# Sub-Optimal Search Algorithms for PTS Phase Selection


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***Abstract*— PTS method is a well-known method which can reduce the PAPR in OFDM systems. The conventional PTS techniques can provide good PAPR reduction performance; however, the search complexity of the original PTS method increases exponentially with the number of the sub-blocks. In this paper, we proposed two algorithms to drastically reduce search complexity while slightly degrade the PAPR reduction performance.**



*Keywords*— **OFDM, PAPR, PTS, Low Complexity, Phase Selection.**


## 1. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is an attractive technique for achieving high-bit-rate wireless data communication [1-3]. It has attracted a lot of attentions especially in the field of wireless communications, and it been adopted as the standard transmission technique in the wireless LAN systems and the terrestrial digital broadcasting system. One of the major drawback of the OFDM system is the high peak-to-average power ratio (PAPR), which may cause high out-of-band radiation when the OFDM signal passed through a radio frequency power amplifier. Consequently, high PAPR is one of the most important implementation challenges for OFDM system designers.

In order to reduce the PAPR, several techniques have been proposed [4-7]. Among these methods, the partial transmit sequence (PTS) [5] is the most attractive scheme because of good PAPR reduction performance and no restrictions to the number of the subcarriers. PTS method divides the input data block into disjoint sub-blocks and recombines them by using phase factors. The sub-blocks are then added to form the OFDM symbol for transmission. The objective of the PTS scheme is to select optimal phase factors for each sub-block set.

PTS method significantly reduces the PAPR, but unfortunately, finding the optimal phase factors is a highly search complex problem. In order to reduce the search complexity, the selection of the phase factors is limited to a set of finite number of elements. The exhaustive search algorithm (ESA) is then employed to find the best phase factor. However, the search complexity increases exponentially with the number of sub-blocks.

In order to reduce the search complexity, many extensions of PTS schemes have been proposed recently. However, for all these searching methods, either the PAPR reduction is suboptimal or the complexity is still high. For example, iterative flipping algorithm (IFA) has been proposed in [8] to reduce the PAPR with less computation complexity and implementation complexity. Using Hamming distance and Hamming weight to find phase factors has also been proposed in [9]. However, those algorithms are suboptimal. Therefore, we proposed two algorithms to reduce search complexity and those algorithms still have good performance.

The remainder of paper is organized as follows. section 2 introduces the PAPR in OFDM systems and the principles of the PTS method. Our proposed algorithms is presented in section 3. The results of simulations are shown in section 4. Finally, the conclusion is given in section 5.

## 2. PAPR AND PTS METHOD

### 2.1. PAPR in OFDM Systems

In an OFDM system, we denote $\mathbf{X}=[X_0,X_1,...,X_{N-1}]$ as the input data block of length $N$. Then the complex baseband OFDM signal consisting of $N$ subcarriers is given by

$$x(t)=\frac{1}{\sqrt{N}}\sum_{n=0}^{N-1}X_n\cdot e^{j2\pi\Delta ft},\ 0\le t<NT \tag{1}$$

where $j=\sqrt{-1}$, $\Delta f$ denotes the subcarrier spacing, and $NT$ denotes the data sequence period, $X_n$ denotes the modulated symbols. Then the PAPR of continuous-time signal $x(t)$ is defined as the ratio of maximum instantaneous power and average power, given by

$$PAPR=\frac{\max\limits_{0\le t<NT}|x(t)|^2}{E\left[|x(t)|^2\right]} \tag{2}$$

where $E[\bullet]$ denote expectation operation. We can also represent PAPR of discrete-time signal as

$$PAPR=\frac{\max\limits_{0\le k<N}|x_k|^2}{E\left[|x_k|^2\right]} \tag{3}$$

## 2.2. OFDM System with PTS to Reduce the PAPR

The PTS method is introduced in this section and the structure of PTS is shown in Fig. 1.

