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A Phase Rotation Sequence Estimation based on Minimum Euclidean Distance of Fourth-power Constellation for Blind Selected Mapping

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Abstract Blind selected mapping (blind SLM) is an effective peak-to-average power ratio (PAPR) reduction technique, in which the phase rotation sequence which has been used at the transmitter side is blindly estimated. The Euclidean distance between the received symbols after de-mapping and the original QAM constellation is used for the estimation, which requires high computational complexity. In this paper, we introduce a phase rotation sequence estimation based on the minimum Euclidean distance of the fourth-power constellation. The use of fourth-power constellation reduces symbol candidates in minimum Euclidean distance calculation and hence, contributes to complexity reduction. A set of phase rotation sequences constructed by random selection from {0°,45°} or {0°,135°} are also introduced to further reduce the complexity. Computer simulation result confirms that the proposed phase rotation and sequence estimation technique can reduce the computational complexity without degrading the uncoded bit error rate (BER) performance for the given PAPR reduction.

Keyword Peak-to-average power ratio (PAPR), OFDM, single carrier (SC), selected mapping (SLM)

1. Introduction

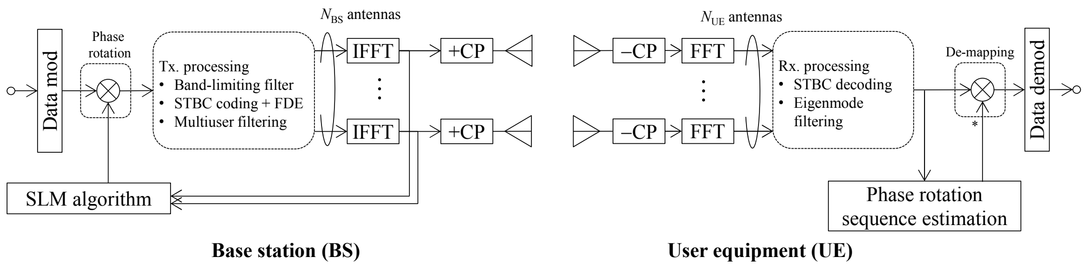
Recently, the development of the fifth-generation (5G) mobile communications systems [1] is intensified aiming at inauguration of 5G communications services in around 2020. Even in 5G systems, the low peak-to-average power ratio (PAPR) waveform design remains important, in particular, for battery-powered user equipments (UEs). Single-carrier (SC) signals have lower PAPR compared to orthogonal frequency division multiplexing (OFDM) signals [2]. However, PAPR reduction technique is also necessary for SC transmission since PAPR of SC signals increases when a transmit filtering is employed [3].

Selected mapping (SLM) [4] is an efficient and simple PAPR reduction scheme. SLM selects the waveform having the lowest PAPR among many candidates generated by applying phase rotation to the original transmit signal. The SLM [4], originally proposed for OFDM, requires side information transmission. SLM without side information (called blind SLM) compatible with both SC and OFDM was proposed in [5]. Its applications to space-time block coded transmit diversity (STBC-TD) and multiuser multi-input multi-output (MU-MIMO) were discussed in [6] and [7], respectively.

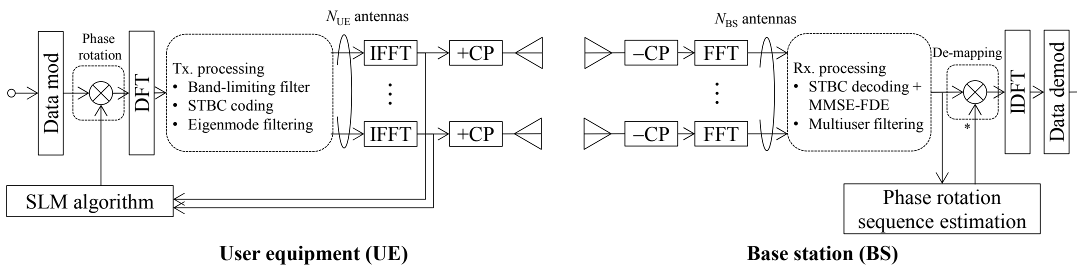
The blind SLM in [5-7] employs a maximum likelihood (ML) phase rotation sequence estimation based on minimum Euclidean distance between the de-mapped symbols and the original signal constellation. The ML estimation works effectively, but the ML estimation requires high computational complexity. A 2-step phase rotation sequence estimation based on Viterbi algorithm [8] was proposed to reduce the computational complexity, however, its complexity reduction capability is obvious only when the number of phase rotation sequences is larger than the number of subcarriers.

