Selective Mapping for Broadband Single-Carrier Transmission Using Joint Tx/Rx MMSE-FDE

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***Abstract*—Joint transmit/receive frequency-domain equalization based on minimum mean-square error criterion (joint Tx/Rx MMSE-FDE) is a promising frequency-domain-based technique suppressing inter-symbol interference (ISI) in broadband single-carrier (SC) transmission. However, it changes the frequency-domain spectrum shape and hence increases the peak-to-average power ratio (PAPR) of transmit signal. In this paper, we introduce the selective mapping (SLM), which is typically used in orthogonal frequency division multiplexing (OFDM), to reduce PAPR of SC transmission using joint Tx/Rx MMSE-FDE. The SLM is applied to the frequency-domain SC signal after applying the transmit FDE (Tx-FDE) and prior to inverse fast Fourier transform (IFFT). It is shown, by computer simulation, that the SLM can reduce PAPR without significant effect on bit-error rate (BER). Performance of SLM is also evaluated in various phase-rotation sequence types, in both deterministic and random points of views, and number of candidates. Computational complexity of the proposed transmission scheme employing SLM algorithm is also discussed.**

Keywords*— Single-carrier (SC) transmission; selective mapping (SLM); frequency-domain equalization (FDE); peak-to-average power ratio (PAPR)

I. INTRODUCTION

Broadband wireless channel is characterized as a frequency-selective fading channel, in which inter-symbol interference (ISI) degrades system performance in terms of bit-error rate (BER) [1]. Orthogonal frequency division multiplexing (OFDM) is robust against fading, but its high peak-to-average power ratio (PAPR) property is the main drawback. On the other hand, single-carrier (SC) transmission [2] is more attractive for uplink communication in LTE-Advanced (LTE-A) system because of lower PAPR compared to OFDM, while the use of frequency-domain equalization (FDE) can effectively suppress the impact of ISI [3].

SC signal can be generated by inserting discrete Fourier transform (DFT) into conventional OFDM transmitter [4]. By using this approach, frequency-domain signal processing, where transmit and receive filtering can be done as simple one-tap multiplication, is approachable. Flexibility of signal manipulation, e.g. mapping, is also able to be simply implemented. An example of such proposed algorithms taking advantage of the use of frequency-domain processing is SC transmission using joint transmit/receive frequency-domain equalization based on minimum mean-square error (MMSE) criterion (joint Tx/Rx MMSE-FDE) [5].

Joint Tx/Rx MMSE-FDE has been introduced as a promising technique suppressing ISI by pre-multiplying transmit FDE (Tx-FDE) weight to the frequency-domain signal to be transmitted. As a merit of ISI suppression, SC transmission using joint Tx/Rx MMSE-FDE provides better BER performance compared to the original SC, however, it also increases PAPR of the transmit signal because of changes in frequency-domain spectrum shape [6]. This implies necessity of PAPR reduction algorithm.

There exist many proposed PAPR reduction algorithms for OFDM such as clipping [7], pulse shaping [8], and tone reservation [9]. Selective mapping (SLM) [10] is an example among those algorithms which reduce the PAPR of OFDM signal by multiplying a selected phase-rotation sequence to the frequency-domain signal before transmission. When employing joint Tx/Rx MMSE-FDE, the SC signal is transformed by DFT into the frequency-domain signal for applying Tx-FDE. This brings us to introduce a SLM algorithm to broadband SC transmission using joint Tx/Rx MMSE-FDE in order to reduce the PAPR.

In this paper, the proposed SLM is applied to the frequency-domain SC signal after Tx-FDE prior to inverse fast Fourier transform (IFFT). We expect a better BER performance as a result of ISI reduction by joint Tx/Rx MMSE-FDE, while the PAPR is simultaneously reduced by SLM. We also show that conventional Tx-FDE in [5] can be used without any modifications in case that the phase-rotation sequence has unit magnitude. To confirm this, performance evaluation, in terms of BER and PAPR, is done by computer simulation with various types of phase-rotation sequences and number of candidates.

