t \le NT} |x_t|^2}{E[|x_t|^2]} \tag{2}$$

where $E[\cdot]$ denotes the expectation. For efficient approximation of x_t and its PAPR, only NL samples of x_t called L-times oversampled OFDM signals are considered, where L denotes the oversampling factor and is an integer larger than or equal to 1. L-times oversampled OFDM symbols $\mathbf{X} = [X_0, X_1, \cdots, X_{LN-1}]^T$ are obtained by padding consecutive $(L-1)N$ zero symbols $\underbrace{[0, 0, \cdots, 0]}_{(L-1)N}^T$ to middle or end of the original OFDM symbols. L-times oversampled OFDM symbols $\mathbf{X}$ are transformed to L-times oversampled OFDM signals $\mathbf{x} = [x_0, x_1, \cdots, x_{LN-1}]^T$ of which an element x_n is represented by

$$x_n = \frac{1}{\sqrt{LN}} \sum_{k=0}^{LN-1} X_k e^{\frac{j2\pi kn}{LN}}, \ 0 \le n \le LN-1. \tag{3}$$

It is well known that the OFDM signals $\mathbf{x}$ can be interpreted as the inverse discrete Fourier transform (IDFT) of the OFDM symbols $\mathbf{X}$. In OFDM systems, OFDM signals are generated by using the IFFT operation, which reduces computational complexity of the IDFT.

It is generally known that the PAPR of the original OFDM signal can be precisely estimated from not Nyquist-rate sampled OFDM signal with $L = 1$ but oversampled OFDM signal with $L = 4$. The PAPR of the L-times oversampled OFDM signals $\mathbf{x}$ is calculated as

$$PAPR(\mathbf{x}) = \frac{\max\limits_{n=0}^{LN-1} |x_n|^2}{E[|x_n|^2]}. \tag{4}$$

In general, OFDM signals have the high PAPR value and this high PAPR is one of the most significant problems in OFDM systems. The conventional PTS scheme can resolve the PAPR problem of OFDM signals as in the following subsection.

B. Conventional PTS Scheme

In the first stage of the conventional PTS scheme, input OFDM symbols $\mathbf{X}$ are partitioned into V disjoint sub-blocks $\mathbf{X}_v = [X_{v,0}, X_{v,1}, \cdots, X_{v,N-1}]^T, 0 \le v \le V-1$, satisfying the condition such that

$$\mathbf{X} = \sum_{v=0}^{V-1} \mathbf{X}_v. \tag{5}$$

By applying IFFT to each sub-blocks, the sub-signals $\mathbf{x}_v = [x_{v,0}, x_{v,1}, \cdots, x_{v,N-1}]^T, 0 \le v \le V-1$ are generated. In order to generate candidate OFDM signals, each sub-signals is differently multiplied by the phase rotating factor $b_v = e^{j\phi_v}$, where $\phi_v \in [0, 2\pi)$ for $v = 0, \cdots, V-1$. The phase rotating factor is usually an element of the finite set given as $b_v \in \{e^{j2\pi l/W} \mid l = 0, 1, \cdots, W-1\}$, where W is the number of allowed phase rotating factors. The phase rotating vectors to generate the candidate OFDM signals are represented by $\mathbf{b}^{(u)} = [b_0^{(u)}, b_1^{(u)}, \cdots, b_{V-1}^{(u)}]$, $u = 0, 1, \cdots, U-1$. By using the u-th phase rotating vector, the u-th candidate OFDM signal $\mathbf{x}^{(u)}$ is generated as

$$\begin{aligned} \mathbf{x}^{(u)} &= [x_0^{(u)}, x_1^{(u)}, \cdots, x_{N-1}^{(u)}]^T \\ &= \sum_{v=0}^{V-1} b_v^{(u)} \mathbf{x}_v, \qquad u = 0, 1, \cdots, U-1 \end{aligned} \tag{6}$$

