Blind Selected Mapping Techniques for Space-Time Block Coded Filtered Single-Carrier Signals

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***Abstract*—Single-carrier with frequency-domain equalization and space-time block coded transmit diversity (SC-FDE/STTD) is a promising broadband transmission technique achieving spatial and frequency diversity gains. SC signal has low peak-to-average power ratio (PAPR) property, but its PAPR increases if transmit filtering and/or high-level data modulation are used. Recently, we proposed a blind selected mapping (blind SLM) which does not require the side-information sharing between transmitter and receiver. Maximum likelihood detection (MLD) is employed. Our previous studies considered the single-antenna transmission case (i.e., without transmit diversity) only. In this paper, we extend the blind SLM technique to SC-FDE/STTD. Phase rotation of the transmit signal is carried out as a linear precoder prior to STTD encoder. Performance evaluation is done by computer simulation to show that the blind SLM provides a low-PAPR signal and a good BER performance without side-information sharing even for SC-FDE/STTD.**

***Index Terms*—Single-carrier, selected mapping, peak-to-average power ratio, space-time block coded transmit diversity**

I. Introduction

Small-cell network using distributed antennas (DA) [1] is a promising candidate for the fifth-generation (5G) mobile communication systems as it achieves high spectrum efficiency (SE) and energy efficiency (EE) simultaneously. Space-time block coded transmit diversity (STTD) [2] is a cooperative multi-input multi-output (MIMO) transmission utilizing DAs to acquire spatial diversity gain. The use of single-carrier with frequency-domain equalization (SC-FDE) [3] and STTD, called SC-FDE/STTD, achieves both frequency and spatial diversity gains, which consequently results in its robustness against frequency-selective fading [4].

SC signal has low peak-to-average power ratio (PAPR). The low-PAPR transmit signal waveform contributes to low input peak power into power amplifier (PA), resulting in low power consumption in the PA. However, the PAPR of SC signal becomes higher when transmit filtering (including band-limiting filtering and transmit equalization) and/or higher data modulation level are used. Increasing the transmit filter roll-off factor reduces the PAPR of SC signal, but it widens the signal bandwidth and degrades the SE. Selected mapping (SLM) [5] is a promising solution which effectively reduces the PAPR without signal distortion but with small overhead bits (side-information). We recently proposed a blind SLM technique which does not require transmission of side-information, in which the phase rotation is applied either in frequency domain (called FD-SLM) [6] or in time domain (called TD-SLM) [7]. At the receiver, maximum-likelihood detection (MLD) is employed for data detection without side-information. It was confirmed in [6-7] that the blind SLM provides a low-PAPR signal and a good bit-error rate (BER) performance without side-information sharing between transmitter and receiver. In our study of blind SLM [6-7], we considered the single-antenna case only and the PAPR reduction in the case of SC-FDE/STTD was left as our future study. It should be noted that the number of side-information bits increases with the number of transmit antennas [8].

This problem motivates us to extend the blind SLM to SC-FDE/STTD. In this paper, we utilize the structure of STTD encoding/decoding matrices which contain complex-conjugate operation only [9]. The phase rotation (both in FD-SLM and TD-SLM) can be applied as a linear precoder prior to STTD encoder without major changes in encoding and decoding matrices. Phase rotation pattern can be selected as either a common pattern based on minimax criterion, or different patterns for each transmit block. At the receiver, MLD is applied after STTD decoder for detecting the data without requiring the side-information. PAPR and BER of the SC-FDE/STTD using blind SLM are evaluated by computer simulation to show that a low-PAPR signal is provided while achieving both frequency and spatial diversity gains.

The rest of this paper is organized as follows. Sect. II introduces the blind SLM for SC-FDE/STTD. Sect. III shows the transmitter model of SC-FDE/STTD using blind SLM. Sect. IV shows the receiver model with MLD. Sect. V shows the simulation results, and Sect. VI concludes the paper.