The input data $\mathbf{X}$ is partitioned into $M$ disjoint sets, and $\tilde{X}_i$ is the $i$-th sub-block with length $N$, where $i$=1,2,...,$M$, i.e.:

$$\mathbf{X}=\sum_{i=1}^{M}\tilde{X}_i. \tag{4}$$

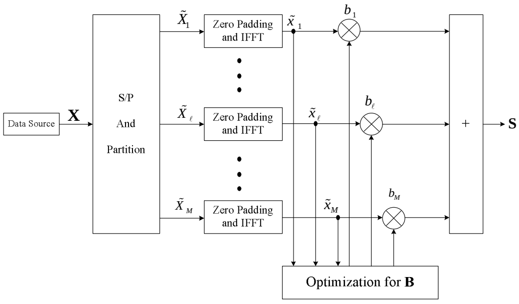


Fig. 1 The diagram of PTS structure

In general, for PTS scheme, the known sub-block partitioning methods can be classified into three categories: adjacent partition, interleaved partition and pseudo-random partition. In this paper, we choose adjacent partition. Then, each $\tilde{X}_i$ passes IFFT operation. We assume that

$$\tilde{x}_i=IFFT\{\tilde{X}_i\},\ \forall i. \tag{5}$$

Let

$$\begin{aligned}\tilde{\mathbf{x}}&=\begin{bmatrix}\tilde{x}_1 & \tilde{x}_2 & \cdots & \tilde{x}_M\end{bmatrix}^T\\ &=\begin{bmatrix}\tilde{x}_1\\ \tilde{x}_2\\ \vdots\\ \tilde{x}_M\end{bmatrix}=IFFT\left\{\begin{bmatrix}\tilde{X}_1\\ \tilde{X}_2\\ \vdots\\ \tilde{X}_M\end{bmatrix}\right\}\\ &=\begin{bmatrix}IFFT\{\tilde{X}_1\}\\ IFFT\{\tilde{X}_2\}\\ \vdots\\ IFFT\{\tilde{X}_2\}\end{bmatrix}\\ &=\begin{bmatrix}\tilde{x}_{1,0} & \tilde{x}_{1,1} & \cdots & \tilde{x}_{1,N-1}\\ \tilde{x}_{2,0} & \tilde{x}_{2,1} & \cdots & \tilde{x}_{2,N-1}\\ \vdots & \vdots & \vdots & \vdots\\ \tilde{x}_{M,0} & \tilde{x}_{M,1} & \cdots & \tilde{x}_{M,N-1}\end{bmatrix}.\end{aligned} \tag{6}$$

and

$$\mathbf{B}=\begin{bmatrix}b_1\\ b_2\\ \vdots\\ b_M\end{bmatrix}=\begin{bmatrix}e^{j\theta_1}\\ e^{j\theta_2}\\ \vdots\\ e^{j\theta_M}\end{bmatrix}, \tag{7}$$

where $\theta_i$ is a phase factor and $\theta_i\in[0,2\pi)$.

The output signal $\mathbf{S}$ is the combination of $\tilde{\mathbf{x}}$ and $\mathbf{B}$. The form of $\mathbf{S}$ is:

$$\begin{aligned}\mathbf{S}&=\tilde{\mathbf{x}}^T\mathbf{B}\\ &=\begin{bmatrix}\tilde{x}_{1,0} & \tilde{x}_{1,1} & \cdots & \tilde{x}_{1,N-1}\\ \tilde{x}_{2,0} & \tilde{x}_{2,1} & \cdots & \tilde{x}_{2,N-1}\\ \vdots & \vdots & \vdots & \vdots\\ \tilde{x}_{M,0} & \tilde{x}_{M,1} & \cdots & \tilde{x}_{M,N-1}\end{bmatrix}^T\begin{bmatrix}e^{j\theta_1}\\ e^{j\theta_2}\\ \vdots\\ e^{j\theta_M}\end{bmatrix}.\end{aligned} \tag{8}$$

Assume that $\boldsymbol{\theta}=[\theta_1\ \theta_2\ ...\ \theta_M]^T$ is the optimal phase vector for input signal. $\boldsymbol{\theta}$ must satisfy:

$$\hat{\boldsymbol{\theta}}=\begin{bmatrix}\hat{\theta}_1\\ \hat{\theta}_2\\ \vdots\\ \hat{\theta}_M\end{bmatrix}=\arg\min_{\theta_i\in[0,2\pi)}\max_{\ell=0,1,...,N-1}\left\{\frac{\left|\sum_{i=1}^{M}\tilde{x}_{i,\ell}e^{j\theta_i}\right|^2}{\frac{1}{N}\sum_{k=0}^{N-1}\left|\sum_{i=1}^{M}\tilde{x}_{i,k}e^{j\theta_i}\right|^2}\right\} \tag{9}$$

where $e^{j\hat{\theta}_i}$ is the optimum rotation of $i$-th sub-block.

