Efficient PAPR Reduction Scheme for OFDM-NOMA Systems Based on DSI & Precoding Methods

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Abstract

Orthogonal Frequency Division Multiplexing (OFDM)-based Non-Orthogonal Multiple Access (NOMA) systems are one of the promising technologies to be implemented in novel broadcasting and communication systems. One of the main challenges in OFDM-based systems is the high peak to average power ratio (PAPR), which degrades the system efficiency. Partial transmit sequence (PTS) and precoding approaches are among the best PAPR reduction schemes. The PAPR reduction in precoding methods is not as good as PTS-based ones. On the other hand, PTS-based methods are computationally complex and mostly degrade the system BER. In this paper, an efficient PAPR reduction scheme, based on the precoding and dummy sequence insertion (DSI) techniques, is proposed to overcome the previous limitations. Furthermore, a novel dummy sequence generation procedure is devised for the proposed method. We demonstrate that the developed scheme has a remarkably better PAPR reduction and BER performance than the precoding and PTS-based approaches. Moreover, it is computationally much less complex compared to the PTS methods.

Keywords: Orthogonal Frequency Division Multiplexing (OFDM), Non-Orthogonal Multiple Access (NOMA), OFDM-NOMA, Peak-to-Average Power Ratio (PAPR), Precoding, Dummy Sequence Insertion (DSI)

1. Introduction

The advent of new technologies, such as real-time remote healthcare, autonomous transportation, and remote education, calls for higher data rates, better cell coverage, ultra-reliability, and massive connectivity in novel communication and broadcasting systems [1]. OFDM [2] is one of the promising schemes that enables communication systems to reach higher data rates by mitigating the intersymbol interference (ISI) caused by frequency selective channels. This system utilizes several orthogonal carriers to transmit the user's data. For approaching the assigned goals, novel communication and broadcasting systems need more spectral resources. Since the available wireless frequency resources are overcrowded and limited, the systems should utilize the available time-frequency resources more efficiently.

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Power-domain NOMA [3] allows communication systems to transmit different users' data over the same time-frequency resources, with different power levels. In this approach, users with stronger channel conditions get a lower fraction of power than users with weaker channel conditions to ensure all users experience equal quality. At the receiver, each user detects its data using the successive interference cancelation (SIC) technique. Also, cooperative-NOMA [4] technique could be utilized to improve the cell coverage and reliability of the system by creating additional data streams between users within the same NOMA cluster. A hybrid power-domain NOMA scheme is proposed in [5] to improve the achievable sum rate and system bit error rate (BER) over the frequency selective channel. Additionally, this scheme improves system reliability and ensures the target data rate for cell-edge users.

OFDM-NOMA systems have the advantages of OFDM and NOMA techniques in terms of higher data rates, mitigating ISI caused by multipath propagation of wireless channels, spectral efficiency, user fairness, and massive connectivity. However, the design of a suitable OFDM-NOMA system needs to deal with the disadvantages of both concepts, such as high error propagation, sensitivity to Carrier Frequency Offset (CFO), and high PAPR. In this article, we mainly focus on high PAPR, which is one of the main limitations of OFDM-based systems. High PAPR makes the High Power Amplifier (HPA) work in its saturation region. This leads to in-band distortion, and out-band radiation [6]. Therefore, lowering the PAPR of the signal before the HPA is necessary to minimize the undesirable effects.