The rest of this paper is organized as follows. System model of the proposed SC transmission is illustrated in Section II. PAPR reduction using SLM algorithm is discussed in Section III. The FDE weights derivation for joint Tx/Rx MMSE-FDE is shown in Section IV. Section V is performance evaluation, and finally Section VI concludes this paper.

II. TRANSMISSION SYSTEM MODEL

We assume single-user N_c-length block transmission with N_g-length of cyclic prefix insertion. DFT and its inverse operation are used in the proposed transmission scheme for reaching frequency-domain processing. Unlike the conventional SC, transmit filtering is replaced by transmit FDE (Tx-FDE) weight, and the SLM module is equipped after Tx-FDE but before IFFT. The proposed SC transceiver is depicted by Fig. 1.

A. Transmitter

We begin with a block of N_c data-modulated symbols $\mathbf{d}=[d(0),d(1),\ldots,d(N_c-1)]^T$. The block $\mathbf{d}$ is transformed into frequency domain by N_c-point fast Fourier transform (FFT), yielding the frequency-domain signal vector as $\mathbf{D}=\mathbf{F}_{N_c}\mathbf{d}$, where $\mathbf{D}=[D(0),D(1),\ldots,D(N_c-1)]^T$, and N_c-point DFT matrix is given by

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

Next, $\mathbf{D}$ is multiplied by Tx-FDE weight matrix, represented by $\mathbf{W}_t=diag[W_t(0),\ldots,W_t(N_c-1)]$. A selected phase-rotation sequence matrix $\mathbf{P}_{\hat{u}}=diag[P_{\hat{u}}(0),\ldots,P_{\hat{u}}(N_c-1)]$ is applied to the frequency-domain signal to be transmitted afterward. Note that $\mathbf{W}_t$ and $\mathbf{P}_{\hat{u}}$ can be simply implemented as one-tap multiplication due to the diagonality. The derivations of $\mathbf{W}_t$ and $\mathbf{P}_{\hat{u}}$ will be discussed in details in the next section.

After that, N_c-point IFFT, represented by $\mathbf{F}_{N_c}^H$, is applied for transforming the signal back to time domain. Time-domain signal $\mathbf{s}=[s(0),s(1),\ldots,s(N_c-1)]^T$ after passing through all processes can be expressed as

$$\mathbf{s}=\mathbf{F}_{N_c}^H\mathbf{P}_{\hat{u}}\mathbf{W}_t\mathbf{D}=\mathbf{F}_{N_c}^H\mathbf{P}_{\hat{u}}\mathbf{W}_t\mathbf{F}_{N_c}\mathbf{d}\,. \tag{2}$$

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.

B. Receiver

The transmission is conducted under independent L-path block Rayleigh fading channel [1]. Chanel response is given by

$$h(\tau)=\sum_{l=0}^{L-1}h_l\delta(\tau-\tau_l)\,, \tag{3}$$

where h_l and τ_l are complex-valued path gain and time delay of the l-th path, respectively. $\delta(\cdot)$ is the delta function. The received signal after CP removal, $\mathbf{r}=[r(0),r(1),\ldots,r(N_c-1)]^T$, is

$$\mathbf{r}=\sqrt{\frac{2E_s}{T_s}}\mathbf{h}\mathbf{s}+\mathbf{n}\,, \tag{4}$$

where E_s and T_s are symbol energy and symbol duration, respectively. $\mathbf{s}$ is obtained from (2), and $\mathbf{n}$ is noise vector in which element is zero-mean additive white Gaussian noise (AWGN) having the variance $2N_0/T_s$ with N_0 being the one-sided noise power spectrum density. In addition, channel response matrix $\mathbf{h}$ is a circular matrix representing time-domain channel impulse response, which is

$$\mathbf{h}=\begin{bmatrix}h_0 & & & h_{L-1} & \cdots & h_1\\ h_1 & \ddots & & & \ddots & \vdots\\ \vdots & & h_0 & \mathbf{0} & & h_{L-1}\\ h_{L-1} & & h_1 & \ddots & & \\ & \ddots & \vdots & & \ddots & \\ \mathbf{0} & & h_{L-1} & \cdots & \cdots & h_0\end{bmatrix}. \tag{5}$$

