

BER Reduction for PSK-Based MM-OFDM-BIM with Two-Level Symbol Power

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Abstract—Phase-shift keying (PSK)-based multi-mode orthogonal frequency division multiplexing with block-wise index modulation (MM-OFDM-BIM) has been proposed as a novel OFDM-IM technique with adjustable-reliability transmission and low detection complexity. Although the phase normalization for PSK can reduce the complexity of the index modulation (IM) detection, the essence of constant symbol power of PSK limits its bit error rate (BER) performance. We propose a new constellation design to reallocate the original symbols into two-level power for PSK-based MM-OFDM-BIM while retaining the merit of phase normalization. By adopting the optimum power parameter, the numerical results demonstrate that the proposed scheme can effectively improve BER performance without increasing computational complexity.

Keywords—Index modulation; Multiple-mode OFDM; Phase-shift keying; Convolution code; Constellation design

I. INTRODUCTION

Orthogonal frequency division multiplexing with index modulation (OFDM-IM) techniques [1], [2] have garnered great attention in recent years due to its exceptional flexibility with a trade-off between spectral efficiency (SE) and energy efficiency (EE) and high compatibility with the popular OFDM systems such as 3GPP Long Term Evolution (LTE)/New Radio (NR) and IEEE 802.11/16 series standards [3]. The basic concept of OFDM-IM is using the subcarrier index modulation (SIM) [4] on OFDM for transmitting extra bits via subcarrier indices. In [5], the subcarrier indices are distinguished by different subcarrier activation patterns (SAPs) which originate from a group of on-off mode subcarriers within the IM blocks. By controlling the number of activated subcarriers in an IM block, the SE and EE of the system can be easily modified to adapt the various communication scenarios in 5G and beyond systems [6], [7].

Since the SE requirement becomes more crucial for next-generation wireless networks, several all-subcarrier-activated techniques, such as dual-mode (DM)- [8] and multiple-mode (MM)- [9] OFDM-IM, have been proposed to focus on the SE improvement. However, these schemes only improve SE over good signal-to-noise ratio (SNR) environment, especially with channel coding in consideration. Conversely, these techniques' performances even degrade with low SNR. To this, MM-OFDM with block-wise IM (MM-OFDM-BIM) [10] has been

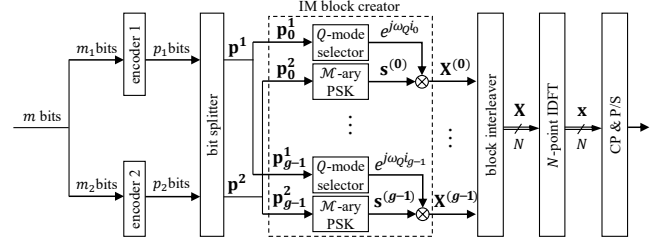
proposed to enhance BER performance according to SNR, with an adjustable cost of SE. Because of the phase-shift keying (PSK) in place, a low-complexity detector of MM-OFDM-BIM has been designed using the phase normalization. Although PSK benefits from reducing the detection complexity, its minimized Euclidean distance of the constellation points for each symbol cannot be increased due to the constant symbol power. The PSK forces the constellation to be located on a circle that implies room for BER improvement.

In this paper, we propose an improved version of PSK-based MM-OFDM-BIM by redesigning its constellation pattern with two-level symbol power. Such a concept of constellation design is similar to quadrature phase-shift keying (QPSK) for DM-OFDM-IM [11], but the proposed allows the alphabets in different IM mode can be transmitted with the same symbol power. This proposed scheme significantly ameliorates the BER performance without any increase of computational complexity, while retaining the merit of phase normalization. Through the Monte Carlo simulations, we identify the optimum allocation for two-level symbol power and evaluate the BER improvement over a frequency-selective fading channel for the proposed scheme.

