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Selected Mapping Algorithm for PAPR Reduction of Space-Frequency Coded OFDM Systems Without Side Information

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***Abstract*—Selected mapping (SLM) is a well-known technique for peak-to-average-power ratio (PAPR) reduction of orthogonal frequency-division multiplexing (OFDM) systems. In this technique, different representations of OFDM symbols are generated by rotation of the original OFDM frame by different phase sequences, and the signal with minimum PAPR is selected and transmitted. To compensate for the effect of the phase rotation at the receiver, it is necessary to transmit the index of the selected phase sequence as side information (SI). In this paper, an SLM technique is introduced for the PAPR reduction of space-frequency-block-coded OFDM systems with Alamouti coding scheme. Additionally, a suboptimum detection method that does not need SI is introduced at the receiver side. Simulation results show that the proposed SLM method effectively reduces the PAPR, and the detection method has performance very close to the case where the correct index of the phase sequence is available at the receiver side.**

***Index Terms*—Orthogonal frequency-division multiplexing (OFDM), peak-to-average-power ratio (PAPR), selected mapping (SLM), space frequency block coded (SFBC), spatial diversity.**

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I. Introduction

Orthogonal frequency-division multiplexing (OFDM) is a well-known technique for transmission of high rate data over broadband frequency-selective channels [1]. One of the drawbacks of OFDM systems is high peak-to-average-power ratio (PAPR), which leads to the saturation of the high-power amplifier. Thus, a high-dynamic-range amplifier is needed, which increases the cost of the system. The frequency-domain symbols of an OFDM frame is denoted by $\mathbf{X} = [X(0), X(1), \ldots, X(N_c-1)]^T$, where N_c is the number of subcarriers. It is assumed that $X(k) \in \mathcal{C}$, where $\mathcal{C}$ is the set of constellation points. The vector $\mathbf{x} = [x(0), x(1), \ldots, x(N-1)]^T$ contains the time-domain samples of the complex baseband OFDM signal as given by

$$x(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N_c-1} X(k) e^{j\frac{2\pi nk}{N}} \tag{1}$$

where $j=\sqrt{-1}$, and N/N_c is the oversampling ratio. It is clear that $\mathbf{x} = \text{IFFT}_N\{\mathbf{X}\}$, where $\text{IFFT}_N\{\}$ is the N-point inverse fast Fourier transform (IFFT) operation. The PAPR of the OFDM frame is defined by

$$\text{PAPR}(\mathbf{x}) = \frac{\max_n \left\{|x(n)|^2\right\}}{E\left\{|x(n)|^2\right\}} \tag{2}$$

where $E\{.\}$ is the mathematical expectation. According to (1), the time-domain samples are the sum of N_c independent terms. When N_c is large, based on the central limit theorem, the time-domain samples have a Gaussian distribution; thus, they may have large amplitudes [2]. To overcome this problem, some algorithms have been proposed, which reduce the PAPR of the baseband OFDM signal [3]–[16]. Some of these methods need side information (SI) to be transmitted to the receiver, such as partial transmit sequence [3], [4] and selected mapping (SLM) [5]–[7]. Some other methods do not need SI, such as clipping and filtering [8], [9], tone reservation [10], [11], block coding [12], [13], and active constellation extension [14].

In the SLM method, D different representations of the OFDM frame are generated, and that with minimum PAPR is transmitted. If the vectors $[\phi^d(0), \phi^d(1), \ldots, \phi^d(N_c-1)]^T$, $d = 0, 1, \ldots, D-1$, are D random phase sequences with the length of N_c and $\mathbf{b}^d = [e^{j\phi^d(0)}, e^{j\phi^d(1)}, \ldots, e^{j\phi^d(N_c-1)}]^T$, then D representations of the signal $\mathbf{x}$ are

$$\mathbf{x}^d = \text{IFFT}_N\{\mathbf{X} \otimes \mathbf{b}^d\}, \qquad 0 \le d \le D-1 \tag{3}$$

where $\otimes$ is element-by-element production. The index of the optimum phase sequence is

$$\bar{d} = \underset{d\in\{0,1,\ldots,D-1\}}{\arg\min} \left\{\text{PAPR}\{\mathbf{x}^d\}\right\}. \tag{4}$$

To reduce the complexity of the application of different phase sequences, often, phases $\phi^d(k)$ are randomly chosen from $\{0, \pi\}$. This means that $b^d(k) \in \{\pm 1\}$, and it is enough to change the sign of the symbols $X(k)$ before IFFT operation. The signal $\mathbf{x}^{\bar{d}}$ is transmitted, and the index of selected phase sequence $\bar{d}$ is sent to the receiver as SI. If the SI is received with an error, the OFDM frame will be lost; thus, this information must be protected by coding. Several SLM methods have been proposed for single-antenna OFDM systems, which do not need explicit SI [15]–[17]. Some of these algorithms pay a penalty for the transmission power [15], [16]. The drawback of the method introduced in [17] is that the phase sequences must be chosen such that $b^d(k)X(k) \notin \mathcal{C}$. This leads to the increase in the complexity of the transmitter, because the coefficients $b^d(k)$ can no longer be chosen from the set $\{\pm 1\}$.

