Modified Blind Selected Mapping for OFDM/Single-carrier Signal Transmission

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***Abstract*—Blind selected mapping (blind SLM) is an effective peak-to-average power ratio (PAPR) reduction technique which does not require side information sharing. In the blind SLM, the receiver employs phase rotation sequence estimation, which can be carried out using maximum-likelihood (ML) estimation or 2-step sequence estimation using Viterbi algorithm. ML estimation typically requires much higher computational complexity. Recently, we showed that the use of codebook generated from a 2-level phase rotation set $\{0°, 135°\}$ and the ML phase rotation sequence estimation based on the fourth-power QAM constellation requires much less complexity compared to the conventional blind SLM with 3-level phase rotation set $\{0°, 120°, 240°\}$ and ML estimation based on original QAM constellation. However, for a large number of phase rotation sequences, the computational complexity still remains high. In this paper, in order to further reduce the computational complexity, we propose a combined use of a 2-level phase rotation set $\{0°, 135°\}$ and the 2-step sequence estimation. The use of 2-level phase rotation set significantly reduces the number of branches and states in the Viterbi algorithm and hence, leads to complexity reduction. Simulation results confirm that the blind SLM using the 2-level phase rotation set and the 2-step sequence estimation has less computational complexity while achieving similar BER to the ML sequence estimation.**

***Index Terms*—PAPR, OFDM, single carrier, SLM**

I. Introduction

The design of low peak-to-average power ratio (PAPR) waveforms remains important even in the fifth-generation (5G) mobile communication systems, especially for user equipments (UEs) [1]. Single carrier (SC) signals generally have lower PAPR than orthogonal frequency division multiplexing (OFDM) signals [2]. However, PAPR of SC signals increases due to transmit processing such as band-limit filtering and precoding [3], indicating that PAPR reduction is also necessary for SC transmission.

We have been studying a PAPR reduction technique called blind selected mapping (blind SLM) [4]. It can lower the PAPR effectively by multiplying the original data symbol sequence with a phase rotation sequence. The phase rotation sequence is selected from a codebook of random sequences generated from a set $\{0°, 120°, 240°\}$ [4]. The receiver needs to estimate the phase rotation sequence which has been used at the transmitter side in order to carry out data de-modulation. The blind SLM in [4] is compatible with both OFDM and SC signals. Its applications to multiple-input multiple-output (MIMO) transmissions, such as space-time block coded transmit diversity (STBC-TD) and multiuser MIMO (MU-MIMO), were discussed in [5] and [6], respectively.

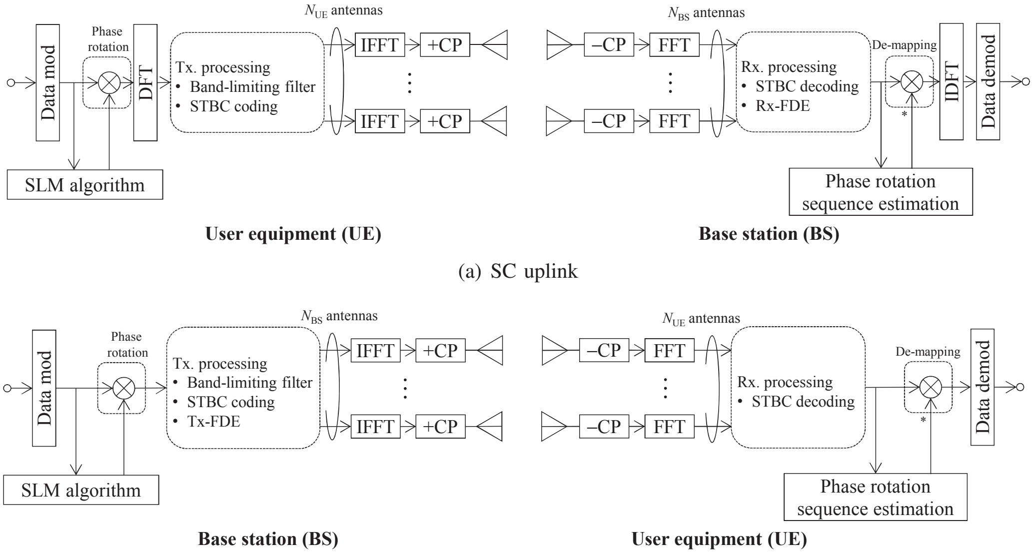
The phase rotation sequence estimation in [4-6] is the maximum likelihood (ML) estimation based on minimum Euclidean distance between the de-mapped symbols and the original signal constellation. It works effectively but requires high computational complexity. A 2-step phase rotation sequence estimation using Viterbi algorithm [7] was proposed to reduce the complexity, but its capability is obvious only when the number of phase rotation sequences is high. Recently, we showed that the use of codebook generated from a 2-level phase rotation set $\{0°, 135°\}$ and the ML phase rotation estimation based on the fourth-power QAM constellation [8,9] requires less computational complexity than those of [4-7]. This is because the number of signal points in the fourth-power constellation is much less than the original QAM, leading to reduced number of candidates in minimum Euclidean distance search. It is shown in [9] that the new ML phase rotation sequence estimation achieves less computational complexity while keeping the same bit error rate (BER) as the conventional blind SLM in [4-6]. However, since the complexity of the ML estimation in [9] exponentially increases as the number of phase rotation sequences increases, the complexity becomes higher than the 2-step estimation in [7].

