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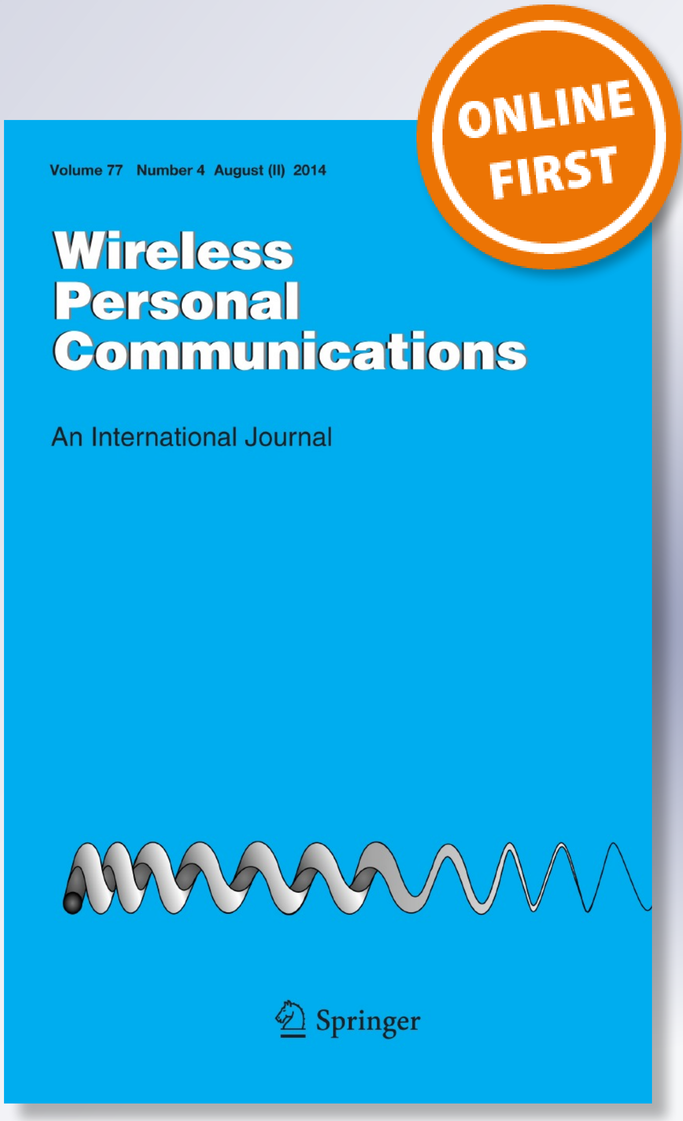

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PAPR Reduction in MIMO-OFDM Systems: Spatial and Temporal Processing

Bashir Reza Karimi · Mojtaba Beheshti · Mohammad Javad Omidi

Abstract Multiple-input multiple-output orthogonal frequency division multiplexing (MIMO-OFDM) technology is a promising solution for next generation wireless communications, due to high bandwidth efficiency, resistance to RF interference, and robustness to multipath fading. A major drawback of OFDM is its high peak-to-average power ratio (PAPR) which results in non-linearities in the output signal. In this paper, two methods based on spatial/temporal processing are proposed to reduce the PAPR of MIMO-OFDM systems. These methods divide the OFDM block at each transmit antenna into some subblocks. Then, spatial and temporal processing in the form of circular shifting or interleaving are applied to generate different candidate sequences. Finally, for each transmit antenna the candidate sequence with the lowest PAPR is chosen for transmission. Compared to the conventional PAPR reduction schemes such as ordinary partial transmit sequences (O-PTS), the proposed methods require lower computational complexity and have superior PAPR reduction performance.

Keywords Multiple-input multiple-output (MIMO) · Orthogonal frequency division multiplexing (OFDM) · Peak-to-average power ratio (PAPR) · Spatial and temporal processing

1 Introduction

Orthogonal frequency division multiplexing (OFDM) technique [1] has received considerable attention in wireless communications because of the following advantages [2,3]: a) immunity

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to multipath fading, inter-symbol interference, and co-channel interference; b) high spectral efficiency in supporting broadband communication; c) lower implementation complexity compared to the single carrier solution. These advantages are ensured by dividing a wide-band frequency selective channel into parallel narrowband flat fading subchannels. OFDM has been adopted in a wide range of applications including asymmetric digital subcriber line (ADSL), wireless local area networks such as IEEE 802.11, and wireless metropolitan area networks such as IEEE 802.16. It has also been considered for regional area networks using cognitive radio technology as in IEEE 802.22 and the next generation of mobile communication (long-term evolution (LTE)) [4]. The combination of multiple-input multiple-output (MIMO) technology and OFDM is a viable solution for future wireless communication systems. MIMO methods can improve the error performance, signal quality, and system capacity [5]. Therefore, MIMO-OFDM has attracted lots of interests in various wireless systems and standards [6].

The time domain MIMO-OFDM signal is a summation of several orthogonal waveforms. So, large peaks which are much higher than the average of the signal may happen. This phenomenon is called high peak-to-average power ratio (PAPR), which is a major drawback of MIMO-OFDM systems at the transmitter side [7]. The high PAPR increases the complexity of digital-to-analog converters (DACs). Also, it can cause non-linear distortion due to non-linear devices such as DACs, mixers, and high power amplifiers (HPAs). The non-linear distortion introduces spectral spreading, inter-modulation, and changes in the signal constellation. Using expensive power amplifiers with large linear ranges is a solution to mitigate the PAPR problem. However, high linearity implies low power efficiency and the amplification process would be inefficient. PAPR reduction schemes based on signal modification have also been proposed to deal with the PAPR problem.

Various PAPR reduction methods for single-input single-output (SISO)-OFDM systems have been addressed in the literature [8–17]. These methods can be classified into distortion and distortionless ones. Distortion methods, such as amplitude clipping [8], clipping and filtering [9], and nonlinear companding transform [10], lead to in-band distortion, peak regrowth, or out-of-band radiation which results in bit error rate (BER) performance degradation. Many distortionless methods have also been proposed which achieve PAPR reduction at the expense of transmit power increase, data rate loss, and computational complexity increase. These methods are based on coding [11], tone reservation [12], tone injection [13], active constellation extension [14], partial transmit sequence (PTS) [15], selective mapping (SLM) [16], or interleaving [17]. Among them, PTS is a promising method since it is simple to implement, introduces no distortion in the transmitted signal, and has good PAPR reduction performance. However, PTS suffers from high computational complexity due to several inverse fast Fourier transform (IFFT) operations, and requires side information transmission. Any error in the detection of side information can damage the entire data block and degrade the BER performance drastically.