II. Blind SLM Technique for SC-FDE/STTD

Assuming that an N_c-length time-domain transmit block is represented by $\mathbf{s} = [s(0), s(1), \ldots, s(N_c-1)]^T$, PAPR of $\mathbf{s}$ calculated over a V-times oversampled block, which is

$$\mathrm{PAPR}(\mathbf{s}) = \frac{\max\{|s(n)|^2, n = 0, \frac{1}{V}, \frac{2}{V}, \ldots, N_c - 1\}}{\frac{1}{N_c}\sum_{n=0}^{N_c-1}|s(n)|^2}. \quad (1)$$

In STTD encoder [10], J frequency-domain transmit blocks are encoded into N_t parallel stream with Q frequency-domain blocks. J and Q are as a function of N_t and their relationship is summarized in Table I. Assuming the j-th time-domain input transmit block of STTD encoding $\mathbf{x}_j = [x_j(0), \ldots, x_j(N_c -$

$1)]^T, j = 0 \sim J-1$ and its corresponding frequency-domain block $\mathbf{X}_j = [X_j(0), \ldots, X_j(N_c-1)]^T$, the STTD encoder is expressed by

$$\mathbf{S}_{N_t} = \begin{cases} \mathbf{X}_0 \quad \text{if } N_t = 1, \\ \frac{1}{\sqrt{2}} \begin{bmatrix} \mathbf{X}_0 & -\mathbf{X}_1^* \\ \mathbf{X}_1 & \mathbf{X}_0^* \end{bmatrix} \quad \text{if } N_t = 2 \\ \frac{1}{\sqrt{3}} \begin{bmatrix} \mathbf{X}_0 & -\mathbf{X}_1^* & -\mathbf{X}_2^* & \mathbf{0} \\ \mathbf{X}_1 & \mathbf{X}_0^* & \mathbf{0} & -\mathbf{X}_2^* \\ \mathbf{X}_2 & \mathbf{0} & \mathbf{X}_0^* & \mathbf{X}_1^* \end{bmatrix} \quad \text{if } N_t = 3 \\ \frac{1}{\sqrt{4}} \begin{bmatrix} \mathbf{X}_0 & -\mathbf{X}_1^* & -\mathbf{X}_2^* & \mathbf{0} \\ \mathbf{X}_1 & \mathbf{X}_0^* & \mathbf{0} & -\mathbf{X}_2^* \\ \mathbf{X}_2 & \mathbf{0} & \mathbf{X}_0^* & \mathbf{X}_1^* \\ \mathbf{0} & \mathbf{X}_2 & -\mathbf{X}_1 & \mathbf{X}_0 \end{bmatrix} \quad \text{if } N_t = 4 \end{cases}. \quad (2)$$

In addition, the STTD encoder was developed to support N_t=5 and 6 [10], but we consider N_t=1 to 4 in this paper. From (2), the output signal of STTD encoder is either an original or a complex-conjugated version of $\mathbf{X}_j$. Since the PAPR of the time-domain waveforms obtained from $\mathbf{X}_j$ and that of $\mathbf{X}_j^*$ are identical, applying the phase rotation as a linear precoder to J transmit blocks prior to STTD encoding achieves the same PAPR performance as applying the phase rotation to $N_t \times Q$ output blocks after STTD encoding. This idea can reduce the computational complexity occurred by unnecessary PAPR calculation at the transmitter.

In this paper, we introduce a $JN_c \times JN_c$ phase rotation matrix $\mathbf{P}_{\hat{u}} = diag[\mathbf{P}_{\hat{u},0}, \mathbf{P}_{\hat{u},1}, \ldots, \mathbf{P}_{\hat{u},J-1}]$ as a linear precoder representing SLM. Each submatrix $\mathbf{P}_{\hat{u},j}$, $j = 0 \sim J-1$ is selected from U phase rotation patterns in a codebook $\{\mathbf{P}_u = diag[P_u(0), \ldots, P_u(N_c-1)];\ u = 0 \sim U-1\}$, which are randomly generated except that the first pattern is set to an identity matrix $\mathbf{I}_{N_c}$. The codebook is generated only one time and then is used for all transmit blocks. The phase rotation patterns generation in blind FD-SLM [6] and blind TD-SLM [7] are difference, i.e. $P_u(k) \in \{1, -1\}$ (binary phase rotation) for FD-SLM and $P_u(n) \in \{1, \exp(j2\pi/3), \exp(-j2\pi/3)\}$ (polyphase rotations) for TD-SLM, respectively. We consider two pattern selection criteria in the blind SLM as follows.