where U is the number of candidate OFDM signals to be generated. Since all the first phase rotating factors $b_v^{(0)}, 0 \le v \le V-1$ are usually fixed to 1, $U = W^{V-1}$ candidate OFDM signals are generated in the conventional PTS scheme [8]. In the last stage of the conventional PTS scheme, among U candidate OFDM signals, the optimal OFDM signal $\mathbf{x}^{(u_{\text{opt}})}$ which has the minimum PAPR value is selected for the transmitted OFDM signal, where u_{opt} denotes the index of the optimal OFDM signal, that is,

$$u_{\text{opt}} = \arg \min_{u=0}^{U-1} PAPR(\mathbf{x}^{(u)}). \tag{7}$$

Although the conventional PTS scheme can achieve considerable PAPR reduction with simple but highly effective method, there are some disadvantages. High computational

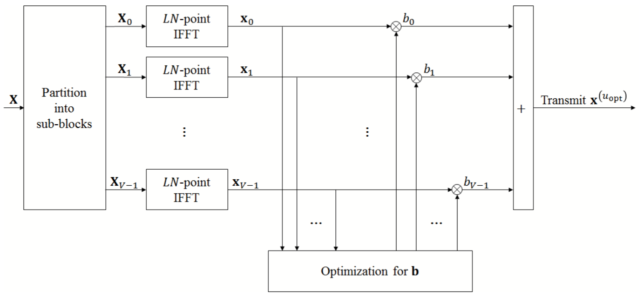


Fig. 1. A block diagram of the conventional PTS scheme.

complexity is the main disadvantage of the conventional PTS scheme. In the conventional PTS scheme, most of the computational complexity comes from V IFFTs and the generation and PAPR calculation of U candidate OFDM signals. In the following subsection, recently proposed low-complexity PTS schemes [17], [18] which mainly reduce the computational complexity required for calculation of candidate OFDM signals are introduced. These low-complexity PTS schemes have the same idea that only a few dominant samples of OFDM signals are used for the PAPR calculation instead of all the samples. They reduce the computational complexity considerably while maintaining the same PAPR reduction performance (optimum) compared with the conventional PTS scheme.

C. Low-Complexity PTS Schemes Using Dominant Time-Domain Samples

RC-PTS [17] was the firstly proposed low-complexity PTS scheme using dominant time-domain samples. This PTS scheme utilizes only a part of time-domain samples to estimate the peak power of each candidate OFDM signal, which are called dominant samples. In RC-PTS, dominant samples are selected by using a metric Q_n given as

$$Q_n = \sum_{v=0}^{V-1} |x_{v,n}|^2. \tag{8}$$

For an index n, if Q_n is greater than or equal to a pre-set threshold γ_Q, the time-domain sample $\mathbf{x}_n$ is selected for a dominant sample. The set of indices of the dominant samples is denoted by

$$\mathbf{S}_Q(\gamma_Q) = \{n \mid Q_n \geq \gamma_Q, \quad 0 \leq n \leq LN-1\}. \tag{9}$$

Among all time-domain samples, only the dominant samples with indices in $\mathbf{S}_Q(\gamma_Q)$ are multiplied by the corresponding phase rotating vectors and used to estimate the PAPR of each candidate OFDM signal. Then, in RC-PTS, the OFDM signals $\mathbf{x}^{(u_{\text{opt}})}$ selected for transmission is the candidate signal with index u_{opt} which is represented by

$$u_{\text{opt}} = \arg\min_{u=0}^{U-1} \max_{n \in \mathbf{S}_Q(\gamma_Q)} \left| \sum_{v=0}^{V-1} b_v^{(u)} x_{v,n} \right|^2. \tag{10}$$

Although the number of candidate OFDM signals of RC-PTS is the same as that of the conventional PTS scheme, the computational complexity for calculating of each candidate OFDM signals of RC-PTS is much less than that of the conventional PTS scheme. Therefore, it is clear that RC-PTS significantly reduces the computational complexity compared with the conventional PTS scheme.