There are three major groups of PAPR reduction methods in OFDM-based systems. The first and simplest group is signal distortion approaches, such as [7], which mostly lowers the signal PAPR by cutting the time-domain OFDM signal. Due to the distortion property of these methods, and considering the high error propagation in SIC, they are not suitable choices for PAPR reduction in OFDM-NOMA systems. The other set of PAPR reduction methods is called signal scrambling. Selective Mapping (SLM) [8] and Partial Transmit Sequences (PTS) [9] are some of the examples of this set. PTS-based methods reduce the PAPR value by dividing the signal into different disjoint subblocks and multiplying each subblock by a proper phase rotation factor, and finally, recombining them to generate a signal with lower PAPR. These methods are computationally complex and need an extensive search to find appropriate phase rotation factors. Furthermore, they require to send the selected phase rotation factors as side information to the receiver, which degrades system BER. The proposed method has the same PAPR reduction as the conventional PTS (C-PTS) method but with lower complexity. In [10] and [11], the authors lowered the complexity of the PTS method by optimizing the selection of phase rotation factors. In [12], the authors suggested a PTS method based on the dominant time-domain samples and defines two novel criteria for selecting them. [13] recommended a low computational complexity PTS method based on particle swarm optimization (PSO). In [14], the authors proposed the DSI-PTS PAPR reduction method, based on the PTS and DSI [15] techniques. Finally, [16] surveyed the PTS-based PAPR reduction methods in OFDM systems. In the recent works, DSI-based techniques have attracted great attention due to their low complexity. [17] combined the DSI and subcarrier group modulation (SGM) to improve the spectral efficiency. [18] proposed a semi-dummy sequence insertion technique, in which the dummy sequences are generated by Fourier transform of a number of the data modulated symbols. [19] suggested Blind Dummy-Zero-Insertion (DZI) technique based on the DSI method to reduce the PAPR and the used power. In [20] the authors proposed the combination of SLM, PTS, and DSI schemes. [21] investigated the DSI technique on multiple input multiple output OFDM (MIMO-OFDM) systems and proposed a multiuser precoding matrix to remove the interference caused by the inserted dummy sequence. However, most of those techniques have limited PAPR reduction performance and need an extra process for creating the dummy sequence for each data signal. Among the prior works, DSI-PTS is the most worthwhile DSI-based technique. The last group of PAPR reduction approaches includes precoding methods, such as Discrete Cosine Matrix Transform (DCMT) [22], Discrete Sine Matrix Transform (DSMT) [23], and Discrete Hartley Matrix Transform (DHMT) [24]. In [25], the authors combined Walsh-Hadamard transform (WHT) [26] and Zadoff chu transform (ZCT) [27] matrices to form a novel PAPR reduction matrix transform, named T-transform. These methods lower the aperiodic autocorrelation value among the modulated data symbols before the IFFT calculation, which leads to lower PAPR [28]. Precoding methods are among the promising ways for PAPR reduction in OFDM-NOMA systems. [29] studies the PAPR reduction in OFDM-NOMA systems using the precoding methods. Precoding methods have no distortion effect on the signal and do not require to transmit any side information, which leads to better BER performance. However, their PAPR reduction is not as good as signal scrambling methods.

Due to the SIC in OFDM-NOMA systems, they have high error propagation. PTS-based techniques need to transmit extra information, which degrades the system BER. Therefore, they are not a suitable choice for PAPR reduction in OFDM-NOMA systems. On the other hand, the PAPR reduction performance of the precoding methods is not as good as the PTS-based ones. In this paper, we propose a hybrid PAPR reduction scheme that can considerably reduce the signal PAPR and solve the previous limitations. The proposed method is based on the DSI and precoding methods. It needs no side information to be transmitted and does not degrade the BER performance of the system. Also, we offer a novel dummy sequence generation technique to improve the PAPR reduction of the method. We suggest a special way of implementing the DSI method in order to minimize unnecessary operations on each iteration. We will demonstrate that the suggested scheme has better PAPR reduction compared to the precoding methods. Moreover, it outperforms PTS-based methods in terms of PAPR reduction and BER performance with much less computational complexity. The rest of the paper is arranged as follows: The model for the OFDM-NOMA system is described in section 2. In section 3, we review the DSI, precoding, and PTS-based PAPR reduction methods. The proposed scheme and its characteristics will be described in section 4. The computational complexity of the proposed method is presented and compared with PTS-based approaches in section 5. Performance evaluation of the suggested technique in terms of Aperiodic-Autocorrelation property, power spectral density (PSD), PAPR reduction, and BER performance is presented in section 6, and section 7 concludes the paper.