Using several transmitter antennas, one can improve the data rate or bit error rate (BER) of wireless systems. In spatial multiplexing systems, independent symbols are transmitted from several antennas, and this leads to the increase in data rate. In [18], a simplified SLM method has been introduced for PAPR reduction of spatially multiplexed OFDM systems. If spatial diversity techniques are used in wireless systems with several transmitter antennas, the BER can be reduced. The space-time codes to achieve the full transmission diversity have been introduced in [19] and [20]. Through a combination of spatial diversity and OFDM techniques, a higher capacity can be achieved over broadband multipath fading wireless channels [21], [22]. Two possible combinations of spatial diversity and OFDM techniques are space-time-block-coded (STBC) OFDM and space-frequency-block-coded (SFBC) OFDM systems. Both combinations suffer from high-PAPR problem. In [23], the clipping and filtering method has been used for PAPR reduction in SFBC-OFDM systems with two transmitter antennas, and an iterative method has been proposed to compensate for the effect of clipping noise at the receiver. In [24], the polyphase interleaving and inversion (PII) method has been proposed for SFBC-OFDM systems with two transmitter antennas and Alamouti space frequency coding scheme. In addition, in [25], a low-complexity version of the SLM method has been proposed for PAPR reduction of SFBC-OFDM systems.

In this paper, it is shown that the simplified SLM technique can be applied to SFBC-OFDM systems with two transmitter antennas and Almouti coding scheme without changing the orthogonality of space frequency coding. In this method, the optimum phase sequence is applied to the OFDM frames of two antennas such that the SFBC structure remains constant. Then, it will be shown that, at the receiver side, the optimum phase sequence can be detected without SI. Detection of phase sequence index does not need any extra transmission power or extra computation complexity of the transmitter and only uses the intrinsic redundancy of the SFBC code. In this paper, the optimum receiver for blind detection of the phase sequence index $(\bar{d})$ is introduced, and a low-complexity suboptimum receiver is proposed. Simulation results show that the performance of the proposed blind detection method depends on the OFDM frame length, and its symbol error rate (SER) is very close to the case where the phase sequence index is available at the receiver side.

The remainder of this paper is organized as follows: In Section II, the system model of the SFBC-OFDM system with two transmitter antennas is introduced. In Section III, an SLM technique for PAPR reduction of this system is proposed. In Section IV, the optimum method for detection of the phase sequence index is introduced, and a suboptimum algorithm with lower complexity is proposed. Section V includes the simulation results that show the effect of the proposed method in PAPR reduction and SER.

II. Space-Frequency Block Coded Orthogonal Frequency-Division Multiplexing System Model

In SFBC-OFDM systems with two transmitter antennas, the frequency-domain vector transmitted from the pth antenna is denoted by $\mathbf{X}_p = [X_p(0), X_p(1), \ldots, X_p(N_c-1)]$. The vectors $\mathbf{X}_1$ and $\mathbf{X}_2$ can be generated from the original OFDM frame $\mathbf{X}$ as follows [21]:

$$\begin{pmatrix} X_1(2\nu) & X_1(2\nu+1) \\ X_2(2\nu) & X_2(2\nu+1) \end{pmatrix} = \underbrace{\begin{pmatrix} X(2\nu) & X(2\nu+1) \\ X^*(2\nu+1) & -X^*(2\nu). \end{pmatrix}}_{\mathbf{C}} \quad \nu = 0, 1, \ldots, N_c/2-1. \tag{5}$$

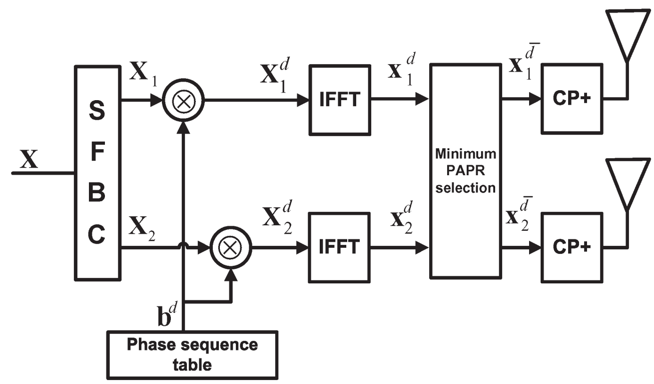

Fig. 1. Block diagram of the SFBC-OFDM transmitter with two transmitter antennas and the SLM method for PAPR reduction.