The received signal is transformed into frequency domain by N_c-point FFT, obtaining the frequency-domain received signal $\mathbf{R}$ as

$$\begin{aligned}\mathbf{R}&=\sqrt{\frac{2E_s}{T_s}}\mathbf{F}_{N_c}\mathbf{h}\mathbf{s}+\mathbf{F}_{N_c}\mathbf{n}\\ &=\sqrt{\frac{2E_s}{T_s}}\mathbf{F}_{N_c}\mathbf{h}\mathbf{F}_{N_c}^H\mathbf{P}_{\hat{u}}\mathbf{W}_t\mathbf{D}+\mathbf{F}_{N_c}\mathbf{n}\,.\\ &=\sqrt{\frac{2E_s}{T_s}}\mathbf{H}_c\mathbf{P}_{\hat{u}}\mathbf{W}_t\mathbf{D}+\mathbf{N}\end{aligned} \tag{6}$$

Here, we also define the frequency-domain channel response as $\mathbf{F}_{N_c}\mathbf{h}\mathbf{F}_{N_c}^H=diag[H_c(0),\ldots,H_c(N_c-1)]\equiv\mathbf{H}_c$.

Receive FDE (Rx-FDE) $\mathbf{W}_r=diag[W_r(0),\ldots,W_r(N_c-1)]$ is applied at the receiver as a joint cooperation with Tx-FDE to reduce the impact from frequency-selective fading. The received signal after equalization $\hat{\mathbf{D}}=\mathbf{W}_r\mathbf{R}$ is finally transformed back into time domain by N_c-point IFFT. Therefore, the time-domain equalized received signal before demodulation $\hat{\mathbf{d}}=[\hat{d}(0),\hat{d}(1),\ldots,\hat{d}(M-1)]^T$ is given by

$$\begin{aligned}\hat{\mathbf{d}}&=\mathbf{F}_{N_c}^H\mathbf{W}_r\mathbf{R}\\ &=\sqrt{\frac{2E_s}{T_s}}\mathbf{F}_{N_c}^H\mathbf{W}_r\mathbf{H}_c\mathbf{P}_{\hat{u}}\mathbf{W}_t\mathbf{D}+\mathbf{F}_{N_c}^H\mathbf{W}_r\mathbf{N}\,.\end{aligned} \tag{7}$$

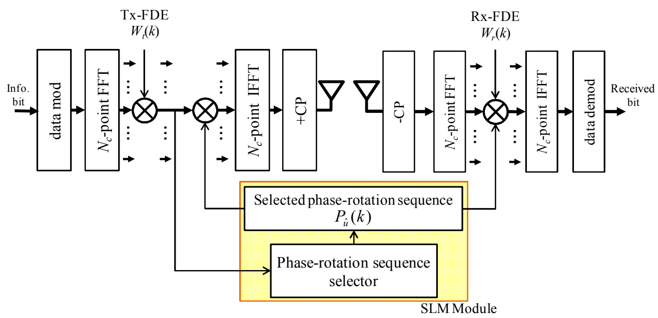

Fig. 1. System model of SC transmission using joint Tx/Rx FDE and SLM.

