PAPR Reduction of Multicarrier Faster-than-Nyquist Signaling by PTS Based on Tree

Biao Cai\textsuperscript{1}, Aijun Liu\textsuperscript{1*}, Chao Gao\textsuperscript{2}, Xiaohu Liang\textsuperscript{1}

\textsuperscript{1}College of Communication Engineering, Army Engineering University, Nanjing, Jiangsu, 210007, China

\textsuperscript{2}Beijing Institute of System Engineering, Beijing, 100101, China

\textsuperscript{*}Corresponding author’s e-mail: liuaj.cn@163.com

Abstract. One of the common drawbacks of multicarrier communication systems is high peak-to-average power ratio (PAPR). Due to the overlapped structure of multicarrier faster-than-Nyquist (MFTN) signaling, conventional PAPR reduction methods show poor performance. In this paper, we investigate the PAPR reduction of MFTN signaling. Specifically, as one of the most attractive methods, partial transmit sequence (PTS) method is exploited in MFTN signaling system. First, PTS method is combined with alternative signal (AS) to overcome the overlapped structure of MFTN signaling. Then, a tree based PTS method combined with AS is proposed, and all the candidate signal sequences can be generated by adding the difference signal sequence on the last candidate signal sequence, which can reduce the computational complexity significantly. Simulation results show that the proposed method does not have performance degradation on PAPR reduction.

1. Introduction

With the fast development of wireless communications technology, spectrum resource has become more and more scarce. The researches in spectral efficiency transmission schemes are attractive and growing rapidly. A most common way is simply increasing the modulation level, which will result in decrease in energy efficiency. An alternative way is breaking the classical Nyquist transmission theory, in other word, sending the symbol faster than Nyquist criterion, to arrive at faster-than-Nyquist signalling [1-4]. By extending the idea of FTN to both time and frequency domain, multicarrier faster-than-Nyquist (MFTN) [5-6] signaling can be obtained. It has mentioned in [5-6] that the rate improvement is higher than conventional single carrier FTN and can reach as high as 100% when combined with some iterative detection algorithms. Duo to the high spectral efficiency of MFTN signaling, it has been considered as one of the potential promising techniques for future satellite and optical communication systems [7-10].

Since that time interval packing of adjacent symbols and frequency spacing packing of adjacent subcarrier for conventional Nyquist multicarrier signaling systems exist in MFTN signaling, the inevitable intersymbol interference (ISI) and intercarrier interference (ICI) are introduced. And current researches about MFTN mainly focus on the Mazo limit [11] and signal detection algorithms.

However, similar to other multicarrier signals, MFTN exhibits high peak-to-average power ratio (PAPR) which is one of the main drawbacks. It will lead to MFTN signaling very sensitive to the nonlinear devices, such as high power amplifier (HPA) and digital-to-analog converter (DAC). When the high peak of MFTN signaling reaches the nonlinear region of nonlinear device, in-band distortion...
and out-of-band radiation occur which will severely degrade the system performance. Therefore, PAPR reduction method is important for MFTN signaling.

Currently, the researches investigating the PAPR of MFTN signaling is not enough yet. In [12-13], the PAPR problem of single carrier FTN is investigated. The analysis and results show that the ISI of single carrier FTN increase the envelope fluctuation of it. In consideration of the fact that ISI and ICI exist in MFTN signaling, MFTN signaling suffers from higher PAPR than conventional orthogonal frequency division multiplexing (OFDM) [5]. In [14], a two-stage phase rotation partial transmit sequence (PTS) method has been proposed for PAPR reduction in MFTN signaling. It utilizes the discrete Fourier transform (DFT) implementation structure in MFTN signaling transmitter and shown better PAPR reduction performance than directly application of conventional PTS method. However, a two-stage phase rotation optimization will dramatically increase the number of candidate signal sequences, which results in extremely high computational complexity, and increase the side information, which degrades the spectral efficiency. On the other hand, although the PAPR reduction has been widely investigated and a lot of PAPR reduction methods have been proposed in conventional Nyquist orthogonal multicarrier transmission systems, such as OFDM, the directly adoption of them will result in poor performance on account of the special signal structure of MFTN signaling.