II. PSK-BASED MM-OFDM-BIM WITH CONVENTIONAL CONSTELLATION

A. PSK-Based MM-OFDM-BIM

The block diagram of the PSK-based MM-OFDM-BIM transmitter [10] is illustrated in Fig. 1. The m incoming bits are partitioned into m_1 bits for IM and m_2 bits for constellation mapping. As a coded OFDM-IM technique, both bit streams are first encoded by convolution code with the code rate γ_1 and γ_2 before IM block creator, respectively. Assuming $\gamma_1 = \gamma_2 = \gamma$, we have $p_1 = m_1/\gamma$ and $p_2 = m_2/\gamma$ bits to be fed into g IM



blocks, where each occupies $n = N/g$ subcarriers. In MM-OFDM-BIM, all subcarriers in the β -th IM block are employed with the same mode, $\beta \in \{0, \dots, g-1\}$. Thus the bit vector for all Q -mode index selectors can be presented as

$$\mathbf{p}^1 = [\mathbf{p}_0^1 \dots \mathbf{p}_{g-1}^1], \quad (1)$$

where $\mathbf{p}_\beta^1 = [b_\beta^1(0) \dots b_\beta^1(\log_2 Q - 1)]$, $b_\beta^1 \in \{0, 1\}$. Note that the value of Q assumes a power of 2 owing to the mapping of the effective bits. The bit vector for all $\mathcal{M}$ -ary PSK symbol mappers is given by

$$\mathbf{p}^2 = [\mathbf{p}_{0,0}^2 \dots \mathbf{p}_{0,n-1}^2 \dots \mathbf{p}_{g-1,0}^2 \dots \mathbf{p}_{g-1,n-1}^2], \quad (2)$$

where $\mathbf{p}_{\beta,\alpha}^2 = [b_{\beta,\alpha}^2(0) \dots b_{\beta,\alpha}^2(\log_2 \mathcal{M} - 1)]$, $b_{\beta,\alpha}^2 \in \{0, 1\}$, and $\alpha \in \{0, \dots, n-1\}$ is the order of subcarriers in the β -th IM block. Then, the output symbol vector of the β -th IM block, $\mathbf{X}^{(\beta)} = [X_0^{(\beta)}, \dots, X_{n-1}^{(\beta)}]^T$, is gained by

$$\mathbf{X}^{(\beta)} = \mathbf{s}^{(\beta)} e^{j\omega_Q i_\beta}, \quad (3)$$

where $i_\beta \in \{0, \dots, Q-1\}$ is the selected mode from $\mathbf{p}_\beta^1 \rightarrow i_\beta$, $\omega_Q = \omega_{\mathcal{M}}/Q$ is the margin angle between each mode, $\omega_{\mathcal{M}} = 2\pi/\mathcal{M}$ is the rotational-symmetric angle of $\mathcal{M}$ -ary PSK constellation, $\mathbf{s}^{(\beta)} = [s_0^{(\beta)}, \dots, s_{n-1}^{(\beta)}]^T$ is the input PSK symbol vector from $\mathbf{p}_{\beta,\alpha}^2 \rightarrow \mathbf{s}^{(\beta)}$, where $\bullet \rightarrow \bullet$ and $[\cdot]^T$ denotes bit mapping operation and transposition, respectively. Then the interleaved frequency components of the OFDM block are defined as

$$\mathbf{X} \triangleq [X_0^{(0)}, \dots, X_{n-1}^{(0)}, \dots, X_{n-1}^{(g-1)}, \dots, X_{n-1}^{(g-1)}]^T. \quad (4)$$

After the N -point inverse discrete Fourier transform (DFT) of $\mathbf{X}$, the time-domain OFDM block is given by

$$\mathbf{x} \triangleq [x_0, \dots, x_{N-1}]^T = \frac{1}{\sqrt{N}} \mathbf{W}_N^H \mathbf{X}, \quad (5)$$

where $\mathbf{W}_N$ is the N -point DFT matrix which follows $\mathbf{W}_N^H \mathbf{W}_N = N\mathbf{I}_N$, and $[\cdot]^H$ denotes Hermitian transposition. The transmitted baseband signals are prepared after the insertion of the P -length cyclic prefix (CP) at the beginning of $\mathbf{x}$, as $[x_{N-P}, \dots, x_{N-1}, x_0, \dots, x_{N-1}]^T$, and the parallel-to-serial conversion (P/S).