Since $\theta_i\in[0,2\pi)$, it becomes extremely difficult to find the optimal phase vector. Typically, the phase factors are constrained to a finite set. For example, if we use the phase set $\theta_i\in\left\{0,\frac{\pi}{2},\pi,\frac{3\pi}{2}\right\}$, we can only find suboptimal phase vector $\hat{\boldsymbol{\theta}}_{sub}$. And $\hat{\boldsymbol{\theta}}_{sub}$ can be expresses as

$$\hat{\boldsymbol{\theta}}_{sub}=\begin{bmatrix}\hat{\theta}_1\\ \hat{\theta}_2\\ \vdots\\ \hat{\theta}_M\end{bmatrix}=\arg\min_{\theta_{sub_i}\in\left[0,\frac{\pi}{2},\pi,\frac{3\pi}{2}\right]}\max_{\ell=0,1,\ldots,N-1}\left\{\frac{\left|\sum_{i=1}^{M}\tilde{x}_{i,\ell}e^{j\theta_{sub_i}}\right|^2}{\frac{1}{N}\sum_{k=0}^{N-1}\left|\sum_{i=1}^{M}\tilde{x}_{i,k}e^{j\theta_{sub_i}}\right|^2}\right\} \tag{10}$$

Then, the transmitted signal is $\hat{\mathbf{S}}=\tilde{\mathbf{x}}^T\hat{\mathbf{B}}$, i.e.:

$$\hat{S}_k=\sum_{i=1}^{M}\tilde{x}_{i,k}e^{j\hat{\theta}_i}, k=0,2,\ldots,N-1. \tag{11}$$

In Eq. (10), it is obvious that finding a best phase factor set is still a complex and difficult problem when $M$ is large. Therefore, we proposed suboptimal search algorithms for PTS phase selection in the next section.

# 3. PROPOSED ALGORITHM

## 3.1. Observation and Definition of New Phase Factor

Form Eq. (6), $\tilde{\mathbf{x}}$ is a $M\times N$ matrix. The $i$-th row of $\tilde{\mathbf{x}}$ is the $i$-th sub-block after passing IFFT. $\tilde{x}_{i,k}$ is the $k$-th sample value of the $i$-th sub-block after IFFT, $i$=1, 2, ..., $M$ and $k$=1, 2, ..., $N$-1, and each $\tilde{x}_i$ is 1×$N$ matrix.

The output signal can be expressed as

$$\mathbf{S}=IFFT\{X\}=\left[\sum_{i=1}^{M}\tilde{x}_{i,0}\quad\sum_{i=1}^{M}\tilde{x}_{i,1}\quad\cdots\quad\sum_{i=1}^{M}\tilde{x}_{i,N-1}\right], \tag{12}$$

If multiplication between each sub-block and phase factor is not processed. The $k$-th element of 1×$N$ output matrix is the $k$-th sample output. **S** is output signal without passing PAPR reduction scheme. For all samples, let the index of the sample which has maximum value be $\hat{k}$.

$$\hat{k}=\arg\max_{0\le k\le N-1}\left\{\left|\sum_{i=1}^{M}\tilde{x}_{i,k}\right|\right\} \tag{13}$$

The original PAPR can be expressed as

$$PAPR_{orignal}=\frac{\left|\sum_{i=1}^{M}\tilde{x}_{i,\hat{k}}\right|^2}{\frac{1}{N}\sum_{k=0}^{N-1}\left|\sum_{i=1}^{M}\tilde{x}_{i,k}\right|^2} \tag{14}$$

For all sub-blocks of the $\hat{k}$-th sample, let the index of the sub-block which has the maximum value be $\hat{i}$.

$$\hat{i}=\arg\max_{0\le i\le M}\left\{\left|\sum_{i=1}^{M}\tilde{x}_{i,\hat{k}}\right|\right\} \tag{15}$$