In this paper, we address the PAPR problem of MFTN signaling, propose a effective PAPR reduction method and focus on reducing the computational complexity of it. The main contributions of this paper can be listed as follows.

1) In order to handle the overlapped structure of MFTN signaling, PTS combined with alternative signal (AS) [15], which is referred to by PTS-AS in the following paper, is introduced.

2) The replication computation in PTS-AS is analyzed. Then, tree based PTS-AS (T-PTS-AS) is proposed to remove the replication computation. The simulation results demonstrate that T-PTS-AS method can dramatically reduce the computational complexity while achieving the same PAPR reduction performance as PTS-AS method.

The paper is organized as follows. In section II, the system model of MFTN signaling and the refined PAPR are briefly introduced. The conventional PTS method and PTS-AS are presented in section III. Section IV analyzes the replication computation in PTS-AS and gives the detailed description of T-PTS-AS method. The computational complexity and PAPR reduction performance are evaluated in Section V. Section VI is the conclusion.

2. System Model and PAPR
Let us consider a baseband MFTN system with $K$ subcarriers. The data streams are modulated to form the independent identically distributed (i.i.d) transmit symbols $\{a_{lk}\}$, where $l$ is the time index and $k$ is the subcarrier index, belonging to given $M$-th order complex constellation with zero-means. The transmit symbols pass through the base pulse $g(t)$ and regularly shift in both time and frequency domain. Then, the baseband MFTN signaling is given by

$$s(t) = \sum_{l=0}^{\infty} \sum_{k=0}^{K-1} a_{lk}g(t-l\tau T)e^{-j2\pi Fk(t-l\tau T)}$$  \hspace{1cm} (1)$$

where $g(t)$ is $T$-orthogonal base pulse satisfying the equation defined by

$$\int_{-\infty}^{\infty} g(t-mT)g(t-nT)dt = 0, m \neq n. \quad F \text{ is the minimum orthogonal frequency spacing between adjacent subcarriers and unequal for different shaping pulse. For instance, } F=1/T \text{ for sinc pulse and } F=(1+\beta)/T \text{ for root raised cosine pulse with roll-off factor } \beta. \text{ Moreover, } \tau \text{ and } \nu \text{ are the time interval packing ratio and frequency spacing packing ratio, respectively. For MFTN signaling systems, } \tau, \nu \in [0,1] \text{ which makes the time interval } \tau T \text{ and frequency spacing } \nu F \text{ smaller than that of the conventional orthogonal multicarrier transmission systems to achieve higher spectral efficiency. And MFTN signaling will be reduced to conventional Nyquist situation when } \tau \cdot \nu = 1.$$
It is worthy of note that the receiver of MFTN signaling is omitted in this paper, since we focus on the PAPR reduction which is localized in the transmitter.

Generally, PAPR is a frequently used metric to measure the sensitivity of transmitted signal with non-constant envelop to the nonlinear devices. Similar to [14], by dividing the transmitted signal into equal length time interval with length $\tau T$, and focus on the PAPR of MFTN in per symbol period $\tau T$, then, we can obtain the PAPR for each interval by

$$PAPR_p = 10\log_{10} \frac{\max_{t \leq (p+1)\tau T} |s(t)|^2}{E[|s(t)|^2]}, \quad p = 0,1,\cdots$$

in which $p$ is the time index, $E[\cdot]$ stands for the expectation operation.

PAPR in (2) can only represent the PAPR in a given interval. In general, probability distribution of PAPR is used to evaluate the PAPR performance of a transmission system. One of the commonly used probability distribution function is complimentary cumulative distribution function (CCDF) which is the probability that the PAPR exceeds a given level $PAPR_0$ defined by

$$CCDF = Pr(PAPR > PAPR_0)$$

where $Pr(\cdot)$ represents the probability of event in brackets.