As shown in Fig. 1, the vectors $\mathbf{X}_1$ and $\mathbf{X}_2$ are passed through the IFFT operation to yield the time-domain samples $x_1(n)$ and $x_2(n)$, $0 \le n \le N-1$. It is noteworthy that the orthogonality of the space frequency matrix $\mathbf{C}$ in (5) leads to full diversity at the receiver side [20], i.e.,

$$\mathbf{C}\mathbf{C}^H = \left(|X(2\nu)|^2 + |X(2\nu+1)|^2\right)\mathbf{I}_2 \tag{6}$$

where $\mathbf{I}_n$ is the $n \times n$ identity matrix. The PAPR of the pth antenna is defined by

$$\text{PAPR}\{\mathbf{x}_p\} = \frac{\max_n \left\{|x_p(n)|^2\right\}}{E\left\{|x_p(n)|^2\right\}}, \qquad p = 1, 2 \tag{7}$$

where $E\{.\}$ is the mathematical expectation. The overall PAPR of the SFBC-OFDM system is defined by

$$\text{PAPR} = \max_{p \in \{1,2\}} \text{PAPR}\{\mathbf{x}_p\}. \tag{8}$$

III. Selected Mapping for Peak-to-Average Power Ratio Reduction of Space-Frequency Block Coded Orthogonal Frequency-Division Multiplexing Systems

Vectors $\mathbf{X}_1$ and $\mathbf{X}_2$ can be multiplied by D different phase sequences to yield the minimum PAPR representation, but the SFBC structure shown in (5) must remain constant, so that full diversity can be achieved. In [18], simplified SLM for PAPR reduction of spatial multiplexed OFDM has been proposed. In this scheme, the OFDM frames of the antennas are simultaneously modified with the same single-phase sequence. This leads to reduction in the number of bits that must be transmitted as SI. In this paper, this approach is used for SFBC-OFDM systems. It is shown that, if this method is used for the SFBC-OFDM system, then the intrinsic redundancy of space frequency coding can be used to detect the index of the phase sequence at the receiver side without SI. Based on this approach, D different representations of the signals $\mathbf{x}_1$ and $\mathbf{x}_2$ are generated as follows:

$$\begin{aligned} \mathbf{x}_1^d &= \text{IFFT}_N\{\mathbf{X}_1 \otimes \mathbf{b}^d\} \\ \mathbf{x}_2^d &= \text{IFFT}_N\{\mathbf{X}_2 \otimes \mathbf{b}^d\}, \qquad 0 \le d \le D-1. \end{aligned} \tag{9}$$

In [26], it has been shown that a simple and optimal choice for the phase sequences is a random selection of 0 and π with equal probabilities. In this case, the complexity of the phase rotation is very low, because $b^d(k) = e^{j0} = +1$ or $b^d(k) = e^{j\pi} = -1$; thus, the multiplication of the frequency-domain vectors by the vector $\mathbf{b}^d$ is via the sign change of some of the symbols. The pairs $[b^d(2\nu), b^d(2\nu+1)]$, $0 \le \nu < N_c/2$, $0 \le d < D-1$, can take the values $[+1, -1]$,

$[-1, +1]$, $[-1, -1]$, and $[+1, +1]$ with equal probabilities. After the multiplication of $\mathbf{X}_1$ and $\mathbf{X}_2$ by $\mathbf{b}^d$, the matrix $\mathbf{C}$ in (5) is changed to

$$\mathbf{C}^d = \begin{pmatrix} b^d(2\nu)X(2\nu) & b^d(2\nu+1)X(2\nu+1) \\ b^d(2\nu)X^*(2\nu+1) & -b^d(2\nu+1)X^*(2\nu). \end{pmatrix}. \quad (10)$$

It is noteworthy that matrix $\mathbf{C}^d$ is also orthogonal, i.e.,

$$\mathbf{C}^d(\mathbf{C}^d)^H = \left(|X(2\nu)|^2 + |X(2\nu+1)|^2\right)\mathbf{I}_2. \quad (11)$$

By keeping the space frequency code orthogonal, full diversity can be achieved at the receiver side. It is shown in the next section that it is possible to detect the index of the optimum phase sequence $(\bar{d})$ at the receiver without SI.

IV. Blind Detection of $\bar{d}$

At the receiver side, it is necessary to determine the index of the optimum phase sequence $\bar{d}$ to correctly detect the transmitted symbols $X(k), 0 \le k \le N_c - 1$. After removing the cyclic prefix and applying FFT, the received vector $\mathbf{Y} = [Y(0), Y(1), \ldots, Y(N_c-1)]^T$ can be described as follows:

$$Y(k) = H_1(k)X_1^{\bar{d}}(k) + H_2(k)X_2^{\bar{d}}(k) + V(k), \qquad 0 \le k \le N_c - 1 \quad (12)$$

where $H_p(k)$ and $V(k)$ are the baseband equivalent coefficient of the channel between the pth transmitter antenna and the receiver antenna and the noise component at the kth subcarrier, respectively. The complex baseband noise is assumed to be white zero-mean Gaussian. It is assumed that the channel coefficients are known at the receiver side. Our goal is to determine $\bar{d}$ from $\{Y(k)\}_{k=0}^{N_c-1}$. SFBC-OFDM systems are often used in the channels that are flat over several subcarriers. If the channel coefficients are assumed to be the same for two adjacent subcarriers $(H_p(2\nu) = H_p(2\nu+1), p = 1, 2, 0 \le \nu \le N_c/2 - 1)$, then, from (5), (9), and (12), it can be seen that

$$\begin{aligned} Y(2\nu) &= H_1(2\nu)b^{\bar{d}}(2\nu)X(2\nu) \\ &\quad + H_2(2\nu)b^{\bar{d}}(2\nu)X^*(2\nu+1) + V(2\nu) \\ Y(2\nu+1) &= H_1(2\nu)b^{\bar{d}}(2\nu+1)X(2\nu+1) \\ &\quad - H_2(2\nu)b^{\bar{d}}(2\nu+1)X^*(2\nu) + V(2\nu+1). \end{aligned} \quad (13)$$