A. Minimax criterion

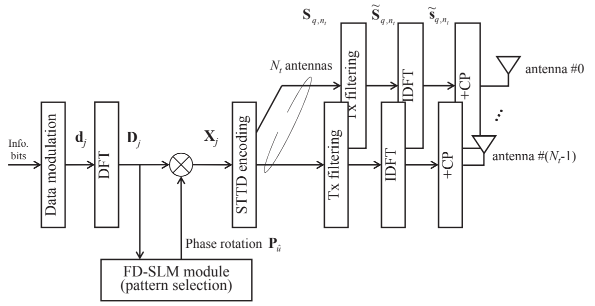
SLM based on minimax criterion selects the phase rotation in order to minimize the maximum PAPR value among multiple transmit blocks. Phase rotation pattern selection of SLM for SC-FDE/STTD based on the minimax criterion is given by

$$\mathbf{P}_{\hat{u},j} = \begin{cases} \arg\min\limits_{u=0\sim U-1}\left(\max\limits_{j=0\sim J-1} \text{PAPR}\left(\mathbf{F}_{N_c}^H \mathbf{P}_{u,j} \mathbf{H}_T \mathbf{F}_{N_c} \mathbf{d}_j\right)\right) \\ \qquad\qquad \text{for FD-SLM} \\ \arg\min\limits_{u=0\sim U-1}\left(\max\limits_{j=0\sim J-1} \text{PAPR}\left(\mathbf{F}_{N_c}^H \mathbf{H}_T \mathbf{F}_{N_c} \mathbf{P}_{u,j} \mathbf{d}_j\right)\right) \\ \qquad\qquad \text{for TD-SLM} \end{cases}. \quad (3)$$

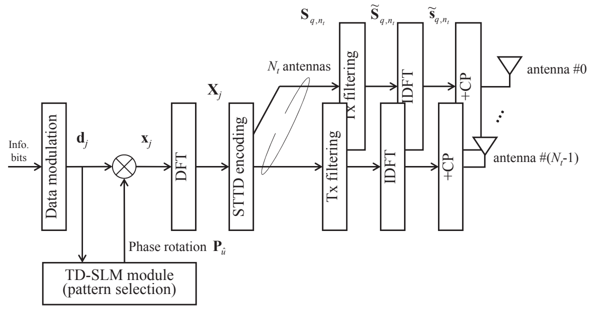
The selection algorithm in (3) returns the same phase rotation pattern for all J transmit blocks. The definition of the remaining matrices representation is described in Sect. III.

TABLE I
STTD ENCODING PARAMETERS

N_t	J	Q	$R_{\text{STTD}} = J/Q$
1	1	1	1
2	2	2	1
3	3	4	0.75
4	3	4	0.75



(a) Blind FD-SLM



(b) Blind TD-SLM

Fig. 1. SC-FDE/STTD transmitter with blind SLM.

B. Block-by-block minimization criterion

SLM based on block-by-block minimization criterion selects the phase rotation pattern which minimizes the PAPR of each individual transmit block. Phase rotation pattern selection of SLM techniques for STTD-SC-FDE based on the block-by-block minimization criterion can be expressed by

$$\mathbf{P}_{\hat{u},j} = \begin{cases} \arg\min\limits_{u=0\sim U-1}\left(\text{PAPR}\left(\mathbf{F}_{N_c}^H \mathbf{P}_{u,j} \mathbf{H}_T \mathbf{F}_{N_c} \mathbf{d}_j\right)\right) \\ \qquad\qquad \text{for FD-SLM} \\ \arg\min\limits_{u=0\sim U-1}\left(\text{PAPR}\left(\mathbf{F}_{N_c}^H \mathbf{H}_T \mathbf{F}_{N_c} \mathbf{P}_{u,j} \mathbf{d}_j\right)\right) \\ \qquad\qquad \text{for TD-SLM} \end{cases}. \quad (4)$$

It can be seen from (4) that the selected phase rotation patterns can be different for the particular j-th transmit block, at which we can expect to achieve the same degree of freedom in PAPR reduction as that of the SLM in single-antenna case.

III. TRANSCEIVER MODELS

A. Transmitter model with blind SLM

Transmitter system models of SC-FDE/STTD using blind FD-SLM and blind TD-SLM are illustrated by Figs. 1(a) and 1(b), respectively. Point-to-point transmission using N_t transmit antennas and N_r receive antennas are assumed. We begin with J data-modulated transmit blocks $\mathbf{d} = [\mathbf{d}_0, \mathbf{d}_1, \ldots, \mathbf{d}_{J-1}]$ where $\mathbf{d}_j = [d_j(0), \ldots, d_j(N_c-1)]^T$ represents the j-th block. The j-th frequency-domain input signal of STTD encoder after applying SLM techniques in Sect. II, $\mathbf{X}_j = [X_j(0), \ldots, X_j(N_c-1)]^T, j = 0 \sim J-1$, is expressed by