To remedy the above complexity problem in the blind SLM, we introduce an ML phase rotation sequence estimation based on minimum Euclidean distance of the fourth-power constellation. The use of the fourth-power constellation can reduce the number of candidates in the Euclidean distance calculation and hence, contributes to computational complexity reduction. It is recommended in [9] that the use of the fourth-power constellation together with a set of phase rotations {0°,45°} or {0°,135°} can further reduce the complexity. Note that Ref. [9] uses the fourth-power constellation and the above phase rotation sets for embedding the side information into data transmission (i.e., Ref. [9] uses the phase rotations {0°, 180°} to generate waveform candidates, selects the one with the lowest PAPR as transmit signal, then embeds the side information by applying 45° phase rotation to some subcarriers). In this paper, performance evaluation of the blind SLM using the above phase rotation sets and the phase rotation sequence estimation based on the fourth-power constellation is carried out by computer simulation in aspects of PAPR, BER and computational complexity. The simulation results confirm that the proposed ML phase rotation sequence estimation can reduce the estimation complexity without degrading the uncoded BER performance for the given PAPR reduction.

The rest of this paper is organized as follows. Section 2 provides an overview of blind SLM. Sect. 3 introduces a new set of phase rotations and the ML phase rotation sequence estimation using the fourth-power constellation. Sect. 4 shows computer simulation results and discussion. Finally, Sect. 5 concludes the paper.



(a) OFDM downlink



(b) SC uplink

Fig. 1 Transceivers equipped with blind SLM.

2. Overview of the conventional blind SLM

Here, we briefly describe the concept of blind SLM in [5-7]. For simplicity, we describe only the signal representation for STBC-TD and MU-MIMO, where the representation for single-antenna transmission (SISO) is obtained by setting the number of base station (BS) antennas (N_{BS}) and UE antennas (N_{UE}), J, Q, and STBC coding rate $R_{STBC}=J/Q$, to be 1. In single-user STBC-TD, we assume that STBC-TD with transmit filtering is used in the OFDM downlink, while the STBC-TD without transmit filtering is used in SC uplink. Similarly, OFDM and SC are respectively adopted for downlink and uplink transmission in MU-MIMO, where the filtering weights are already derived and discussed in [10]. The number of transmit antennas N_t becomes N_{BS} for downlink and N_{UE} for uplink, respectively. The transceiver system models equipped with blind SLM can be depicted by Fig. 1.

2.1. SLM algorithm

Assuming that a time-domain transmit waveform is $\{s(n);n=0{\sim}N_c-1\}$, PAPR is calculated over a V-times oversampled block and is given by