The maximum-likelihood (ML) detection for $\bar{d}$ and $\mathbf{X}$ is (14), shown at the bottom of the page. The new variables $Z^d(2\nu)$, $Z^d(2\nu+1)$ and $f^d(\nu)$ are defined by

$$\begin{aligned} Z^d(2\nu) &= b^d(2\nu)X(2\nu) \\ Z^d(2\nu+1) &= b^d(2\nu)X(2\nu+1) \\ f^d(\nu) &= \frac{b^d(2\nu+1)}{b^d(2\nu)} = b^d(2\nu)b^d(2\nu+1). \end{aligned} \quad (15)$$

Based on these definitions, (13) can be rewritten as

$$\begin{aligned} Y(2\nu) &= H_1(2\nu)Z(2\nu) + H_2(2\nu)Z^*(2\nu+1) + V(2\nu) \\ Y(2\nu+1) &= H_1(2\nu)f^{\bar{d}}(\nu)Z(2\nu+1) - H_2(2\nu)f^{\bar{d}}(\nu)Z^*(2\nu) \\ &\quad + V(2\nu+1) \end{aligned} \quad (16)$$

and the ML detection can be modified as (17), shown at the bottom of the page.

The complexity of ML detection is on the order of DA^{N_c}, where A is the number of the constellation points [$A = 16$ for 16-quadrature amplitude modulation (QAM)]. To reduce the complexity of this detection method, a suboptimum algorithm is proposed in this paper. In the first step of the proposed method, the vectors $[Z(2\nu), Z(2\nu+1), f(2\nu)]$ are independently detected for different ν's. The detected vector $[\dot{Z}(2\nu), \dot{Z}(2\nu+1), \dot{f}(2\nu)]$ is given by

$$\left[\dot{Z}(2\nu), \dot{Z}(2\nu+1), \dot{f}(\nu)\right] = \underset{\substack{\hat{f}(\nu)\in\{\pm1\} \\ \hat{Z}(2\nu)\in\mathcal{C} \\ \hat{Z}(2\nu+1)\in\mathcal{C}}}{\arg\min}\ M_\nu \left(Y(2\nu), Y(2\nu+1), \hat{Z}(2\nu), \hat{Z}(2\nu+1), \hat{f}(\nu)\right) \quad (18)$$

where

$$\begin{aligned} M_\nu &\left(Y(2\nu), Y(2\nu+1), \hat{Z}(2\nu), \hat{Z}(2\nu+1), \hat{f}(\nu)\right) \\ &= \left|Y(2\nu) - H_1(2\nu)\hat{Z}(2\nu) - H_2(2\nu)\hat{Z}^*(2\nu+1)\right|^2 \\ &\quad + \left|Y(2\nu+1) - H_1(2\nu)\hat{f}(\nu)\hat{Z}(2\nu+1)\right. \\ &\quad \left. + H_2(2\nu)\hat{f}(\nu)\hat{Z}^*(2\nu)\right|^2. \end{aligned} \quad (19)$$

It can easily be seen that

$$\begin{aligned} M_\nu &\left(Y(2\nu), Y(2\nu+1), \hat{Z}(2\nu), \hat{Z}(2\nu+1), \hat{f}(\nu)\right) \\ &= \left(|H_1(2\nu)|^2 + |H_2(2\nu)|^2 - 1\right)\left|\hat{Z}(2\nu)\right|^2 \\ &\quad + \left|\hat{Z}(2\nu) - H_1(2\nu)Y^*(2\nu) + H_2^*(2\nu)\hat{f}(\nu)Y(2\nu+1)\right|^2 \\ &\quad + \left(|H_1(2\nu)|^2 + |H_2(2\nu)|^2 - 1\right)\left|\hat{Z}(2\nu+1)\right|^2 \\ &\quad + \left|\hat{Z}(2\nu+1) - H_2^*(2\nu)Y(2\nu) - H_1^*(2\nu)\hat{f}(\nu)Y(2\nu+1)\right|^2. \end{aligned} \quad (20)$$

$$\left[d^{(\mathrm{ML})}, \mathbf{X}^{(\mathrm{ML})}\right] = \underset{\substack{0\le\hat{d}<D \\ \hat{\mathbf{X}}\in\mathcal{C}^{N_c}}}{\arg\min} \sum_{\nu=0}^{\nu=N_c/2-1} \left\{\left|Y(2\nu) - H_1(2\nu)b^{\hat{d}}(2\nu)\hat{X}(2\nu) - H_2(2\nu)b^{\hat{d}}(2\nu)\hat{X}^*(2\nu+1)\right|^2 + \left|Y(2\nu+1) - H_1(2\nu)b^{\hat{d}}(2\nu+1)\hat{X}(2\nu+1) + H_2(2\nu)b^{\hat{d}}(2\nu+1)\hat{X}^*(2\nu)\right|^2\right\} \quad (14)$$