$$\mathbf{X}_j = \begin{cases} \mathbf{P}_{\hat{u},j}\mathbf{F}_{N_c}\mathbf{d}_j \text{ for FD-SLM} \\ \mathbf{F}_{N_c}\mathbf{P}_{\hat{u},j}\mathbf{d}_j \text{ for TD-SLM} \end{cases}, \tag{5}$$

where the selected phase rotation pattern $\mathbf{P}_{\hat{u},j}, j = 0 \sim J-1$ is selected based on (3) or (4). $\mathbf{F}_{N_c}$ represents discrete Fourier transform (DFT) matrix and is expressed by

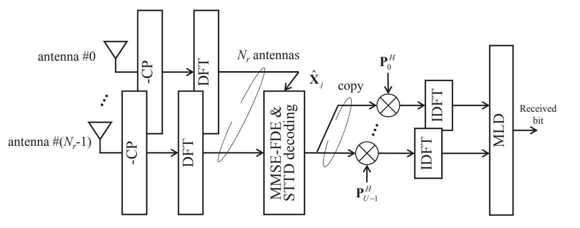
$$\mathbf{F}_{N_c} = \frac{1}{\sqrt{N_c}} \begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{\frac{-j2\pi(1)(1)}{N_c}} & \cdots & e^{\frac{-j2\pi(1)(N_c-1)}{N_c}} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & e^{\frac{-j2\pi(N_c-1)(1)}{N_c}} & \cdots & e^{\frac{-j2\pi(N_c-1)(N_c-1)}{N_c}} \end{bmatrix}, \tag{6}$$

and its Hermitian transpose $\mathbf{F}_{N_c}^H$ is inverse DFT (IDFT).

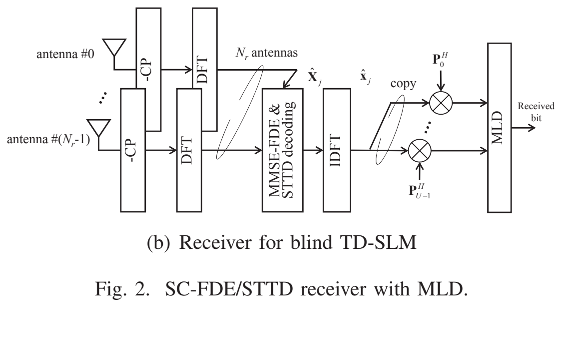
Then, $\mathbf{X}_j$ is used in STTD encoding described by (2), yielding the STTD encoding output as N_t parallel streams of Q blocks $\mathbf{S} = [\mathbf{S}_0, \mathbf{S}_1, \ldots, \mathbf{S}_{Q-1}]$ where $\mathbf{S}_q = [\mathbf{S}_{q,0}, \ldots, \mathbf{S}_{q,n_t}, \ldots, \mathbf{S}_{q,N_t-1}]^T, q = 0 \sim Q-1, n_t = 0 \sim N_t-1$ and $\mathbf{S}_{q,n_t} = [S_{q,n_t}(0), \ldots, S_{q,n_t}(N_c-1)]^T$. Each frequency-domain output block is then multiplied by transmit filtering matrix $\mathbf{H}_T = diag[H_T(0), \ldots, H_T(N_c-1)]$, obtaining filtered signal $\tilde{\mathbf{S}}_{q,n_t} = \mathbf{H}_T\mathbf{S}_{q,n_t}$ for all $q = 0 \sim Q-1$ and $n_t = 0 \sim N_t-1$. The transmit filtering considered in this paper is square-root raised cosine (SRRC) filtering, where its filter transfer function is expressed by [11]

$$H_T(k) = \begin{cases} 1, \text{ if } \frac{N_c}{2} - \frac{(1-\alpha)N_c}{2} \le k < \frac{N_c}{2} + \frac{(1-\alpha)N_c}{2} \\ 0, \text{ if } k < \frac{N_c}{2} - \frac{(1-\alpha)N_c}{2} \text{ or } k > \frac{N_c}{2} + \frac{(1-\alpha)N_c}{2}, \\ \cos\left(\frac{\pi}{2\alpha}\left(\frac{|k-N_c/2|}{N_c}\right) - \frac{1-\alpha}{2}\right), \text{ otherwise} \end{cases} \tag{7}$$