The maximum power of the n-th sample among U candidate OFDM signals in the conventional PTS scheme is denoted by V_n, which is represented by

$$\begin{aligned} V_n &= \max_{u=0}^{U-1} \left| x_n^{(u)} \right|^2 \\ &= \max_{u=0}^{U-1} \left| \sum_{v=0}^{V-1} b_v^{(u)} x_{v,n} \right|^2, \qquad n = 0, 1, \cdots, LN-1 \end{aligned} \tag{11}$$

where $b_v^{(u)}$ and the metric V_n constitute the n-th phase rotating vector $\mathbf{b}^{(u)} = [b_0^{(u)}, b_1^{(u)}, \cdots, b_{M-1}^{(u)}]$ and $\mathbf{V} = [V_0, V_1, \cdots, V_{N-1}]$, respectively. In other words, V_n indicates the upper bound of the powers obtained from the n-th samples of the candidate OFDM signals. It implies that greater V_n means higher probability that peak power of the candidate OFDM signals occurs at the n-th sample. Therefore, V_n is the proper metric to select dominant samples which determine the PAPRs of the candidate OFDM signals.

After the RC-PTS had been proposed, the LC-PTS [18] utilizing the metric V_n was proposed to further reduce the computational complexity of the conventional PTS scheme and the RC-PTS. Instead of directly using V_n, the LC-PTS uses estimation of V_n with significantly less computational complexity. In the LC-PTS, estimation of V_n is obtained by doing only a few search over proper phase rotating vectors, instead of doing full search over all phase rotating vectors. In order to accurately estimate V_n of the n-th time-domain sample for all candidate OFDM signals, the n-th time-domain samples of the v-th sub-block except the first sub-block $x_{v,n}, 1 \leq v \leq V-1$ are rotated to a finite number of sub-planes of the complex plane on $x_{0,n}$ is located.

In the LC-PTS, the proposed metric P_n to estimate the maximum power V_n is calculated by

$$P_n = \max_{c=0}^{C-1} \left| x_n^{(K_n(c))} \right|^2 \tag{12}$$

where $K_n(c), 0 \leq c \leq C-1$ are indices of a small number of candidate signals selected by properly rotating time-domain sub-samples. The index set of dominant samples obtained by using P_n is denoted by $\mathbf{S}_P(\gamma_P)$, which is represented by

$$\mathbf{S}_P(\gamma_P) = \{n \mid P_n \geq \gamma_P, \quad 0 \leq n \leq LN-1\} \tag{13}$$

where γ_P is a pre-set threshold. Then, only the sub-samples with the indices in $\mathbf{S}_P(\gamma_P)$ are used for calculating the PAPR of each candidate OFDM signal. Since P_n is a good approximation of V_n, the LC-PTS can achieve a substantial reduction of computational complexity in calculating the PAPRs of candidate OFDM signals while accurately estimating the PAPRs. In the LC-PTS, the OFDM signal $\mathbf{x}^{(u_{\text{opt}})}$ selected for transmission is the candidate signal with index u_{opt} which is represented by

$$u_{\text{opt}} = \arg \min_{u=0}^{U-1} \max_{n \in \mathbf{S}_P(\gamma_P)} \left| \sum_{m=0}^{M-1} b_m^{(u)} x_{m,n} \right|^2. \tag{14}$$

III. Proposed PTS Scheme

A. Notations

For easy understanding of the proposed PTS scheme, notations for several sub-planes of 2-dimensional complex plane are defined. These sub-planes are utilized to select a candidate in the 2-dimensional plane. First of all, eight half-planes $\{P_i\}, 0 \le i \le 7$ are defined as in Fig. 2(a). Also, eight quadrants $\{Q_i\}, 0 \le i \le 7$ are defined as in Fig. 2(b). In addition, eight $45°$-sub-planes $\{R_i\}, 0 \le i \le 7$ are defined as in Fig. 2(c). Note that the variable i used in these sub-planes is integer to indicate indices of the sub-planes.