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

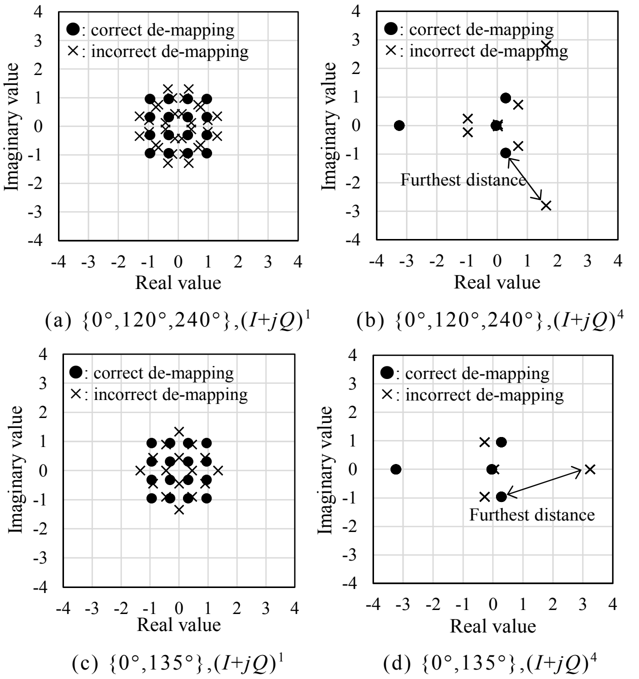
In STBC-TD transmission, the j-th block of an N_c-length data block $\{d_j(n);n=0\sim N_c-1,j=0\sim J-1\}$ is phase-rotated by multiplying with the phase rotation sequence $\{\Phi_{\hat{m}}(n);n=0\sim N_c-1\}$, yielding the phase rotated block $\{d_{j,\hat{m}}(n);n=0\sim N_c-1,j=0\sim J-1\}$. In the SC uplink, $\{d_{j,\hat{m}}(n)\}$ is transformed to frequency components block $\{D_{j,\hat{m}}(k);k=0\sim N_c-1\}$ by N_c-point discrete Fourier transform (DFT). In the OFDM downlink, we simply obtain $\{D_{j,\hat{m}}(k)\}=\{d_{j,\hat{m}}(k)\}$. Then $\{D_{j,\hat{m}}(k)\}$ are passed through transmit signal processing e.g. STBC coding and/or transmit filtering, obtaining the frequency-domain transmit signal at the n_t-th transmit antenna ($n_t=0{\sim}N_t-1$) and the q-th timeslot as $\{S_{n_t,q,\hat{m}}(k);k=0\sim N_c-1\}$ and its corresponding time-domain waveform after applying inverse DFT (IDFT) as $\{s_{n_t,q,\hat{m}}(n);n=0\sim N_c-1\}$. If we assume that $N_{UE}=2$, $S_{n_t,q,\hat{m}}(k)$ can be described by the following matrix representations.

$$\begin{bmatrix} S_{0,0,\hat{m}}(k) & S_{0,1,\hat{m}}(k) \\ S_{1,0,\hat{m}}(k) & S_{1,1,\hat{m}}(k)\end{bmatrix}=\sqrt{\frac{2E_s}{T_s}}\begin{bmatrix} D_{0,\hat{m}}(k) & -D^*_{1,\hat{m}}(k) \\ D_{1,\hat{m}}(k) & D^*_{0,\hat{m}}(k)\end{bmatrix} \quad \text{for SC uplink,} \tag{2a}$$

$$\begin{bmatrix} S_{0,0,\hat{m}}(k) & S_{0,1,\hat{m}}(k) \\ \vdots & \vdots \\ S_{N_{BS}-1,0,\hat{m}}(k) & S_{N_{BS}-1,1,\hat{m}}(k)\end{bmatrix}=\sqrt{\frac{2E_s}{T_s}}\mathbf{W}_T(k)\begin{bmatrix} D_{0,\hat{m}}(k) & -D^*_{1,\hat{m}}(k) \\ D_{1,\hat{m}}(k) & D^*_{0,\hat{m}}(k)\end{bmatrix} \quad \text{for OFDM downlink,} \tag{2b}$$

where $\mathbf{W}_T(k)$ is the transmit filtering [6]. E_s and T_s are symbol energy and symbol duration, respectively.

In MU-MIMO transmission, information sequence to be transmitted for the u-th user ($u=0{\sim}U-1$) is data-modulated into G streams of N_c-length block $\{\mathbf{d}_u(n);n=0{\sim}N_c-1\}$ with $\mathbf{d}_u(n)=[d_{u,0}(n),\ldots,d_{u,g}(n),\ldots,d_{u,G-1}(n)]^T$. $\{\mathbf{d}_u(n)\}$ is then multiplied by the selected phase rotation sequence to obtain $\{\mathbf{d}_{u,\hat{m}(u)}(n)=\Phi_{\hat{m}(u)}(n)\mathbf{d}_u(n)\}$. In the SC uplink, $\{\mathbf{d}_{u,\hat{m}(u)}(n)\}$ is transformed to frequency components block $\{\mathbf{D}_{u,\hat{m}(u)}(k);k=0\sim N_c-1\}$ by N_c-point DFT. In the OFDM downlink, we simply get $\{\mathbf{D}_{u,\hat{m}(u)}(k)\}=\{\mathbf{d}_{u,\hat{m}(u)}(k)\}$. An $N_t\times 1$ frequency component vector at the k-th subcarrier can be expressed by