where α is filter roll-off factor. After that, $\tilde{\mathbf{S}}_{q,n_t}$ is transformed back into time domain by N_c-point IDFT, yielding the q-th time-domain transmit block to be transmitted at the n_t-th antenna $\tilde{\mathbf{s}}_{q,n_t} = [\tilde{s}_{q,n_t}(0), \ldots, \tilde{s}_{q,n_t}(N_c-1)]^T$ as $\tilde{\mathbf{s}}_{q,n_t} = \mathbf{F}_{N_c}^H\tilde{\mathbf{S}}_{q,n_t}$. Finally, the last N_g samples of transmit block are copied as a cyclic prefix (CP) and inserted into the guard interval (GI), then a CP-inserted signal block of $N_g + N_c$ samples is transmitted.



(a) Receiver for blind FD-SLM



(b) Receiver for blind TD-SLM

Fig. 2. SC-FDE/STTD receiver with MLD.

B. Receiver models with MLD

The receiver models of SC-FDE/STTD using blind FD-SLM and blind TD-SLM are illustrated by Figs. 2(a) and 2(b), respectively. The wireless propagation channel is assumed to be a symbol-spaced L-path frequency-selective block Rayleigh fading channel [4], where its impulse response between the n_t-th transmit antenna and the n_r-th DA (i.e. receive antenna) is

$$h_{n_r,n_t}(\tau) = \sum_{l=0}^{L-1} h_{n_r,n_t,l}\delta(\tau - \tau_{n_r,n_t,l}), \tag{8}$$

where $h_{n_r,n_t,l}$ and $\tau_{n_r,n_t,l}$ are complex-valued path gain and time delay of the l-th path, respectively. In addition, $h_{n_r,n_t,l}$ is assumed to be the same for Q encoded block in this paper for simplicity. The q-th time-domain received block at the n_r-th DA $\mathbf{r}_{q,n_r} = [r_{q,n_r}(0), \ldots, r_{q,n_r}(N_c-1)]^T$ is given by

$$r_{q,n_r}(n) = \sqrt{\frac{2E_s}{T_s}} \sum_{n_t=0}^{N_t-1}\sum_{l=0}^{L-1} h_{n_r,n_t,l}\tilde{s}_{q,n_t}(n-\tau_{n_r,n_t,l}) + z_{q,n_r}(n), \tag{9}$$

where E_s is symbol energy, and $z_{q,n_r}(n)$ is zero-mean additive white Gaussian noise (AWGN) having the variance $2N_0/T_s$ with T_s is symbol duration and N_0 being the one-sided noise power spectrum density. After CP removal, $r_{q,n_r}(n)$ is transformed into frequency domain by N_c-point DFT, yielding the frequency-domain received signal vector at the n_r-th DA and the q-th received block $\mathbf{R}_{q,n_r} = [R_{q,n_r}(0), \ldots, R_{q,n_r}(N_c-1)]^T$ as

$$R_{q,n_r}(k) = \sqrt{\frac{2E_s}{T_s}} \sum_{n_t=0}^{N_t-1} H_{n_r,n_t}(k)\tilde{S}_{q,n_t}(k) + Z_{q,n_r}(k). \tag{10}$$

Frequency-domain channel response between the n_t-th transmit antenna and the n_r-th DA and noise at the n_r-th DA and the q-th received block are given by

$$H_{n_r,n_t}(k) = \sum_{l=0}^{L-1} h_{n_r,n_t,l}\exp(-j2\pi k\tau_{n_r,n_t,l}/N_c), \tag{11a}$$

$$Z_{q,n_r}(k) = \frac{1}{N_c}\sum_{n=0}^{N_c-1} z_{q,n_r}(n)\exp(-j2\pi kn/N_c). \tag{11b}$$

Next, FDE based on minimum MSE criterion (MMSE-FDE) [3] is carried out, yielding the equalized signal $\tilde{\mathbf{R}}_{(N_t\times Q)} = \mathbf{W}_{(N_t\times N_r)}^H\mathbf{R}_{(N_r\times Q)}$, where $\mathbf{W} = [\mathbf{W}_0, \ldots, \mathbf{W}_{n_r}, \ldots, \mathbf{W}_{N_r-1}]^T$ and $\mathbf{W}_{n_r} = [\mathbf{W}_{n_r,0}, \ldots, \mathbf{W}_{n_r,n_t}, \ldots, \mathbf{W}_{n_r,N_t-1}]$ represents the MMSE-FDE weight matrix. The FDE weight at the k-th frequency index is given by [12]

$$W_{n_r,n_t}(k) = \frac{H_T(k)H_{n_r,n_t}(k)}{\sum_{n_r=0}^{N_r-1}\sum_{n_t=0}^{N_t-1}|H_T(k)H_{n_r,n_t}(k)|^2 + \frac{E_s}{N_0}^{-1}}. \tag{12}$$