We then consider a slowly time-varying multipath Rayleigh fading channel with impulse response $\mathbf{h} = [h_0, \dots, h_{L-1}]^T$, where L is the number of the channel taps ($L \leq P$) and each entry of $\mathbf{h}$ is a circularly symmetric complex Gaussian random variable with the distribution of $\mathcal{CN}(0, \sigma_L^2)$, where $\sum \sigma_L^2 = 1$. Assuming a perfect estimation of the channel response, the received signal of the β -th IM block after the CP removal, serial-to-parallel conversion (S/P), N -point DFT, equalization and de-interleaving is expressed as

$$\mathbf{Y}^{(\beta)} \triangleq [Y_0^{(\beta)}, \dots, Y_{n-1}^{(\beta)}]^T = \Pi^{(\beta)} \{\text{diag}\{\mathbf{X}\} \mathbf{H} + \mathbf{W}\}, \quad (6)$$

where $\text{diag}\{\cdot\}$ denotes a diagonal matrix, $\Pi^{(\beta)}\{\cdot\}$ is the de-interleaving operator for the β -th IM block, $\mathbf{H}$ is the frequency domain channel transfer function coefficients obtained by the N -point DFT of the zero-padded channel impulse response $\mathbf{h}^0 = [h_0, \dots, h_{L-1}, 0, \dots, 0]^T$, and $\mathbf{W}$ is the frequency-domain expression of AWGN with power N_0 .

At the receiver, a two-stage block detector with phase normalization is utilized for low-complexity and reliable IM bits detection [10]. In the first stage—mode detection, the $\mathbf{p}_\beta^1$ information is obtained by the detected index $\hat{i}_\beta$ from (7), where $\mathbf{H}^{(\beta)} = \Pi^{(\beta)}\{\mathbf{H}\}$. The constellation of each IM block is then rotated back to mode-0 by $\hat{\mathbf{Y}}^{(\beta)} = \mathbf{Y}^{(\beta)} e^{-j\omega_Q i_\beta}$. In the second stage, $\mathcal{M}$ -ary PSK demapping is performed on the detected symbol vector $\hat{\mathbf{s}}^{(\beta)} = [\hat{s}_0^{(\beta)}, \dots, \hat{s}_{n-1}^{(\beta)}]^T$ to gain the $\mathbf{p}_{\beta,\alpha}^2$ information, where

$$\hat{\mathbf{s}}^{(\beta)} = \arg \min_{s^{(\beta)} \in \{\mathcal{S}_0, \dots, \mathcal{S}_{\mathcal{M}-1}\}} \|\hat{\mathbf{Y}}^{(\beta)} - \mathbf{X}^{(\beta)} \mathbf{H}^{(\beta)}\|^2, \quad (8)$$

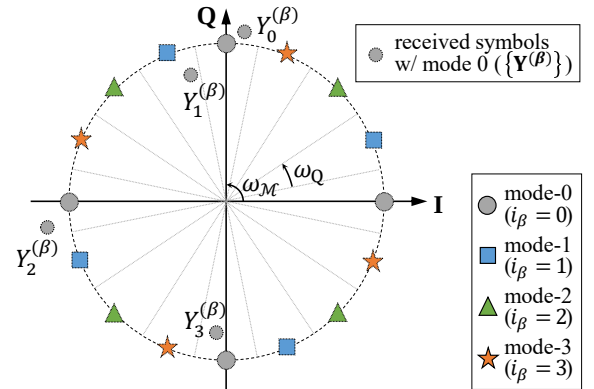
and $\{\mathcal{S}_0, \dots, \mathcal{S}_{\mathcal{M}-1}\}$ is the set of the alphabet of $\mathcal{M}$ -ary PSK.

Therefore, the SE of MM-OFDM-BIM can be expressed as

$$\eta(Q, \mathcal{M}, n) = \frac{m_1 + m_2}{N} = \frac{\gamma(p_1 + p_2)}{N} = \gamma \left[\frac{\log_2(Q)}{n} + \log_2(\mathcal{M}) \right]. \quad (9)$$

B. Conventional Signal Constellation

According to (3), it is evident that the constellation of MM-OFDM-BIM essentially follows the $Q\mathcal{M}$ -ary PSK. Fig. 2 illustrates an example of the constellation and received symbols of conventional MM-OFDM-BIM for $(Q, \mathcal{M}, n) = (4, 4, 4)$.



$$\hat{i}_\beta = \arg \min_{\{i_\beta^{(0)}, i_\beta^{(1)}\}} \begin{cases} \min_{i_\beta^{(0)} \in \{0, Q\}} \|\mathbf{Y}^{(\beta)} - \mathcal{S}_0 e^{j\omega_Q i_\beta^{(0)}} \mathbf{H}^{(\beta)}\|^2 & \text{(ML detection for mode 0)} \\ \min_{i_\beta^{(1)} \in \{1, \dots, Q-1\}} \|\mathbf{Y}^{(\beta)} - \mathcal{S}_0 e^{j\omega_Q i_\beta^{(1)}} \mathbf{H}^{(\beta)}\|^2 & \text{(ML detection for mode 1 to } Q-1) \end{cases} \quad (7)$$