(a) {0°,120°,240°},$(I+jQ)^1$ (b) {0°,120°,240°},$(I+jQ)^4$

(c) {0°,135°},$(I+jQ)^1$ (d) {0°,135°},$(I+jQ)^4$

Fig. 2 Received signal after de-mapping.

$$\mathbf{S}_{u,\hat{m}(u)}(k)=\sqrt{2E_s/T_s}\,\mathbf{W}_{T,u}(k)\mathbf{D}_{u,\hat{m}(u)}(k) \tag{3}$$

where $\mathbf{W}_{T,u}(k)$ is transmit filtering, which is either eigenmode transmit filtering for SC uplink or minimum mean square error (MMSE) based multiuser filtering for OFDM downlink [10]. $\{\mathbf{S}_{u,\hat{m}(u)}(k)\}$ is then transformed back into time domain by IDFT to obtain the transmit waveforms through N_t antennas as $\{\mathbf{s}_{u,\hat{m}(u)}(n)\}$ with $\mathbf{s}_{u,\hat{m}(u)}(n)=[s_{u,\hat{m}(u),0}(n),...,s_{u,\hat{m}(u),n_t}(n),...,s_{u,,\hat{m}(u)N_t-1}(n)]^T$.

In the case of SC uplink STBC-TD without transmit filtering (i.e., employing band-limiting filter only), the PAPR of signals before and after STBC coding are exactly the same. This is because the STBC coding employs only complex conjugate operations [6]. Therefore, we can select an individual phase rotation sequence for each of $\{d_j(n)\}$. The selected sequence for the j-th data block, $\{\Phi_{\hat{m}(j)}(n)\}$ with the corresponding sequence index $\hat{m}(j)$, is determined by

$$\hat{m}(j)=\arg\min_{m=0\sim M-1}\left(\text{PAPR}\left(\{\Phi_m(n)d_j(n)\}\right)\right), \tag{4}$$

where $\{\Phi_m(n);n=0\sim N_c-1,m=0\sim M-1\}$ is the m-th phase rotation sequence in a predefined codebook and is generated randomly as $\Phi_m(n)\in\{e^{j0},e^{j(2\pi/3)},e^{j(4\pi/3)}\}$, except the first pattern is defined as $\{\Phi_0(n)=1;n=0\sim N_c-1\}$. Note that the above phase rotation set is not optimal but sufficient to achieve blind estimation at the receiver [5].

Meanwhile, Eq. (4) is not available for STBC-TD with transmit filtering and MU-MIMO transmission since the signals before and after transmit filtering have different PAPR. In this case, a selection criterion which minimizes the maximum PAPR value (called Mini-max criterion) among all N_t transmit antennas is used. A common phase rotation with the corresponding sequence index can be defined as follows.

$$\hat{m}=\arg\min_{m=0\sim M-1}\left(\max_{\substack{n_t=0\sim N_t-1\\ q=0\sim Q-1}}\text{PAPR}\left(\{s_{n_t,q,m}(n)\}\right)\right)\ \text{for STBC-TD}, \tag{5a}$$

$$\hat{m}(u)=\arg\min_{m=0\sim M-1}\left(\max_{n_t=0\sim N_t-1}\text{PAPR}\left(\{s_{u,n_t,m}(n)\}\right)\right)\ \text{for MU-MIMO}. \tag{5b}$$

Note that Eq. (5) becomes independent from u for OFDM downlink MU-MIMO since all data streams are multiplied with the same transmit filtering. The selection criterion in Eq. (5) is sub-optimal and hence, PAPR increases when N_t increases. However, it can keep the phase rotation estimation simple and no major changes on filtering weights calculation is required.