For a complex value x and a 2-dimensional complex sub-plane S, $x \in S$ denotes that x is located in S. For two 2-dimensional complex sub-planes S_1 and S_2, $S_1 \cup S_2$ denotes the combination sub-plane of S_1 and S_2. $i\%j$ denotes the modulo operation that finds the remainder after division of i by j.

B. Proposed Selection Method of Candidate Samples for Dominant Samples

In this subsection, an advanced low-complexity PTS scheme (AL-PTS) using dominant samples is proposed. AL-PTS uses an adaptive selection method of candidate time-domain samples to select dominant samples.

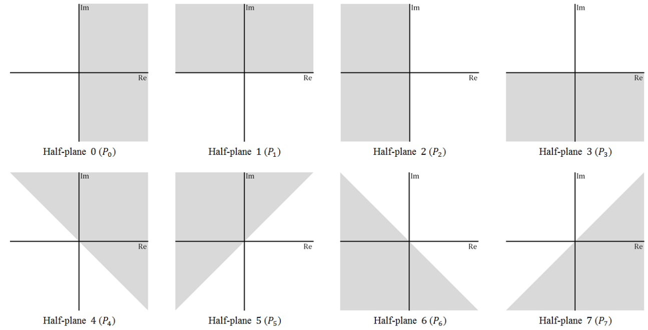
The detailed procedures of the proposed adaptive selection method of candidate time-domain samples to select the dominant samples are described as follows. Suppose that the first time-domain sub-sample $x_{0,n}$ of original OFDM signal sample x_n is located in R_i. For each n, select the index set of candidate samples, $K_n(c)$ by using one selection method adaptively among below three ones **S1**, **S2**, and **S3**. Note that size of $K_n(c)$, that is, the number of candidate samples for each n can be different each other, contrary to that in the LC-PTS.

S1) If all sub-samples $x_{v,n}, 1 \le v \le V-1$ except $x_{0,n}$, satisfy below condition (**C1**), select two candidate samples of which all the sub-samples are located in the $135°$-sub-planes $R_{(i\%8)} \cup Q_{((i+1)\%8)}$ and $R_{(i\%8)} \cup Q_{((i+6)\%8)}$, respectively.

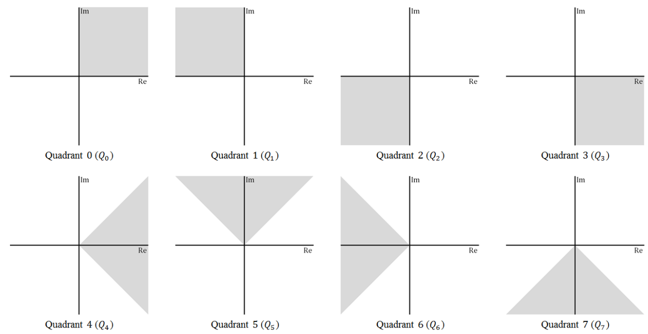
- **C1)** One or more $x_{v,n}$ are located in $R_{((i+2)\%8)} \cup R_{((i+6)\%8)}$ and the others are located in $R_{(i\%8)} \cup R_{((i+4)\%8)}$.

S2) If all $x_{v,n}$ except $x_{0,n}$ do not satisfy **C1** but one of the below 5 conditions (**C2-1–C2-5**), select only one candidate sample.

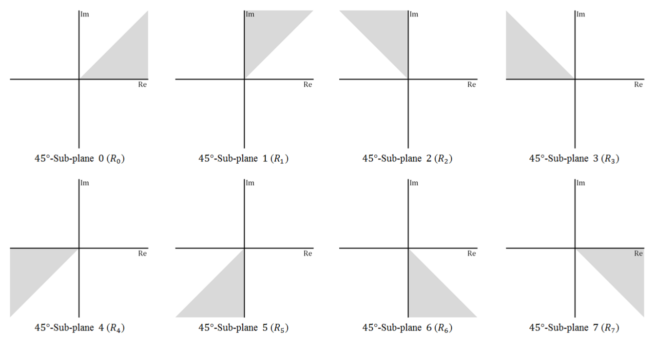
- **C2-1)** All $x_{v,n}$ except $x_{0,n}$ are located in $Q_{(i\%8)} \cup Q_{((i+4)\%8)}$.
  - **S2-1)** In case of **C2-1**, select the candidate sample of which all the sub-samples are located in the quadrant $Q_{(i\%8)}$.



