A New Blind SLM Scheme with Low Complexity of OFDM Signals

Hyun-Seung Joo, Seok-Joong Heo, Hyun-Bae Jeon, Jong-Seon No
Dept. of Electrical Eng. and Computer Science, INMC
Seoul National University, Seoul 151-744, Korea
email: {joohs,hsjbest,lucidream}@ccl.snu.ac.kr, jsno@snu.ac.kr

Dong-Joon Shin
Dept. of Electronics and Computer Eng.
Hanyang University, Seoul 133-791, Korea
email: djshin@hanyang.ac.kr

***Abstract*—In this paper, we propose a new blind SLM scheme with low complexity using m-sequence as a phase sequence in OFDM. The proposed scheme significantly reduces the computational complexity for searching the side information of the phase sequence and decoding the alternative symbol sequence at the receiver. After generating alternative symbol sequences using a set of phase sequences, the side information for each alternative symbol sequence is embedded in it through block partitioning and phase rotation. The proposed method does not need additional inverse fast Fourier transform (IFFT) and has the same PAPR reduction performance compared to the conventional SLM scheme. In order to find the side information, a maximum likelihood (ML) decoder with lower complexity is derived, which guarantees lower detection failure probability of side information compared to the conventional blind SLM scheme.**

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is an efficient method for high speed data transmission in the multipath fading. In addition, it is possible to significantly enhance the throughput by adapting the data rate per subcarrier from decomposition of a wideband channel. However, a major disadvantage of OFDM is the high peak to average power ratio (PAPR) of the transmitter's output signal, where the range of PAPR is proportional to the number of subcarriers. Due to the high PAPR feature, an OFDM signal may suffer from significant inter-modulation and undesired out-of-band radiation when it passes through nonlinear devices, e.g., a high power amplifier (HPA) [1] - [3].

Several PAPR reduction schemes have been proposed such as clipping and filtering [1], coding [4], selected mapping (SLM), partial transmit sequence (PTS) [5], and tone reservation (TR) [6]. SLM, one of symbol scrambling techniques, can reduce the peak power of OFDM signals without signal distortion. The key idea of SLM is that the OFDM signal with the smallest PAPR is selected for transmission from several alternative OFDM signals which are obtained by applying inverse fast Fourier transform (IFFT) to each alternative symbol sequence that is an input symbol sequence multiplied by one of the phase sequences.

In SLM, the side information must be transmitted to enable the receiver to search the original OFDM symbol sequence. Such side information causes slight degradation in bandwidth efficiency. Moreover, wrong detection of side information at the receiver results in critical degradation of bit error rate (BER). For this reason, the side information must be highly protected not to affect the error performance of OFDM systems. Several blind SLM (BSLM) schemes to eliminate the need for side information have been studied [7] - [10]. An ML decoder is derived for the BSLM scheme using random phase sequences [8], which shows good BER performance. But, the conventional BSLM schemes result in large decoding complexity.

In this paper, we propose a new BSLM scheme with low decoding complexity. The proposed scheme reduces detection failure of the side information compared to the conventional BSLM scheme. After generating alternative symbol sequences through multiplying an input symbol sequence by phase sequences, the side information for each alternative symbol sequence is embedded in it through block partitioning and phase rotation. A maximum likelihood (ML) decoder with low computational complexity is derived to find the side information at the receiver. The proposed scheme gives negligible detection failure probability of side information and has the same PAPR reduction performance compared to the conventional SLM scheme.

II. CONVENTIONAL BLIND SLM SCHEME

In the OFDM systems, an input symbol sequence $\mathbf{X} = [X_0, X_1, \cdots, X_{N-1}]$ is given as a vector of complex-valued symbols with the time duration T_s. After splitting the serial data into parallel data streams, all substreams are summed by applying IFFT. The discrete time OFDM signal after IFFT is given as

$$x_k = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} X_n e^{j2\pi \frac{n}{N} k}, \qquad 0 \le k \le N-1 \qquad (1)$$

where X_n is the input data symbol loaded on the n-th subcarrier and N is the number of the subcarriers.