It is known that PTS method is implemented on discrete time signals. For the convenience of the analysis and derivation in the following paper, the expressions in (1) and (2) need to be transformed into discrete forms. By sampling the MFTN signaling with sample period $cT$ and defining $s[n] = s(nT_c)$, we can obtain the discrete time MFTN signaling by

$$s(n) = \sum_{l=0}^{\infty} \sum_{k=0}^{K-1} a_{l,k} g[n-l\tau N] e^{j2\pi n k / JK}$$

where $g[n]$ is the sampled baseband pulse with finite support $[-L_s / 2, L_s / 2]$. According to [15], $J \geq 4$ is sufficient to make the PAPR of discrete time MFTN signal approximate the true PAPR. Then, the corresponding PAPR can be defined by

$$PAPR_j = 10\log_{10} \frac{\max_{l\leq\frac{\pi}{2\tau T}} |s[n]|^2}{E[|s[n]|^2]}$$
3. PTS and PTS-AS in MFTN Signaling

3.1. PTS Method

In PTS method, the $l$-th input data block $\mathbf{a}_l = [a_{l,0}, a_{l,1}, \ldots, a_{l,K_l}]$ is partitioned into $U$ disjoint subblocks, which are represented by $\{a^{(u)}_l, u = 0, 1, \ldots, U-1\}$. Hence,

$$\mathbf{a}_l = \sum_{u=0}^{U-1} a^{(u)}_l$$

where $a^{(u)}_l = [a^{(u)}_{l,0}, a^{(u)}_{l,1}, \ldots, a^{(u)}_{l,K_l}]$ with $a^{(u)}_{l,k} = a_{l,k}$ or 0. Then, MFTN symbols in the data subblock $a^{(u)}_l$ pass through the baseband pulse to arrive at $a^{(u)}_{l,k} g[n-l\tau N]$. After that, modulated with the corresponding subcarriers, we have

$$a^{(u)}_{l,k} g[n-l\tau N] e^{j2\pi/k \phi(n-l\tau N)}$$

in which $-L_s/2 + l\tau N \leq n \leq L_s/2 + l\tau N$. Then, add all the modulated subcarriers in (9), the signal subsequence corresponding to $u$-th subblock in $l$-th data block can be obtained by

$$s^{(u)}_l[n] = \sum_{k=0}^{K_l} a^{(u)}_{l,k} g[n-l\tau N] e^{j2\pi/k \phi(n-l\tau N)}$$

All these signal subsequences corresponding to the $l$-th data block are rotated by the independent phase rotation factors $\{\phi_u, u = 0, 1, \ldots, U-1\}$. In practical implementation, the values of phase rotation factors are fetched from a finite set defined by $\phi_u \in \{e^{j2\pi/w}, w = 0, 1, \ldots, W-1\}$. There are $U$ phase rotation factors for each data block, by combining them together, we can define the phase rotation vectors by $\mathbf{b}^{(c)} = [b^{(c)}_0, b^{(c)}_1, \ldots, b^{(c)}_{U-1}]$, $c = 0, 1, \ldots, C-1$, where $C = W^{U-1}$ is the number of all possible phase rotation vectors. And the candidate signal sequence corresponding to a given phase rotation vector $\mathbf{b}^{(c)}$ is given by

$$s^{(c)}_l[n] = \sum_{u=0}^{U-1} b^{(c)}_u s^{(u)}_l[n]$$

The objective is to optimally combine the $U$ signal subsequences to obtain the optimal MFTN signal sequence with lowest PAPR. And the process can be formulated as

$$\min_{\mathbf{b}^{(c)}} \max_{-L_s/2 + l\tau N \leq n \leq L_s/2 + l\tau N} \|s^{(c)}_l[n]\|^2.$$ 

And the exhaustive search is adopted here to find the optimal phase rotation vector.

3.2. PTS-AS Method

Due to the overlapped structure of MFTN signaling, the PTS method in previous subsection show poor PAPR reduction performance in MFTN signaling. In order to give a practicable and effective method, the PTS-AS method is proposed which is composed by PTS and AS method. The procedures of generating the candidate signal sequence of PTS-AS method is same as that of PTS method by (11). Therefore, those procedures are omitted here.