2.2. Phase rotation sequence estimation

Phase rotation sequence estimation is employed after the receive signal processing such as STBC decoding and/or receive filtering. Phase rotation sequence estimation is done by calculating Euclidean distance between the de-mapped signal (i.e. multiplied by the complex conjugate of phase rotation sequence) and original constellation. If the de-mapping is done correctly, the de-mapped signal should be very close to the original constellation and hence, its Euclidean distance from the nearest QAM symbol is very small. The phase rotation sequence associated with the de-mapped signal having the minimum averaged Euclidean distance is selected.

In STBC-TD, assuming the j-th time-domain received block before de-mapping is $\{\hat{d}_j(n);n=0\sim N_c-1,j=0\sim J-1\}$, the phase rotation sequence estimation can be expressed as

$$\tilde{m}(j)=\arg\min_{m=0\sim M-1}\left(\sum_{n=0}^{N_c-1}\min_{c\in\Psi_{\text{mod}}}\left\|\Phi_m^*(n)\hat{d}_j(n)-c\right\|\right), \tag{6}$$

where Ψ_{mod} represents the original data modulated constellation (i.e. QAM mapping) and $\|\cdot\|$ represents Euclidean norm. Note that Eq. (6) is independent from j in the case of OFDM downlink. On the other hand, in MU-MIMO transmission, the time-domain received vector of the u-th UE is denoted by $\{\hat{\mathbf{d}}_u(n);n=0\sim N_c-1\}$ with $\hat{\mathbf{d}}_u(n)=[\hat{d}_{u,0}(n),...,\hat{d}_{u,g}(n),...,\hat{d}_{u,G-1}(n)]^T$. Then, the phase rotation sequence estimation can be expressed as

$$\tilde{m}(u)=\arg\min_{m=0\sim M-1}\left(\sum_{g=0}^{G-1}\sum_{n=0}^{N_c-1}\min_{c\in\Psi_{\text{mod}}}\left\|\Phi_m^*(n)\hat{d}_{u,g}(n)-c\right\|\right). \tag{7}$$

Finally, the received symbols prior to hard decision is obtained by $\{\tilde{d}_j(n)=\Phi_{\tilde{m}(j)}^*(n)\hat{d}_j(n);n=0\sim N_c-1,j=0\sim J-1\}$ for STBC-TD and $\{\tilde{\mathbf{d}}_u(n)=\Phi_{\tilde{m}(u)}^*(n)\hat{\mathbf{d}}_u(n);n=0\sim N_c-1\}$ for MU-MIMO, respectively.

3. Proposed blind SLM

The blind SLM in [5-7] requires high complexity. Although the 2-step phase rotation sequence estimation [8] can reduce the complexity at the receiver, its complexity reduction capability becomes obvious only if M is large. To improve the blind SLM, we introduce a new set of phase rotation with a phase rotation sequence estimation based on minimum Euclidean distance of the fourth-power symbols. Prior to introducing the new phase rotation and phase estimation algorithm, we mention that there is no major changes on the signal representations and the SLM algorithms in [5-8], also in Eqs. (4) and (5). The modifications are needed only at the codebook design and the phase rotation estimation.

Here, we assume the ML sequence estimation and use the fourth-power constellation (i.e., $(I+jQ)^4$) instead of the original constellation (i.e., $(I+jQ)$) [9]. Figs. 2(a) and 2(b) show the comparison of the original and the fourth-power constellation assuming the phase rotations of $\Phi_m(n)\in\{e^{j0},e^{j(2\pi/3)},e^{j(4\pi/3)}\}$ (in other words, 3-value phase rotations of {0°,120°,240°}). Moreover, [9] indicates that $\Phi_m(n)\in\{e^{j0},e^{j(\pi/4)}\}$ or $\Phi_m(n)\in\{e^{j0},e^{j(3\pi/4)}\}$ (in other words, 2-value phase rotations of {0°,45°} or {0°,135°}) are attractive since they also enlarge the Euclidean distance between correct and incorrect de-mapped symbols, which may contribute to BER improvement. Their constellation can be depicted by Figs. 2(c) and 2(d). It is observed from Fig. 2 that the number of symbol candidates significantly reduces, leading to computational complexity reduction, even the fourth-power operations is also used. Note that we cannot use a set $\Phi_m(n)\in\{e^{j0},e^{j(\pi/4)},e^{j(3\pi/4)}\}$ since it causes an overlap between correct and incorrect de-mappings. By employing the above concept, we modify the phase rotations and the phase estimation as follows.