III. PROPOSED CONSTELLATION FOR PSK-BASED MM-OFDM-BIM

A. Proposed Signal Constellation

From the conventional signal constellation for PSK-based MM-OFDM-BIM described in Section II-B, it is clear to see the room for BER improvement owing to the dense constellation on the unit circle. For an acceptable BER performance, the cases of $Q\mathcal{M} \leq 16$, e.g., $(Q, \mathcal{M}) = (2, 4)$, $(2, 8)$, $(4, 2)$, $(4, 4)$, $(8, 2)$, are investigated in this paper, where $(Q, \mathcal{M}) = (2, 2)$ is not discussed since its optimum constellation still follows [10].

Aiming to increase the minimized Euclidean distance of the conventional constellation points, we propose the two-level symbol power while keeping the merit of phase normalization on reducing the detection complexity of MM-OFDM-BIM. By redesigning (3), the proposed IM block creator is given as

$$\mathbf{X}^{(\beta)} = \mathbf{s}^{(\beta)} A_{i_\beta} e^{j\omega_Q i_\beta}, \quad (10)$$

where $A_{i_\beta} \in (0, 1]$ is the assigned amplitude for mode- i_β that is normalized to mode-0. Considering the uniformity of the constellation distribution, the value of A_{i_β} is defined as

$$A_{i_\beta} = \begin{cases} 1, & \text{if } i_\beta \text{ is even} \\ A_{in}, & \text{if } i_\beta \text{ is odd} \end{cases} \quad (11)$$

That is, the constellation of even modes is still fixed on the unit circle and of odd modes is designed to lie on an inner circle with amplitude A_{in} . Fig. 3 illustrates the same example given in Fig. 2 using the proposed constellation design. Note that the constellation of conventional PSK-based MM-OFDM-BIM is a special case of (10) and (11) when $A_{in} = 1$.

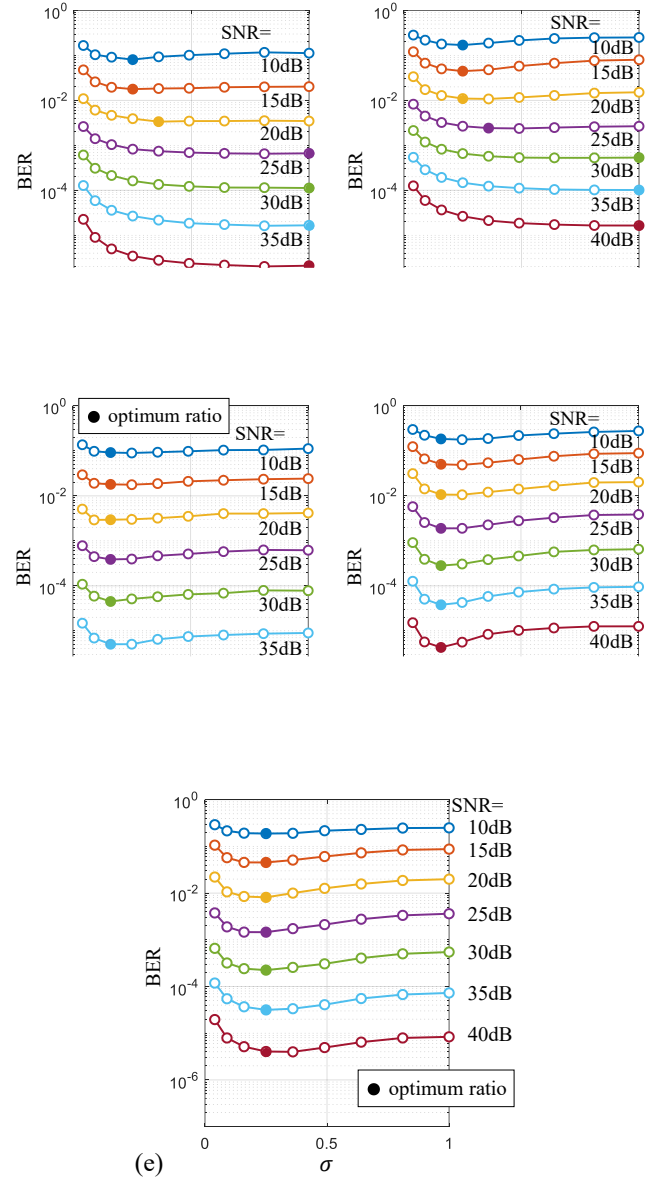
B. Power Ratio Optimization

Since the total power is fixed for a given SNR, the value of A_{in} controls the power ratio $\sigma = A_{in}^2 / 1^2 = A_{in}^2$ that determines the BER performance of the proposed constellation. Here, we analyze the optimum power ratio for each $(Q, \mathcal{M})$ parameters at different SNRs through a numerical simulation shown in Fig. 4.