For an input symbol sequence $\mathbf{X}$, SLM generates U independent alternative symbol sequences and transmits the OFDM signal with the minimum PAPR. U alternative symbol sequences are generated by multiplying $\mathbf{X}$ by U phase sequences. Suppose that U phase sequences are given by

$$\mathbf{P}^u = [e^{j\phi_0^u}, e^{j\phi_1^u}, ..., e^{j\phi_{N-1}^u}] \qquad (2)$$

where $\phi_n^u \in [0, 2\pi)$ and $u \in \{1, 2, ..., U\}$. Let $\mathbf{a} \otimes \mathbf{b}$ represent the componentwise multiplication of vectors $\mathbf{a}$ and $\mathbf{b}$. For an input symbol sequence $\mathbf{X}$, $\mathbf{X} \otimes \mathbf{P}^{\tilde{u}}$ with the minimum PAPR among U alternative symbol sequences $\mathbf{X} \otimes \mathbf{P}^u$, $1 \le u \le U$, is selected for transmission. The index $\tilde{u}$ of the selected phase sequence should be transmitted to the receiver in the SLM scheme.

In order to eliminate transmission of the side information, BSLM scheme where the indices are embedded into alternative symbol sequences without loss of data rate was proposed in [8]. BSLM enables the receiver to distinguish the selected phase sequence from others without transmitting any side information. We assume that the receiver receives $\mathbf{R} = \mathbf{X} \otimes \mathbf{P}^{\tilde{u}} + \hat{\mathbf{N}}$ and computes $\mathbf{R} \otimes \mathbf{P}^{u*}$ for $u = 1, 2, ..., U$, where $\mathbf{R} = [R_0, R_1, \cdots, R_{N-1}]$ is a received symbol sequence, $\hat{\mathbf{N}} = [\hat{N}_0, \hat{N}_1, \cdots, \hat{N}_{N-1}]$ is a noise vector, and $\mathbf{P}^{u*}$ is the complex conjugate of $\mathbf{P}^u$. To derive the input symbol sequence without the side information $\tilde{u}$ from $\mathbf{R}$ at the receiver, U phase sequences in the BSLM scheme should have the following properties:

- The set of $\mathbf{P}^u$'s is fixed and known a priori;
- $\mathbf{X} \otimes \mathbf{P}^u$ and $\mathbf{X} \otimes \mathbf{P}^v$ are sufficiently different for $u \neq v$.

To decode the input symbol sequence at the receiver, each phase ϕ_n^u for all n and u must satisfy the condition of $X_n e^{j\phi_n^u} \notin \mathcal{Q}$, where $\mathcal{Q}$ is a given symbol constellation for X_n. The set of $\mathbf{P}^u$ should be chosen readily to ensure this condition. If $u \neq \tilde{u}$ and channel noise is not considered from decoding OFDM symbol, each element of $\mathbf{R} \otimes \mathbf{P}^{u*}$ will not be a symbol in the constellation $\mathcal{Q}$. Therefore, by using this result, a simplified ML decoder for the BSLM scheme can be derived to recover OFDM symbol without the side information [8]. Specially, when the Euclidian distance between any $\mathbf{P}^u$ and $\mathbf{P}^v$ is very large, BER performance of the ML decoder is expected to be very good.

The received symbol R_n after the FFT demodulation at the receiver can be written as

$$R_n = G_n X_n e^{j\phi_n^{\tilde{u}}} + \hat{N}_n \tag{3}$$

where G_n is the frequency response of the fading channel at the nth subcarrier and $\hat{N}_n$ is an additive white complex Gaussian noise (AWGN) sample at the nth subcarrier. Without the side information of $\tilde{u}$, the optimal ML decoder computes the decision metric for decoding the received symbol sequence. The optimal metric of the ML decoder is given as

$$D_{opt} = \min_{\substack{[\hat{X}_0, \hat{X}_1, \ldots, \hat{X}_{N-1}] \subseteq \mathbf{Q}^N \\ \mathbf{P}^u, u \in \{1,2,\ldots,U\}}} \sum_{n=0}^{N-1} |R_n e^{-j\phi_n^u} - G_n \hat{X}_n|^2 \tag{4}$$

where $|\cdot|$ denote the absolute value of a complex number. This algorithm has to search all q^N symbols for q-ary constellation that is repeated for each of $\mathbf{P}^1, \mathbf{P}^2, \ldots, \mathbf{P}^U$. Consequently, the overall computational complexity to compute (4) is Uq^N $|\cdot|^2$ operations. Since the computational complexity from real additions is negligible compared to that of $|\cdot|^2$ operations, the additions will be ignored. Since the decoder exhaustively searches all symbol sequences, it can be performed only for small N. Thus, a suboptimal decoding method with reduced complexity should be derived.