$$\left[d^{(\mathrm{ML})}, \mathbf{X}^{(\mathrm{ML})}\right] = \underset{\substack{0\le\hat{d}<D, \hat{\mathbf{Z}}\in\mathcal{C}^{N_c} \\ f^{\hat{d}}(\nu)=b^{\hat{d}}(2\nu)b^{\hat{d}}(2\nu+1)}}{\arg\min} \sum_{\nu=0}^{\nu=N_c/2-1} \left\{\left|Y(2\nu) - H_1(2\nu)\hat{Z}(2\nu) - H_2(2\nu)\hat{Z}^*(2\nu+1)\right|^2 + \left|Y(2\nu+1) - H_1(2\nu)f^{\hat{d}}(\nu)\hat{Z}(2\nu+1) + H_2(2\nu)f^{\hat{d}}(\nu)\hat{Z}^*(2\nu)\right|^2\right\} \quad (17)$$

The first two terms of the preceding equation depend on $\hat{Z}(2\nu)$ and $\hat{f}(\nu)$, and the last two terms depend on $\hat{Z}(2\nu+1)$ and $\hat{f}(\nu)$. To minimize M_ν in (20), $\dot{Z}(2\nu)$ and $\dot{Z}(2\nu+1)$ must be chosen as

$$\dot{Z}(2\nu) = Q\left(\underbrace{\frac{H_1(2\nu)Y^*(2\nu) - H_2^*(2\nu)\hat{f}(\nu)Y(2\nu+1)}{|H_1(2\nu)|^2 + |H_2(2\nu)|^2}}_{\zeta(2\nu)}\right)$$
$$\dot{Z}(2\nu+1) = Q\left(\underbrace{\frac{H_2^*(2\nu)Y(2\nu) + H_1^*(2\nu)\hat{f}(\nu)Y(2\nu+1)}{|H_1(2\nu)|^2 + |H_2(2\nu)|^2}}_{\zeta(2\nu+1)}\right) \tag{21}$$

where $Q(.)$ is the mapping to the nearest constellation point. The two equations in (21) depend on $\hat{f}(\nu)$; thus, $\dot{Z}^+(2\nu)$ and $\dot{Z}^+(2\nu+1)$ are defined by inserting $\hat{f}(\nu) = +1$, and $\dot{Z}^-(2\nu)$, and $\dot{Z}^-(2\nu+1)$ are defined by inserting $\hat{f}(\nu) = -1$ in (21). It can easily be seen that, for the minimization of M_ν, the parameter $\dot{f}(\nu)$ must be chosen as

$$\dot{f}(\nu) = \text{sign}\Big(M_\nu\big(Y(2\nu), Y(2\nu+1), \dot{Z}^-(2\nu), \dot{Z}^-(2\nu+1), -1\big) - M_\nu\big(Y(2\nu), Y(2\nu+1), \dot{Z}^+(2\nu), \dot{Z}^+(2\nu+1), +1\big)\Big) \tag{22}$$

where $sign(.)$ is the sign function. This is equivalent to

$$\dot{f}(\nu) = \text{sign}\Big(\left|\dot{Z}^-(2\nu) - \zeta^-(2\nu)\right|^2 + \left|\dot{Z}^-(2\nu+1) - \zeta^-(2\nu+1)\right|^2 - \left|\dot{Z}^+(2\nu) - \zeta^+(2\nu)\right|^2 - \left|\dot{Z}^+(2\nu+1) - \zeta^+(2\nu+1)\right|^2\Big) \tag{23}$$

where $\zeta(2\nu)$ and $\zeta(2\nu+1)$ are defined in (21). After the detection of $\dot{f}(\nu)$ for $\nu = 0, 1, \ldots, N_c/2 - 1$, the vector $\dot{\mathbf{f}} = [\dot{f}(0), \dot{f}(1), \ldots, \dot{f}(N_c/2-1)]^T$ is constructed. In the second step, this vector must be mapped to the closest sequence among $\mathbf{f}^d$, $0 \le d \le D-1$, i.e.,

$$\widetilde{d} = \underset{0 \le d \le D-1}{\arg\min}\ dist(\dot{\mathbf{f}}, \mathbf{f}^d) \tag{24}$$

where $dist(\mathbf{A}, \mathbf{B})$ is the Hamming distance between the vectors $\mathbf{A}$ and $\mathbf{B}$, and the vectors $\mathbf{f}^d$, $d = 0, 1, \ldots, D-1$ have been calculated and stored at the receiver as

$$\mathbf{f}^d = \left[f^d(0), f^d(1), \ldots, f^d\left(\frac{N_c}{2} - 1\right)\right]^T$$
$$f^d(\nu) = b^d(2\nu)b^d(2\nu+1). \tag{25}$$

Now, based on the definitions of (15), the detected symbols $\widetilde{X}(k)$ can be derived as (26), shown at the bottom of the page.