Meanwhile, the 2-step sequence estimation using Viterbi algorithm in [7] was designed and evaluated using 3-level phase rotation set. The results of [7] encourage us to use Viterbi algorithm in the sequence estimation. Hence, in this paper, to further reduce the computational complexity, we apply 2-level phase set $\{0°, 135°\}$ to the blind SLM with 2-step sequence estimation using Viterbi algorithm and the fourth-power QAM constellation. The use of fourth-power constellation reduces the number of possible paths to be considered during survival path searching. Furthermore, the use of 2-level phase set significantly reduces the number of branches and states in the Viterbi algorithm compared to [7]. These reasons lead to complexity reduction. In this paper, the new blind SLM is called a *modified blind SLM* in short.

Performance evaluation of the modified blind SLM is carried out in terms of PAPR, BER and computational complexity by computer simulation assuming single-user transmission, both OFDM downlink and SC uplink, using single-antenna (SISO) and STBC-TD. The simulation results confirm that the blind SLM using the 2-level phase rotation set and the 2-step sequence estimation achieves low computational complexity even when the number of phase rotation sequences is high, while keeping the BER similar to the ML estimation.

II. Overview of conventional blind SLM in [4-7]

Here, we briefly describe the concept of blind SLM in [4-7]. For simplicity, STBC-TD signal representation is described, where the representation for SISO is obtained by setting the number of base station (BS) antennas (N_{BS}) and UE antennas (N_{UE}), STBC coding parameters J and Q [5] to be 1. The



(a) SC uplink

(b) OFDM downlink

Fig. 1. Transceiver system model with blind SLM.

number of transmit antennas (N_t) becomes N_{BS} for OFDM downlink and N_{UE} for SC uplink, respectively. 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. The transceiver models with blind SLM can be depicted by Fig. 1.

A. SLM algorithm

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

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

In STBC-TD transmission, information sequence to be transmitted is data-modulated and is divided into J blocks, obtaining the j-th block of N_c-length data symbol $\{d_j(n); n = 0 \sim N_c - 1, j = 0 \sim J - 1\}$. $\{d_j(n)\}$ is phase-rotated by the selected phase rotation sequence $\{\Phi_{\hat{m}(j)}(n); n = 0 \sim N_c - 1\}$, obtaining the phase-rotated block $\{d_{j,\hat{m}}(n); n = 0 \sim N_c - 1, j = 0 \sim J - 1\}$. $\{d_{j,\hat{m}}(n)\}$ is then transformed into frequency-domain components block $\{D_{j,\hat{m}(j)}(k); k = 0 \sim N_c - 1\}$ by N_c-point DFT. After that, $\{D_{j,\hat{m}(j)}(k)\}$ are passed through transmit processing such as STBC coding, obtaining the frequency-domain signal at the n_t-th transmit antenna ($n_t = 0 \sim N_t - 1$) and the q-th timeslot ($q = 0 \sim Q - 1$) as $\{S_{n_t,q,\hat{m}(j)}(k); k = 0 \sim N_c - 1\}$ and its time-domain signal after inverse DFT (IDFT) as $\{s_{n_t,q,\hat{m}(j)}(n); n = 0 \sim N_c - 1\}$. If we assume that N_{UE}=2 while N_{BS} is arbitrary, J=Q=2 and $S_{n_t,q,\hat{m}(j)}(k)$ can be described by the following matrix representations.