The maximum sample can be re-written as

$$\begin{aligned}\sum_{i=1}^{M}\tilde{x}_{i,\hat{k}}&=\tilde{x}_{1,\hat{k}}+\tilde{x}_{2,\hat{k}}+\cdots+\tilde{x}_{M,\hat{k}}\\&=\left|\tilde{x}_{1,\hat{k}}\right|e^{j\angle\tilde{x}_{1,\hat{k}}}+\left|\tilde{x}_{2,\hat{k}}\right|e^{j\angle\tilde{x}_{2,\hat{k}}}+\cdots+\left|\tilde{x}_{M,\hat{k}}\right|e^{j\angle\tilde{x}_{M,\hat{k}}},\end{aligned} \tag{16}$$

We define the phase set

$$\theta_{\hat{k}}=\left\{\angle\tilde{x}_{i,\hat{k}}:i=1,2,\ldots,\hat{i}-1,\hat{i}+1,\ldots,M\right\}, \tag{17}$$

let

$$\theta_{d_t}=\left\{\angle\tilde{x}_{i,\hat{k}}-\angle\tilde{x}_{\hat{i},\hat{k}}:i=1,2,\ldots,\hat{i}-1,\hat{i}+1,\ldots,M\right\}, \tag{18}$$

where $t$ is the decreasing order that arranges the index of the sample values, so $t$=1 indicates the max-value sample. The simulation the PDF of $\theta_{d_t}$ with difference orders of sample is shown in Fig. 2 [10].

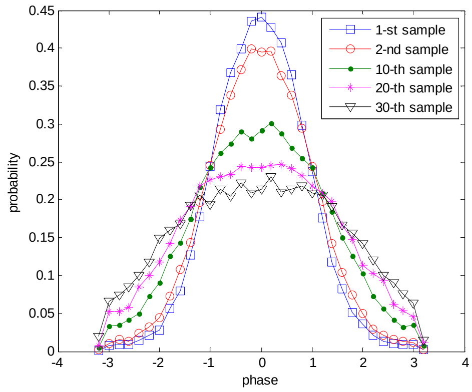


Fig. 2 PDF of $\theta_{d_t}$ for difference orders of sample

In Fig. 2, we observes that the phases of $\tilde{x}_{i,\hat{k}}$ are very close and that is why the maximum amplitude value is on the $\hat{k}$-th sample value of original signal after passing IFFT. Because the phases of $\hat{k}$-th sample value of each sub-block after IFFT are very close, we try to use this property and the amplitudes of $\tilde{x}_{i,\hat{k}}$, $i$=1, 2, ..., $M$, $i\neq\hat{i}$ to reduce the amplitude of $\tilde{x}_{\hat{i},\hat{k}}$.Now, we need to choose proper phase factor to reduce PAPR. Since $\angle\tilde{x}_{i,\hat{k}}$ is very close to $\angle\tilde{x}_{\hat{i},\hat{k}}$ for $i\neq\hat{i}$, we define a new phase factor.

We define that each $\angle\tilde{x}_{i,\hat{k}}$ rotates the phase with $\left\{0,\frac{\pi}{2},\pi,\frac{3\pi}{2}\right\}$, respectively, for $i\neq\hat{i}$. Then we can obtain four phase factor for some $i$:

$$\left\{ \angle\tilde{x}_{i,\hat{k}} \quad \angle\tilde{x}_{i,\hat{k}}+\frac{\pi}{2} \quad \angle\tilde{x}_{i,\hat{k}}+\pi \quad \angle\tilde{x}_{i,\hat{k}}+\frac{3\pi}{2} \right\}, \tag{19}$$

and let

$$\phi_j = \angle\tilde{x}_{i,\hat{k}} + j\frac{\pi}{2},\ j = 0,1,2,3 \tag{20}$$

then, we select one of $\phi_j$ which is nearest to the reverse direction of $\angle\tilde{x}_{\hat{i},\hat{k}}$.

$$\phi_i = \arg\min_{0\le j\le 3}\left(\angle\tilde{x}_{\hat{i},\hat{k}} + \pi - \phi_j\right). \tag{21}$$

We give an example in Fig. 3.