The steps of the proposed algorithm for the blind detection of $\bar{d}$ and the OFDM frame $\mathbf{X}$ can be summarized as

- Calculate $\dot{Z}^+(2\nu)$, $\dot{Z}^-(2\nu)$, $\dot{Z}^+(2\nu+1)$, and $\dot{Z}^-(2\nu+1)$ for $\nu = 0, 1, \ldots, N_c/2 - 1$ using (21).
- Evaluate $\dot{f}(\nu)$, $\nu = 0, 1, \ldots, N_c/2 - 1$, using (22), and construct the vector $\dot{\mathbf{f}}$.
- Determine $\widetilde{d}$ using (24).
- Determine the transmitted symbols $\{\widetilde{X}(k)\}_{k=0}^{k=N_c-1}$ using (26).

It is noteworthy that, in the proposed method, the complexities of FFT operation, channel estimation, and synchronization do not change, and only the SFBC decoding is done twice: once for the calculation of the $\dot{Z}^+(k)$ and the another for that of $\dot{Z}^-(k)$. To calculate FFT with length Nc, $3N_c\log_2(N_c)$ real additions (RAs) and $2N_c\log_2(N_c)$ real multiplications (RMs) are required [27]. Based on (21), the SFBC decoder for every ν requires 26 RAs and 32 RMs. Thus, the SFBC decoding for all the subcarriers needs $13N_c$ RAs and $16N_c$ RMs. Therefore, the additional complexity at the receiver can be calculated as

$$\begin{aligned}\text{Percentage of additional RAs}(\%) &= \frac{\text{SFBC RAs}}{\text{SFBC RAs} + \text{FFT RAs}} \\ &= \frac{13}{13 + 3\log_2(N_c)} \times 100\% \\ \text{Percentage of additional RMs}(\%) &= \frac{\text{SFBC RMs}}{\text{SFBC RMs} + \text{FFT RMs}} \\ &= \frac{16}{16 + 2\log_2(N_c)} \times 100\%.\end{aligned} \tag{27}$$

For example, for $N_c = 512$, 35% RAs and 50% RMs are added to the receiver. If the complexity of channel estimation and synchronization algorithms are taken into account, the denominators in (27) are increased, and the additional complexity percentage of the receiver is reduced.

V. Simulation Results

The performance of the proposed method has been evaluated for two different OFDM frame lengths $N_c = 128$ and $N_c = 512$. The symbols $X(k)$ are chosen from the 16-QAM constellations.

A. Performance in PAPR Reduction

The performance of the proposed method in PAPR reduction is evaluated by the complementary cumulative density function (CCDF) of the PAPR, which is defined as

$$\text{CCDF}(\text{PAPR}_0) = \Pr\{\text{PAPR} \ge \text{PAPR}_0\} \tag{28}$$

$$\left(\widetilde{X}(2\nu), \widetilde{X}(2\nu+1)\right) = \begin{cases} \left(\dot{Z}^+(2\nu), \dot{Z}^+(2\nu+1)\right) & \text{if } b^{\widetilde{d}}(2\nu) = +1, f^{\widetilde{d}}(\nu) = +1 \\ \left(-\dot{Z}^+(2\nu), -\dot{Z}^+(2\nu+1)\right) & \text{if } b^{\widetilde{d}}(2\nu) = -1, f^{\widetilde{d}}(\nu) = +1 \\ \left(\dot{Z}^-(2\nu), \dot{Z}^-(2\nu+1)\right) & \text{if } b^{\widetilde{d}}(2\nu) = +1, f^{\widetilde{d}}(\nu) = -1 \\ \left(-\dot{Z}^-(2\nu), -\dot{Z}^-(2\nu+1)\right) & \text{if } b^{\widetilde{d}}(2\nu) = -1, f^{\widetilde{d}}(\nu) = -1 \end{cases} \tag{26}$$

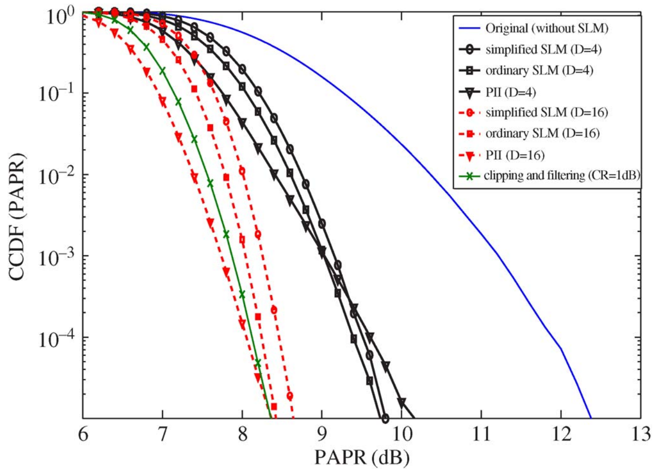

Fig. 2. PAPR reduction performance of the ordinary SLM, simplified SLM, PII, and clipping and filtering methods for the SFBC-OFDM system with two transmitter antennas and $N_c = 128$ for different values of D.