First extensions of PAPR reduction methods to MIMO case are given in [18,19]. Recently, several articles addressed the PAPR reduction for space frequency (time) block coding MIMO-OFDM systems [20–22]. Also, some PTS-based methods have been introduced to improve the PAPR statistics for MIMO-OFDM. Ordinary PTS (O-PTS) and alternate PTS (A-PTS) are proposed in [23] and [24], respectively. However, these methods do not utilize the inherent potential of MIMO transmission and they still suffer from high computational complexity.

This paper is motivated by the need to efficient PAPR reduction approaches for MIMO-OFDM systems. The contributions of the paper are as follows: (1) A PAPR reduction method based on spatial and temporal shifting is proposed for MIMO-OFDM systems; (2) A PAPR

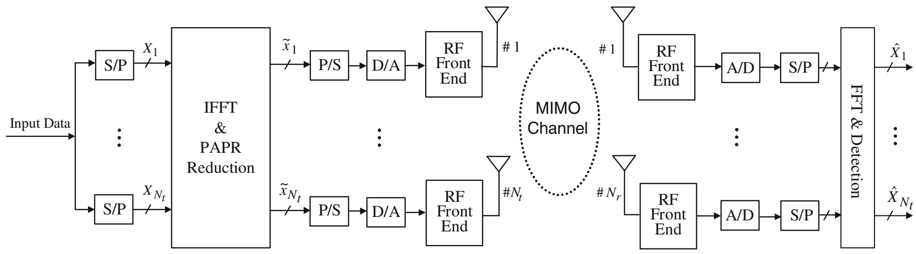

Fig. 1 MIMO-OFDM system model

reduction method is developed for MIMO-OFDM systems using spatial and temporal interleaving. The main idea of the proposed methods is that the OFDM block of each transmit antenna is divided into several subblocks. Then, the order of subblocks across all transmit antennas as well as the order of elements of subblocks are changed by circular shifting or interleaving operations. Therefore, the phases of orthogonal waveforms at each transmit antenna are changed and the PAPR is reduced effectively.

This paper is organized as follows: in Sect. 2, system model is introduced and PAPR is formulated. Sections 3 and 4 introduce the conventional PAPR reduction methods for SISO and MIMO-OFDM systems, respectively. New PAPR reduction methods for MIMO-OFDM systems are proposed in Sect. 5. The simulation results are presented in Sect. 6. Finally, conclusions are drawn in Sect. 7.

Notation: Bold lower (upper) case letters denote time domain (frequency domain) vectors. Operators $\mathcal{E}\{\cdot\}$ and IFFT$\{\cdot\}$ stand for expectation and inverse fast Fourier transform (IFFT) of order N, respectively. The mth subblock at the tth transmit antenna in the time domain is $\mathbf{x}_t^{(m)}$. The nth element of subblock $\mathbf{x}_t^{(m)}$ is denoted by $x_{t,n}^{(m)}$. The c times circular shift of vector $\mathbf{x}$ is obtained by function circ$(\mathbf{x}, c)$. Also, $\mathbf{x}_t^{(m,c)}$ represents the c times circular shift of the mth subblock at the tth transmit antenna. The optimized vector for mth subblock at the tth transmit antenna is denoted by $\tilde{\mathbf{x}}_t^{(m)}$. The smallest (biggest) integer higher (lower) than the constant a is $\lceil a \rceil$ ($\lfloor a \rfloor$). Function mod(x, y) returns the remainder on dividing x by y with the result having the same sign as x.

2 System Model and PAPR Formulation

A spatial multiplexing scheme with N_t transmit and N_r receive antennas is considered as in Fig. 1. The input data stream is partitioned into parallel substreams and each substream is mapped into complex quadrature phase-shift keying (QPSK) symbols and arranged into OFDM blocks of length N. Each block is transformed to the time domain by the IFFT operation. The resulting blocks are then converted to continuous time signals and transmitted over the MIMO channel.

If the OFDM block at the tth transmit antenna is $\mathbf{X}_t = [X_{t,0}, X_{t,1}, \ldots, X_{t,N-1}]^T$, the resulting continuous time signal is expressed as

$$x_t(t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_{t,k} e^{j2\pi kt/T} \quad , 0 \leq t \leq T \tag{1}$$

where T is the time duration of an OFDM block. Corresponding to $x_t(t)$, a discrete time signal $\mathbf{x}_t = [x_{t,0}, x_{t,1}, \ldots, x_{t,N-1}]^T$ is considered where $x_{t,n} = x_t(\frac{nT}{N})$. The PAPR of the transmitted OFDM signal in one symbol period T for the tth transmit antenna is defined as [25]

$$\mathrm{PAPR}_t = 10\log_{10}\frac{\max_{n=0,\ldots,N-1}\{|x_{t,n}|^2\}}{\mathcal{E}\{|x_{t,n}|^2\}} \tag{2}$$

The discrete time signal $\mathbf{x}_t$ contains only N samples of the continuous time signal $x_t(t)$. Therefore, the actual PAPR of $x_t(t)$ might be higher than that of $\mathbf{x}_t$. To obtain a more accurate approximation of the actual PAPR, an oversampling factor L is considered, where L is a positive integer. The oversampled discrete time signal is the LN-point IFFT of OFDM block with $(L-1)N$ zero padding. It is shown in [25] that $L = 4$ is sufficient to achieve accurate PAPR approximation. The PAPR of L-times oversampled discrete time signal for the tth transmit antenna can be expressed as

$$\mathrm{PAPR}_{t,\,over} = 10\log_{10}\frac{\max_{n=0,\ldots,LN-1}\{|x_{t,n}|^2\}}{\mathcal{E}\{|x_{t,n}|^2\}} \tag{3}$$