(a) Phase rotation sequence generation

Phase rotation sequence generation is simply modified by randomly generating the predefined codebook as $\Phi_m(n)\in\{e^{j0},e^{j(\pi/4)}\}$ or $\Phi_m(n)\in\{e^{j0},e^{j(3\pi/4)}\}, m=1\sim M-1$, except the first sequence as $\{\Phi_0(n)=1, n=1\sim N_c-1\}$ for representing the original waveform.

(b) Phase rotation sequence estimation

By substituting the fourth-power constellation in the ML estimation equations, Eqs. (6) and (7) can be rewritten as follows.

$$\tilde{m}(j)=\arg\min_{m=0\sim M-1}\left(\sum_{n=0}^{N_c-1}\min_{c\in\Psi_{\mathrm{mod}}^4}\left\|\left(\Phi_m^*(n)\hat{d}_j(n)\right)^4-c\right\|\right) \quad \text{for STBC-TD} \tag{8a}$$

$$\tilde{m}(u)=\arg\min_{m=0\sim M-1}\left(\sum_{g=0}^{G-1}\sum_{n=0}^{N_c-1}\min_{c\in\Psi_{\mathrm{mod}}^4}\left\|\left(\Phi_m^*(n)\hat{d}_{u,g}(n)\right)^4-c\right\|\right) \quad \text{for MU-MIMO} \tag{8b}$$

where Ψ_{mod}^4 is the fourth-power modulated constellation. Eq. (8) applies the fourth-power operation to the received signal, meaning that the effect of noise also increases. However, larger error magnitude when the de-mapping is incorrect can be obtained and hence, the phase rotation sequence estimation can work effectively even in the low-SNR region.

Table 1 Simulation parameters.

Data transmission	Data modulation	16QAM
	No. of subcarriers	N_c=128
	CP length	N_g=16
SLM algorithm	Phase rotation type	Random
	No. of phase sequences	M=1~256
User equipment	Channel estimation	Ideal
	No. of UE antennas	N_{UE}=2
Base station	Channel estimation	Ideal
	No. of BS antennas	N_{BS}=4
	Transmit filter	MRT-FDE
	Receive filter	MMSE-FDE
Channel	Fading	Frequency-selective block Rayleigh
	Power delay profile	Symbol-spaced, 16-path uniform

In addition, by observing Fig. 2(d) and referring [9], the difference of real value is relatively larger than that of imaginary value. Therefore, it is sufficient to use one-dimension distance instead of two-dimension distance and consequently Eq. (8) can be simplified as

$$\tilde{m}(j)=\arg\min_{m=0\sim M-1}\left(\sum_{n=0}^{N_c-1}\min_{c\in\Psi_{\mathrm{mod}}^4}\left|\mathrm{Re}\left\{\left(\Phi_m^*(n)\hat{d}_j(n)\right)^4\right\}-\mathrm{Re}\{c\}\right|\right) \quad \text{for STBC-TD} \tag{9a}$$

$$\tilde{m}(u)=\arg\min_{m=0\sim M-1}\left(\sum_{g=0}^{G-1}\sum_{n=0}^{N_c-1}\min_{c\in\Psi_{\mathrm{mod}}^4}\left|\mathrm{Re}\left\{\left(\Phi_m^*(n)\hat{d}_{u,g}(n)\right)^4\right\}-\mathrm{Re}\{c\}\right|\right) \quad \text{for MU-MIMO} \tag{9b}$$

The phase rotation sequence estimation in Eq. (9) employs one-dimension distance calculation, which can also contribute to computational complexity reduction.