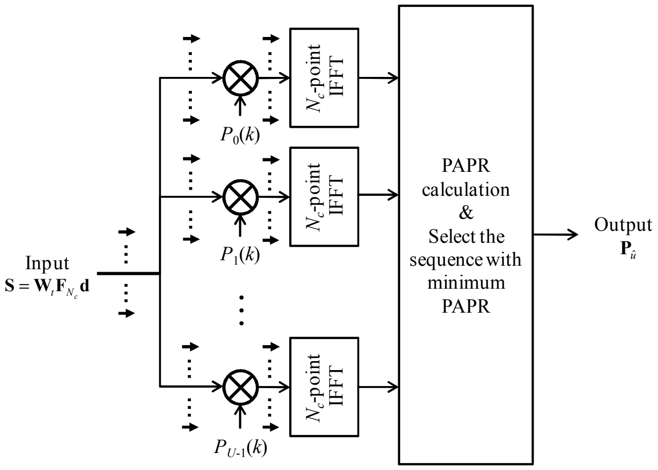

Fig. 2. SLM Module.

III. PAPR Reduction Using SLM

SLM [10] has been introduced as a PAPR reduction technique for OFDM with relatively small overhead and without distortion on waveform. It typically exploits in frequency domain, and hence, leads to the possibility to employ SLM in SC transmission. In this paper, SLM is applied to the frequency-domain SC signal after Tx-FDE prior to IFFT.

Regarding to the definition of **s** in (2), PAPR of the signal to be transmitted, calculated over a block of transmission, is expressed by

$$PAPR = \frac{\max\left\{|s(n)|^2\right\}}{E[|s(n)|^2]}, n = 0, \frac{1}{O.S.}, \frac{2}{O.S.}, ..., N_c - 1, \tag{8}$$

where $O.S.$ represents oversampling factor (we assume $O.S.$=4 in this paper).

Assuming that the frequency-domain signal after applying Tx-FDE is $\mathbf{S} = \mathbf{W}_t \mathbf{F}_{N_c} \mathbf{d}$. A set of U different phase-rotation sequences $\mathbf{P}_u$ is defined as candidates. Note that the first candidate $\mathbf{P}_0$ is set to be all "1" sequence as a representative of original signal, where the other candidates are generated either in a deterministic approach such as Walsh-Hadamard sequences [11], or in a random approach as $\{P_u(k) = \pm 1 \mid k = 0, \ldots, N_c - 1\}$. We restrict the phase-rotation sequences to be real number for avoiding the increase of complex-valued multiplication, and to be unit-magnitude for limiting the transmit power constraint. The instantaneous PAPR of transmit block $\mathbf{s}_u = \mathbf{F}_{N_c}^H \mathbf{P}_u \mathbf{S}$ for all u are calculated, and the phase-rotation sequence index $\hat{u}$, whose provides the lowest PAPR among those candidates, is selected by

$$\hat{u} = \arg \min_{u=0,1,\ldots,U-1} PAPR(\mathbf{s}_u = \mathbf{F}_{N_c}^H \mathbf{P}_u \mathbf{S}). \tag{9}$$

In addition, the algorithm can be simply illustrated by Fig. 2. A frequency-domain signal is multiplied by different phase-rotation sequences before the IFFT process. Then, the PAPR of signal corresponding to those sequences are calculated. Finally, the sequence with the lowest PAPR is selected and used for the actual transmission.

IV. Tx-FDE and Rx-FDE Weights

Performance of SC using joint Tx/Rx MMSE-FDE in aspect of ISI reduction has been exhaustively discussed in [5]. In this paper, we expect to use the proposed algorithm in [5] without major modifications even though SLM is employed.

A. Rx-FDE Weight

Similar to [5], FDE weights calculation begins from the receiver. Employing the MMSE criterion where the mean-square error (MSE) is calculated between frequency-domain transmitted and received signals. An error vector **e** is given by

$$\begin{aligned} \mathbf{e} &= \hat{\mathbf{D}} - \sqrt{\frac{2E_s}{T_s}} \mathbf{D} \\ &= \sqrt{\frac{2E_s}{T_s}} (\mathbf{W}_r \mathbf{H}_c \mathbf{P}_{\hat{u}} \mathbf{W}_t - \mathbf{I}_{N_c}) \mathbf{D} + \mathbf{W}_r \mathbf{N} \end{aligned}. \tag{10}$$