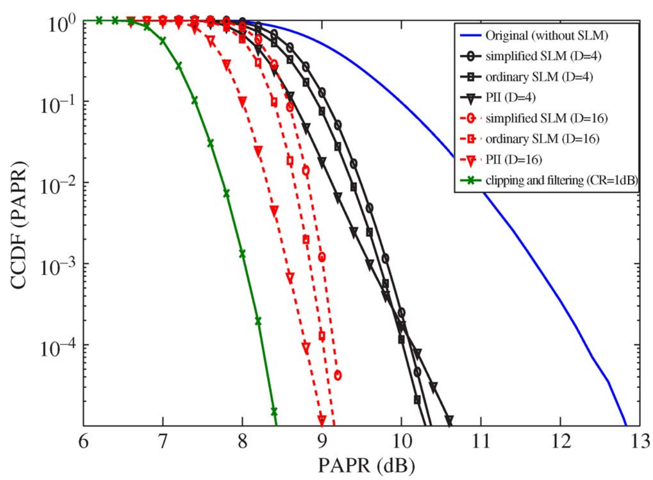

Fig. 3. PAPR reduction performance of the ordinary SLM, simplified SLM, PII, and clipping and filtering methods for the SFBC-OFDM system with two transmitter antennas and $N_c = 512$ for different values of D.

where $\Pr\{A\}$ is the probability of the event A. To find the peak values and to estimate the PAPR of the analog signal, the oversampling ratio of 4 has been used. To estimate the CCDF of the PAPR, 10^6 OFDM frames have been generated in a Monte Carlo simulation. In this section, nonblind ordinary SLM and PII methods [24] have also been simulated. In the ordinary SLM method, the frame $\mathbf{X}(k)$ is multiplied by the sequences $\mathbf{b}^d$ before the SFBC encoder. In the PII method, the frame $\mathbf{X}(k)$ is partitioned into D subblocks called polyphases. Then, the even and odd elements of each subblock are interleaved, and their signs are inverted such that the PAPR is minimized. Figs. 2 and 3 show the performances of the proposed method for $N_c = 128$ and $N_c = 512$, respectively. As can be seen from Fig. 2, at the probability of 10^{-5}, the simplified SLM method reduces the PAPR by 2.6 and 3.7 dB for $D = 4$ and 16, respectively. As can be seen, the performance of the simplified SLM is very close to that of the PII and ordinary SLM methods (for $D = 4$ even its performance is better than that of the PII method). For $N_c = 512$, the application of the simplified SLM method leads to PAPR reductions of 2.5 and 3.5 dB for $D = 4$ and 16, respectively. In this case, the performance degradation in comparison with PII and ordinary SLM methods is less than 0.3 dB (for $D = 4$, the simplified SLM method outperforms the PII method). In Figs. 2 and 3, the performance of the clipping and filtering introduced in [23] is also plotted. As shown in the figures, the clipping and filtering method is 0.3 and 0.9 dB better than the simplified SLM method in PAPR reduction for $N_c = 128$ and $N_c = 512$, respectively. In the next section. it is seen that the clipping and filtering method leads to a considerable increase in SER at the receiver side.

B. SER Performance

To evaluate the SER performance, a five-tap frequency-selective channel with impulse response of $h_p(t) = \sum_{l=1}^{L=5} h_p(l)\delta(t - \Delta(l))$ is assumed between the pth transmitter antenna and the receiver antenna [17]. The tap delays $\Delta(l)$ are $[0, 0.0025, 0.005, 0.01, 0.01, 0.015, 0.025] \times T$, and the average path gains $10 \log E\{|h_p(l)|^2\}$ are [0, -4, -8, -16, -24, -39] dB, where T is the OFDM symbol duration. The error probabilities $\dot{P}$, $\widetilde{P}(d|\bar{d})$, and $\widetilde{P}$ are defined by

$$\begin{aligned} \dot{P} &= \Pr\left\{\dot{f}(\nu) \neq f^{\bar{d}}(\nu)\right\} \\ \widetilde{P}(d|\bar{d}) &= \Pr\left\{\widetilde{d} = d|\bar{d}\right\} \\ \widetilde{P} &= \Pr\{\widetilde{d} \neq \bar{d}\}. \end{aligned} \tag{29}$$

It is clear that

$$\begin{aligned} \widetilde{P} &= \sum_{\bar{d}=0}^{\bar{d}=D-1} \Pr\{\widetilde{d} \neq \bar{d}|\bar{d}\} \Pr\{\bar{d}\} \\ &= \frac{1}{D} \sum_{\bar{d}=0}^{\bar{d}=D-1} \sum_{d=0, d\neq\bar{d}}^{D-1} \widetilde{P}(d|\bar{d}) \\ &= \frac{1}{D} \sum_{\bar{d}=0}^{\bar{d}=D-1} \sum_{d=0, d\neq\bar{d}}^{D-1} (\dot{P})^{dist(\mathbf{f}^d, \mathbf{f}^{\bar{d}})}. \end{aligned} \tag{30}$$

Thus, the probability of error detection of $\bar{d}$ depends on the Hamming distance of the sequences $\mathbf{f}^d$, $0 \leq d \leq D - 1$. If these vectors are randomly generated as described in Section III, then the Hamming distance of the sequences increases when the OFDM frame length is increased; hence, the detection error is reduced.