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

$$\begin{bmatrix} S_{0,0,\hat{m}(j)}(k) & S_{0,1,\hat{m}(j)}(k) \\ \vdots & \vdots \\ S_{N_{\text{BS}}-1,0,\hat{m}(j)}(k) & S_{N_{\text{BS}}-1,1,\hat{m}(j)}(k) \end{bmatrix} \quad \text{for downlink,}$$
$$= \sqrt{\frac{2E_s}{T_s}} \mathbf{W}_T(k) \begin{bmatrix} D_{0,\hat{m}(j)}(k) & -D^*_{1,\hat{m}(j)}(k) \\ D_{1,\hat{m}(j)}(k) & D^*_{0,\hat{m}(j)}(k) \end{bmatrix} \quad (3)$$

where $\mathbf{W}_T(k)$ is the transmit filtering [5]. E_s and T_s are symbol energy and symbol duration, respectively.

In SC uplink STBC-TD without transmit filtering (i.e., employing band-limiting filter only), the PAPR remains the same as original transmit block after applying STBC coding since the coding matrix employs only complex conjugate operation [6]. Hence, we can select a phase rotation sequence for each of $\{d_j(n)\}$ prior to applying STBC coding. The selected phase rotation sequence for the j-th data block, $\{\Phi_{\hat{m}(j)}(n)\}$ with the sequence number $\hat{m}(j)$, is determined by

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

where $\{\Phi_m(n); n = 0 \sim N_c - 1, m = 0 \sim M - 1\}$ is the m-th phase sequence in a codebook and is generated randomly as $\Phi_m(n) \in \{e^{i0}, e^{i3\pi/4}\}$ (equivalent to $\{0°, 135°\}$), except the first sequence is defined as $\{\Phi_0(n) = e^{j0}; n = 0 \sim N_c - 1\}$.

Meanwhile, Eq. (4) is not available for STBC-TD with transmit filtering due to matrix multiplication. In this case, selection of a phase rotation sequence which minimizes the maximum PAPR value (Mini-max) among all N_t (=N_{BS} for OFDM downlink) transmit antennas is used. The selected phase rotation sequence for all J data blocks, $\{\Phi_{\hat{m}(j)}(n)\}$ with the sequence number $\hat{m}(j) = \hat{m}$, is determined by

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

The selection in Eq. (5) is sub-optimal and therefore PAPR increases when N_t increases. However, the phase rotation estimation at the receiver is kept simple and no major changes on filtering weights calculation is required. Note that the criterion in Eq. (5) also can be used for MU-MIMO [6].

B. Phase rotation sequence estimation

After the receive signal processing, phase rotation sequence estimation is carried out by calculating Euclidean distance between the de-mapped signal (i.e. multiplied by $\{\Phi^*_{\hat{m}(j)}(n); n = 0 \sim N_c - 1\}$) 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 distance from the nearest QAM symbol is very small. The phase rotation sequence associated with the de-mapped signal having the minimum averaged distance is selected.

Assuming the j-th time-domain received block after employing receive processing (i.e., MMSE-FDE, STBC decoding and IDFT for SC uplink, and only STBC decoding for OFDM downlink) and before de-mapping is $\{\hat{d}_j(n); n = 0 \sim N_c - 1, j = 0 \sim J-1\}$, the estimated phase rotation sequence can be found by

$$\tilde{m}(j) = \arg\min_{m=0\sim M-1}\left(\sum_{n=0}^{N_c-1}\min_{\mathbb{C}\in\mathbf{\Psi}_{\text{mod}}}\left|\Phi_m^*(n)\hat{d}_j(n)-\mathbb{C}\right|^2\right), \quad (6)$$

where $\mathbf{\Psi}_{\text{mod}}$ is the original constellation (e.g. QAM mapping). Eq. (6) can be carried out based on either ML [4] or Viterbi algorithm [7]. Meanwhile, when the phase rotation set $\{0°, 135°\}$ is used, we can relax Eq. (5) by considering the distance between the fourth-power of the de-mapped symbols and the fourth-power of QAM symbols [8,9], which is

