Abstract—Partial transmit sequence (PTS) is one of the most well-known peak-to-average power ratio (PAPR) reduction techniques proposed for orthogonal frequency-division multiplexing (OFDM) systems. The main drawbacks of the conventional PTS (C-PTS) are high computational complexity and transmission of several side information bits. A new PTS with simple detector is proposed in this paper to deal with these drawbacks of C-PTS. The candidates can be generated through cyclically shifting each sub-block sequence in time domain and combining them in a recursive manner. At the receiver, by utilizing the natural diversity of phase constellation for different candidates, the detector can successfully recover the original signal without side information. Numerical simulation shows that the proposed scheme performs very well in terms of PAPR. The probability of detection failure of the side information demonstrates that the detector could work without any side information with high reliability. The proposed scheme achieves almost the same bit error rate (BER) performance as the C-PTS with perfect side information, under both additive white Gaussian noise (AWGN) channel and Rayleigh fading channel.

Index Terms—Computational complexity, orthogonal frequency division multiplexing (OFDM), partial transmit sequences (PTS), peak-to-average power ratio (PAPR), side information (SI).

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is an attractive technique for wireless high-rate data transmission due to the minimizing effect over frequency-selective fading channels [1]. As such, OFDM has been chosen for European Digital Audio Broadcasting (DAB), Digital Video Broadcasting (DVB), WLAN standards (802.11), WiMax (802.16) and is being considered for the long term evolution of 3GPP. However, OFDM has some drawbacks in the transmission system. One of the major problems of the OFDM system is that OFDM signal has higher peak to-average-power ratio (PAPR) than single carrier signal because OFDM signal is the sum of many narrowband signals in the time domain [2]. The high PAPR can cause inter-modulation and out-of-band radiation due to power amplifier nonlinearity. In order to combat the problem, the transmission amplifier must operate within its linear region to prevent spectral distortion and the degradation of the bit error rate (BER). High linearity normally implies low efficiency and large power dissipation, which is prohibitive for use in portable wireless applications [3]. Therefore, it is highly desirable to reduce the PAPR of an OFDM signal.

OFDM PAPR reduction has been a subject of intense research in the past decade. Many methods have been proposed including clipping of the OFDM signal [4], coding techniques [5], active constellation extension (ACE) [6], companding transform [7], [8], tone reservation (TR) [9], [10], tone injection (TI) [11], partial transmit sequence (PTS) [12]–[16], selective mapping (SLM) [16]–[20] and various combinations of the above. Among them, SLM and PTS are two promising techniques because they are simple to implement, no distortion in the transmitted signal and can significantly improve the statistics of the PAPR. However, the conventional SLM and PTS suffer from higher computational complexity due to several N-dimensional inverse fast Fourier transform (IFFT) operations, where N is the number of subcarriers. At the same time in order to recover the original OFDM signal successfully, the transmitter has to send the selected signal index, called side information, to the receiver using extra subcarrier. It will degrade the OFDM system’s spectrum efficiency. Worst still, the BER performance of the OFDM systems can possibly be degraded significantly since any error in the detection of side information can damage the entire data block. Several proposals that suggested techniques for eliminating the need of side information have appeared in the literature [21]–[25]. However, these proposed techniques rely on some reference symbols that increase the transmission power or with the potential of higher complexity for the detector.

We are motivated to deal with the two drawbacks of the conventional PTS together—complexity and side information. In this paper, interleaved PTS making use of the combination of cyclically shifting sub-block sequences is proposed to generate new candidates. Our scheme can reduce the computational complexity without any side information. The advantages of the proposed scheme are described as follows. Firstly, with the interleaved partition method, reduction of computational complexity can be achieved by using Cooley-Tukey FFT algorithm [26]. Secondly by utilizing cyclically shifted sub-block sequences and IFFT property, the proposed scheme will generate a set of candidates with different phase constellation without multiplication according to the different shifting number of the sub-block. Furthermore, we can utilize the reduced complexity recursive method in our previous work to further decrease the computational complexity [14]. The detector at the receiver can distinguish which candidate has been transmitted through the minimum Hamming distances of the receiver signals (after FFT)
to the phase constellation for each sub-block without side information. Because the cyclically shifting operates on each sub-block individually, hence the overall detected number only increases linearly with the number of sub-block rather than exponentially as in C-PTS.