If the frequency domain symbols $\{X_{t,\,k}\}_{k=0,\ldots,\,N-1}$ are assumed statistically independent, based on the central limit theorem, the distribution of time domain samples $x_{t,n}$, can be assumed to be Gaussian. This leads to a high PAPR at the tth transmit antenna [23]. When multiple transmit antennas are used, the worst-case peak-to-average power over all transmit antennas is considered. Therefore, the PAPR of the MIMO-OFDM system can be defined as

$$\mathrm{PAPR}_{\mathrm{MIMO}} = \max_{t=1,\ldots,N_t} \mathrm{PAPR}_t \tag{4}$$

To evaluate the performance of different PAPR reduction methods, the complementary cumulative distribution function (CCDF) of the PAPR is usually considered. The CCDF shows the probability that the PAPR of a system exceeds a certain threshold PAPR_0. Assuming Gaussian distributed time domain samples $x_{t,n}$, the CCDF of the PAPR for SISO-OFDM system is expressed as [23]

$$\mathcal{P}\left(\mathrm{PAPR}_{\mathrm{SISO}} > \mathrm{PAPR}_0\right) = 1 - \left(1 - e^{-\mathrm{PAPR}_0}\right)^N \tag{5}$$

In case of MIMO-OFDM systems, there are NN_t discrete time samples. Hence, the CCDF is obtained as [23]

$$\mathcal{P}\left(\mathrm{PAPR}_{\mathrm{MIMO}} > \mathrm{PAPR}_0\right) = 1 - \left(1 - e^{-\mathrm{PAPR}_0}\right)^{NN_t} \tag{6}$$

Equation (6) shows that the PAPR problem becomes more serious as the number of transmit antennas is increased. So, the need to an efficient PAPR reduction method for the MIMO case is evident.

3 PAPR Reduction for SISO-OFDM

In this section, two existing methods for reduction of PAPR in SISO-OFDM systems are explained. First is partial transmit sequence (PTS) method with good PAPR reduction performance, but it suffers from high computational complexity. Second method is an improved version of PTS in which, the computational complexity is reduced without any performance degradation. In these methods, each frequency domain OFDM block $\mathbf{X}$ is partitioned into M disjoint subblocks of length N. In each subblock, the subcarriers that are related to other

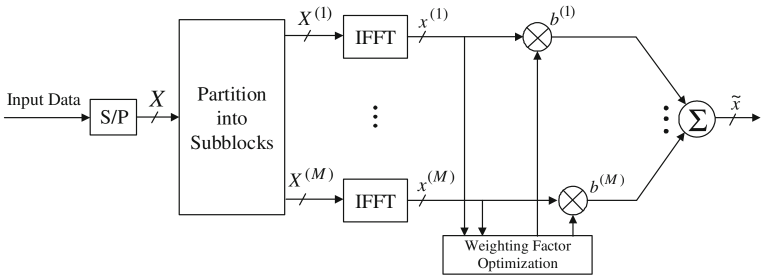

Fig. 2 The block diagram of PTS method

subblocks are set to zero. Therefore, the frequency domain OFDM block can be obtained by adding subblocks $\mathbf{X}^{(m)} = [X_0^{(m)}, X_1^{(m)}, \ldots, X_{N-1}^{(m)}]^T$, for $m = 1, \ldots, M$

$$\mathbf{X} = \sum_{m=1}^{M} \mathbf{X}^{(m)} \tag{7}$$

Using M IFFT operations, the partitioned subblocks $\{\mathbf{X}^{(m)}\}_{m=1,\ldots,M}$ are transformed to time domain subblocks $\{\mathbf{x}^{(m)}\}_{m=1,\ldots,M}$ which are called partial sequences. The Eq. (7) can be written in the time domain as

$$\mathbf{x} = \sum_{m=1}^{M} \mathbf{x}^{(m)} = \sum_{m=1}^{M} \mathrm{IFFT}\{\mathbf{X}^{(m)}\} \tag{8}$$

The common approach of the introduced methods in this section is to generate new candidate sequences $\tilde{\mathbf{x}}$ from the original data sequence $\mathbf{x}$. Then, the candidate sequence with the lowest PAPR is chosen for transmission.

3.1 Partial Transmit Sequence (PTS)

In PTS method [15], the partial sequences are independently rotated by phase factors $\{b^{(m)}\}_{m=1,\ldots,M}$ and combined together to generate a set of candidate sequence

$$\tilde{\mathbf{x}} = \sum_{m=1}^{M} b^{(m)} \mathbf{x}^{(m)} \tag{9}$$

where $b^{(m)} = \exp(j\varphi^{(m)})$, and $\varphi^{(m)} \in [0, 2\pi)$. The PTS method is shown in Fig. 2. The phase factors $b^{(m)}$ are usually limited to a W element set, i.e., $b^{(m)} \in \{b_1, b_2, \ldots, b_W\}$. For instance, they can be chosen from $\{\pm 1\}$ $(W = 2)$ or $\{\pm 1, \pm j\}$ $(W = 4)$. Without any performance degradation, the phase factor of the first partial sequence can be set to one $(b^{(1)} = 1)$. Hence, $W^{(M-1)}$ candidate sequences could be generated. The candidate with the lowest PAPR is obtained by exhaustive search of all candidate sequences which induces high computational complexity. The optimized phase factors $\{b_{\text{opt}}^{(m)}\}_{m=1,\ldots,M}$ should be send to the receiver as the side information to recover the original OFDM block $\mathbf{X}$. So, PTS method requires to transmit $\lceil \log_2(W^{(M-1)}) \rceil$ side information bits.

3.2 Low Copmlexity PTS

The low complexity PTS method generates new candidate sequences by circularly shifting some of the partial sequences [26]. Considering c circular shifts for the mth partial sequence, the alternative partial sequence can be written as

$$\mathbf{x}^{(m,c)} = \text{circ}(\mathbf{x}^{(m)}, c) = [x_c^{(m)}, \ldots, x_{N-1}^{(m)}, x_0^{(m)}, \ldots, x_{c-1}^{(m)}] \tag{10}$$

The number of circular shifts is restricted to a set with C elements, i.e., $c \in \{c_1, \ldots, c_C\}$. When the first partial sequence is kept unchanged, there are $C^{(M-1)}$ candidate sequences. If the mth partial sequence is shifted c times, the new candidate sequence is obtained as

$$\widetilde{\mathbf{x}} = \mathbf{x} - (\mathbf{x}^{(m)} - \mathbf{x}^{(m,c)}) \tag{11}$$

All candidate sequences corresponding to c circular shifts can be generated using (11) for $m = 2, 3, \ldots, M$ recursively. Finally, the candidate sequence with the lowest PAPR is transmitted. Compared to PTS, this method requires no multiplication. So, the computational complexity is reduced. Also, due to the IFFT property for different number of shifts, the detector can determine which shift number is operated on the mth subblock and the transmitter will not require to transmit any side information.