4. Performance evaluation

Simulation parameters are summarized in Table 1. Single-user STBC-TD is assumed in this paper, while channel coding is not considered for simplicity. Path loss and shadowing loss are not considered. Note that the performance evaluation of blind SLM in MU-MIMO transmission is left as future works. Phase rotation codebook are generated based on random approach as $\Phi_m(n)\in\{e^{j0},e^{j(2\pi/3)},e^{j(4\pi/3)}\}$ (referred as conventional blind SLM [5]), $\Phi_m(n)\in\{e^{j0},e^{j(\pi/4)}\}$ (referred as {0°,45°}) and $\Phi_m(n)\in\{e^{j0},e^{j(3\pi/4)}\}$ (referred as {0°,135°}). Performance evaluation is done and discussed in terms of PAPR, BER and computational complexity, and then compared with the conventional blind SLM in [5-7], i.e. blind SLM using the conventional phase rotation set and the phase sequence estimation based on original QAM constellation.

4.1. PAPR vs computational complexity

PAPR performance is evaluated by measuring the PAPR value at complementary cumulative distribution function (CCDF) equals 10^{-3}, called $\mathrm{PAPR}_{0.1\%}$. Computational complexity is evaluated by counting the number of real-valued addition operations and assuming that the complexity of real-valued multiplication is approximately 3 times of real-valued addition [11]. The total complexity is summarized in Table 2. In addition, since we are considering only the complexity of phase rotation sequence estimation which is the major part of blind SLM receiver, the complexity shown in Table 2 is identical for both OFDM downlink and SC uplink transmissions. Note that the complexity of 2-step estimation is calculated based on 27-state trellis diagram [8].

Fig. 3 shows the $PAPR_{0.1\%}$ versus total computational complexity of OFDM downlink and SC uplink STBC-TD using blind SLM. PAPR reduces when M increases in all transmission schemes, but the total complexity also increases. The use of 2-step estimation can reduce the complexity while maintaining the same PAPR as that of conventional blind SLM in [5], but the complexity reduction capability becomes obvious only when M>64. The proposed phase rotation sequence estimation in Eq. (9) can reduce the complexity even when M≤64 due to less number of symbol candidates in the Euclidean distance calculation. However, the phase rotation set {0°,45°} reduces the PAPR reduction capability of blind SLM. This is because the narrow range of phase rotation makes the phase-rotated waveform remains similar to the original waveform and hence, the PAPR of phase-rotated waveform is not much different from the original one.

Meanwhile, the use phase rotation {0°,135°} together with the phase rotation sequence estimation based on Eq. (9) can keep PAPR the same as that of conventional blind SLM and with less computational complexity. Assuming M=64, the proposed blind SLM can lower the PAPR by 2.7 dB for OFDM downlink and 2.9 dB for SC uplink, respectively. Its complexity is only 35% of the conventional blind SLM with ML estimation [5] and 50% of the conventional blind SLM with 2-step estimation [8]. Note that the use of fourth-power constellation will also be able to reduce the complexity of 2-step estimation. We leave it as our future studies since it involves many design parameters such as the maximum number of states.

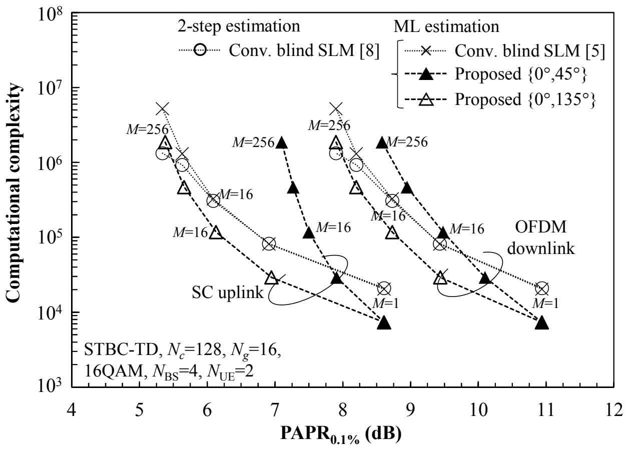


Fig. 3 $PAPR_{0.1\%}$-complexity tradeoff.

Table 2 Computational complexity per transmit block (16QAM).