The rest of this paper is organized as follows. In Section II, the PAPR problem of a typical OFDM system is formulated and then the principle of the conventional PTS (C-PTS) scheme is explained. The main idea of our proposed cyclically shifting interleaved PTS approach, the detector without side information as well as the candidate analysis are presented in Section III. In Section IV, the PAPR, detection failure and BER results are presented and discussed. Finally, conclusions are drawn in Section V.

II. OFDM SYSTEM WITH PTS TO REDUCE PAPR

In OFDM system, a block of \( N \) symbols, \( \mathbf{X} = [X(0), \ldots, X(N - 1)]^T \), is formed with each symbol modulating by one set of subcarrier. Then, an OFDM signal is obtained by summing up all the \( N \) modulated independent subcarriers, where \( N \) is the number of subcarriers. The \( N \) subcarriers are chosen to be orthogonal such that the adjacent subcarrier separation \( \Delta f = 1/T \) where \( T \) is the OFDM signal duration. The mathematical representation of the OFDM signal can be written as

\[
x(t) = \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} X(l)e^{j2\pi l\Delta f t}, \quad 0 \leq t \leq T \tag{1}
\]

The subcarrier vector \( \mathbf{X} \) is formed according to a certain modulation scheme such as phase shift keying (PSK) or quadrature amplitude modulation (QAM). Thus, \( \mathbf{X} \) is a vector of \( N \) constellation symbols from a constellation \( \mathcal{C} \).

The PAPR of OFDM signal in one symbol period in (1) is defined as the ratio between the maximum instantaneous power and its average power, which can be written as

\[
PAPR = 10\log_{10}\frac{\max_{0 \leq t \leq T}|x(t)|^2}{P_{av}} \tag{2}
\]

where \( P_{av} \) is the average power of \( x(t) \) and it can be computed in the frequency domain because IFFT is a (scaled) unitary transformation.

The transmitted discrete time signal \( x[n] \) is usually generated by sampling the continuous time signal \( x(t) \). If sampling at critical sampling or Nyquist rate, \( x[n] \) might have an overly optimistic PAPR value because it might lose some information at the peak of \( x(t) \). Hence, \( x[n] \) is usually oversampled by a factor \( L \) to have a better estimation of the PAPR value of continuous time signal \( x(t) \). The oversampling by the factor \( L \) can be realized by inserting \( (L-1)\Delta f \) zeros in the middle of the \( N \)-point frequency-domain signal \( \mathbf{X} \) and passing the new \( LN \)-point data sequence through the \( LN \)-point IFFT unit. Therefore, the oversampled IFFT output can be expressed as

\[
x[n] = \frac{1}{\sqrt{LN}} \sum_{l=0}^{LN-1} \tilde{X}(l)e^{j2\pi ln/N}, \quad 0 \leq n \leq LN - 1 \tag{3}
\]

where \( \tilde{X} = [X(0), \ldots, X(N/2 - 1), 0, \ldots, 0, X(N/2), \ldots, X(N - 1)]^T \). It is shown \( L = 4 \) is sufficient to capture the peak information of \( x(t) \) [27].

In a typical OFDM system with PTS approach to reduce the PAPR, the input data block in \( \mathbf{X} \) is divided by means of a certain partitioning scheme into \( M \) disjoint sub-blocks, which are represented by the vectors \( \{\mathbf{X}_m, m = 1, 2, \ldots, M\} \). Therefore, we can get

\[
\mathbf{X} = \sum_{m=1}^{M} \mathbf{X}_m \tag{4}
\]

where all the subcarrier positions which are presented in another block must be zero so that the sum of all the sub-blocks constitutes the original signal. Then, the sub-blocks \( \mathbf{X}_m \) are transformed into \( M \) time-domain partial transmit sequence \( \mathbf{x}_m \), which can be represented as

\[
\mathbf{x}_m = \text{IFFT}_{LN \times N}\{\mathbf{X}_m\} \tag{5}
\]