The MSE, denoted by e, is expressed as $e = tr[E(\mathbf{e}\mathbf{e}^H)]$, where $tr[\cdot]$ represents trace operation. $\mathbf{W}_r$, whose each element is represented by $W_r(k)$, is obtained by solving $\partial(e)/\partial \mathbf{W}_r = 0$ for a given $\mathbf{W}_t$. That is

$$\mathbf{W}_r = \frac{(\mathbf{H}_c \mathbf{P}_{\hat{u}} \mathbf{W}_t)^H}{\mathbf{H}_c \mathbf{P}_{\hat{u}} \mathbf{W}_t (\mathbf{H}_c \mathbf{P}_{\hat{u}} \mathbf{W}_t)^H + (E_s / N_0)^{-1} \mathbf{I}_{N_c}}. \tag{11}$$

Since all the matrix in (11) are diagonal, $W_r(k)$ can be directly obtained as

$$W_r(k) = \frac{H^*(k) P_{\hat{u}}^*(k) W_t^*(k)}{|H(k) P_{\hat{u}}(k) W_t(k)|^2 + (E_s / N_0)^{-1}}. \tag{12}$$

We can observe from (12) that the receiver needs to know the information of $\mathbf{W}_t$ and $\mathbf{P}_{\hat{u}}$ used at the transmitter. In a practical system, side information is transmitted together with the data in order to inform the receiver that which phase-rotation sequence is used for a particular transmission period.

B. Tx-FDE Weight

The error vector **e**, and corresponding MSE e, is rewritten by substituting the derived $\mathbf{W}_r$ back into (10), obtaining

$$e = \left(\frac{E_s}{N_0}\right)^{-1} tr\left[E\left(\left\{\mathbf{H}_c \mathbf{P}_{\hat{u}} \mathbf{W}_t \mathbf{W}_t^H \mathbf{P}_{\hat{u}}^H \mathbf{H}_c^H + \left(\frac{E_s}{N_0}\right)^{-1} \mathbf{I}_{N_c}\right\}^{-1}\right)\right]. \tag{13}$$

The entire matrix are diagonal, therefore, (13) is rewritten as

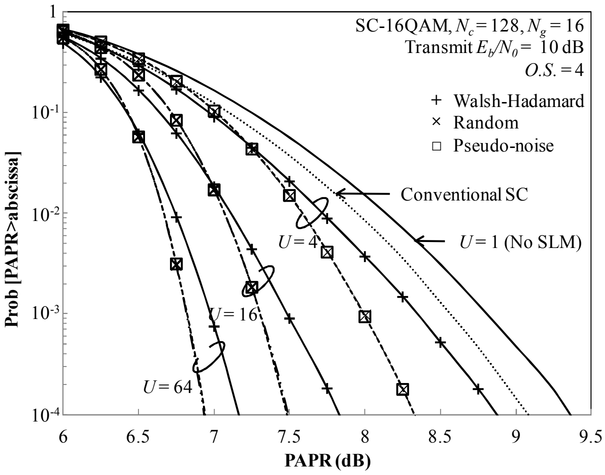

Fig. 3. PAPR performance of SC using joint Tx/Rx FDE and SLM.

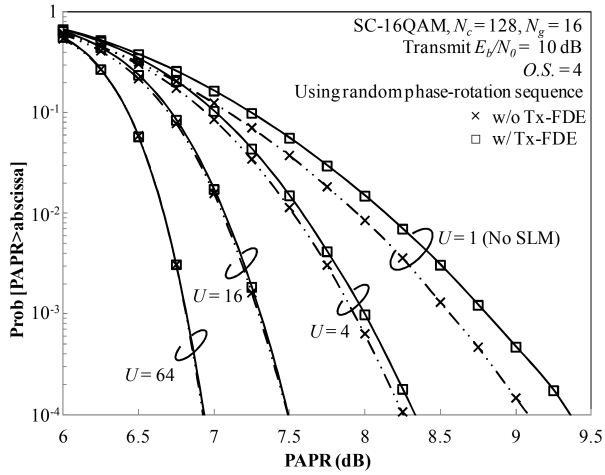

Fig. 4. PAPR performance of SC using SLM.