Assume that R_n is detected into the nearest constellation point $\hat{X}_n$. That is, a soft decision value is made for each subcarrier unlike the optimal ML decoder and the nearest constellation point from R_n is saved for each subcarrier. This process is repeated for $1 \le u \le U$. Therefore, the suboptimal decision metric of the ML decoder [8] can be written as

$$D_{so} = \min_{\mathbf{P}_u, u \in \{1,2,\ldots,U\}} \sum_{n=0}^{N-1} \min_{\hat{X}_n \in Q} |R_n e^{-j\phi_n^u} - G_n \hat{X}_n|^2. \tag{5}$$

The input symbol sequence is recovered from the symbol sequence having D_{so}. Though the complexity of suboptimal decoder is reduced to UqN, it is sill not practical. Thus, we propose a new BSLM scheme with low computational complexity in the next section.

III. New Blind SLM Scheme

Instead of randomly selecting each element ϕ_n^u of the phase sequence, we use an m-sequence and its cyclic shifts with zero padding as the phase sequences $\mathbf{P}^u = [P_0^u, P_1^u, \cdots, P_{N-1}^u]$, where $P_i^u \in \{+1, -1\}$, $0 \le i \le N-1$, and $u \in \{1, 2, \cdots, U\}$ in the proposed BSLM scheme. In fact, $\mathbf{P}^u$ corresponds to rows of cyclic Hadamard matrix constructed from m-sequences. It is known that m-sequences satisfy the optimal condition for the phase sequence in the SLM schemes [11].

In order to embed the side information of the phase sequences into alternative symbol sequences without loss of data rate, block partitioning and phase rotation are used in the proposed BSLM schemes as follows. Assume that U phase sequences are used. The u-th alternative symbol sequence $\mathbf{X} \otimes \mathbf{P}^u$ is divided into L subblocks as $[\mathbf{X}_1^u, \mathbf{X}_2^u, \cdots, \mathbf{X}_L^u]$, where $\mathbf{X}_v^u = [X_{\frac{N(v-1)}{L}} P_{\frac{N(v-1)}{L}}, X_{\frac{N(v-1)}{L}+1} P_{\frac{N(v-1)}{L}+1}, \cdots, X_{\frac{Nv}{L}-1} P_{\frac{Nv}{L}-1}]$ is the v-th subblock with size N/L. U L-tuple phase rotation vectors are defined as

$$\mathbf{W}^u = [w_1^u, w_2^u, \cdots, w_L^u] \tag{6}$$

where $1 \le u \le U$ and $w_v^u \in \{0, 1\}$. Each element in the v-th subblock of the u-th alternative symbol sequence is multiplied by $e^{\theta \cdot w_v^u}$, where $0 < \theta \le \pi/2$ and $1 \le v \le L$. In other words, the phase of each element in the v-th subblock of the u-th alternative symbol sequence is rotated by $\theta \cdot w_v^u$. It is equivalent to use two signal constellations such that each element in the subblocks with $w_v^u = 0$ is modulated by using the signal constellation $\mathbf{S}_0$ and each element in the subblocks with $w_v^u = 1$ is modulated by using the signal constellation $\mathbf{S}_1$ which is obtained by rotating the signals in $\mathbf{S}_0$ by θ. Then the u-th alternative symbol sequence can be written as

$$\mathbf{X}^u = [\mathbf{X}_1^u e^{j\theta w_1^u}, \mathbf{X}_2^u e^{j\theta w_2^u}, \cdots, \mathbf{X}_L^u e^{j\theta w_L^u}]. \tag{7}$$

Now, we have to derive the optimal θ and construct U L-tuple phase rotation vectors which maximize the detection probability of the index of the selected phase sequence at the receiver. It is not difficult to derive the following design criteria for the phase rotation vectors:

- The size of subblock should be as large as possible;
- The Euclidian distance between alternative symbol sequences obtained by applying phase rotation vectors should be as large as possible.

Then U biorthogonal vectors of length $L = U/2$ satisfy the above criteria. Note that U biorthogonal vectors of length $U/2$ are $U/2$ orthogonal vectors of length $U/2$ and their bit-complement vectors. Actually, $L = log_2 U$ gives the largest size of subblock which satisfies the first criterion. But, U rotation vectors of length $L = log_2 U$ do not satisfy the largest Euclidian distance between alternative symbol sequences obtained by applying rotation vectors compared to U biorthogonal vectors of length $L = U/2$. Since the Euclidian distance between alternative symbol sequences is more important than the size of subblock, we will use U biorthogonal vectors of length $U/2$ as rotation vectors. On the other hand, θ must be chosen such that the minimum Euclidean distance between $\mathbf{S}_0$ and $\mathbf{S}_1$ can be maximized. For QPSK, we choose $\theta = \pi/2$.