After that, these partial sequences are independently rotated by the phase factors \( \mathbf{P} = [P_1, \ldots, P_M] \) and combined together to create a set of candidates

\[
\mathbf{R} = \sum_{m=1}^{M} P_m \mathbf{x}_m = \sum_{m=1}^{M} \text{IFFT}\{P_m \odot \mathbf{x}_m\} \tag{6}
\]

Suppose that alphabet \( \mathcal{R} \) denotes the set of the value of \( P_m \), \( m = 1, 2, \ldots, M \), and \( K \) is the number in set \( \mathcal{R} \). Usually, the choice of \( \mathcal{R} \) from \( \{\pm 1\} (K = 2) \) or \( \{\pm 1, \pm j\} (K = 4) \) is interesting since no actual multiplication is performed to rotate the phase. Therefore, total \( O = K^{(M-1)} \) candidates could be generated in C-PTS scheme.

Finally, the candidate with the lowest PAPR is chosen by exhaustive search of the candidates for transmission. The principle structure of the C-PTS scheme is shown in Fig. 1.

In the C-PTS scheme, \( \lfloor\log_2(O)\rfloor \) bits of side information to indicate the optimized phase sequence \( \mathbf{P}^* \) (which is the phase sequence with the lowest PAPR) have to be communicated explicitly to the receiver in order to recover the original symbol \( \mathbf{X} \).
III. PROPOSED PTS SCHEME AND DETECTOR

A. Proposed PTS Scheme

The aim of the proposed PTS scheme is to lower the complexity of the transmitter and detector while the receiver can recover the original signal without side information.

It is shown in ref [12] that, though with the worst PAPR performance when compared with the pseudo-random and adjacent partition method, the interleaved partition method can reduce the computational complexity by using Cooley-Tukey FFT algorithm. Ref [15] proves that the candidates in the conventional interleaved partitioning PTS scheme are not fully independent, which leads to performance inferior to the adjacent partitioning PTS scheme. After enhancing the independence among the candidates, the PAPR performance of the amended interleaved partitioning PTS scheme can be improved to achieve similar performance as the adjacent partitioning PTS scheme. The result in [15] inspires us using the interleaved partition method in PTS scheme rather than pseudo-random and adjacent partition method considering the tradeoff of complexity and PAPR performance.

After obtaining the time domain signal \( x_m(m = 1, \ldots, M) \) in (5), instead of applying different phase vectors \( P \) on \( x_m \) as the C-PTS scheme, the proposed scheme generates new candidates by cyclically shifting some of signals \( x_m \) and combining them together.

\[
\hat{x} = \sum_{m=1}^{M} \hat{x}_m(k)
\]

where the shifting number is \( k \) for sub-block signals \( x_m \) and \( \hat{x}_m(k) \) is

\[
\hat{x}_m(k) = \text{circular}(x_m, k) = [x_m(k), \ldots, x_m(N-1), x_m(0), \ldots, x_m(k-1)]
\]

We apply the mechanism of our previous study in [14] to calculate the new candidates quickly. For example given signal \( x_1 = \sum_{m=1}^{M} x_m \), we can compute the relative signal \( x_2 \) from \( x_1 \) quickly by equ. (9) instead of using equ. (7) when they have only different in sub-block \( m \).

\[
x_2 = x_1 - (x_m - \hat{x}_m(k))
\]

Finally, change sub-blocks signals from 1 to \( M \), we can generate all candidates by use of equ. (9) recursively. The candidate with the lowest PAPR is chosen for transmission.

The following are the advantages of the shifting technique in the proposed PTS scheme. First, no multiplication is required. Second, by utilizing the property of IFFT for different shift number \( k \), cyclically shifted signal will have distinct phase constellation in some of frequency domain signals. Thus, in the receiver, the detector can determine which shift number \( k \) operated on the sub-block \( m \) according to the phase constellation of the received signal (after FFT). As a consequence, the receiver will not require the side information. We will explain it in more details below.