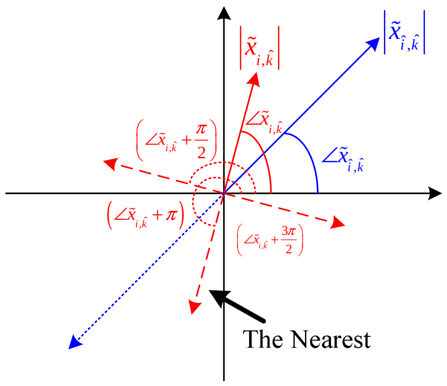


Fig. 3 Illustration of selecting $\phi_i$

Form Eq. (19), each sub-block has been calculated phase vector, except for the $\hat{i}$ -th sub-block.

## 3.2. A Sub-Optimal Search Algorithm for PTS Phase Selection

Although $\sum_{i=1}^{M}\tilde{x}_{i,\hat{k}}$ can be reduced, but some peak value may be generated with greater peak value. We should consider that it is not necessary to multiply all sub-blocks by the corresponding calculated phase vectors.

Thus, the data of the *i*-th sub-block after passing IFFT would be multiplied by $e^{j\phi_i}$ or $e^{j0}\left(=1\right)$ for $i \ne \hat{i}$ , the $\hat{i}$ -th sub-block after passing IFFT would be multiplied by 1. $\theta_i$ can be redefined as:

$$\theta_i = \begin{cases}\phi_i \\ 0\end{cases},\ i \ne \hat{i};\ \theta_i = 0,\ i = \hat{i}\ \cdot \tag{23}$$

Finally, we calculate all results of PAPR and select the proper phase vectors such that the transmitted signal has minimum PAPR. Each sub-block of $\tilde{x}$ has a new phase factor which was shown in eq. (23). The modified PTS structure is shown in Fig. 4.

Then, we proposed an algorithm (Algorithm I) as follows:

1. Input signal in frequency domain needs to be partitioned into $M$ disjoint sub-blocks, $\mathbf{X} = \sum_{i=1}^{M}\tilde{X}_{i}.$ .
2. Each $\tilde{X}_i$ passes IFFT.
3. Let $\hat{k} = \arg\max_{0\le k\le N-1}\left\{\left|\sum_{i=1}^{M}\tilde{x}_{i,k}\right|\right\}$.
4. Let $\hat{i} = \arg\max_{0\le i\le M}\left\{\left|\tilde{x}_{i,\hat{k}}\right|\right\}$.
5. Keep $\hat{i}$ -th sub-block unchanged. Each of the other *M*-1 sub-blocks has tow choice: unchanged or rotating to the nearest reverse direction of $\tilde{x}_{\hat{i},\hat{k}}$.
So, there are totally $2^{M-1}$ possible candidates.
6. Calculate the PAPR for each candidate, and select the phase vector which brings the minimum PAPR in this iteration.
7. With the new minimum PAPR signal, we can repeat step 3 to step 6 until the best minimum PAPR signal is found.

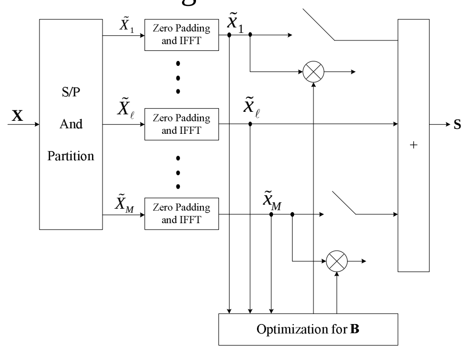


Fig. 4 The modified PTS structure

## 3.3. A Modified Sub-Optimal Search Algorithm for PTS Phase Selection

Although the performance of Algorithm I is not as good as ESA, the computational complexity is reduced. Furthermore, we found that there is rarely better minimum PAPR signal found after *iter*=3. Therefore, we proposed a modified sub-optimal search algorithm for PTS phase selection (Algorithm II).

The algorithm is shown as follows:

1. Input signal in frequency domain needs to be partitioned into $M$ disjoint sub-blocks, $\mathbf{X} = \sum_{i=1}^{M}\tilde{X}_{i}.$ .