$$e = \sum_{k=0}^{N_c-1} \frac{(E_s/N_0)^{-1}}{|H_c(k)|^2 |P_{\hat{u}}(k)|^2 |W_t(k)|^2 + (E_s/N_0)^{-1}}. \tag{14}$$

However, $|P_{\hat{u}}(k)|^2 = 1$ as a result from the restriction of phase-rotation sequence discussed in Section III. This yields the MSE in (14) is exactly the same as in [5], and further implies that the Tx-FDE weight in [5] can be used in this paper without any required modifications. Therefore, $\mathbf{W}_t$, whose each element is represented by $W_t(k)$, is

$$W_t(k) = \max\left[\left\{\sqrt{\frac{(E_s/N_0)^{-1/2}}{\sqrt{\kappa}|H_c(k)|} - \frac{(E_s/N_0)^{-1}}{|H_c(k)|^2}}\right\}, 0\right], \tag{15}$$

where κ is a parameter selected to satisfy the transmit power constraint $\frac{1}{N_c}\sum_{k=0}^{N_c-1}|W_t(k)|^2 = 1$. It can be observed that the computation of $W_t(k)$ needs only the channel information, which means that the receiver itself can generate $W_t(k)$ for computing its own $W_r(k)$.

V. Performance Evaluation

Numerical and simulation parameters are summarized in Table I. We assume 16-QAM block transmission with the number of available subcarriers N_c=128. System performance is evaluated by PAPR, BER, and computational complexity.

TABLE I. SIMULATION PARAMETERS

Transmitter	Data modulation	16-QAM
	FFT/IFFT block size	N_c = 128
	Cyclic prefix length	N_g = 16
SLM Module	Phase-rotation sequence type	Walsh-Hadamard, Long PN, Random
	No. of candidates	U = 1 (no SLM) ~ 64
Channel	Fading	Frequency-selective block Rayleigh fading
	Power delay profile	16-path uniform
Receiver	FDE	joint Tx/Rx-FDE, MMSE-FDE
	Channel estimation	Ideal
	Phase-sequence information	Ideal

A. PAPR Performance

Fig. 3 shows the comparison of PAPR performance of SC transmission using joint Tx/Rx MMSE-FDE and SLM with various types of phase-rotation sequences and number of candidates. Transmission using joint Tx/Rx MMSE-FDE increases PAPR as indicated in the figure. It can be observed that PAPR decreases as U increases. Also, SLM using random phase-rotation sequence gives lower PAPR compared to one using Walsh-Hadamard sequence [11], for example, 0.9 dB and 0.5 dB reduction of the PAPR at the probability of occurrence of 0.1% ($PAPR_{0.1\%}$) for SLM employing random sequence and Walsh-Hadamard sequence, respectively. This observation has been confirmed by [12], however, random phase-rotation sequences lead to difficulty for side information sharing between transmitter and receiver.

In order to attain the randomness under a deterministic approach, we alternatively introduce a set of phase-rotation sequences generated from 4095-bit pseudo-noise (PN) sequence [13], where the transmitter and the receiver are able to share the information by using codebook. As a result, performance of SLM using PN sequences is exactly the same as using random sequences. In addition, it should be mentioned that up to $\log_2 U$ bits are required to be allocated as side information in practical environment.

In Fig. 4, PAPR performance comparison between SC transmission using conventional rectangular filtering and using Tx-FDE is illustrated. It can be seen that although there exists a slight performance gap when U is small, there is no difference of PAPR performance when U is large. Note that the transmission using Tx-FDE can reduce the residual ISI and hence, the better BER performance is achievable. The BER performance is discussed in the next subsection.