B. Phase Diversity Analysis

In order to avoiding the side information, the detector must be able to estimate which candidate had been transmitted without the help of additional information. It also means that every candidate in our proposed PTS scheme will have some difference from the others.

We now explain why our proposed PTS scheme can generate distinct candidates. Let take the candidates \( x_2 \) and \( x_1 \) in equ. (9) as an example. It is obviously that the difference of \( x_2 \) and \( x_1 \) is only the sub-block \( m \), where \( x_m \) in \( x_1 \) is substituted by \( \hat{x}_m(k) \) in \( x_2 \).

Consider the relationship of \( x_m \) and \( \hat{x}_m(k) \) in equ. (8) and using the linear property of Fourier transform, \( \hat{x}_m(k) \) can be expressed as

\[
\hat{x}_m(k) = \text{IFFT}(P_m(k) \odot X_m)
\]

where

\[
P_m(k) = \begin{bmatrix} e^{j2\pi k w_0/LN} & \cdots & e^{j2\pi k w_{LN-1}/LN} \\ \vdots & \ddots & \vdots \\ e^{j2\pi k w_{(LN-2)/LN}} & \cdots & \cdots & e^{j2\pi k w_{(LN-1)/LN}} \end{bmatrix}, \quad 0 \leq n \leq LN-1
\]

\( P_m(k) \) is called the rotation phase vector for the \( m \)th sub-block of candidate \( x_2 \). For \( m \)th sub-block, we have interest in those positions that \( P_m(k) \) reacting on the subcarriers of \( X_m \) are nonzero only, such as the subcarrier \( s_m = m + LN \cdot I \), \( I = 0, \ldots, N/M \), because of the interleaved partition method. In theory, if there is any difference between \( P_m(k) \) and \( P_m(0) \) in subset \( S_m \), where \( P_m(0) \) can be seen as the phase vector action on \( x_m \), the detector can verify which candidate had been transmitted between \( x_2 \) and \( x_1 \). In fact under the influence of the noise, the detector will become more reliable with the increase of Hamming distance between \( P_m(k) \) and \( P_m(0) \).

According to the equ. (10), new candidates in equ. (7) can be rewritten as

\[
\hat{x} = \sum_{m=1}^{M} \text{IFFT}(P_m(k) \odot X_m)
\]

which has almost the same expression as the C-PTS scheme in equ. (6) except the different phase set in \( \hat{x}_m(k) \) and \( P_m(0) \), respectively. For the sub-block \( m \), the set of shift number \( k \) is denoted as \( \vartheta(m, k) \), the length of \( \vartheta(m, k) \) is \( K \) and the relative phase set is denoted as \( \vartheta(\hat{x}_m(k)) \).

C. Detector Without Side Information

The objective of the detector is to determine which phase vector \( P_m^* \) in the phase set \( \vartheta(\hat{x}_m(k)) \) has been selected for the \( m \)th (1 \( \leq m \leq M \)) sub-block.

Assume \( x = [x(0), \ldots, x(LN-1)] \) is the received signal after the FFT demodulation at the receiver. Referring to the interleaved partitioning PTS, we first divide \( x \) into \( M \) sub-block in the same manner as the original signal \( X \). For example if \( X(I) \) belongs to sub-block \( X_m \), then put \( I \) into subset \( R_m \).

It is well-known that the process of rotation phase vector \( P_m^* \) on the sub-block signal \( X_m \) in the PTS scheme will not impact the sub-block signals \( X_I \) when \( I \neq m \). It means that the estimation of the action phase vector \( P_m^* \) for each sub-block \( m \) can be
is the subcarriers set of signal with the same number of sub-blocks, respectively. The sub-block signals to obtain numbers of candidates which only increases linearly with $K$ and $M$ rather than exponentially in the transmitter.