(a) Eight half-planes of complex plane.



(b) Eight quadrants of complex plane.



(c) Eight $45°$-sub-planes of complex plane.

Fig. 2. Sub-planes of complex plane.

- **C2-2)** All $x_{v,n}$ except $x_{0,n}$ are located in $Q_{((i+3)\%8)} \cup Q_{((i+7)\%8)}$.
  - **S2-2)** In case of **C2-2**, select the candidate sample of which all the sub-samples are located in the quadrant $Q_{((i+7)\%8)}$.
- **C2-3)** At least two $x_{v,n}$ except $x_{0,n}$ are located in $R_{((i+1)\%8)} \cup R_{((i+5)\%8)}$ and $R_{((i+3)\%8)} \cup R_{((i+7)\%8)}$, respectively and the others are located in $(R_{((i+2)\%8)} \cup R_{((i+6)\%8)})^c$.
  - **S2-3)** In case of **C2-3**, select the candidate sample of which all the sub-samples are located in the $135°$-subplane $Q_{(i\%8)} \cup R_{((i+7)\%8)}$.
- **C2-4)** At least two $x_{v,n}$ except $x_{0,n}$ are located in $R_{((i+1)\%8)} \cup R_{((i+5)\%8)}$ and $R_{((i+2)\%8)} \cup R_{((i+6)\%8)}$, respectively and the others are located in $(R_{((i+3)\%8)} \cup R_{((i+7)\%8)})^c$.

S2-4) In case of **C2-4**, select the candidate sample of which all the sub-samples are located in the $135°$-subplane $Q_{((i+1)\%8)} \cup R_{(i\%8)}$.

C2-5) At least two $x_{v,n}$ except $x_{0,n}$ are located in $R_{((i+2)\%8)} \cup R_{((i+6)\%8)}$ and $R_{((i+3)\%8)} \cup R_{((i+7)\%8)}$, respectively and the others are located in $(R_{((i+1)\%8)} \cup R_{((i+5)\%8)})^c$.

S2-5) In case of **C2-5**, select the candidate sample of which all the sub-samples are located in the $135°$-subplane $Q_{((i+7)\%8)} \cup R_{((i+6)\%8)}$.

S3) Otherwise, select four candidate samples of which all the partial samples are located in the half-planes $P_{(i\%8)}$, $P_{((i+1)\%8)}$, $P_{((i+2)\%8)}$, and $P_{((i+7)\%8)}$, respectively.

Note that the above selection method is for $W = 2$ and that for $W = 4$ is similar with that for $W = 2$.

In the proposed PTS, the proposed metric T_n to estimate the maximum power V_n is computed as

$$T_n = \max_{c=0}^{C_n-1} \left| x_n^{(K_n(c))} \right|^2 \tag{15}$$

where the number of candidate sample C_n has a different value for each n. The index set of dominant samples obtained by using T_n is denoted by $\mathbf{S}_T(\gamma_T)$, which is represented by

$$\mathbf{S}_T(\gamma_T) = \{n \mid T_n \geq \gamma_T, \quad 0 \leq n \leq LN - 1\} \tag{16}$$

where γ_T is a pre-set threshold. Then, only the sub-samples with the indices in $\mathbf{S}_P(\gamma_P)$ are used to calculate the PAPR of each candidate OFDM signal. Finally, the OFDM signal $\mathbf{x}^{(u_{\text{opt}})}$ selected for transmission in AL-PTS is the candidate signal with index u_{opt} given by

$$u_{\text{opt}} = \arg\min_{u=0}^{U-1} \max_{n \in \mathbf{S}_T(\gamma_T)} \left| \sum_{m=0}^{M-1} b_m^{(u)} x_{m,n} \right|^2. \tag{17}$$