	No. of real-valued multiplications	No. of real-valued additions
Conv. blind SLM (ML) [5]	$M\times(36N_c+1)$	$M\times(51N_c)$
Conv. blind SLM (2-step) [8]	$38\times N_{total\text{-}branch}$	$(51\times N_{total\text{-}branch})+MN_c$
Proposed blind SLM (ML)	$M\times(15N_c+1)$	$M\times(12N_c)$

Remarks: $N_{total\text{-}branch}$ is the number of branches used in Viterbi algorithm per one transmit block (maximum is $729\times N_c$)

4.2. BER performance

Figs. 4(a) and 4(b) shows the uncoded BER performance of STBC-TD with the proposed blind SLM using phase rotation sets {0°,45°} and {0°,135°}, together with the phase rotation sequence estimation using the fourth-power constellation, as a function of average received bit energy-per-noise power spectrum density (E_b/N_0). The BER performances of transmission using the conventional blind SLM [5] and without blind SLM are also plotted for comparison. ML phase rotation sequence estimation is assumed. It is seen that the use of blind SLM, both the conventional blind SLM in [5] and the proposed blind SLM, achieves the same BER performance compared to the transmission without SLM when the received E_b/N_0 is sufficiently high (for example E_b/N_0>0 dB).

The proposed blind SLM using phase rotation set {0°,45°} achieves worse PAPR performance than that of conventional SLM. On the other hand, the proposed blind SLM using phase rotation set {0°,135°} can maintain the PAPR reduction capability as the same as [5] but with less computational complexity in the phase rotation sequence estimation. As a result, the use of phase rotation set {0°,135°} with the phase rotation sequence estimation using the fourth-power constellation is more attractive.

5. Conclusion

A blind SLM technique consisting of 2-value phase rotation sets {0°,45°} or {0°,135°} and phase rotation sequence estimation using the fourth-power constellation was introduced in this paper. The use of fourth-power constellation can reduce the number of symbol candidates in minimum Euclidean distance computation, leading to computational complexity reduction. Computer simulation results assuming STBC-TD transmission confirmed that the proposed blind SLM with phase rotation sets {0°,135°} can reduce the PAPR of OFDM downlink (SC uplink) by 2.7 dB (2.9 dB) when M=64, while the computational complexity of phase rotation sequence estimation is only 35% of the conventional blind SLM using ML estimation, and 50% of 2-step estimation. It was also confirmed that there is no significant BER degradation compared to the transmission without blind SLM when the received E_b/N_0>0 dB.

In addition, the proposed blind SLM can be applied to MU-MIMO transmission without major modification. The performance evaluation of MU-MIMO and the study of

2-step phase rotation estimation based on the fourth-power constellation are also left as our future works.

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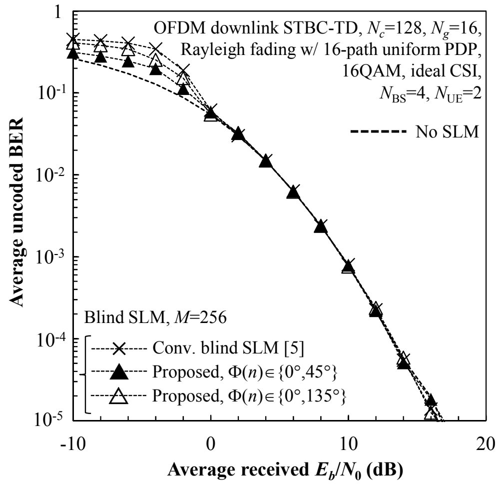
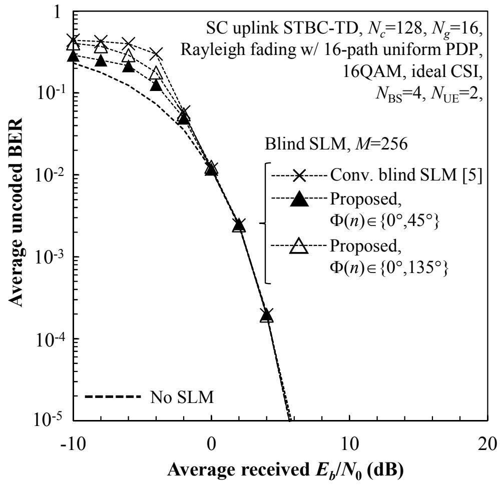


(a) OFDM downlink



(b) SC uplink

Fig. 4 BER performances.