B. BER Performance

As abovementioned in Section IV, and assuming that the phase-rotation sequence used for transmission is perfectly known at the receiver, we expect that there is no significant effect on BER. Fig. 5 shows the BER as a function of average received bit energy-to-noise power spectrum density ratio E_b/N_0=0.25(E_s/N_0)(1+N_g/N_c).

The simulation result confirms our assumption as it shows that the BER performance of SC using joint Tx/Rx FDE and SLM is exactly the same as one without SLM. It also clarifies that there is no change on BER performance even though U is changed. This indicates the proposed SC transmission in this paper, i.e., SC using joint Tx/Rx MMSE-FDE and SLM, can provide better BER and PAPR than conventional SC by sacrificing only small amount of information bits for providing the side information. In addition, note that the BER performance of SC using joint Tx/Rx MMSE-FDE with SLM does not change with various phase-rotation sequences since they are restricted to be unit-magnitude.

C. Computational Complexity

In this paper, we evaluate the computational complexity by computing the number of complex-valued multiplications. SLM requires high computational complexity as referenced in many previous articles because of additional IFFT blocks.

Fig. 6 shows the computational complexity as a function of U for transmission schemes with and without SLM. Note that the computational complexity of joint Tx/Rx MMSE-FDE is not included in this complexity evaluation since it depends on channel condition, and the computational complexity is independent of type of phase-rotation sequence. It can be observed that the computational complexity increases logarithmically as U increases. This is because the number of complex-valued multiplications per N_c-point IFFT operation equals $N_c\log_2(N_c)$.

VI. CONCLUSION

In this paper, we proposed the SLM for broadband SC transmission using joint Tx/Rx MMSE-FDE. The proposed SLM is applied to the frequency-domain SC signal after Tx-FDE prior to IFFT. The simulation results confirmed that the better BER performance is achieved as a result of ISI reduction by joint Tx/Rx MMSE-FDE while the PAPR is simultaneously reduced by the proposed SLM.

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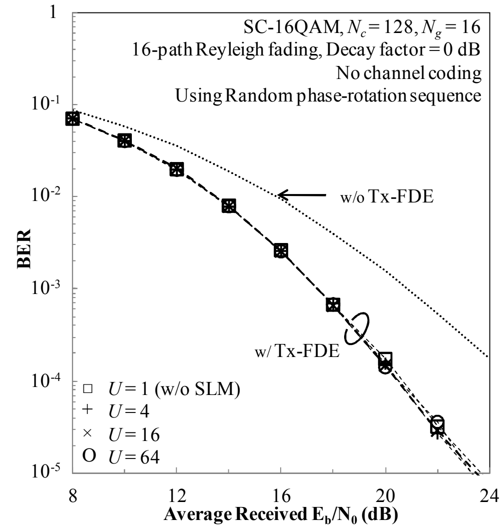

Fig. 5. BER performance of SC using joint Tx/Rx FDE and SLM.

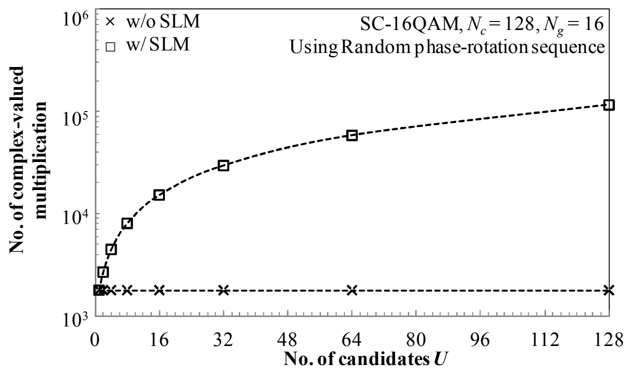

Fig. 6. Computational complexity of SLM in the proposed SC transmission