$$\tilde{m}(j)=\arg\min_{m=0\sim M-1}\left(\sum_{n=0}^{N_c-1}\min_{\mathbb{C}\in\mathbf{\Psi}_{\text{mod}}^4}\left|\Re\{(\Phi_m^*(n)\hat{d}_j(n))^4\}-\Re\{\mathbb{C}\}\right|\right), \quad (7)$$

where $\mathbf{\Psi}_{\text{mod}}^4$ is a set of fourth-power QAM constellation. The size of $\mathbf{\Psi}_{\text{mod}}^4$ is generally much less than that of $\mathbf{\Psi}_{\text{mod}}$ because of an existence of complex-conjugated pairs in $\mathbf{\Psi}_{\text{mod}}$. The above fact contributes to complexity reduction. Although Eq. (7) achieves low-complexity phase rotation sequence estimation, it still needs to compute the 2D-distance for all phase rotation sequences, then the computational complexity is high when M is large.

III. 2-Step Phase Rotation Estimation Using the Fourth-Power Constellation

A modification of 2-step sequence estimation by considering the fourth-power constellation is expected to keep complexity of blind SLM receiver low in every M. Here, we describe the modified 2-step sequence estimation by dividing this session to 2 parts; Viterbi algorithm and sequence verification. Since the phase rotation sequence estimation is done for one received block, we ignore the index j in STBC-TD for simplicity. Here below, the Viterbi algorithm is used to estimate the phase rotation only, where the data decision is not included.

A. Viterbi algorithm

Firstly, we define an objective function with objective metric ϵ based on Eq. (7). By neglecting the codebook and assuming that $\mathbf{\Phi}_{\text{SLM}}$ is a set of possible phase rotation $\{e^{i0}, e^{i3\pi/4}\}$, ϵ is expressed by

$$\epsilon = \sum_{n=0}^{N_c-1}\min_{\substack{\phi(n)\in\mathbf{\Phi}_{\text{SLM}}\\ \mathbb{C}\in\mathbf{\Psi}_{\text{mod}}^4}}\left(\left|\Re\{(\phi^*(n)\hat{d}(n))^4\}-\Re\{\mathbb{C}\}\right|\right), \quad (8)$$

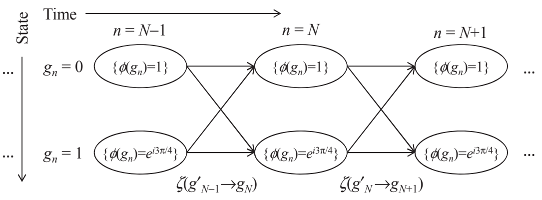


Fig. 2. Illustration of 2-state trellis.

Then, ϵ at time index $n = N$, $0 \le N \le N_c - 1$ is written by

$$\begin{aligned}\epsilon(N) &= \sum_{n=0}^{N}\min_{\substack{\phi(n)\in\mathbf{\Phi}_{\text{SLM}}\\ \mathbb{C}\in\mathbf{\Psi}_{\text{mod}}^4}}\left|\Re\{(\phi^*(n)\hat{d}(n))^4\}-\Re\{\mathbb{C}\}\right| \\ &= \epsilon(N-1)+\min_{\substack{\phi(N)\in\mathbf{\Phi}_{\text{SLM}}\\ \mathbb{C}\in\mathbf{\Psi}_{\text{mod}}^4}}\left|\Re\{(\phi^*(N)\hat{d}(N))^4\}-\Re\{\mathbb{C}\}\right|\end{aligned}. \quad (9)$$

By using Eq. (9), we can search an optimal phase rotation sequence $\{\Phi_{\text{opt}}(n); n = 0 \sim N_c - 1\}$ by using Viterbi algorithm [7,10]. We assume that g_n represents the g-th state ($g = 0 \sim G_{\max} - 1$) at the time index n. The first term and the second term in Eq. (9) can be considered as accumulated path metric entering a state g_N and a branch metric from g'_N to g_{N+1}, respectively. Here, we define the path metric and branch metric as $\epsilon(g_N)$ and $\zeta(g'_N \rightarrow g_{N+1})$. A trellis diagram when $G_{\max} = 2$ can be shown in Fig. 2 as an example.