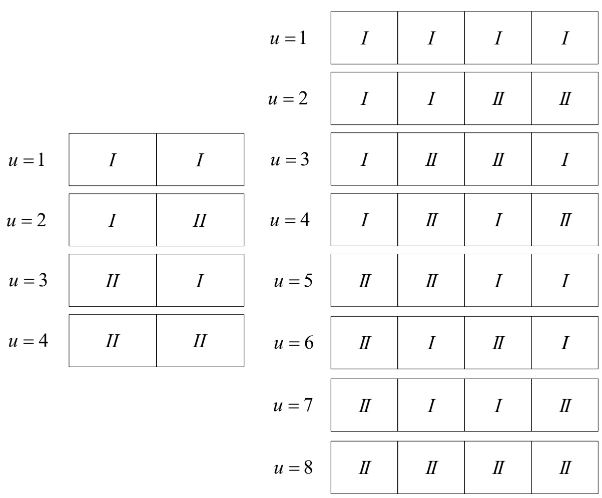

Fig. 1. Examples of block partitioning and phase rotation to embed side information for $U = 4$ and 8.

The biorthogonal vectors for $U = 4$ and 8 are given as

$$\text{For } U = 4,\ \{(0\ 0), (0\ 1), (1\ 0), (1\ 1)\}.$$

$$\begin{aligned} \text{For } U = 8,\ \{&(0\ 0\ 0\ 0), (0\ 0\ 1\ 1), \\ &(0\ 1\ 1\ 0), (0\ 1\ 0\ 1), \\ &(1\ 1\ 0\ 0), (1\ 0\ 1\ 0), \\ &(1\ 0\ 0\ 1), (1\ 1\ 1\ 1)\}. \end{aligned} \tag{8}$$

Fig. 1 shows the examples of the block partitioning by U biorthogonal vectors of length $L = U/2$ with $U = 4$ and 8, where I and II correspond to $w_v^u = 0$ and $w_v^u = 1$, respectively.

We derive a ML decoder with low complexity instead of the conventional BSLM decoder. Specially, two metrics for each subblock corresponding to $\mathbf{S}_0$ and $\mathbf{S}_1$ should be derived.

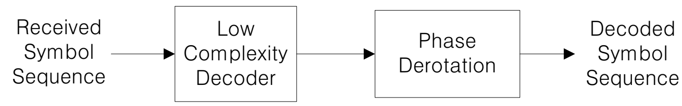

Fig. 2. Block diagram of the receiver for new BSLM.

Let R_n be detected into the nearest constellation point on $\mathbf{S}_0$ similarly to the suboptimal ML decoder of the conventional BSLM scheme. In other words, a soft decision value for each subcarrier is calculated and the nearest constellation point from R_n is saved for each subcarrier. This process is repeated for $1 \leq v \leq L$. On the other hand, the points on the signal constellation $\mathbf{S}_1$ must be considered by the ML decoder. Using R_n derotated by θ, this process is also repeated for $1 \leq v \leq L$ for decoding the symbols on $\mathbf{S}_1$. Therefore, the metric of the ML decoder in each subblock can be written as

$$D_{v,w_v^u} = \sum_{n=\frac{2N(v-1)}{U}}^{\frac{2Nv}{U}-1} \min_{\hat{X}_n \in \mathbf{S}_0} |R_n e^{-j\theta w_v^u} - G_n \hat{X}_n|^2 \tag{9}$$

where $w_v^u \in \{0, 1\}$ and $v \in \{1, 2, \cdots, \frac{U}{2}\}$. Two different metrics for each subblock in the proposed BSLM scheme are calculated according to w_v^u, that is, $D_{v,0}$ and $D_{v,1}$.