E. Analysis of Computational Complexity for the C-PTS and the Proposed PTS

It is well known that there are three partition methods for C-PTS scheme [12]: interleaved, adjacent and pseudo-random. Among them, pseudo-random partitioned C-PTS (denoted as PRC-PTS) scheme can obtain the best PAPR performance but the computational complexity is relative higher. By using the Cooley-Tukey FFT algorithm, interleaved partitioned C-PTS (denoted as IC-PTS) scheme has the lowest computational complexity but it has the worst PAPR performance because the generated candidates are not fully independent [15].

In this section, the overall computational complexity for the PRC-PTS, IC-PTS and the proposed PTS scheme are analyzed. For the sake of fairness of comparison, the result is based on the same number of candidates $K^{(M-1)}$ for different PTS schemes.

\textbf{Computational Complexity for the Transmitter:} It is known that a $LN$-point IFFT requires $LN/2 \log_2(LN)$ numbers of complex multiplication ($\text{tim}_{\text{im}}$) and $LN \log_2(LN)$ numbers of complex addition ($\text{tim}_{\text{add}}$).

The PRC-PTS scheme requires $M$ IFFT operations with $M$ sub-blocks. Therefore, the $\text{tim}_{\text{im}}$ and $\text{tim}_{\text{add}}$ for the PRC-PTS scheme are $MLN/2 \log_2(LN)$ and $MLN \log_2(LN)$, respectively. In addition, $K^{(M-1)} (M-1) LN$ complex additions are required for combining the $M$ sub-block signals to obtain the $K^{(M-1)}$ candidates and searching for the minimum PAPR out of them. Thus, the total $\text{tim}_{\text{im}}$ and $\text{tim}_{\text{add}}$ for the transmitter part of the PRC-PTS scheme are $MLN/2 \log_2(LN/M)$ and $MLN \log_2(LN/M) + K^{(M-1)} (M-1) LN$, respectively.

Compared with the PRC-PTS scheme, the IC-PTS scheme utilizes Cooley-Tukey FFT algorithm, which can reduce the $\text{tim}_{\text{im}}$ and $\text{tim}_{\text{add}}$ of the $M$ IFFT operations to $LN/2 \log_2(LN/M)$ and $LN \log_2(LN/M)$, respectively. Thus, the total $\text{tim}_{\text{im}}$ and $\text{tim}_{\text{add}}$ for the transmitter part of the IC-PTS scheme are $LN/2 \log_2(LN/M)$ and $LN \log_2(LN/M) + K^{(M-1)} (M-1) LN$, respectively.

As for the proposed PTS scheme, the $\text{tim}_{\text{im}}$ and $\text{tim}_{\text{add}}$ are $LN/2 \log_2(LN/M)$ and $LN \log_2(LN/M)$, respectively, which are the same as PRC-PTS scheme. Additional $K^{(M-1)} 2LN$ complex additions are required by the proposed PTS scheme to combine the $M$ sub-block signals to obtain the $K^{(M-1)}$ candidates and search for the minimum PAPR out of them. Thus, the total $\text{tim}_{\text{im}}$ and $\text{tim}_{\text{add}}$ for the transmitter part of the proposed PTS scheme are $LN/2 \log_2(LN/M)$ and $LN \log_2(LN/M) + K^{(M-1)} 2LN$, respectively.

\textbf{Computational Complexity for the Receiver:} The computational complexity of the receiver for the proposed PTS and C-PTS include two parts—1) a $LN$-point FFT operation and 2) the detector. The FFT operation for the PRC-PTS, IC-PTS and the proposed PTS scheme requires $LN/2 \log_2(LN)$ $\text{tim}_{\text{im}}$ and $LN \log_2(LN)$ $\text{tim}_{\text{add}}$, respectively. The detector part requires $(Q + 2)N$ $\text{tim}_{\text{im}}$ and $2QN$ $\text{tim}_{\text{add}}$, respectively, for both the PRC-PTS and IC-PTS scheme where $Q$ is the size of constellation for PSK or QAM modulation. Whereas for the proposed

As for the detector of receiver, we can distributive estimate the shift number $k$ for every sub-block $m$ and only require calculating $K$ candidates. Thus, the overall candidates to be considered are $KM$, which only increases linearly with $K$ and $M$ rather than exponentially in the transmitter.