After applying the MMSE-FDE, STTD decoding is carried out for acquiring the spatial diversity gain. The j-th frequency-domain received block after STTD decoding $\hat{\mathbf{X}}_j, j = 0 \sim J-1$ is obtained by the following decoding techniques.

$$\hat{\mathbf{X}}_{N_t} = \begin{cases} \tilde{\mathbf{R}}_{0,0} \quad \text{if } N_t = 1 \\ \begin{bmatrix} \hat{\mathbf{X}}_0 \\ \hat{\mathbf{X}}_1 \end{bmatrix} = \begin{bmatrix} \tilde{\mathbf{R}}_{0,0} + \tilde{\mathbf{R}}^*_{1,1} \\ \tilde{\mathbf{R}}_{0,1} - \tilde{\mathbf{R}}^*_{1,0} \end{bmatrix} \quad \text{if } N_t = 2 \\ \begin{bmatrix} \hat{\mathbf{X}}_0 \\ \hat{\mathbf{X}}_1 \\ \hat{\mathbf{X}}_2 \end{bmatrix} = \begin{bmatrix} \tilde{\mathbf{R}}_{0,0} + \tilde{\mathbf{R}}^*_{1,1} + \tilde{\mathbf{R}}^*_{2,2} \\ \tilde{\mathbf{R}}_{0,1} - \tilde{\mathbf{R}}^*_{1,0} + \tilde{\mathbf{R}}^*_{3,2} \\ \tilde{\mathbf{R}}_{0,2} - \tilde{\mathbf{R}}^*_{2,0} + \tilde{\mathbf{R}}^*_{3,1} \end{bmatrix} \quad \text{if } N_t = 3 \\ \begin{bmatrix} \hat{\mathbf{X}}_0 \\ \hat{\mathbf{X}}_1 \\ \hat{\mathbf{X}}_2 \end{bmatrix} = \begin{bmatrix} \tilde{\mathbf{R}}_{0,0} + \tilde{\mathbf{R}}^*_{1,1} + \tilde{\mathbf{R}}^*_{2,2} + \tilde{\mathbf{R}}_{3,3} \\ \tilde{\mathbf{R}}_{0,1} - \tilde{\mathbf{R}}^*_{1,0} - \tilde{\mathbf{R}}_{2,3} + \tilde{\mathbf{R}}^*_{3,2} \\ \tilde{\mathbf{R}}_{0,2} + \tilde{\mathbf{R}}_{1,3} - \tilde{\mathbf{R}}^*_{2,0} + \tilde{\mathbf{R}}^*_{3,1} \end{bmatrix} \quad \text{if } N_t = 4 \end{cases} . \quad (13)$$

In the conventional SLM [5], the j-th received block before de-modulation $\tilde{\mathbf{d}}_j = [\tilde{d}_j(0), \ldots, \tilde{d}_j(N_c-1)]^T$ is obtained by multiplying the complex-conjugated version of selected phase rotation pattern (i.e. de-mapping) to $\hat{\mathbf{X}}_j$ for FD-SLM, or $\hat{\mathbf{x}}_j = \mathbf{F}^H_{N_c}\hat{\mathbf{X}}_j$ for TD-SLM, yielding $\tilde{\mathbf{d}}_j = \mathbf{F}^H_{N_c}\mathbf{P}^H_{\hat{u},j}\hat{\mathbf{X}}_j$ and $\tilde{\mathbf{d}}_j = \mathbf{P}^H_{\hat{u},j}\mathbf{F}^H_{N_c}\hat{\mathbf{X}}_j$ for FD-SLM and TD-SLM, respectively. However, side-information transmission is needed for sharing the information of $\hat{u}$ to the receiver, which degrades the SE.