4 PAPR Reduction for MIMO-OFDM

In this section, three known approaches for reduction of PAPR in MIMO-OFDM systems are discussed. The first two methods which use PTS, consider each transmit antenna separately. The third method is also based on PTS but uses a cooperative scheme to jointly consider all transmit antennas and reduce the PAPR of MIMO system.

4.1 Ordinary PTS

Ordinary PTS (O-PTS) method is derived by applying PTS to each of the transmit antennas [23]. In O-PTS, MN_t IFFT operations are needed to obtain the partial sequences at all transmit antennas. The same number of subblocks (M) and allowed phase factors (W) are considered for all transmit antennas. The candidate sequence at the tth transmit antenna is expressed as

$$\widetilde{\mathbf{x}}_t = \sum_{m=1}^{M} b_t^{(m)} \mathbf{x}_t^{(m)} \tag{12}$$

where $\{b_t^{(m)}\}_{\substack{m=1,\ldots,M \\ t=1,\ldots,N_t}}$ are chosen from the set $\{\pm 1\}$ ($W = 2$) or $\{\pm 1, \pm j\}$ ($W = 4$). To determine the candidate with the lowest PAPR at each transmit antenna, $W^{(M-1)}$ different candidate sequences should be searched. So, this method induces very high computational complexity. Also, $N_t \lceil \log_2(W^{(M-1)}) \rceil$ side information bits should be transmitted.

4.2 Alternate PTS

Due to high computational complexity of O-PTS, alternate PTS (A-PTS) has been proposed [24]. In A-PTS, starting from the first subblock, every alternate subblock is kept unchanged and the phase factors are optimized only for even subblocks. Therefore, the computational complexity is reduced significantly at the cost of slight performance degradation. In this

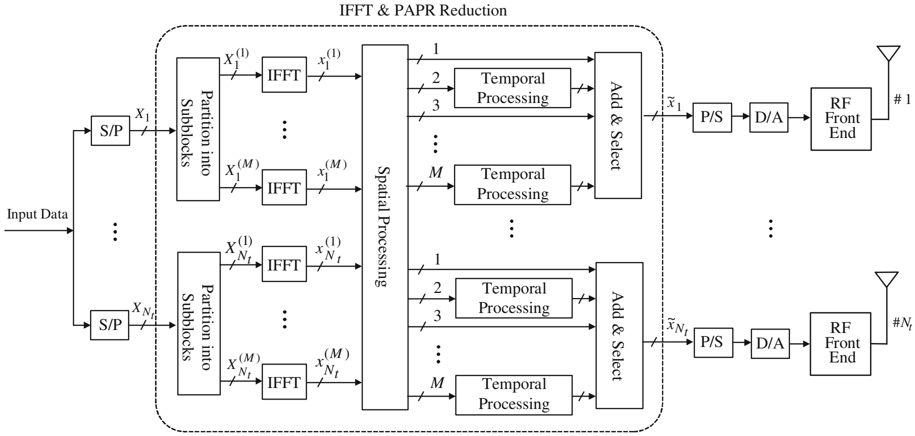

Fig. 3 The block diagram of proposed methods at the transmitter

method, $N_t W^{(M/2)}$ candidate sequences are generated and the number of side information bits is reduced to $N_t \lceil \log_2(W^{(M/2)}) \rceil$ compared to O-PTS.

4.3 Cooperative PTS

The aim of cooperative PTS (Co-PTS) method [27] is to compensate for the performance degradation in A-PTS approach. A-PTS reduces the computational complexity by decreasing the number of candidate sequences. To compensate for the performance degradation and increase the number of candidate sequences, Co-PTS uses spatial circular permutation for the same odd subblocks across all transmit antennas. Then, the PTS method is applied to even subblocks at each transmit antenna, separately.

This cooperative method generates $N_t^{(M/2)}[N_t W^{(M/2)}]$ candidate sequences for all transmit antennas. Thus, it requires $\lceil \log_2(N_t^{(M/2)}) \rceil + N_t \lceil \log_2(W^{(M/2)}) \rceil$ side information bits and only the number of complex additions is increased compared to A-PTS.

5 The Proposed Methods

In this section, two methods for reduction of PAPR in MIMO-OFDM systems are proposed. Figure 3 illustrates the general structure of the proposed methods. The OFDM block at each transmit antenna is partitioned into disjoint subblocks and these frequency domain subblocks are transformed to the time domain by IFFT operation. Then, spatial and temporal processing are applied to the time domain subblocks across all transmit antennas. Finally, a candidate sequence with the best PAPR statistics is chosen for each transmit antenna.

5.1 Spatial and Temporal Shifting (STS)

In this subsection, a method based on circular shifting is proposed. In this method, first, spatial circular shifting is used to generate different subblocks at each transmit antenna. Then, by applying temporal circular shifting to the subblocks of each transmit antenna, different candidate sequences are generated. Finally, by searching all candidate sequences at each transmit antenna, the sequence with the lowest PAPR is obtained for that transmit antenna. This method is called spatial and temporal shifting (STS).

Consider the MIMO-OFDM system in Fig. 1 with $N_t = 4$ transmit antennas and $M = 4$ subblocks at each transmit antenna. For spatial shifting, first, the odd subblocks of different antennas in the vertical direction are sequentially placed in a vector as $[\mathbf{x}_1^{(1)}, \mathbf{x}_1^{(3)}, \mathbf{x}_2^{(1)}, \mathbf{x}_2^{(3)}, \mathbf{x}_3^{(1)}, \mathbf{x}_3^{(3)}, \mathbf{x}_4^{(1)}, \mathbf{x}_4^{(3)}]$. Then, this vector is shifted to rearrange the odd subblocks. For instance, if two times circular shift are applied ($c = 2$), the new vector $[\mathbf{x}_4^{(1)}, \mathbf{x}_4^{(3)}, \mathbf{x}_1^{(1)}, \mathbf{x}_1^{(3)}, \mathbf{x}_2^{(1)}, \mathbf{x}_2^{(3)}, \mathbf{x}_3^{(1)}, \mathbf{x}_3^{(3)}]$ is obtained. Finally, the new odd subblocks of each antenna are picked up one-by-one from this vector and placed beside the corresponding even subblocks. In this example, new subblocks of the first and second antennas will be $[\mathbf{x}_4^{(1)}, \mathbf{x}_1^{(2)}, \mathbf{x}_4^{(3)}, \mathbf{x}_1^{(4)}]$ and $[\mathbf{x}_1^{(1)}, \mathbf{x}_2^{(2)}, \mathbf{x}_1^{(3)}, \mathbf{x}_2^{(4)}]$, respectively.