D. Candidate Analysis

In the transmitter, it is well-known that the total numbers of candidates of the C-PTS scheme are $K^{(M-1)}$, which exponentially increases with the number of sub-block $M$ and phase set $K$. Whereas, the proposed scheme can generate more numbers of candidates ($K^{(M)}$) with the same number of sub-block phase set as the C-PTS scheme if the shift numbers $k$ for all sub-blocks are not equal. For example when the shift set for each sub-block is 2, one shifting number $k$ is 0 (i.e. no shifting) and another shifting number $k$ is $k_m$ for sub-block $m$. Accordingly, the proposed scheme can generate $2^M$ independently candidates rather than $2^{(M-1)}$ independently candidates in the C-PTS scheme, where independently candidates means that each of candidates have the different PAPR.
and are given. It is shown in Table I that this observation can be explained for the PRC-PTS scheme, the computational complexity reduction ratio (CCRR) of the proposed PTS scheme over the PRC-PTS and IC-PTS schemes is defined as

\[
\text{CCRR} = \left( \frac{1 - \frac{\text{Complexity of the proposed PTS scheme}}{\text{Complexity of the PRC/IC-PTS scheme}}} \right) \times 100\%
\]

The CCRRs of the proposed PTS scheme over the PRC-PTS and IC-PTS schemes for typical values of \( N, M \) and \( Q \) are given in Table I, where \( K = 2 \) and \( L = 4 \). It is shown in Table I that the CCRRs of the proposed PTS scheme over the PRC-PTS and IC-PTS schemes is reduced rapidly with the increase of \( N \) and \( M \) especially for large \( M \). This observation can be explained that the complexity on the search of minimum PAPR value in the transmitter increases significantly with the increase of \( M \).

The proposed PTS scheme for reduction of the complexity is more efficient with the increase of \( M \).

### IV. Simulation Results

To illustrate the effectiveness of the proposed scheme, we consider several simulation results to evaluate the performance in terms of PAPR reduction, estimation error probability of the detector as well as bit error rate (BER) at the detector output. The results of the simulation are based on the transmission of \( 3 \times 10^6 \) randomly generated OFDM symbols with the carriers \( N = 256 \) under the condition of an oversampling factor \( L = 4 \).
Both QPSK and 16-QAM modulation techniques are examined here. The complementary cumulative density function (CCDF) of the PAPR is used to measure the performance. The CCDF of the PAPR is defined as

$$\text{CCDF}(\text{PAPR}(x[n])) = Pr(\text{PAPR}(x[n]) > \text{PAPR}_0)$$

where \(\text{PAPR}_0\) is a certain threshold value that is usually given in decibels relative to the root mean square (RMS) value.

The CCDFs of the proposed interleaved PTS scheme with sub-block \(M = 4, 8\) and shifting set \(K = 2\) or \(K = 4\) for each sub-block are shown in Fig. 3 for both QPSK and 16-QAM modulation. We also plot the CCDFs of the original OFDM system without PAPR reduction, the PRC-PTS and the IC-PTS scheme for each value of \(M\) for comparison purpose.

We observe that the performance of our proposed PTS scheme in terms of PAPR reduction is better than that of the PRC-PTS and IC-PTS scheme for all configurations. This is because our proposed PTS scheme has the ability to generate more candidates with the same parameters. For example, when \(K = 2\), the proposed PTS scheme will generate 16 and 256 candidates, respectively, when \(M = 4\) and 8. For the PRC-PTS and IC-PTS scheme, only 8 and 128 candidates can be obtained when \(M = 4\) and 8. The performance improvement of the proposed PTS scheme relative to that of the IC-PTS scheme are found to be 0.75 dB and 0.4 dB for \(M = 4\) and \(M = 8\) with \(K = 2\) when CCDF = 0.1%. When compared with PRC-PTS scheme, the PAPR performance of the proposed PTS scheme also improves about 0.2 dB and 0.15 dB with the same parameters.