IV. Performance Analysis

A. Computational Complexity

This subsection compares the computational complexity of the conventional PTS scheme, RC-PTS [17], LC-PTS [18], and the proposed PTS scheme (AL-PTS). For comparison, the ratio p_γ is defined as the ratio between the number of selected dominant samples and the number of all time-domain samples, that is

$$p_\gamma = \frac{N_\gamma}{LN}. \tag{18}$$

Table I compares the computational complexity after IFFT computations of the conventional PTS, RC-PTS, LC-PTS, and AL-PTS by using parameters N, L, V, and W. Note that the parameter p_γ or N_γ is used for representing the computational complexity of PTS, RC-PTS, LC-PTS, and AL-PTS. Also, the variables C and $C_n, 0 \leq n \leq LN - 1$ are used for representing the computational complexity of LC-PTS and AL-PTS, respectively.

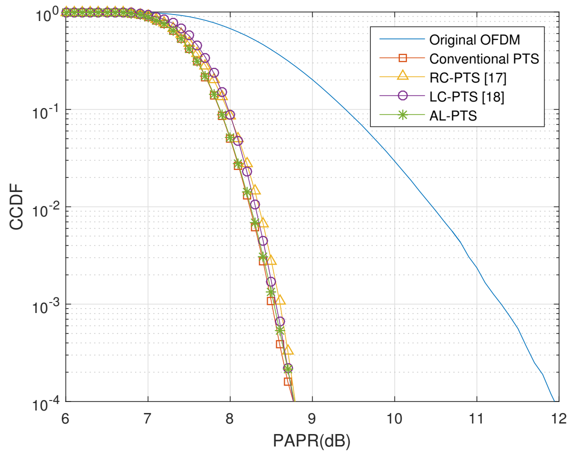
TABLE I
COMPUTATIONAL COMPLEXITY AFTER IFFT COMPUTATION IN PTS SCHEMES

PTS scheme	**Number of complex multiplications**
Conventional PTS	LNV
RC-PTS	$LNV + p_\gamma LNU = LNV + N_\gamma U$
LC-PTS	$LNC + p_\gamma LNU = LNC + N_\gamma U$
AL-PTS	$\sum_{n=0}^{LN-1} C_n + p_\gamma LNU = \sum_{n=0}^{LN-1} C_n + N_\gamma U$