2. System Model

Consider the downlink of an OFDM-NOMA system as depicted in figure 1. The base station utilizes the power-domain NOMA to transmit I users' messages over the same time-frequency resources with different power levels. Note that I is not the total number of users in the network. More users serve via different time-frequency resources. The transmitted

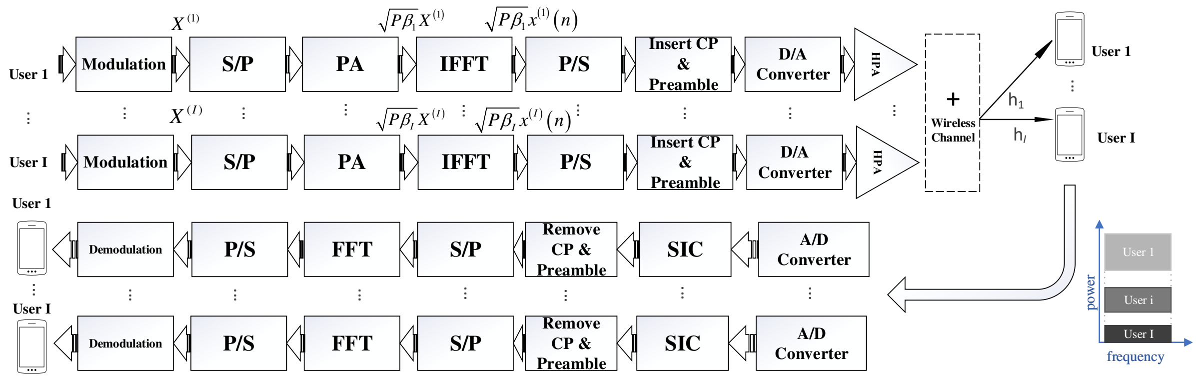


Figure 1: OFDM-NOMA transmitter and receiver

base-band OFDM-NOMA signal can be written as follows:

$$x(n) = \sum_{i=1}^{I} \sqrt{P\beta_i} x^{(i)}(n) \;; 0 \leq n \leq N-1 \;\;, \tag{1}$$

where n is the discrete-time variable, N is the total number of subcarriers, P is the total transmit power, $x^{(i)}(n)$ is the complex base-band OFDM signal of the i^{th} user, and β_i is the fraction of total power assigned to the i^{th} user. Users are ordered according to their channel conditions from the worst to the best. So, $\beta_1 \geq \ldots \geq \beta_i \geq \ldots \geq \beta_I$, which satisfies $\sum_{i=1}^{I} \beta_i = 1$.

Let $\mathbf{X}^{(i)} = \left[X_0^{(i)}, X_1^{(i)}, \ldots, X_{N-1}^{(i)}\right]$ be a block of data symbols for the i^{th} user, which are taken from one of the digital modulation schemes, such as QAM or PSK. Then, the complex base-band OFDM signal for the i^{th} user is as follows:

$$\begin{aligned} x^{(i)}(n) &= IFFT\left\{\mathbf{X}^{(i)}\right\} = \frac{1}{\sqrt{N}} \sum_{z=0}^{N-1} X_z^{(i)} e^{\frac{j2\pi zn}{N}} \\ &= \frac{1}{\sqrt{N}} \left(X_0^{(i)} e^{\frac{j2\pi 0n}{N}} + X_1^{(i)} e^{\frac{j2\pi 1n}{N}} + \ldots + X_{N-1}^{(i)} e^{\frac{j2\pi (N-1)n}{N}} \right) \;; 0 \leq n \leq N-1 \;\;, \end{aligned} \tag{2}$$

PAPR is defined as the ratio between the signal's maximum instantaneous power and its mean value:

$$PAPR = \frac{\max\{|x\,(n)|^2\}}{E\{|x\,(n)|^2\}}, \tag{3}$$

where $E\left\{\mathbf{A}\right\}$ and $max\left\{\mathbf{A}\right\}$ take the mean and maximum value of $\mathbf{A}$, respectively.

Oversampling of the discrete-time OFDM signal is required to get an accurate PAPR value. According to [30], 4-times oversampling is enough to get the PAPR value close to that of the continuous-time signal. Thus, we will use 4-times oversampling for calculating PAPR values.