Fig. 4 shows the SER versus signal-to-noise ratio (SNR) for $N_c = 128$ and $D = 16$ for the case when we have perfect SI at the receiver and when the sequence index is detected based on the proposed method. As can be seen from this figure, for SNRs higher than 14 dB, there is no loss due to error detection of $\bar{d}$. In addition, in this figure, the SER of a single-antenna OFDM system has been plotted. The slope of a two-antenna system at high SNRs is twice that of single-antenna system. This is due to full diversity of space frequency coding. In addition, the SER of the clipped and filtered space-frequency-coded system proposed in [23] has been plotted. As it is apparent, while the clipping and filtering method effectively reduces the PAPR, SER performance is however degraded at the receiver. In [23], a complex iterative algorithm has been proposed to compensate for the clipping effect at the receiver side. Fig. 5 shows the same plots for $N_c = 512$. As can be seen from these figures, for SNR values higher than 10 dB, there is no loss due to error detection of $\bar{d}$.

VI. Conclusion

In this paper, it has been shown that the simplified method that has been previously proposed for spatially multiplexed OFDM systems is suitable for PAPR reduction of SFBC-OFDM systems. In fact, the simplified SLM does not change the orthogonality of space frequency codes. In this method, the same phase sequence is concurrently applied

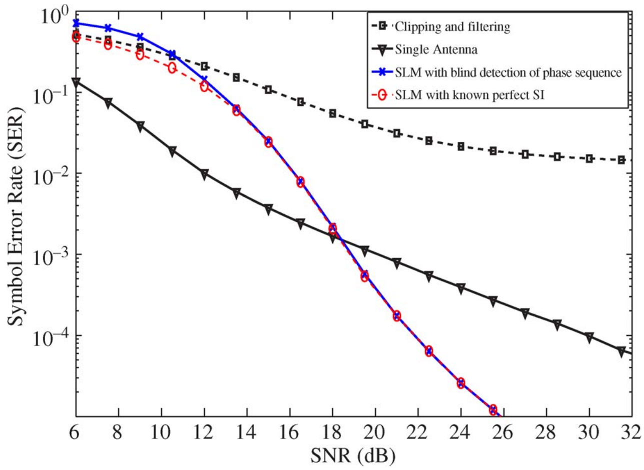

Fig. 4. SER performance of the SFBC-OFDM system with two transmitter antennas and $Nc = 128$ with three different PAPR reduction methods. 1) Simplified SLM with perfect SI ($D = 16$); 2) simplified SLM ($D = 16$) with blind detection of SI; and 3) clipping and filtering, and SER performance of a single-antenna OFDM system.

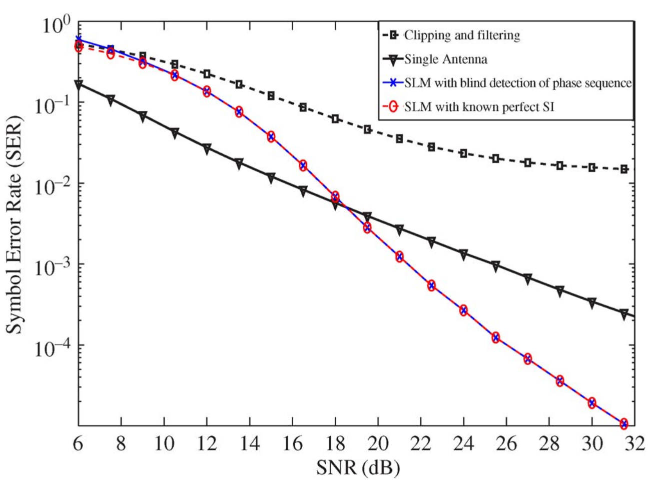

Fig. 5. SER performance of the SFBC-OFDM system with two transmitter antennas and $Nc = 512$ with three different PAPR reduction methods. 1) Simplified SLM with perfect SI ($D = 16$); 2) simplified SLM ($D = 16$) with blind detection of SI; and 3) clipping and filtering, and SER performance of a single-antenna OFDM system.

to the frequency-domain signals for both antennas, and the signal with minimum PAPR has been found and transmitted. Optimum ML detection for the transmitted symbols and the phase sequence index has been introduced, and then, a low-complexity suboptimum detector has been proposed to detect the phase sequence index without SI. Simulation results show that the simplified SLM method effectively reduces the PAPR of SFBC-OFDM system, and the error rate of blind detection of the phase sequence decreases when the number of subcarriers is increased. The detection errors of the proposed method for SNR values higher than 10 and 14 dB are negligible for the OFDM frame lengths of 512 and 128, respectively.

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