After spatial shifting, temporal shifting is performed. In this process, the elements of even subblocks at each transmit antenna are shifted separately. The odd subblocks are kept unchanged in the temporal processing. For $c = 32$ temporal shifts, the second even subblock of the first antenna becomes $[x_{32}^{(2)}, \ldots, x_{N-1}^{(2)}, x_0^{(2)}, \ldots, x_{31}^{(2)}]$. Combining the spatial and temporal shifting, results in a large number of candidate sequences at each transmit antenna. Among them, the sequence which has the lowest PAPR is selected for transmission. The maximum number of allowed spatial shifts for odd subblocks of all transmit antennas is equal to $N_t(M/2)$ and the number of allowed temporal shifts in each subblock is chosen from a set with C elements, i.e., $c \in \{c_1, \ldots, c_C\}$. STS method generates $[N_t(M/2)][N_t C^{(M/2)}]$ candidate sequences and requires $\lceil \log_2(N_t) + \log_2(M/2) \rceil + N_t \lceil \log_2(C^{(M/2)}) \rceil$ side information bits.

In the spatial processing, if circular shifting is separately applied to the even subblocks as well, the number of candidate sequences increases significantly and the PAPR reduction performance will be improved. Therefore, this scheme is called improved STS (I-STS). In this scheme, $[N_t(M/2)]^2[N_t C^{(M/2)}]$ candidate sequences are generated and $2\lceil \log_2(N_t) + \log_2(M/2) \rceil + N_t \lceil \log_2(C^{(M/2)}) \rceil$ side information bits are required.

5.2 Spatial and Temporal Interleaving (STI)

In this subsection, a method for PAPR reduction based on interleaving is proposed. First, spatial interleaving is applied to the subblocks of all transmit antennas. Then, the elements of even subblocks of each transmit antenna are rearranged by temporal interleaving to generate the set of candidate sequences for each antenna. Finally, by exhaustive search, the sequence with the lowest PAPR at each transmit antenna is found. This method is called spatial and temporal interleaving (STI).

Assuming a MIMO-OFDM system with $N_t = 4$ transmit antennas and $M = 4$ subblocks at each transmit antenna, the subblocks of all antennas are placed consecutively to obtain the following vector

$$[\mathbf{x}_1^{(1)}, \mathbf{x}_1^{(2)}, \mathbf{x}_1^{(3)}, \mathbf{x}_1^{(4)}, \ldots, \mathbf{x}_4^{(1)}, \mathbf{x}_4^{(2)}, \mathbf{x}_4^{(3)}, \mathbf{x}_4^{(4)}] \tag{13}$$

For spatial interleaving, this vector is passed through an interleaver function. Here, the interleaver function of [28] is considered as

$$k(i) = \mathrm{mod}\left(N_c * \left(i + s_{\mathrm{spat}} * \left\lfloor \frac{N_c * i}{D} \right\rfloor\right) - (D-1) * \left\lfloor \frac{N_c * i}{D} \right\rfloor, D\right), \quad i = 0, 1, \ldots, (D-1) \tag{14}$$

$x_1^{(1)}$	$x_1^{(2)}$	$x_1^{(3)}$	$x_1^{(4)}$	$x_2^{(1)}$	$x_2^{(2)}$	$x_2^{(3)}$	$x_2^{(4)}$	$x_3^{(1)}$	$x_3^{(2)}$	$x_3^{(3)}$	$x_3^{(4)}$	$x_4^{(1)}$	$x_4^{(2)}$	$x_4^{(3)}$	$x_4^{(4)}$

(a)

row 0:	$x_1^{(1)}$	$x_1^{(2)}$	$x_1^{(3)}$	$x_1^{(4)}$
row 1:	$x_2^{(1)}$	$x_2^{(2)}$	$x_2^{(3)}$	$x_2^{(4)}$
row 2:	$x_3^{(1)}$	$x_3^{(2)}$	$x_3^{(3)}$	$x_3^{(4)}$
row 3:	$x_4^{(1)}$	$x_4^{(2)}$	$x_4^{(3)}$	$x_4^{(4)}$

(b)

$x_1^{(1)}$	$x_2^{(1)}$	$x_3^{(1)}$	$x_4^{(1)}$	$x_3^{(2)}$	$x_4^{(2)}$	$x_1^{(2)}$	$x_2^{(2)}$	$x_1^{(3)}$	$x_2^{(3)}$	$x_3^{(3)}$	$x_4^{(3)}$	$x_3^{(4)}$	$x_4^{(4)}$	$x_1^{(4)}$	$x_2^{(4)}$

(c)

Fig. 4 The concept of interleaving scheme ($D = 16$, $s_{\text{spat}} = 2$, and $N_c = 4$) **a** Non-interleaved subblocks vector, **b** Interleaver matrix, **c** Interleaved subblocks vector

where $k(i)$ denotes the index of an element in the non-interleaved vector (Eq. 13) which is placed in ith position of the interleaved vector. D is the total number of subblocks and other parameters will be defined later.

The concept of interleaving has been shown in Fig. 4. The elements of non-interleaved vector (Fig. 4a) are written into interleaver matrix row-wise, starting in row zero (going from left to right). This matrix has N_c columns. Then, subblocks are read out of the matrix column-wise, starting at a given element in the first column, going down the column, and wrapping around to the top (if necessary). After reading each column, a shift of s row positions is applied. Figure 4b demonstrates the starting element in each column of the interleaver matrix with $s = 2$. The output vector is shown in Fig. 4c. For spatial interleaving, $D = MN_t$, $N_c = M$, and $s = s_{\text{spat}} \in \{0, 1, \ldots, (N_t - 1)\}$.