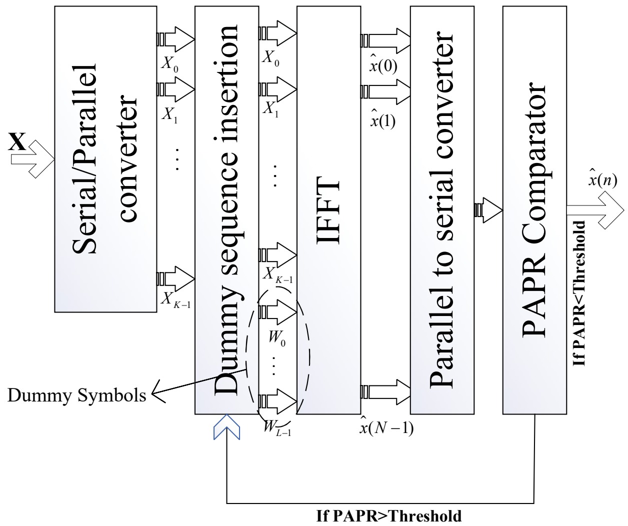


Figure 2: The block diagram of the DSI method

3. Methods

In this section, we will review the DSI technique for PAPR reduction in OFDM-based systems. Then, we will study precoding methods, which is one of the prominent PAPR reduction schemes for OFDM-NOMA systems. Finally, we will investigate PTS-based approaches to compare with our proposed method.

3.1. The Dummy Sequence Insertion Method

As depicted in figure 2, in the DSI method, the input data sequence is combined with a dummy sequence to generate the first candidate signal. Later, the signal passes through the IFFT. If its PAPR value is less than the threshold, the system transmits the signal. Otherwise, another dummy sequence will be generated and added to the input data, and after passing through the IFFT block, the system will recheck PAPR. This process continues until finding an appropriate signal with a PAPR value less than the threshold. If there is no such signal, then the signal with the least PAPR value is chosen to be transmitted. Since the inserted dummy symbols have no value of information , the receiver could easily ignore them.

If $[X_0, X_1, \ldots, X_{K-1}]$ is the modulated input data with a length of K and $\left[W_0^b, W_1^b, \ldots, W_{L-1}^b\right]$ is the inserted dummy sequence with a length of L, then the IFFT input for the i^{th} user will be:

$$\hat{\mathbf{X}}^{(i)} = \left[X_0^{(i)}, X_1^{(i)}, \ldots, X_{K-1}^{(i)}, W_0^b, W_1^b, \ldots, W_{L-1}^b\right], \tag{4}$$

where $K + L = N$ and $L < K$. b is the selected dummy sequence number. The final complex base-band OFDM signal for the i^{th} user is written as follows:

$$\begin{aligned}\hat{x}^{(i)}(n) = \frac{1}{\sqrt{N}} \sum_{z=0}^{N-1} \hat{X}_z^{(i)} e^{\frac{j2\pi zn}{N}} = \ & \frac{1}{\sqrt{N}} \Big(X_0^{(i)} e^{\frac{j2\pi 0n}{N}} + X_1^{(i)} e^{\frac{j2\pi 1n}{N}} + \ldots + X_{K-1}^{(i)} e^{\frac{j2\pi (K-1)n}{N}} \\ & + W_0^b e^{\frac{j2\pi Kn}{N}} + \ldots + W_{L-1}^b e^{\frac{j2\pi (N-1)n}{N}} \Big); \ 0 \le n \le N-1.\end{aligned} \tag{5}$$

Table 1: PRECODING MATRICES

Precoding	Precoding Matrix Elements
Walsh-Hadamard Transform (dimension should be a power of 2)	$\mathbf{W}_1 = 1, \mathbf{W}_{2m} = \begin{bmatrix} \mathbf{W}_m & \mathbf{W}_m \\ \mathbf{W}_m & -\mathbf{W}_m \end{bmatrix}$
Zadoff chu Transform (ZCT)	$P_{i,m} = \frac{1}{\sqrt{N}} z(k) = \frac{1}{\sqrt{N}} \begin{cases} \exp\left(\frac{j\pi}{L} k^2\right) & \text{for } l \text{ even} \\ \exp\left(\frac{j\pi}{L}\left(k^2 + k\right)\right) & \text{for } l \text{ odd} \end{cases}$ $; L = N \times N, \quad k = i \times N + m$
T-transform (dimension should be a power of 2)	$\mathbf{W}_N \times (\mathbf{ZCT})_N$
Discrete Fourier Transform	$P_{i,m} = exp\left(j \frac{2\pi i m}{N}\right)$
Discrete Sine Matrix Transform	$P_{i,m} = \sqrt{\frac{2}{N}}\, \gamma\ sin\left(\frac{(\pi(2i+1)(m+1))}{2N}\right)$ $; \gamma = \begin{cases} \frac{1}{\sqrt{2}} & m = N-1 \\ 1 & \text{Otherwise} \end{cases}$
Discrete Cosine Matrix Transform	$P_{i,m} = \begin{cases} \frac{1}{\sqrt{N}} & i = 0, \quad 0 \leq m \leq N-1 \\ \sqrt{\frac{2}{N}} cos \frac{\pi(2m+1)i}{2N} & 1 \leq i \leq N-1 \\ & 0 \leq m \leq N-1 \end{cases}$
Discrete Hartley Matrix Transform	$P_{i,m} = \frac{1}{\sqrt{N}}\left[\cos\left(\frac{2\pi}{N} i m\right) + \sin\left(\frac{2\pi}{N} i m\right)\right]$