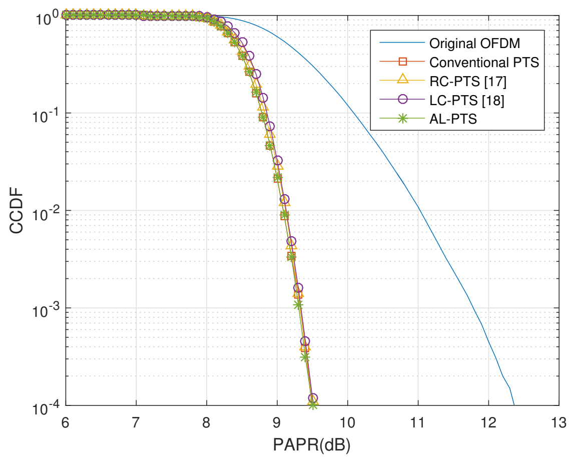
B. Simulation Results

This subsection compares the PAPR reduction performance of the conventional PTS, RC-PTS, LC-PTS, and AL-PTS. Fig. 3 compares the PAPR reduction performance of these PTS schemes for $L = 4$, $M = 8$, and $W = 2$, where $N = 256$ and $N = 1024$ in Fig. 3(a) and Fig. 3(b), respectively. For fair comparison, N_γ and the corresponding p_γ are differently set in RC-PTS, LC-PTS, and AL-PTS, such that they achieve the optimal PAPR reduction performance, which is the same as that of the conventional PTS with their respective minimum N_γ and p_γ. Note that the respective minimum N_γ and p_γ of RC-PTS, LC-PTS, and AL-PTS for the optimal PAPR reduction performance is obtained by exhaustive search. Therefore, RC-PTS, LC-PTS, and AL-PTS all achieve the optimal PAPR reduction performance but their required N_γ and p_γ are all different. In Fig 3(a), N_γs of RC-PTS, LC-PTS, and AL-PTS are 56, 15, and 51, respectively and therefore the corresponding p_γs of them are 0.055, 0.015, and 0.05, respectively. Also, in Fig 3(b), N_γs of RC-PTS, LC-PTS, and AL-PTS are 102, 16, and 164, respectively and therefore the corresponding p_γs of them are 0.025, 0.004, and 0.04, respectively.

Table II compares the computational complexity of the conventional PTS, RC-PTS, PS-PTS, and LA-PTS for $L = 4$, $M = 8$, and $W = 2$, where $N = 256$ and $N = 1024$ in sub-figure Fig. II(a) and Fig. II(b), respectively. As in Fig. 3, in Table II, N_γ and the corresponding p_γ are differently set in RC-PTS, LC-PTS, and AL-PTS, such that they achieve the optimal PAPR reduction performance, which is the same as that of the conventional PTS with their minimum N_γ and p_γ. In Table II(a), compared with the conventional PTS, AL-PTS shows the lowest computational complexity (22.2%) with achieving the same PAPR $8.7\,dB$ at CCDF $= 10^{-4}$, which is optimum as shown in Fig. 3(a). On the other hand, RC-PTS and LC-PTS show the relative computational complexity (55.5%) and (51.5%) with the optimal PAPR, respectively. Also, in Table II(b), compared with the conventional PTS, AL-PTS shows the lowest relative computational complexity (21.3%) with achieving the same PAPR $9.5\,dB$ at CCDF $= 10^{-4}$, which is optimum as shown in Fig. 3(b). On the other hand, RC-PTS and LC-PTS show the relative computational complexity (52.5%) and (50.4%) with the optimal PAPR, respectively.



(a) $N = 256$



(b) $N = 1024$

Fig. 3. PAPR reduction performance of PTS schemes for 16-QAM, $L = 4$, $V = 4$, and $W = 2$.

TABLE II
COMPUTATIONAL COMPLEXITY AFTER IFFT COMPUTATION IN PTS SCHEMES FOR 16-QAM, $L = 4$, $V = 4$, AND $W = 2$

(a) $N = 256$

PTS scheme	**Number of complex multiplications**
Conventional PTS	8192 (100%)
RC-PTS ($N_\gamma = 56$, $p_\gamma = 0.055$)	4216 (55.5%)
LC-PTS ($N_\gamma = 15$, $p_\gamma = 0.015$)	4216 (51.5%)
AL-PTS ($N_\gamma = 51$, $p_\gamma = 0.05$)	1821 (22.2%)

(b) $N = 1024$

PTS scheme	**Number of complex multiplications**
Conventional PTS	32768 (100%)
RC-PTS ($N_\gamma = 102$, $p_\gamma = 0.025$)	17200 (52.5%)
LC-PTS ($N_\gamma = 16$, $p_\gamma = 0.004$)	16512 (50.4%)
AL-PTS ($N_\gamma = 164$, $p_\gamma = 0.04$)	6964 (21.3%)

V. CONCLUSION

In this paper, in order to lower the computational complexity of the PTS schemes, the advanced low-complexity PTS schemes is proposed. Although the proposed PTS scheme uses dominant samples similar to some previous low-complexity PTS schemes, it utilizes more efficient selection method of the dominant samples. The proposed PTS scheme reduces the computational complexity much more compared to the previous PTS schemes while achieving the optimal PAPR reduction performance.

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