We apply MLD for data detection without side-information. Assuming that the received signal-to-noise power ratio (SNR) is sufficiently high, the samples in the received block obtained from correct de-mapping are very close to an original signal constellation. MLD utilizes the above fact by searching a possible de-mapping pattern and corresponding received block with the lowest Euclidean distance from original signal constellation [6]. Note that the use of polyphase rotations is mandatory in the blind TD-SLM in order to avoid symmetric rotation [7]. The j-th received block before data de-modulation obtained from the MLD is given by

$$\tilde{\mathbf{d}}_j = \begin{cases} \arg \min\limits_{\substack{v=0\sim U-1, \\ \tilde{\mathbf{d}}\in\Psi_{\text{mod}}}} \|\mathbf{F}^H_{N_c}\mathbf{P}^H_{v,j}\hat{\mathbf{X}}_j - \tilde{\mathbf{d}}_j\|^2 \text{ for FD-SLM} \\ \arg \min\limits_{\substack{v=0\sim U-1, \\ \tilde{\mathbf{d}}\in\Psi_{\text{mod}}}} \|\mathbf{P}^H_{v,j}\mathbf{F}^H_{N_c}\hat{\mathbf{X}}_j - \tilde{\mathbf{d}}_j\|^2 \text{ for TD-SLM} \end{cases}, \quad (14)$$

where Ψ_{mod} is a set of constellations for a particular modulation level. Note that the MLD in (14) is carried out for an N_c-length block.

IV. Performance Evaluation

Simulation parameters are summarized in Table II. Channel coding is not considered (performance evaluation considering channel coding is left as our future work). Note that the polyphase rotations used in this paper are not an optimal set but sufficient for allowing the blind TD-SLM. In case of transmission with side-information sharing, the required side-information bits are $J \log_2 U$ [8].

A. $PAPR_{0.1\%}$

PAPR performance is evaluated by measuring the PAPR value at complementary cumulative distribution function (CCDF) equals 10^{-3}, called $\text{PAPR}_{0.1\%}$. Fig. 3 shows the $\text{PAPR}_{0.1\%}$ of SC-FDE/STTD using blind SLM technique as a function of the number of phase rotation patterns (U). The $\text{PAPR}_{0.1\%}$ of blind TD-SLM and blind FD-SLM assuming N_t=1 are also shown for comparison. PAPR reduces as U increases as presented in [6-7]. The blind TD-SLM achieves lower PAPR compared to the blind FD-SLM regardless of the phase rotation pattern selection criterion and N_t. This is because the time-domain transmit waveform samples obtained from blind TD-SLM is in a fixed set (for example, 16×3=48 patterns for 16QAM), where the transmit waveform samples obtained from blind FD-SLM is random [7]. This fact can limit the output waveform patterns and results in near-optimal solution than that of FD-SLM at the same U.

TABLE II
Simulation parameters.

Transmitter	Data modulation	16QAM
	No. of subcarriers	N_c=256
	CP length	N_g=16
	Transmit filtering	SRRC (α=0)
SLM module	Phase rotation sequence type	FD-SLM: random binary TD-SLM: random polyphase
Channel	Fading type	Frequency-selective block Rayleigh
	Power delay profile	Symbol-spaced 16-path uniform
Receiver	Channel estimation	Ideal

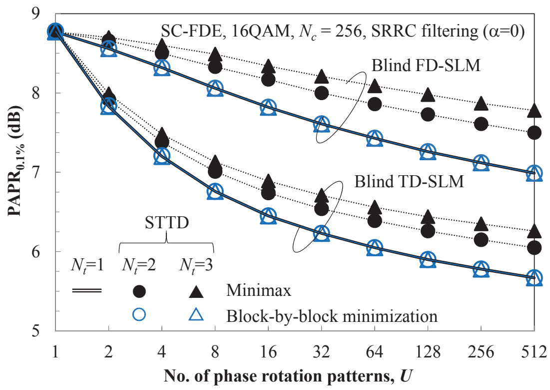
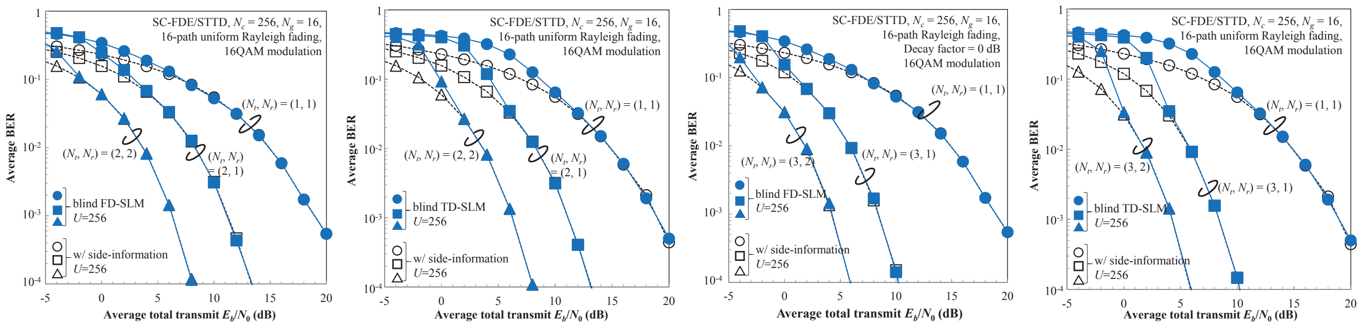