There are several dummy sequence generation techniques to be utilized in the DSI method. For example, complementary sequences and correlation sequences can be used as dummy sequences, or it can be generated by taking an initial all-zero or all-one sequence and constructing new sequences by flipping each of the dummy bits sequentially.

The DSI method does not need to send any side information, and therefore, does not degrade the BER performance. However, it has a low PAPR reduction property, which makes it insufficient as a main PAPR reduction scheme for most of the OFDM-based systems.

3.2. Precoding Method

Figure 3 demonstrates the block diagram of the precoding methods. In this scheme, the modulated input data precodes through multiplying with a precoding matrix (e.g., DSTM, DHMT,...). This multiplication reduces the aperiodic autocorrelation in the input of IFFT, and hence, leads to lower PAPR [28].

Consider $1 \times N$ modulated input data vector of the i^{th} user as $\mathbf{X}^{(i)} = \left[X_0^{(i)}, X_1^{(i)}, \ldots, X_{N-1}^{(i)}\right]$. If we have a $N \times N$ precoding matrix and call it $\mathbf{P}$, then $1 \times N$

precoded input data vector for the i^{th} use will be as follows:

$$\mathbf{D}^{(i)} = \mathbf{X}^{(i)}\mathbf{P} = \left[D_0^{(i)}, D_1^{(i)}, \ldots, D_{N-1}^{(i)}\right] . \tag{6}$$

Table 1 shows some of the well-known PAPR reduction precoding matrices. $P_{i,m}$ is the precoding matrix element in the m^{th} row and n^{th} column, where $0 \leq i, m \leq N-1$.

The receiver multiplies the received signal with the inverse precoding matrix to decode the received data:

$$\widetilde{\mathbf{X}}^{(i)} = \widetilde{\mathbf{D}}^{(i)}\mathbf{P}^{-1} = \left[\widetilde{X}_0^{(i)}, \widetilde{X}_1^{(i)}, \ldots, \widetilde{X}_{N-1}^{(i)}\right] , \tag{7}$$

where $\widetilde{\mathbf{D}}^{(i)}$ is the received signal of the i^{th} user after the SIC and $\mathbf{P}^{-1}$ is the inverse precoding matrix.

The precoding method has acceptable PAPR reduction property and does not distort the signal. It needs no side information to be transmitted, and therefore does not degrade the BER. However, the PAPR reduction property of this method is not as good as the PTS-based methods.

3.3. PTS-based Methods

In the C-PTS method, the block of input data symbols is divided into M disjoint subblocks via the Interleave (IL-PTS), Adjacent (AP-PTS), or Pseudo-Random (PR-PTS) partitioning scheme. Later, each subblock passes through the IFFT and multiplies by a phase rotation factor b_m. Then, the disjoint subblocks recombine to generate the first candidate signal. The system takes the phase rotation factors from the following set:

$$b_m \in \left\{e^{\frac{j2\pi v}{V}} \mid v = 0,\ 1, \ldots, V-1\right\}, \tag{8}$$

where b_m is the phase rotation factor for the m^{th} subblock and V is the number of possible phases of rotation factors. From [31], the phase rotation factors are taken from the set of $\{1, -1\}$ or $\{1, -1, j, -j\}$ to reduce the overall computational complexity. After constructing V^{M-1} candidate signals via multiplying each of the M subblocks by V different phase rotation factors and recombining them, the system transmits the signal with the least PAPR.