The initial path metric of all states at sample index $n = -1$ is set as $\epsilon(g_{-1}) = 0$. At a time index $n = N$ where $0 \le N \le N_c - 1$, the branch metric is expressed by

$$\zeta(g'_N\rightarrow g_{N+1})=\min_{\mathbb{C}\in\mathbf{\Psi}_{\text{mod}}^4}\left|\Re\{(\phi^*(g_{N+1})\hat{d}(N+1))^4\}-\Re\{\mathbb{C}\}\right|. \quad (10)$$

Here, we change the phase rotation as a function of time index $\phi(N+1)$ to $\phi(g_{N+1})$, which is the phase rotation value stored in the state g at time $N + 1$. The path metric entering state g_{N+1} is selected by the following criterion.

$$\epsilon(g_{N+1}) = \min_{g'_N=0\sim G_{\max}-1}\left(\epsilon(g'_N) + \zeta(g'_N \rightarrow g_{N+1})\right). \quad (11)$$

Note that the selection of paths and branches in Eqs. (10) and (11) are repeated until $n = N_c-1$. Once the selection is done until $n = N_c-1$, the surviving path metric which corresponds to an optimal state number $g_{n,\text{opt}}$ and optimal phase sequence $\{\Phi_{\text{opt}}(n) = \phi(g_{n,\text{opt}}); n = 0 \sim N_c - 1\}$ can be determined by backward computation as follows.

$$g_{N_c-1,\text{opt}} = \arg\min_{g_{N_c-1}=0\sim G_{\max}-1}\epsilon(g_{N_c-1}), \quad (12a)$$

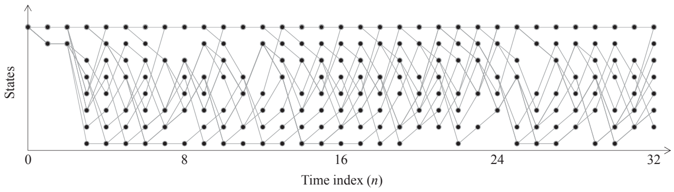
$$g_{n',\text{opt}} = \arg\min_{g_{n'}=0\sim G_{\max}-1}\left(\epsilon(g_{n'})+\zeta(g_{n'}\rightarrow g_{n'+1,\text{opt}})\right), \quad (12b)$$

where $n' = N_c-2, N_c-3, \ldots, 0$. Since the blind SLM uses 2-value phase rotation, $G_{\max}$ can be set as 2^P where P is the time memory of a state (i.e., a particular state g_n is determined as a set of phase rotation for time index $n-P-1, n-P-2, \ldots, n$. The algorithm in Eqs. (10)-(12) can be used without changes but it needs to start from index $n = P - 1$.

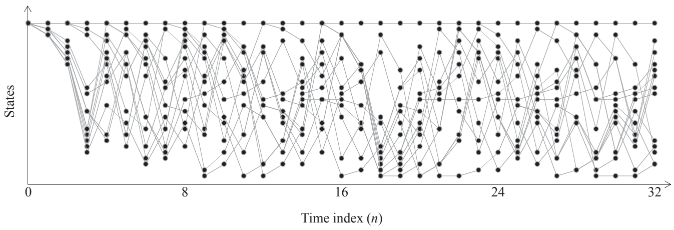
Besides constructing the trellis diagram based on Eqs. (8)-(12), we utilize the phase rotation sequence codebook and remove the redundant branches and states prior to the estimation [7]. Fig. 3 shows an example of a trellis constructed by setting $N_c = 32$ and $M = 16$, where we can observe that the trellis diagram is sparse and the use of 2-level phase set requires less number of branches and states than that of 3-level phase set, consequently requires lower complexity than the full trellis diagram. It was discussed in [7] that an increasing of $G_{\max}$ can improve the estimation accuracy, but also increases the complexity due to many surviving branches and states.

B. Verification and correction

The Viterbi algorithm in Sect.III-*A* is focusing at selecting the path and the corresponding phase rotation sequence which provide the lowest distance from the fourth-order constellation. Therefore, there exists probability that the resultant sequence



(a) 2-level phase rotation set with $G_{\max} = 8$



(b) 3-level phase rotation set with $G_{\max} = 27$

Fig. 3. An example of trellis diagram with reduced branches.