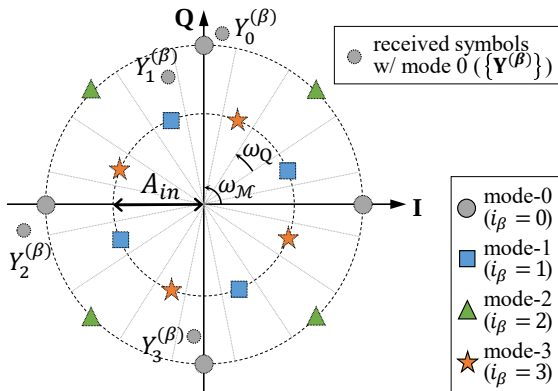
From the viewpoint of constellation circles, it can be observed that the optimum power ratio varies for different

SNRs if only one mode is allocated on a constellation circle, e.g., $(Q, \mathcal{M}) = (2, 4)$, $(2, 8)$ described in Fig. 4(a) and 4(b). The optimum power ratio is small in low SNRs while is raised to one in high SNRs. This is because lowering σ can effectively enhance the BER performance of p_1 bits against low SNRs without the consideration to interfering with other modes. As the SNR increases enough for good detection for p_1 bits, adopting $\sigma = 1$ is the optimum constellation as it maximizes the minimum Euclidean distance for p_2 bits.

For the same reason, the optimum power ratio is always fixed for any SNRs if the constellation circle contains multiple modes, e.g., $(Q, \mathcal{M}) = (4, 2)$, $(8, 2)$, $(4, 4)$ described in Fig. 4(c), 4(d) and 4(e). As seen, the optimum power ratio is $\sigma = 0.16$ for $(Q, \mathcal{M}) = (4, 2)$, $(8, 2)$ and $\sigma = 0.25$ for $(Q, \mathcal{M}) = (4, 4)$. Such a difference is caused by the settings of $(Q, \mathcal{M})$.



(4,4)



According to (9), the overall SE is primarily determined by $\mathcal{M}$, especially for large n . Hence, the optimum value of σ would be reasonably increased with $\mathcal{M}$.

IV. NUMERICAL RESULTS AND DISCUSSION

In this section, the coded BER improvement with optimum power ratio is validated by numerical simulation over AWGN and Rayleigh fading channel for $N = 2048$ and $P = L = 128$. The convolution code ($\gamma = 1/2$, constraint length 7) is adopted in the encoders.