After spatial interleaving, the odd and even subblocks are substituted with the subblocks in the first and second halves of the interleaved vector, respectively. For example, the output vector in Fig. 4c is assigned to different transmit antennas as follows

$$\begin{aligned}
&\text{Antenna 1: } [\mathbf{x}_1^{(1)}, \mathbf{x}_1^{(3)}, \mathbf{x}_2^{(1)}, \mathbf{x}_2^{(3)}]\\
&\text{Antenna 2: } [\mathbf{x}_3^{(1)}, \mathbf{x}_3^{(3)}, \mathbf{x}_4^{(1)}, \mathbf{x}_4^{(3)}]\\
&\text{Antenna 3: } [\mathbf{x}_3^{(2)}, \mathbf{x}_3^{(4)}, \mathbf{x}_4^{(2)}, \mathbf{x}_4^{(4)}]\\
&\text{Antenna 4: } [\mathbf{x}_1^{(2)}, \mathbf{x}_1^{(4)}, \mathbf{x}_2^{(2)}, \mathbf{x}_2^{(4)}]
\end{aligned}$$

Now, temporal interleaving is applied to each transmit antenna separately. To do this, the even subblocks of each transmit antenna is passed through the interleaver function, while, the odd subblocks are kept unchanged. The interleaver function of [28] can be used here, too. However, in this case, $D = N$, $N_c = 8$, and $s = s_{\text{temp}} = \frac{N}{N_c M}(2i - 1)$, $i = 1, \ldots, S$, where S is the number of allowed shift constants in temporal interleaving. By changing the shift constants s_{spat} and s_{temp} in the spatial and temporal processing, the set of candidate sequences for each transmit antenna is obtained. STI method generates $N_t^2 [N_t S^{(M/2)}]$ candidate sequences and requires $2\lceil \log_2(N_t) \rceil + N_t \lceil \log_2(S^{(M/2)}) \rceil$ side information bits.

In the temporal processing, in order to increase the number of candidate sequences and achieve further PAPR reduction performance improvement, the interleaver function can be applied to both odd and even subblocks. This scheme which is called improved STI (I-STI), generates $N_t^2 [N_t S^{(M-1)}]$ candidate sequences and requires $2\lceil \log_2(N_t)\rceil + N_t \lceil \log_2(S^{(M-1)})\rceil$ side information bits.

5.3 Complexity Comparison

In this subsection, the computational complexity of our proposed approaches are compared to those of O-PTS, A-PTS, and Co-PTS. The computational complexity has three main parts: the complexity of IFFT operation ($\mathrm{C_{IFFT}}$), the complexity of optimization process ($\mathrm{C_{op}}$), and the complexity of PAPR calculation for each sequence ($\mathrm{C_{cal}}$). Therefore, the total complexity per transmit antenna is derived as

$$\mathrm{C_{total}} = M \cdot \mathrm{C_{IFFT}} + V \cdot (\mathrm{C_{op}} + \mathrm{C_{cal}}) \tag{15}$$

where M and V are the number of subblocks and candidate sequences at each transmit antenna, respectively. The complexity of IFFT operation is equal to $(NL/2)log_2(NL)$ complex multiplications and $(NL)log_2(NL)$ complex additions. Each complex multiplication is equal to four real multiplications and two real additions and each complex addition is equal to two real additions. The proposed methods require no multiplication. So, the complexity of their optimization processes ($\mathrm{C_{op}}$) only includes complex additions. To calculate the PAPR for each candidate sequence, only the peak power has to be calculated which requires $2NL$ real multiplications and NL real additions.

In STS method, $N_t(M/2)$ different candidate sequences at each transmit antenna are generated by spatial shifting. The temporal processing also generates $C^{(M/2)}$ candidate sequences at each transmit antenna which requires $2LN$ complex additions for each sequence. The number of candidate sequences in the spatial processing is $[N_t(M/2)]^2$ for I-STS approach. Also, the temporal processing of I-STS is similar to that of STS.

In STI method, the spatial processing generates N_t^2 candidate sequences at each transmit antenna. Also, the temporal processing generates $S^{(M/2)}$ candidate sequences at each transmit antenna which requires $2LN$ complex additions. The spatial processing parts of I-STI and STI are the same and the temporal processing of I-STI generates $S^{(M-1)}$ candidate sequences at each transmit antenna. Table 1 compares the total number of real multiplications/additions of different approaches for each transmit antenna.

The computational complexity reduction ratio (CCRR) of our proposed methods over Co-PTS can be defined as [26]

$$\mathrm{CCRR} = \left(1 - \frac{\text{complexity of the proposed method}}{\text{complexity of Co-PTS}}\right) \times 100\% \tag{16}$$

The multiplicative and additive CCRRs of the proposed methods over Co-PTS for typical values of N_t, N, and M is given in Table 2. It can be seen that in most cases the proposed STS and STI methods reduce the computational complexity dramatically compared to Co-PTS. For instance, using STS method, $N_t = 2$, and $M = 4, 8, 16$, 0–93.75 % reduction in the number of multiplications and 23.53–98.99 % reduction in the number of additions can be obtained. When $N_t = 4$ and $M = 4, 8, 16$, STS results in 44.44–99.95 and 61.02–99.96 % reduction improvement in the number of multiplications and additions, respectively. It is also seen that the CCRRs of STS and STI approaches are improved with increasing the number of subblocks.