The main difference of PTS-AS and PTS method is the objective function for finding the optimal phase rotation vector. The sequential optimization is adopted in PTS-AS. To be specific, the previous MFTN symbols are taken into account in finding the optimal candidate signal sequence of current MFTN data block. And a mathematical statement is given below.

For the zero-th MFTN data block, after the generation of all possible candidate signal sequence $s^{(c)}_0[n]$, the one with the lowest PAPR is chosen by
\[ \hat{s}_0[n] = \arg \min_b \max_{L_n \leq L \leq L_n+L_y/2} \left| s_0^b[n] \right|^2 \]  

(13)

where \( \hat{s}_0[n] \) is the chosen signal sequence for the zero-th MFTN data block. Then, \( \hat{s}_0[n] \) is used in the optimization problem for choosing the signal sequence for the first MFTN data block by

\[ \hat{s}_1[n] = \arg \min_b \max_{L_n \leq L \leq L_n+L_y+\tau N} \left| \hat{s}_0[n] + s_i^b[n] \right|^2. \]  

(14)

By that analogy, \( \hat{s}_0[n] \) and \( \hat{s}_1[n] \) are used to find the next optimal signal sequence \( \hat{s}_2[n] \). The procedure above is repeated in the subsequent MFTN symbols. Thus, the general objective function of PTS-AS method can be formulated by

\[ \hat{s}_l[n] = \arg \min_b \max_{L_n \leq L \leq L_n+L_y+\tau N} \left| \sum_{i=0}^{l-1} \hat{s}_i[n] + s_i^b[n] \right|^2. \]  

(15)

Also, the exhaustive search is adopted here.

It is worthy of note that the first factor of phase rotation vector cannot be fixed in the search since the existence of previous symbols in the objective function. Thus, \( C = W^U \) in PTS-AS method. Although the computational complexity of PTS-AS is higher than PTS method, the computational complexity is still acceptable and it overcomes the overlapped structure and reduce the PAPR of MFTN effectively.

4. T-PTS-AS Method

In the PTS-AS method, the generation of a large amount of candidate signal sequences leads to the high computational complexity. However, a lot of replication computations exist in the generation of candidate signal sequences which can be used to reduce the complexity. Actually, similar works has been done in OFDM systems [16-17], but can not directly used to MFTN signaling. In the section, a novel tree based PTS-AS method is proposed for MFTN signaling by building a \( W \)-way tree, listing all the candidate signal sequence into the tree structure and utilizing the correlation between the neighbor leaves to removing the replication computation. The T-PTS-AS can achieve much lower complexity than PTS-AS method.

Here, we firstly analyze the replication computation in PTS-AS method. For the convenience of presentation, we assume that \( W = 2 \), \( U = 4 \), and vector \( s_i^{(n)} \) is used to represent the signal subsequence \( s_i^{(n)}[n] \) in which \( -L_y/2 + \tau n \leq n \leq L_y/2 + \tau n \).

For two given phase rotation vectors \( b = \{+1,+1,+1,+1\} \) and \( b = \{+1,+1,+1,-1\} \), the corresponding candidate signal sequences are

\[ s_i^{(0)} + s_i^{(1)} + s_i^{(2)} + s_i^{(3)} \]  

(16a)

\[ s_i^{(0)} + s_i^{(1)} + s_i^{(2)} - s_i^{(3)} \]  

(16b)

It is easy to see that \( s_i^{(0)} + s_i^{(1)} + s_i^{(2)} \) is computed repeatedly in (16a) and (16b). Actually, the signal sequence in (16b) can be calculated by

\[ (16b) = (16a) - s_i^{(3)} - s_i^{(3)}. \]  

(17)

And this will be called by identical signal replication computation in the following paper.