TABLE I
SIMULATION PARAMETERS

Modulation	Data modulation	16QAM
	FFT/IFFT block size	N_c = 256
	Cyclic prefix length	N_g = 16
Blind SLM parameter	Phase sequence type	Random polyphase
	No. of sequences	M = 1∼1024
	Phase sequence estimation method	Maximum-likelihood, 2-step estimation
	Oversampling factor	V = 8
Channel	Fading type	Frequency-selective block Rayleigh
	Power delay profile	symbol-spaced 16-path uniform
User equipment	No. of UE antennas	N_{UE} = 2
	Channel estimation	Ideal
Base station	No. of BS antennas	N_{BS} = 4
	Rx filter (uplink)	MMSE-FDE
	Tx filter (downlink)	Maximal ratio (MRT)-FDE
	Channel estimation	Ideal

$\{\Phi_{\text{opt}}(n); n = 0 \sim N_c - 1\}$ is not in the predefined codebook due to frequency-selective fading and noise [7]. Here, verification and correction are introduced for checking the similarity between $\{\Phi_{\text{opt}}(n)\}$ and the existing sequences in the codebook. We can use the Hamming distance as the indicator since there is no effect from the difference in rotation angle to the data detection error.

Let $\{\Phi_{\text{opt}}(n); n = 0 \sim N_c - 1\}$ be the resultant phase rotation sequence obtained from the Viterbi algorithm. The estimated phase rotation sequence to be used in de-mapping $\{\Phi_{\tilde{m}}(n); n = 0 \sim N_c - 1\}$, with the sequence number $\tilde{m}$, can be determined by

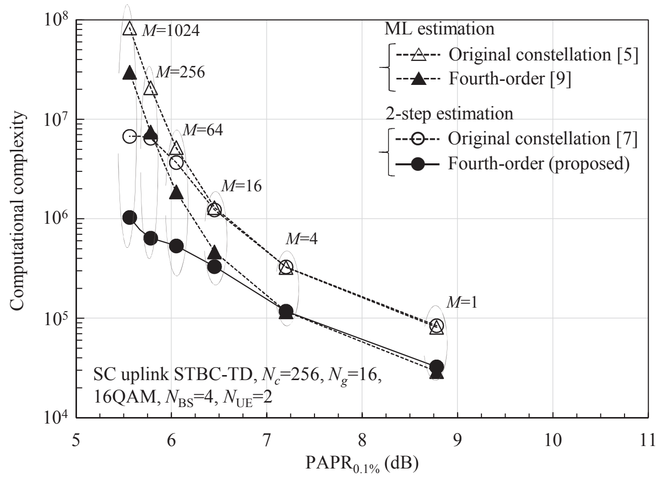
$$\begin{aligned} \tilde{m} &= \arg \min_{m=0\sim M-1} b\left(\{\Phi_{\text{opt}}(n)\}, \{\Phi_m(n)\}\right) \\ &= \arg \min_{m=0\sim M-1} \left(\sum_{n=0}^{N_c-1} \left(\Phi_{\text{opt}}(n) \oplus \Phi_m(n)\right) \right) \end{aligned}, \quad (13)$$

where $b\left(\{A\}, \{B\}\right)$ denotes the Hamming distance between sequence A and B and $\oplus$ is exclusive or operation [11]. Finally, the soft-decision data symbol after de-mapping and before de-modulation $\{\tilde{d}(n); n = 0 \sim N_c - 1\}$ is obtained by $\{\tilde{d}(n) = \Phi^*_{\tilde{m}}(n)\hat{d}(n); n = 0 \sim N_c - 1\}$. Note that when the STBC-TD is used, the estimated phase rotation sequence should be indexed with j as $\{\Phi_{\tilde{m}(j)}(n); n = 0 \sim N_c - 1\}$, with the corresponding sequence number $\tilde{m}(j), j = 0 \sim J-1$.

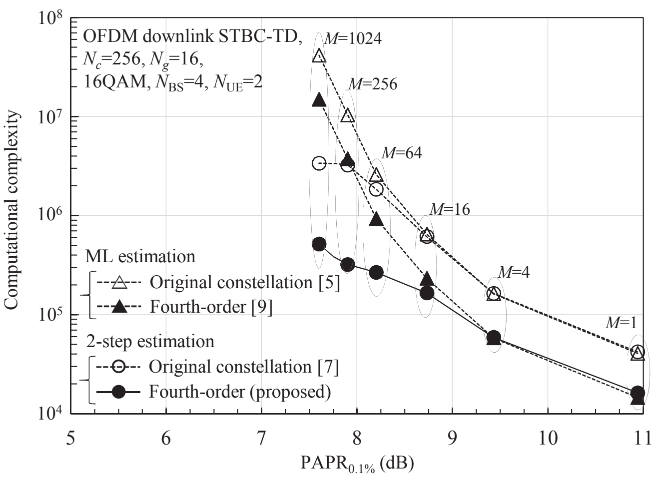
TABLE II
COMPUTATIONAL COMPLEXITY PER ONE TRANSMIT BLOCK (16QAM)