For fairness of comparison of the PAPR performance, Fig. 3 also plots the curves of the proposed PTS scheme with the candidates \(K^{M-1}\), which is the same as the PRC-PTS and IC-PTS scheme. The results shown that the proposed PTS scheme can obtain similar PAPR performance as the PRC-PTS scheme, which is better than the IC-PTS scheme.

Fig. 4 illustrates the probability of detection failure \((P_{df})\) against the signal noise ratio (SNR) of the proposed detector over additive white Gaussian noise (AWGN) and Rayleigh fading channel. The SNR is defined as the ratio between the average energy per transmitted bit \(E_b\) and the one-sided power spectral density \(N_0\) of white Gaussian noise. The parameter \(P_{df}\) represents the probability that the receiver cannot recover the embedding side information. It is worthwhile to note that, if the detector fails to recover the embedding side information of sub-block \(m\), only the original signal \(X_m\) in sub-block \(m\) is lost in the PTS scheme rather than the complete OFDM frame in the SLM scheme.

From Fig. 4, we can observe that the value of \(P_{df}\) depends on the number of sub-block \(M\) and the modulation scheme employed. As \(M\) is increased, the performance of our proposed PTS scheme decreases simply because lower \(M\) allows for better distinction between shifted and non-shifted symbols after transmission through the channel, which makes the occurrence of an erroneous detection event less likely. The results in Fig. 4 also indicate that the performance of detector with QPSK is better than with 16-QAM. This can be explained by the fact that the distinction of constellation symbols in 16-QAM is lower than the QPSK for the same SNR.

We can also see that the \(P_{df}\) rapidly reduces with the increase of SNR, which means the selected shifting number for each sub-block is recovered successfully. Thus, the detector of the proposed scheme is reliable.

Given the excellent decoder performance over AWGN and Rayleigh fading channels, we next investigate the BER performance degradation of the proposed scheme here. Such degradation is due to the occasional failure detection of embedding side information. Fig. 5 demonstrates the reliability of the detector for the proposed PTS scheme. Thus, we can expect that the good BER performance can be maintained in the proposed PTS scheme when compared with the PRC-PTS using perfect side information. The BER performance of the proposed PTS scheme over AWGN channel or Rayleigh fading channel for sub-block \(M = 4\), 8 with QPSK and 16-QAM modulation are presented in Fig. 5. For comparison purpose, we have also plotted the BER curves of the PRC-PTS scheme with perfect (i.e. error-free) or without perfect side information. It is clearly shown that the BER of the proposed PTS scheme is lower than
the C-PTS scheme without perfect side information, which almost overlaps together with the C-PTS scheme with perfect side information except around the region of low SNR.

V. CONCLUSION

An interleaved partitioning PTS scheme making use of the recursive combination of cyclically shifting sub-block sequences and linear property of IFFT is presented in this paper. The proposed scheme will lower the complexity without side information. From the aspect of complexity, firstly, the Cooley-Tukey FFT can be implemented because of the interleaved partition method. Secondly, no multiplication is performed by the cyclically shifting. Thirdly, the reduced complexity recursive method is used to generate new candidates. Finally, the overall detected number at the detector increases linearly with the number of sub-blocks rather than exponentially with the increase of candidates. By utilizing cyclically shifting of the sub-block sequence, a set of candidates with different phase constellation will be generated according to the different shifting number of the sub-block. As such, the detector can distinguish which candidates had been transmitted without any side information. Cyclically shifted sub-block sequences also increase the independence and the total number of candidates compared with the conventional PTS scheme. Simulation results have shown that the proposed scheme is reliable to estimate the selected shifting number of each sub-block. As a result, it achieves almost the same BER performance as the conventional PTS scheme but without any side information both in AWGN and rayleigh fading channels. It may be worthwhile to further study the performance of the proposed PTS in frequency selective channel without perfect channel information in the future.

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Fig. 5. BER performance of the proposed PTS scheme in AWGN and Rayleigh fading channel.