The ML decoder chooses the minimum total metric summed by $U/2$ metrics of all subblocks from U biorthogonal vectors. The minimum Euclidian distance solution yields the alternative symbol sequence. Therefore, the decision metric of the ML decoder is given as

$$D_{\bar{u}} = \min_{W^u} \sum_{v=1}^{\frac{U}{2}} D_{v,w_v^u}. \tag{10}$$

After finding the index $\bar{u}$ and yielding the alternative symbol sequence $\hat{\mathbf{X}} = [\hat{X}_0, \hat{X}_1, \cdots, \hat{X}_{N-1}]$, the input symbol sequence is determined by $\hat{\mathbf{X}} \otimes \mathbf{P}^{\bar{u}}$. Fig. 2 shows the block diagram of the ML decoder for the proposed BSLM.

The value of U affects detection failure probability (DFP) of the side information. DFP is the probability that the receiver cannot recover the side information. If U increases, DFP of both BSLM schemes is degraded in low signal-to-noise ratio (SNR). As the conventional BSLM scheme uses large U, the Euclidian distance between the used phase sequence and the others is reduced. In this case, the search for the selected phase sequence is difficult and the phase sequence detection may be frequently failed. Similarly, large U degrades DFP of the proposed BSLM scheme for the same reason. Due to the effect of block partitioning and phase rotation, DFP of the proposed BSLM scheme is better than that of the conventional BSLM scheme in low SNR. It will be shown from the simulation results in the next section that the proposed BSLM scheme has good performance of the side information detection over the AWGN channels.

Regardless of U, total complexity of the proposed BLSM is only $2qN\ |\cdot|^2$ operations in the OFDM system of q-ary modulation. The decoding complexity reduction ratio (DCRR)

TABLE I
COMPARISON OF DECODING COMPLEXITY OF THE CONVENTIONAL BSLM AND THE PROPOSED BSLM WHEN $U = 4,\ 8$, AND 16.

	$U=4$	$U=8$	$U=16$
Conventional Blind SLM	$4qN$	$8qN$	$16qN$
New Blind SLM	$2qN$	$2qN$	$2qN$
DCRR	50%	75%	87.5%

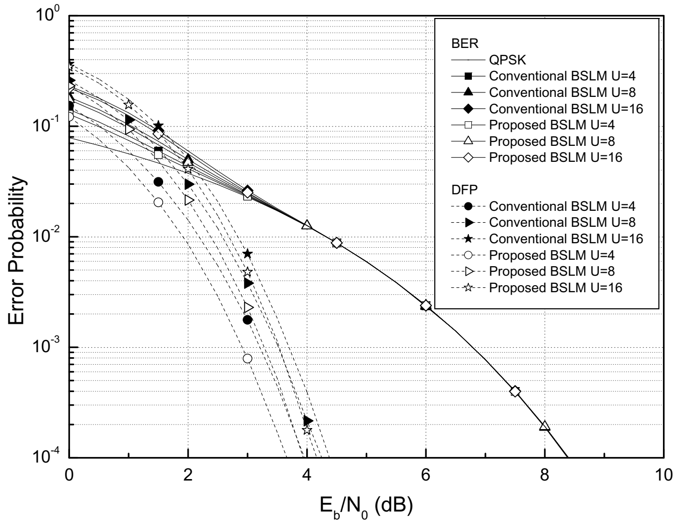

Fig. 3. Comparison of DFP and BER of the proposed BSLM and the conventional BSLM with QPSK and $N = 64$.

of the proposed BSLM over the conventional BSLM is defined as

$$\begin{aligned} \mathrm{DCRR} &= \left(1-\frac{\mathrm{A}}{\mathrm{B}}\right)\times 100 \\ &= \left(\frac{UqN-2qN}{UqN}\right)\times 100 \\ &= \left(\frac{U-2}{U}\right)\times 100 \quad (\%) \end{aligned} \tag{11}$$

where A is the decoding complexity of the conventional BSLM and B is the decoding complexity of the proposed BSLM. Table I compares the decoding complexity of the conventional BSLM and the proposed BSLM when $U = 4,\ 8$, and 16.

IV. SIMULATION RESULTS

The simulation is done in the AWGN channel, QPSK with phase rotator $\theta = \pi/4$, and $N = 64$ and 256. Also, $U = 4,\ 8$, and 16 are considered. The decoding algorithm using (9) and (10) is used for the proposed BSLM scheme.

A. DFP

Figs 3 and 4 show DFP of the conventional and the proposed BSLM schemes with $N = 64,\ 256$, and QPSK. DFP is the probability that the receiver cannot recover the side information. For small N, such as 64, there exists noticeable DFP for both BSLM schemes at low SNR region because the length is too short to obtain reliable metrics from subblocks.