	No. of real-valued multiplications	No. of real-valued additions
ML estimation, original [5]	$M \times (36N_c + 1)$	$M \times (51N_c + 1)$
ML estimation, fourth-order [9]	$M \times (15N_c + 1)$	$M \times (12N_c + 1)$
2-step estimation, original [7]	$38 \times N_{tb}$	$(51 \times N_{tb}) + MN_c$
2-step estimation, fourth-order (Proposed)	$17 \times N_{tb}$	$(12 \times N_{tb}) + MN_c$

Remark: Identical for SC and OFDM, N_{tb} is the number of branches used in Viterbi algorithm for one received block (maximum is $(G_{\max})^2 \times N_c$).



(a) SC uplink



(b) OFDM downlink

Fig. 4. PAPR$_{0.1\%}$ versus computational complexity.

IV. PERFORMANCE EVALUATION

Simulation parameters are summarized in Table I. Single-user SISO or STBC-TD are assumed in this paper, while channel coding is not considered for simplicity. Path loss and shadowing loss are not considered. Phase rotation codebook are generated randomly as $\Phi_m(n) \in \{e^{i0}, e^{i2\pi/3}, e^{i4\pi/3}\}$ for conventional blind SLM using 3-level phase rotation set and the estimation based on original QAM mapping, and $\Phi_m(n) \in \{e^{i0}, e^{i3\pi/4}\}$ for the modified blind SLM. Performance evaluation is discussed and compared with the conventional blind SLM in [4-7] using either ML or 2-step phase rotation sequence estimation, but based on original QAM constellation.

A. PAPR vs. computational complexity

We consider the PAPR value at complementary cumulative distribution function (CCDF) is 10^{-3} and called $\text{PAPR}_{0.1\%}$. Computational complexity is defined by the number of real-valued addition operations and assuming that the complexity of real-valued multiplication is approximately 3 times of real-valued addition [12]. The total complexity of phase rotation sequence estimation is summarized in Table II. The complexity of 2-step estimation based on original constellation is calculated using Viterbi algorithm with $G_{\text{max}} = 27$ [7]. In addition, the complexity of modified blind SLM is evaluated at $G_{\text{max}} = 8$.

Fig. 4 shows the $\text{PAPR}_{0.1\%}$ versus total computational complexity of STBC-TD using the modified blind SLM and the conventional blind SLM in [5], [7] and [9]. Transmission scheme with the tradeoff mark in the bottom-left of Fig. 4 means it provides low PAPR with low-complexity receiver. PAPR reduces when M is large in all schemes, but the total complexity also increases. The use of 2-step estimation with original constellation can reduce the complexity while maintaining the same PAPR as that of conventional blind SLM with ML estimation in [5], but the complexity reduction capability is obvious when $M > 64$.

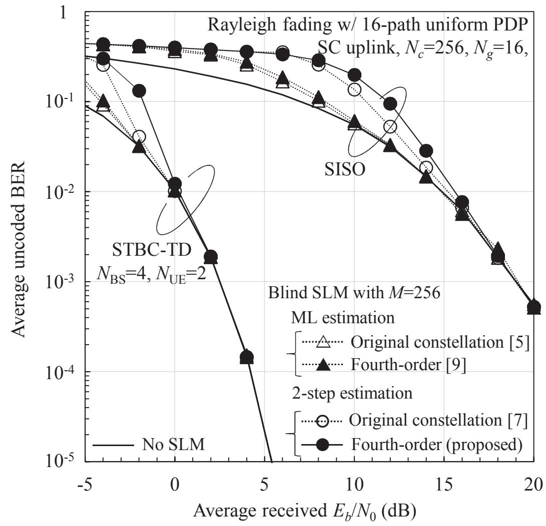
The use of ML estimation with fourth-power constellation [9] can reduce the complexity even when $M \leq 64$ due to less number of signal points in the minimum Euclidean distance calculation. However, the complexity of ML estimation using fourth-order constellation becomes higher than 2-step estimation using original constellation when $M > 64$. This is because the complexity of ML estimation is a function of M, while the complexity of 2-step estimation mostly depends on N_{tb}, which is almost constant when M is large.