Table 1 Complexity of the proposed methods compared to existing ones, $\alpha = 2NL \cdot log_2(NL)$ and $\beta = 3NL \cdot log_2(NL)$

Method	Real multiplications	Real additions
O-PTS	$M\alpha + W^{(M-1)} \cdot 2NL$	$M\beta + W^{(M-1)} \cdot (2NL(M-1) + NL)$
A-PTS	$M\alpha + W^{(M/2)} \cdot 2NL$	$M\beta + W^{(M/2)} \cdot (2NL(M-1) + NL)$
Co-PTS	$M\alpha + N_t^{(M/2)} \cdot W^{(M/2)} \cdot (2NL)$	$M\beta + N_t^{(M/2)} \cdot W^{(M/2)} \cdot (2NL(M-1) + NL)$
STS	$M\alpha + [N_t(M/2)] \cdot C^{(M/2)} \cdot (2NL)$	$M\beta + [N_t(M/2)] \cdot C^{(M/2)} \cdot (4NL + NL)$
I-STS	$M\alpha + [N_t(M/2)]^2 \cdot C^{(M/2)} \cdot (2NL)$	$M\beta + [N_t(M/2)]^2 \cdot C^{(M/2)} \cdot (4NL + NL)$
STI	$M\alpha + N_t^2 \cdot S^{(M/2)} \cdot (2NL)$	$M\beta + N_t^2 \cdot S^{(M/2)} \cdot (4NL + NL)$
I-STI	$M\alpha + N_t^2 \cdot S^{(M-1)} \cdot (2NL)$	$M\beta + N_t^2 \cdot S^{(M-1)} \cdot (4NL + NL)$

Table 2 CCRRs(%) of the proposed methods over Co-PTS for $W = C = S = 4$

Method	$M = 4$	$M = 8$	$M = 16$	$M = 4$	$M = 8$	$M = 16$
	Multiplicative CCRR, $N_t = 2$			*Multiplicative CCRR*, $N_t = 4$		
STS	0	49.23	93.75	44.44	93.66	99.95
STI	0	73.85	98.44	0	93.66	99.98
	Additive CCRR, $N_t = 2$			*Additive CCRR*, $N_t = 4$		
STS	23.53	83.07	98.99	61.02	97.89	99.96
STI	23.53	91.38	99.75	27.12	97.89	99.98

6 Simulations

In the following, simulation results are presented to demonstrate the effectiveness of the proposed methods. In addition, other approaches such as O-PTS [23], A-PTS [24], and Co-PTS [27] are simulated. A MIMO-OFDM system with $N_t = 4$ transmit antennas is considered. QPSK signaling and OFDM modulation with $N = 128$ subcarriers is used. The oversampling factor is considered to be $L = 2$, since the largest PAPR change occurs from $L = 1$ to $L = 2$ [29]. The CCDF of the PAPR is used to measure the performance of different PAPR reduction methods. To obtain the CCDF graphs, 10^5 independent OFDM blocks are randomly generated at each transmit antenna. Also, each OFDM block is partitioned into $M = 4$ subblocks. The set of allowed phase factors in O-PTS, A-PTS, and Co-PTS methods, is considered to be $\{\pm 1\}$. In STS method, the number of temporal shifts are $c = 0$ or 32. For temporal interleaving, the shift constant of the interleaver matrix, s_{spat} is set to be 4, or 12. In all simulation results, the CCDF of the original MIMO-OFDM system without PAPR reduction is also depicted.

Figure 5 shows the CCDF of PAPR for the proposed STS and I-STS methods. It is observed that both of these methods have better PAPR reduction performance than that of Co-PTS, O-PTS, and A-PTS. When CCDF=0.1 %, the performance improvement of STS over Co-PTS, O-PTS, and A-PTS are 0.84, 1.47, and 2.01 dB, respectively. In the case of I-STS, 1.79, 2.42, and 2.96 dB performance improvements are obtained over Co-PTS, O-PTS, and A-PTS, respectively. It is seen that I-STS has much better performance since it generates more candidate sequences with the same parameters.

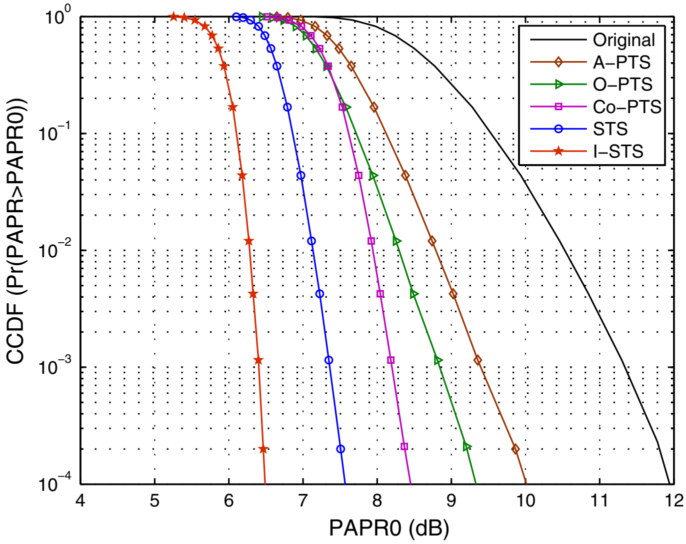

Fig. 5 PAPR reduction performance of proposed STS and I-STS methods

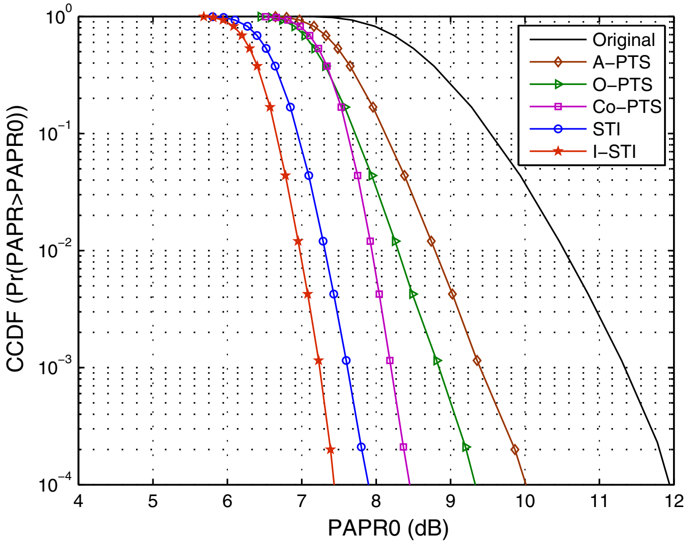

Fig. 6 PAPR reduction performance of proposed STI and I-STI methods

Figure 6 depicts the performance of STI and I-STI methods. It can be seen that these methods achieve better PAPR reduction performance compared to conventional approaches. When CCDF=0.1 %, STI provides 0.59, 1.22, and 1.76 dB performance improvements over Co-PTS, O-PTS, and A-PTS, respectively. With the same parameters, the performance of I-STI improves about 0.96, 1.59, and 2.13 dB over Co-PTS, O-PTS, and A-PTS, respectively.