2. Each $\tilde{X}_i$ passes IFFT.

3. Let $\hat{k} = \arg \max_{0 \le k \le N-1} \left\{ \left| \sum_{i=1}^{M} \tilde{x}_{i,k} \right| \right\}$.

4. Let $\hat{i} = \arg \max_{0 \le i \le M} \left\{ \left| \tilde{x}_{i,\hat{k}} \right| \right\}$.

5. Keep $\hat{i}$ -th sub-block unchanged. Each of the other $M$-1 sub-blocks has tow choice: unchanged or rotating to the nearest reverse direction of $\tilde{x}_{\hat{i},\hat{k}}$.

So, there are totally $2^{M-1}$ possible candidates.
6. Calculate the PAPR for each candidate, and select the phase vector which brings the minimum PAPR in this iteration.
7. Instead of considering the sample with the largest magnitude only, we consider the 2-nd ~*k*-th largest magnitude in step 6.

## 4. NUMERICAL RESULTS

Computer simulations are shown in this section. The simulation parameters are shown as follows. We consider an OFDM system with 128 subcarriers using QPSK modulation. A sub-optimal search algorithm for PTS phase selection we call the proposed Ⅰ.

Fig. 5 and Fig. 6 shows the CCDFs for the PTS method with the ESA, the PTS method with AlgorithmⅠ when *iter*=1, *iter*=2, *iter*=3, and the original OFDM. Fig. 5 is obtained directly from the output of IFFT operation for *k*=128 subcarriers and *M*=4 sub-blocks. Fig. 6 is obtained directly from the output of IFFT operation for *k*=128 subcarriers and *M*=8 sub-blocks. In ESA, four allowed phase factors +1, -1, +j, -j (W = 4) are used, and the PAPR reduction performance is obtained by a Monte Carlo search with WM phase factors.

In Fig. 5, the $10^{-3}$ PAPR of the OFDM signal is 11.11dB, indicating a large PAPR. The $10^{-3}$ PAPR of the ESA is 7.87dB. It is evident that the ESA algorithm can provide better PAPR reduction. However, Algorithm Ⅰ with iteration value *iter*= 1, *iter*=2 and *iter*=3, the $10^{-3}$ PAPRs reduce to 8.58dB, 8.452dB, 8.451dB, respectively. Fig. 5 shows that the performance of our proposed algorithm Ⅰ is much closer to that of the ESA. However, the search complexity is reduced significantly. We list the search complexity in Table 1.

In Fig. 6, the $10^{-3}$ PAPR of the OFDM signal is 11.36dB. The $10^{-3}$ PAPR of the ESA is 6.43dB, it is evident that the ESA algorithm can provide better PAPR reduction. However, Algorithm Ⅰ with iteration value *iter*= 1, *iter*=2 and *iter*=3 when $M$ = 8, the $10^{-3}$ PAPRs reduce to 7.36dB, 7.162dB, 7.162dB, respectively. Form Fig. 6 shows that the performance of Algorithm Ⅰ is much closer to that of the ESA. However, the search complexity is reduced significantly. We list the search complexity in Table 2.

**TABLE 1**
**SEARCH COMPLEXITY OF ESA AND ALGORITHM I($M$=4)**

| | Search complexity | $M$=4, $W$=4 | $10^{-3}$ PAPR |
|---|---|---|---|
| ESA | $W^M$ | 256 | 7.874dB |
| Algorithm Ⅰ *iter*=1 | $2^{M-1}$ | 8 | 8.577dB |
| Algorithm Ⅰ *iter*=2 | $2\times 2^{M-1}$ | 16 | 8.452dB |
| Algorithm Ⅰ *iter*=3 | $3\times 2^{M-1}$ | 24 | 8.451dB |

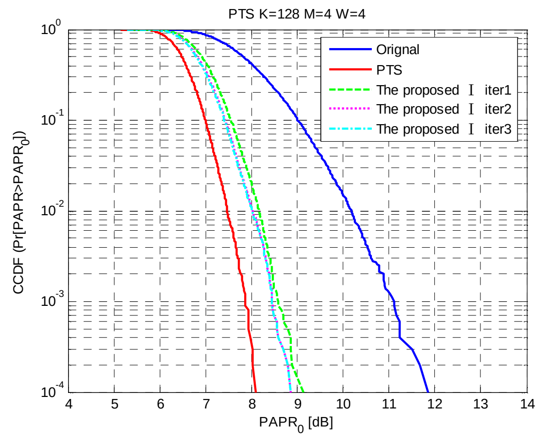


Fig. 5 CCDFs of the PAPR for Algorithm Ⅰ and ESA ($M$=4).