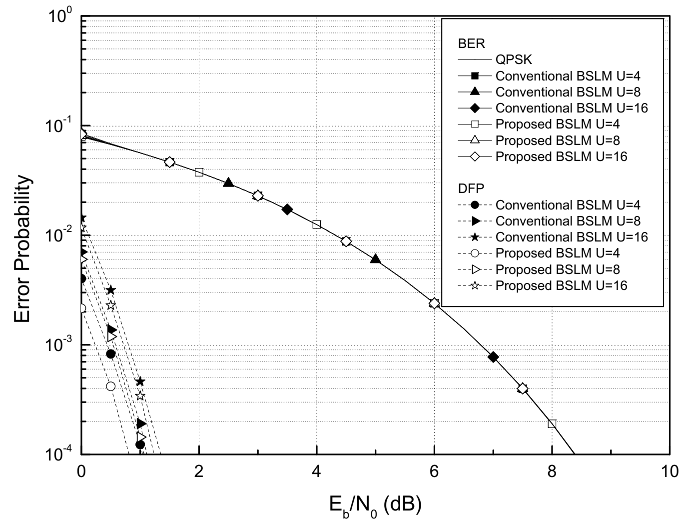

Fig. 4. Comparison of DFP and BER of the proposed BSLM and the conventional BSLM with QPSK and $N = 256$.

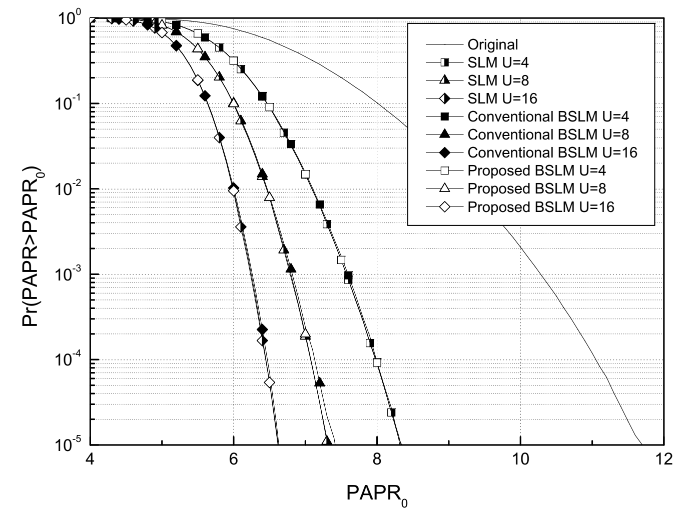

Fig. 5. Comparison of PAPR reduction performance of the proposed BSLM and the conventional BSLM with QPSK and $N = 64$.

When $N = 256$, DFP decreases fast and no detection failure of side information for $N = 1024$ is observed in the proposed BSLM.

B. BER

Figs 3 and 4 also compare the BER performance of the conventional and the proposed BSLM schemes with $N = 64,\ 256$, and QPSK. The total BER P_e at the receiver is given by

$$P_e = P_b \cdot (1 - P_{df}) + P_{b|df} \cdot P_{df} \tag{12}$$

where P_b is the BER when correct detection of side information is performed, $P_{b|df}$ is the BER when the detection of side information is failed, and P_{df} is the DFP [10]. The increased erroneous side information detection degrades total BER. When detection of the side information is failed, half of the bits is the errors in QPSK ($P_{b|df} = 0.5$). But, the BER degradation exists in low SNR for small N and large U. And P_{df} converges to 0 at the BER of 10^{-2} in both BSLM schemes. Since practical OFDM systems require better BER than 10^{-2}, the BER degradation from the detection failure in low SNR can be negligible. In addition, the BER in both BSLM schemes with $N = 256$ is not degraded from the detection failure which is almost negligible.

C. PAPR Performance

Fig. 5 with $N = 64$ shows that the proposed BSLM scheme with the phase rotation and block partitioning has the same PAPR reduction performance compared with that of the conventional SLM scheme.

V. CONCLUSION

In this paper, we proposed a new BSLM with low computational complexity for PAPR reduction of OFDM signals. The proposed BSLM scheme uses block partitioning and phase rotation in order to embed side information into the alternative symbol sequence. The proposed BSLM scheme which has low computational complexity for detecting side information has the same BER performance compared with the conventional BSLM scheme. Performance of the PAPR reduction is also the same as that of the conventional SLM scheme in the AWGN channel.

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