For another two given phase rotation vectors \( b = \{+1,+1,+1,+1\} \) and \( b = \{+1,+1,-1,-1\} \), the corresponding signal sequences are

\[ s_i^{(0)} + s_i^{(1)} - s_i^{(2)} + s_i^{(3)} \]  

(18a)

\[ s_i^{(0)} + s_i^{(1)} - s_i^{(2)} - s_i^{(3)} \]  

(18b)

Also, the signal sequence (18b) can be calculated by

\[ (18b) = (18a) - s_i^{(3)} - s_i^{(3)}. \]  

(19)
Note that the difference signal sequence in (17) and (19) are the same. In other word, the difference signal \( s_l^{(3)} - s_l^{(3)} \) is computed repeatedly. And this will be called by difference signal replication computation in the following paper.

Based on the analysis above, we can construct \( W \)-way tree according to the following rules:

1) There are \( U + 1 \) layers in the tree which are corresponding to interference sequence in current signal sequence brought by the previous MFTN symbols and \( U \) signal subsequence, respectively.

2) Each father node has \( W \) children nodes. And \( W \) nodes value is the fetched from \( \{e^{i2\pi w/W}, w = 0, 1, ..., W - 1\} \) orderly. Since the phase rotation vector of current data block can not change the interference sequence, we can set the root node value to be \( 1 \).

3) Put the nodes in a \( 2W \) period per layer orderly. And, a given \( W \) nodes are symmetry with the last and the next \( W \) nodes.

Fig. 1 exhibits the sketch of \( W \)-way tree structure. All the candidate signal sequences can be obtained by combining the interference signal sequence and signal subsequence which are phase rotated by the factors on the nodes from the root to the leaves. From Fig. 1, we can see that only one phase rotation factor is different on the paths from root to the adjacent leaves. It means that a candidate signal sequence can be generated by utilizing the last candidate signal sequence, as the calculation in (17) and (19). Thus, the candidate signal sequences can be generated one by one, once the difference signal sequences is known, and the identical signal replication computation is removed.

According to the 3) of \( W \)-way tree construction rules above, there are only \( (W - 1) \) difference signal sequence in each layer (except the first layer). We can calculate and store the difference signal sequences for each layer. Thus, the difference signal replication computation can be removed. The difference signal sequence between \( w \)-th node and \( (w - 1) \)-th node in the \( (u + 2) \)-th layer can be calculated by

\[
d_{u,w} = b_{u,w} s_l^{(u)} - b_{u,w-1} s_l^{(u)}
\]

in which \( 0 \leq u \leq U - 1 \) and \( 1 \leq w \leq W - 1 \).

There are \( C = W^U \) leaves (or candidate signal sequence) in the tree and we numbered them from left to right. When calculate the \( (c + 1) \)-th candidate signal sequence, we need to know which difference signal sequence on which layer should added to the \( c \)-th candidate signal sequence. Actually, we can conclude the rule of layer index \( (u + 2) \) and different node index \( n_d \). They satisfy that

\[
(c + 1) = 0 \mod (W^{U-u-1}) \quad \text{and} \quad (c + 1) \neq 0 \mod (W^{U-u})
\]

in which \( 1 \leq c \leq W^U - 1 \), and

\[
n_d = (c + 1) / W^{U-u-1}.
\]
Note that $2W$ period in nodes of each layer, the difference signal sequence can be acquired by

$$d_{u,n_d} = \begin{cases} d_{u,n'_d} & \text{for } n'_d < W \\ -d_{u,(2W-n'_d)} & \text{for } n'_d > W \end{cases}$$

(23)

where $n'_d = n_d \mod (2W)$. Finally, the $(c+1)$-th candidate signal sequence can be calculated by

$$\sum_{i=0}^{L-1} s_i + s_i^c = \sum_{i=0}^{L-1} s_i + s_i^{c-1} + d_{u,n_d}.$$  

(24)

where $\sum_{i=0}^{L-1} s_i$ stands the interference sequence in current signal sequence brought by the previous MFTN symbols. After the generation of all the candidate signal sequence, the optimal signal sequence is chosen according to (15).

The T-PTS-AS method utilizes the structure of $W$-way tree to reduce the computational complexity. The procedures of generation candidate signal sequence in T-PTS-AS after the generation of signal subsequences can be summarized as follows:

1) Calculate the $(W-1)c$ difference signal sequence according to (20).