**TABLE 2**
**SEARCH COMPLEXITY OF ESA AND ALGORITHM I ($M$ = 8)**

| | Search complexity | $M$=8, $W$=4 | $10^{-3}$ PAPR |
|---|---|---|---|
| ESA | $W^M$ | 65536 | 6.428dB |
| Algorithm Ⅰ *iter*=1 | $2^{M-1}$ | 128 | 7.355dB |
| Algorithm Ⅰ *iter*=2 | $2\times 2^{M-1}$ | 256 | 7.162dB |
| Algorithm Ⅰ *iter*=3 | $3\times 2^{M-1}$ | 384 | 7.162dB |

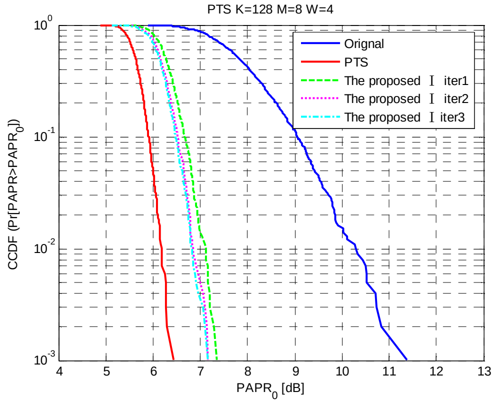


Fig. 6 CCDFs of the PAPR for Algorithm Ⅰ and ESA ($M$=8)..

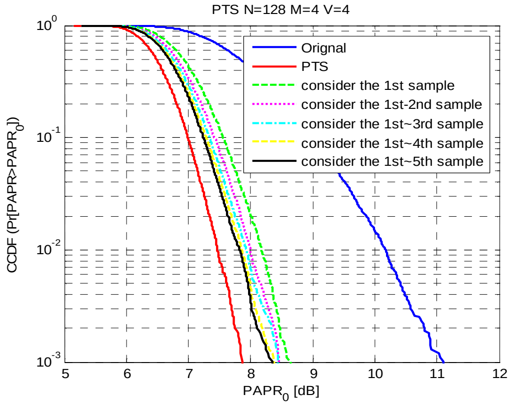


Fig.7 CCDFs of the PAPR for Algorithm II and ESA ($M$=4).

Fig. 7 shows the CCDFs for the PTS method with Algorithm II with the sample value consider the 1st sample , the 1st~2nd, 1st~3rd, 1st~4th and 1st~5th, the $10^{-3}$ PAPRs reduce to 8.642dB, 8.456dB, 8.439dB, 8.373dB, 8.353dB, respectively. Fig. 7 shows that the performance of Algorithm II is much closer to that of the ESA. However, the search complexity is reduced significantly. We list the search complexity in Table 3.

Fig. 8 shows the CCDFs for the PTS method with Algorithm II with the sample value consider the 1st sample , the 1st~2nd, 1st~3rd, 1st~4th and 1st~5th, the $10^{-3}$ PAPRs reduce to 7.704dB, 6.928dB, 6.839dB, 6.734dB, 6.701dB, respectively. Fig. 8 shows that the performance of Algorithm II is much closer to that of the ESA. However, the search complexity is reduced significantly. We list the search complexity in Table 4.

**TABLE 3**
**SEARCH COMPLEXITY OF ESA AND ALGORITHM II ($M$ = 4)**

| | Search complexity | $M$=4, $W$=4 | $10^{-3}$ PAPR |
|---|---|---|---|
| ESA | $W^M$ | 256 | 7.872dB |
| Algorithm II 1st | $2^{M-1}$ | 8 | 8.642dB |
| Algorithm II 1st~2nd | $2\times2^{M-1}$ | 16 | 8.456dB |
| Algorithm II 1st~3rd | $3\times2^{M-1}$ | 24 | 8.439dB |
| Algorithm II 1st~4th | $4\times2^{M-1}$ | 32 | 8.373dB |
| Algorithm II 1st~5th | $5\times2^{M-1}$ | 40 | 8.353dB |