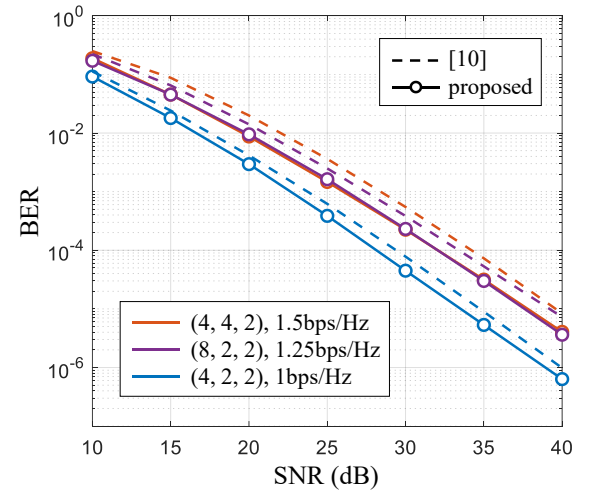
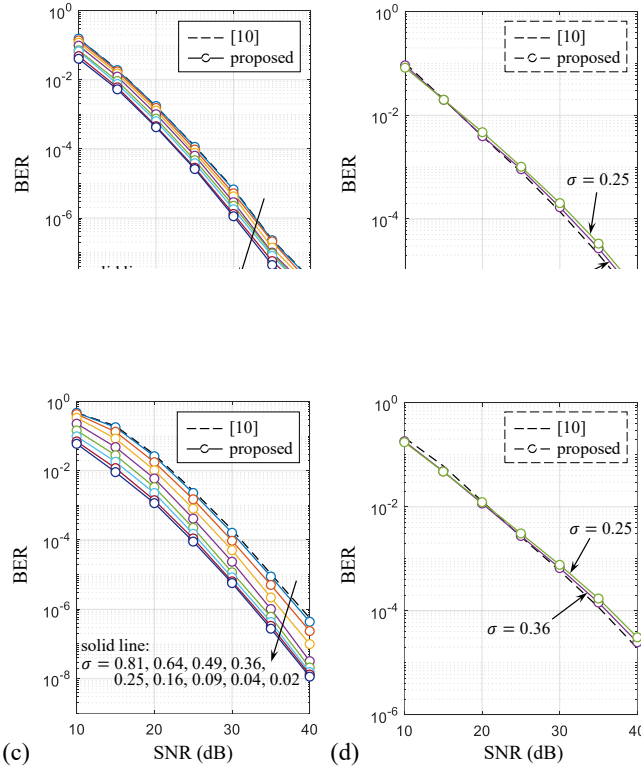
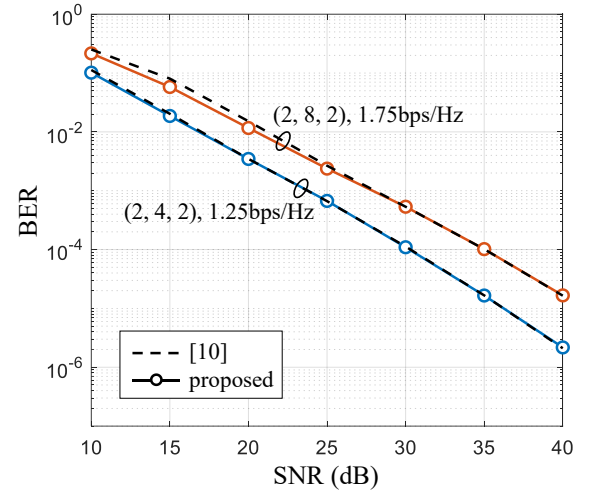
Fig. 5 establishes the coded BER performances of $(Q, \mathcal{M}) = (2, 4), (2, 8)$ for m_1 and m_2 bits, respectively. It can be seen that the BER for m_1 bits (see Fig. 5a and 5c) improves as power ratio increases for both $(Q, \mathcal{M})$ settings. For the m_2 bits (see Fig. 5b and 5d), the BER performance improves when $\text{SNR} \leq 15\text{dB}$ for $(2, 4)$ and $\text{SNR} \leq 25\text{dB}$ for $(2, 8)$ but degrades as SNR increases. These results further implicate the philosophy mentioned in Section III-B. Subsequently, as shown in Fig. 6, the overall BER performance can be improved for low SNRs by selecting the proper optimum power ratios for given SNR indicated in Fig. 4(a) and 4(b). Since the optimum power ratio is one when $\text{SNR} > 15\text{dB}$ for $(2, 4)$ and $\text{SNR} > 25\text{dB}$ for $(2, 8)$, their BER performances still follow the results in [10].

As for the case of $(Q, \mathcal{M}) = (4, 4), (8, 2)$ and $(4, 2)$, which is indicated in Fig. 7 (all use $n = 2$), it can be observed that the proposed scheme markedly improves the BER performance for every SNRs with a fixed optimum power ratio. The improvement for $\mathcal{M} = 4$ is evidently greater than that of $\mathcal{M} =$

2, as $\mathcal{M}$ mainly dominates the BER performance. We denote that the BER improvement of the proposed scheme does not increase any computational complexity since it only alters the design of the constellation.

V. CONCLUSION

In this paper, we have proposed a new constellation design for PSK-based MM-OFDM-BIM to improve its coded BER performance without the increase of computational complexity. By allocating even/odd-mode symbols into two-level power instead of constant one, the phase normalization for reducing the detection complexity can still work in the proposed scheme. The optimum power ratios for several $(Q, \mathcal{M})$ settings over Rayleigh fading channel are also analyzed through numerical simulations. The presented results in terms of BER validates the feasibility of the proposed scheme as a favorable alternative for PSK-based MM-OFDM-BIM.



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