2) Let $c = 0$, calculate the first candidate signal sequence defined by $\sum_{i=0}^{L-1} s_i + s_0^0$.

3) Set $c$ to $c + 1$. Utilizing (21), (22), and (23) to determine the corresponding difference signal sequence. And calculate the $(c+1)$-th candidate signal sequence by (24).

4) If $c < W^{W-1}$, return to step 3).

5. Performance Analysis

5.1. Complexity Analysis

Since the T-PTS-AS focus on reducing the computational complexity after the generation of signal subsequences, the complexities of combining the signal subsequence to construct all the candidate signal sequences are compared here.

For multiplications, when $W = 2$ and $4$, $b \in \{+1,-1\}$ and $b \in \{+1,-1,+j,-j\}$, respectively. Therefore, no multiplication is conducted in phase rotation of signal subsequences in (11) and (20). In other word, no multiplication is needed after the generation of signal subsequences. Thus, the complexity of multiplication will not be compared here. When $W = 8$, multiplications are conducted in half of phase rotation of signal subsequences. And there are $W^{W-1}U$ times and $WU$ times phase rotation of signal subsequence in PTS-AS and T-PTS-AS method, respectively.

For additions, there are $W^W U (L+1)$ additions and $(WU + W^{W-1})(L+1)$ additions in generating all candidate signal sequence in PTS-AS and T-PTS-AS method, respectively.

Computational complexity reduction ratio (CCRR) is often used to evaluate computational complexity. Here, we can define it by

$$CCRR = \left(1 - \frac{\text{complexity of T-PTS-AS}}{\text{complexity of PTS-AS}}\right) \times 100\%$$

(25)

Table I shows CCRR of T-PTS-AS over the PTS-AS with typical SU$ and $W$, from which we can see that T-PTS-AS dramatically reduce the computational complexity.

| $U$ | $W$ | Complex Multiplication (%) | Complex Addition (%) |
|-----|-----|---------------------------|----------------------|
| 2   | 2   | /                         | 64.0625              |
| 4   | 4   | /                         | 73.5352              |
5.2. Complexity Analysis

In this simulation, the modulation scheme is QPSK, and the number of subcarrier $K = 64$. The RRC pulse with $\beta = 0.3$ is employed as baseband pulse. The number of subblocks is $U = 4$.

Fig. 2 exhibits the PAPR reduction performance of conventional PTS, PTS-AS, T-PTS-AS. The $\tau$ and $\nu$ are both fixed to 0.8 in Fig. 2. The CCDF of original MFTN signaling is also shown for comparison. The PTS shows poor performance in MFTN signaling. Even the number of candidate signal sequences is increased, the performance improvement is inapparent. Compared with PTS method, the PTS-AS shows much better performance than PTS method, and it is more suitable for MFTN signaling. Moreover, since no reduction in the number of candidate signal sequences, T-PTS-AS does not have any performance loss when compared with PTS-AS.

Fig. 3 exhibits the PAPR reduction performance in MFTN signaling with $\tau = 0.5$ and $\nu = 0.5$. The simulation result in Fig. 3 is similar to Fig. 2. Even the packing factors vary, the proposed methods show good performance which proves the robustness of them.

| $\tau$ | $\nu$ | PAPR (dB) |
|--------|--------|------------|
| 0.3    | 0.3    | 3.72       |
| 0.5    | 0.5    | 3.52       |

Figure 2. PAPR reduction performance of MFTN signaling, $\tau = 0.8$, $\nu = 0.8$.

Figure 3. PAPR reduction performance of MFTN signaling, $\tau = 0.5$, $\nu = 0.5$. 
6. Conclusion
High PAPR is one of the main drawbacks of MFTN signaling. In this paper, we introduce AS to combined with PTS method which can overcome the overlapped structure of MFTN signaling. Then, the replication computation in PTS-AS is analyzed. Based on that, T-PTS-AS method is proposed in MFTN signaling, which can reduce the complexity significantly. Moveover, T-PTS-AS can keep the same performance as PTS-AS.

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