The modified blind SLM can significantly reduce computational complexity at the receiver compared to the conventional blind SLM for the given PAPR. The use of 2-level phase set can set G_{max} to be less than that of 3-level phase set at the same number of time memory (i.e. $3^3 = 27$ for 3-level phase set but only $2^3 = 8$ for 2-level phase set). Moreover, the use of fourth-order constellation reduces the number of signal points considered in branch metric calculation and consequently contributes to complexity reduction [8,9].

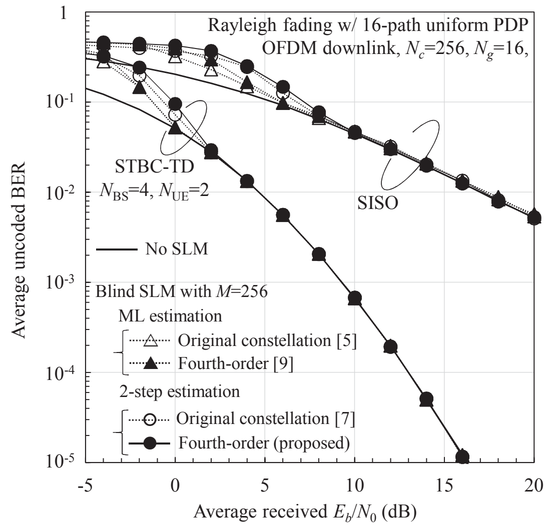
For a given PAPR reduction of 3 dB from the OFDM/SC transmissions without SLM (equivalent to $M = 256$), the required computational complexity of phase rotation estimation in the modified blind SLM is only 3% of the ML estimation using original constellation [5], 9% of the ML estimation using fourth-order constellation [9], and 10% of the 2-step estimation using Viterbi algorithm based on the original constellation [7]. The above result is identical for both OFDM downlink and SC uplink transmissions.

B. BER

Fig. 5 shows the average uncoded BER performance of SISO and STBC-TD with $N_{\text{BS}} = 4$, $N_{\text{UE}} = 2$ and equipped with blind SLM as a function of average received bit energy-per-noise power spectrum density (E_b/N_0). The BER performances of transmission without blind SLM and the conventional blind SLM in [5], [7] and [9] are also plotted for comparison. The number of available phase rotation sequences is set to be $M = 256$.



(a) SC uplink



(b) OFDM downlink

Fig. 5. Uncoded BER performance.

The use of blind SLM with phase rotation sequence estimation achieves worse BER than that of without blind SLM in every scheme when the E_b/N_0 is low, i.e., $E_b/N_0 < 0$ dB (2 dB) for SC uplink (OFDM downlink) STBC-TD. This is because the impact from fading and noise leads to the difficulty in classification between the symbols obtained from correct de-mapping and that of incorrect de-mapping. However, there is no significant BER degradation when either the conventional blind SLM in [5], [7], and [9], or the modified blind SLM is used. This emphasizes the attractiveness of the modified blind SLM since it achieves low-complexity estimation without significant degradation on BER and PAPR.

V. Conclusion

In this paper, we aim at achieving a low-complexity phase rotation sequence estimation even when M is large. We introduced a modified blind SLM using the 2-level phase rotation set $\{0°, 135°\}$ and the 2-step sequence estimation based on the fourth-power constellation. Our modified blind SLM with the 2-level phase rotation set and the 2-step sequence estimation using Viterbi algorithm and the fourth-order constellation achieves low computational complexity due to the following reasons; reduction of branches and states in Viterbi algorithm (contributed by the use of 2-level phase set) and reduction of possible paths to be considered during survival path searching (contributed by the use of fourth-order constellation). Simulation results assuming N_c=256 and $G_{\max}$=8 confirmed that our modified blind SLM can significantly reduce the computational complexity at the receiver to be only 3% of the conventional blind SLM for a given PAPR reduction capability of 3 dB. It was also confirmed that there is no significant BER degradation compared to the transmission without blind SLM when the received $E_b/N_0 > 2$ dB for STBC-TD transmission.

In addition, our modified blind SLM can be applied to MU-MIMO transmission (both SC uplink and OFDM downlink) without major modification.

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