Fig. 3. $\text{PAPR}_{0.1\%}$ versus the number of patterns

It is also observed from Fig. 3 that $\text{PAPR}_{0.1\%}$ increases as N_t increases for both blind FD-SLM and blind TD-SLM if the phase rotation pattern is selected based on minimax criterion. $\text{PAPR}_{0.1\%}$ increases by 0.4 dB and 0.6 dB in blind TD-SLM when N_t=2 and 3, respectively. $\text{PAPR}_{0.1\%}$ also increases by 0.6 dB and 0.9 dB in blind FD-SLM when N_t=2 and 3. The reason is the minimax criterion decreases the degree of freedom in candidate generation in SLM. On the other hand, $\text{PAPR}_{0.1\%}$ does not increase as N_t increases if the phase rotation pattern is selected based on block-by-block minimization criterion. This is because the same degree of freedom is obtained as that of SLM in single-antenna transmission. For example, we can reduce the PAPR of SC-FDE/STTD by 3.2 dB when the blind TD-SLM with U=512 is applied if the block-by-block minimization criterion is used. As a consequence, block-by-block minimization criterion is better than the minimax



(a) Blind FD-SLM, $N_t = 2$ (b) Blind TD-SLM, $N_t = 2$ (c) Blind FD-SLM, $N_t = 3$ (d) Blind TD-SLM, $N_t = 3$

Fig. 4. BER performances.

criterion. Hereafter in this paper, the former is used.

B. BER Performance

BER as a function of total transmit bit energy-to-noise power spectrum density ratio $E_b/N_0 = (1/N_{\text{mod}})(E_s/N_0)(1+N_g/N_c)$ where N_{mod} represents modulation level (4 for 16QAM) of SC-FDE/STTD using blind FD-SLM and blind TD-SLM assuming N_t=2 and 3 are shown in Fig. 4. The blind SLM using block-by-block minimization criterion with $U = 256$ is assumed.

BER improves when either N_t or N_r increases because of higher spatial diversity gain obtained from STTD. BER of the SC-FDE/STTD using either blind FD-SLM or blind TD-SLM degrades compared to the transmission with ideal side-information sharing when the transmit E_b/N_0 is low. This is also consistent with [6-7] as the noise and residual ISI makes the received samples become dispersive even though the de-mapping is carried out correctly. However, there is no difference on BER of transmission using the blind SLM and SLM with ideal side-information sharing when the transmit E_b/N_0 is sufficiently high. This can conclude that SC-FDE/STTD with blind SLM techniques achieve low-PAPR transmit signal waveform without significant BER degradation despite an absence of side-information.

In addition, it is observe that the SC-FDE/STTD using blind FD-SLM achieves better BER than blind TD-SLM at the low transmit E_b/N_0 region (for example, see Figs. 4(a) and 4(b)). The reason can be described by referring the received sample constellations of the blind FD-SLM and the blind TD-SLM in [6-7] as the difference between received sample constellations obtained from correct de-mapping and that of incorrect de-mapping in the blind FD-SLM is more obvious than the blind TD-SLM, which results in more robustness against the noise.

V. Conclusion

Blind SLM for SC-FDE/STTD were described in this paper. Both spatial and frequency diversity gains are obtained by employing orthogonal STTD and FDE. The blind SLM can be implemented as a linear precoder by multiplying phase rotation to the transmit block prior to STTD encoding, if the block-by-block minimization criterion is used in phase rotation pattern selection. Computer simulation results confirmed that PAPR can be lowered by about 3.2 dB regardless of the number N_t of transmit antennas. Blind SLM provides the BER performance very close to SLM with ideal side-information sharing without degrading the spectrum efficiency (SE). The use of MLD requires a high computational complexity and hence, a study on complexity-reduced blind SLM is left as our important future works.

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