**TABLE 4**
**SEARCH COMPLEXITY OF ESA AND ALGORITHM II ($M$ = 8)**

| | Search complexity | $M$=8, $W$=4 | $10^{-2}$ PAPR |
|---|---|---|---|
| ESA | $W^M$ | 65536 | 6.244dB |
| Algorithm II 1st | $2^{M-1}$ | 128 | 7.704dB |
| Algorithm II 1st~2nd | $2\times2^{M-1}$ | 256 | 6.928dB |
| Algorithm II 1st~3rd | $3\times2^{M-1}$ | 384 | 6.839dB |
| Algorithm II 1st~4th | $4\times2^{M-1}$ | 512 | 6.734dB |
| Algorithm II 1st~5th | $5\times2^{M-1}$ | 640 | 6.701dB |

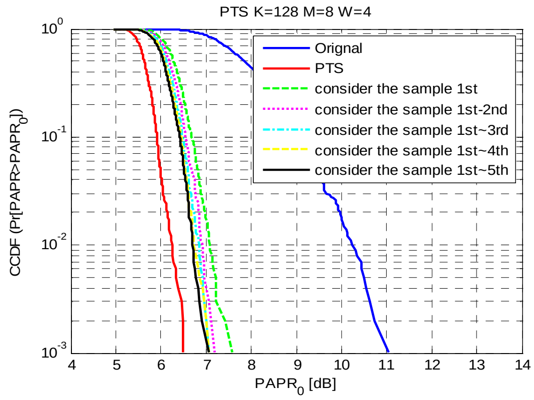

Fig.8 CCDFs of the PAPR for Algorithm II and ESA ($M$=8).

## 5. CONCLUSION

Two algorithms for the PTS method are proposed to reduce the PAPR of OFDM signals in this paper. The simulation results show that those schemes provides good the PAPR reduction performance. And our proposed method reduces the search complexity while keeping good PAPR performance.

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## REFERENCES

[1] R. Van Nee and R. Prasad, *OFDM for Wireless Multimedia Communication*, Artech House Publishers, Boston, MA, 2000.

[2] E. H. M. Alian, H. E. Saffar, and P. Mitran, Jul. 2012"Cross-Band Interference Reduction Trade-Offs in SISO and MISO OFDM-Based Cognitive Radios," *IEEE Transactions on Wireless Communications*, vol. 11, no. 7, pp. 2436-2445.

[3] T. Demeechai, T. G. Chang, and S. Siwamogastham, Oct. 2012, "Performance of a Frequency-Domain OFDM-Timing Estimator," *IEEE Communications Letters*, vol. 16, no. 10, pp. 1680-1683.

[4] R. W. Bauml, R. F. H. Fischer, and J. B. Huber, Oct. 1996, "Reducing the peak-to-average power ratio of multicarrier modulation by selected mapping," *Electron. Letters*, vol. 32, no. 22, pp. 2056-1257.

[5] T. A. Wilkinson and A. E. Jones, Jul. 1995, "Minimization of the peak-to mean envelope power ratio of multicarrier transmission schemes by block coding," in proc. IEEE Vehicular Technology Conference, Chicago, pp. 825-829.

[6] H. Schmidt and K. D. Kammeyer, Oct. 1998, "Reducing the peak to average power ratio of multicarrier signals by adaptive subcarrier selection," *Proceedings of IEEE International Conference on Universal Personal Communications*, vol. 2, pp. 933-937.

[7] S. H. Muller, J. B. Huber, 1997, "OFDM with reduced peak-to-average power ratio by optimum combination of partial transmit sequences," *Electron. Letters*, vol. 33, pp. 368-369.

[8] L. J. Cimiini, Jr. and N. R. Sollenberger, Mar. 2000, "Peak-to-average power ratio reduction of an OFDM signal using partial transmit sequences," *IEEE Commun. Lett.*, vol.4, pp.86-88.

[9] S.H. Han and J. H. Lee, Nov. 2004, "PAPR reduction of OFDM signals using a reduced complexity PTS technique," *IEEE Signal Processing Letters*, vol. 11, no. 11.

[10] Y. R. Tsai, S. J. Hung, Jan. 2008, "PTS with non-uniform phase factors for PAPR reduction in OFDM Systems", IEEE Comm. Lett., vol. 12, no. 1.