Figure 7 compares the PAPR reduction performance of all proposed methods with each other. When CCDF = 0.1 %, it is seen that I-STS provides 0.84, 0.96, and 1.21 dB perfor-

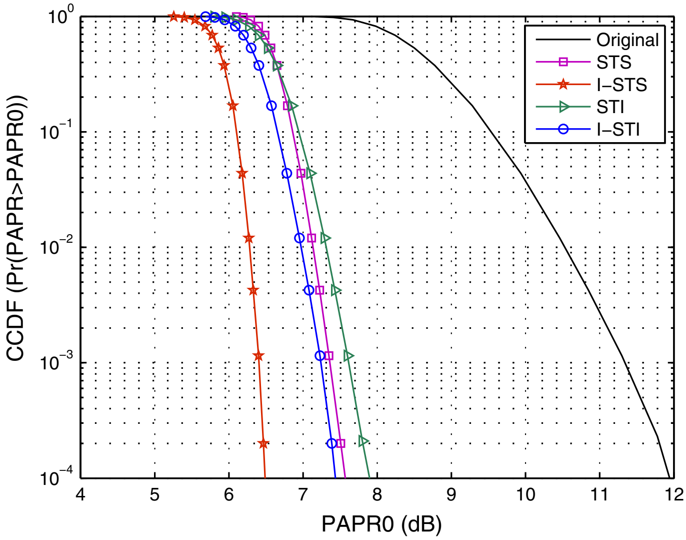

Fig. 7 PAPR reduction performance of all proposed methods

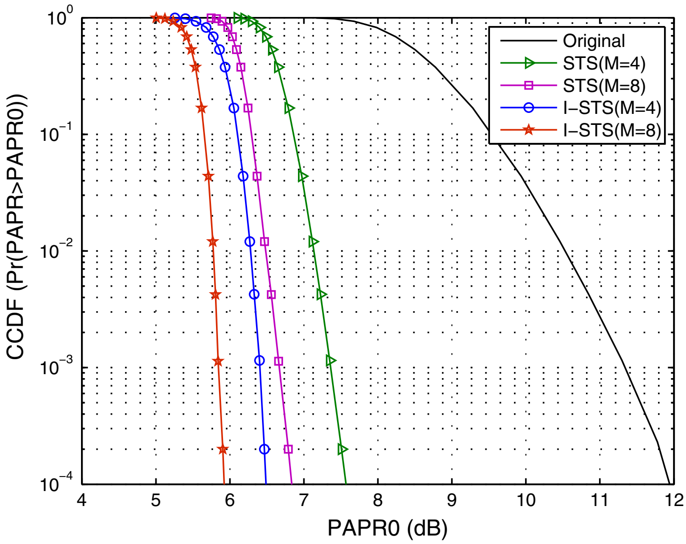

Fig. 8 PAPR reduction performance of proposed STS and I-STS methods for different number of subblocks

mance improvements over I-STI, STS, and STI, respectively. Therefore, I-STS offers the best performance among the proposed methods.

Figure 8 illustrates the CCDF of STS and I-STS methods for different number of subblocks. When CCDF $= 0.1\,\%$, increasing the subblocks from $M = 4$ to $M = 8$ results in 0.69 and 0.55 dB performance improvements are obtained for STS and I-STS, respectively. The same results are obtained for STI and I-STI methods.

Figure 9 shows the CCDF of Co-PTS and STS methods for 16-QAM and 64-QAM constellations. Comparing Figs. 5 and 9 reveales that increasing the constellation size, slightly

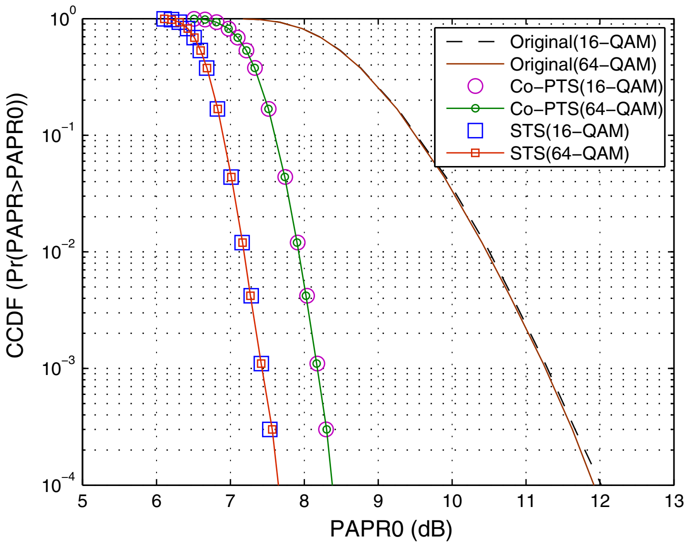

Fig. 9 PAPR reduction performance of Co-PTS and STS methods for various QAM constellations

changes the PAPR behavior. In fact, if the number of subcarriers is considered sufficiently large, the output signal becomes Gaussian following the centeral limit theorem and the PAPR is no longer dependent on mapped data [30].

7 Conclusion

In this paper, a class of PAPR reduction methods for MIMO-OFDM systems is introduced. These methods utilize the inherent potential of MIMO transmission using spatial/ temporal processing. In the spatial processing, all the transmit antennas are considered jointly, while the temporal processing reduces the PAPR of each transmit antenna separately. In most cases, the proposed STS and STI methods induce lower complexity than that of conventional schemes. Furthermore, simulation results show that the proposed methods substantially outperform the Co-PTS, O-PTS, and A-PTS approaches. This performance improvement is because: a) spatial and temporal processing increase the number of candidate sequences at each transmit antenna; b) spatial processing reduces the correlation between subblocks of each transmit antenna; and c) temporal processing at each transmit antenna decreases the correlation between samples of each OFDM block. It is also seen that I-STS method offers better PAPR reduction performance compared to other proposed STS, STI, and I-STI methods.

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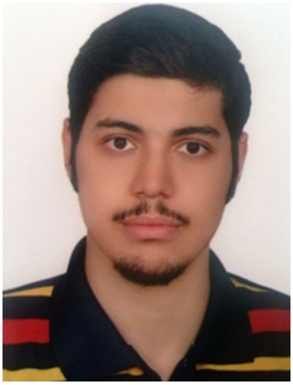

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