Figure 4 shows the block diagram of the PTS method. Although this system has high PAPR reduction property, it needs to send the selected phase rotation factors as side information to the receiver, which reduces the BER performance. Furthermore, the most important

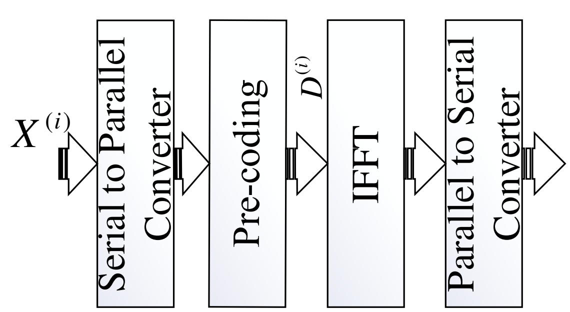


Figure 3: The block diagram of the precoding method

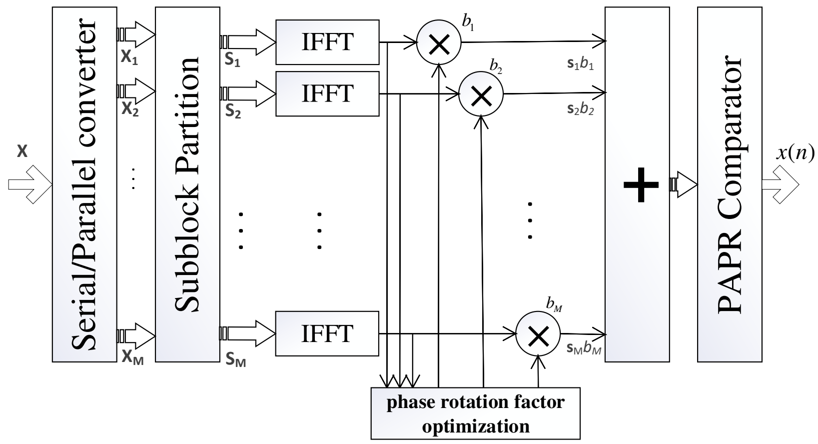


Figure 4: The block diagram of the PTS method

disadvantage of the PTS-based methods is high computational complexity. Therefore, the main focus on the recent PTS-based papers was reducing the complexity [10]-[13]. One of the effective approaches in reducing the complexity of the PTS-based methods is Gray Phase Factor PTS (Gray-PF-PTS) [10]. This method utilizes the Gray code strings to minimize the number of iterations for finding proper phase rotation factors without degrading the PAPR reduction.

4. Proposed Method

One of the main shortcomings of the precoding methods is their lower PAPR reduction compared to the PTS-based ones. Utilizing the DSI technique with the precoding methods could improve the PAPR reduction property of the system. However, the conventional dummy sequence generation techniques, such as complementary sequences, may not have a significant impact on the PAPR reduction when combined with precoding methods. For this reason, we suggest a special dummy sequence generation technique based on the complementary pairs.

An efficient PAPR reduction algorithm based on the DSI and precoding methods is proposed. The OFDM signals generated via the new scheme will have a lower PAPR value, and consequently, it will have a better BER performance. The proposed algorithm will have all the positive points of the precoding methods with a much lower PAPR value. Hence, it outperforms PTS-based methods in terms of PAPR reduction, BER, and computational complexity.

Consider $\mathbf{X}^{(i)} = \left[X_0^{(i)}, X_1^{(i)}, \ldots, X_{K-1}^{(i)}\right]$ is modulated data for the i^{th} user, and $\mathbf{P}$ is a $K \times K$ precoding matrix. The precoded data sequence of the i^{th} user is as follows:

$$\mathbf{D}^{(i)} = \mathbf{P}\mathbf{X}^{(i)} = \left[D_0^{(i)}, D_1^{(i)}, \ldots, D_{K-1}^{(i)}\right]. \tag{9}$$

Also, consider $\mathbf{W}^b = \left[W_0^b, W_1^b, \ldots, W_{L-1}^b\right]$ is the inserted dummy sequence, where b is the selected dummy sequence number and $K + L = N$. The IFFT input block for the i^{th} user is

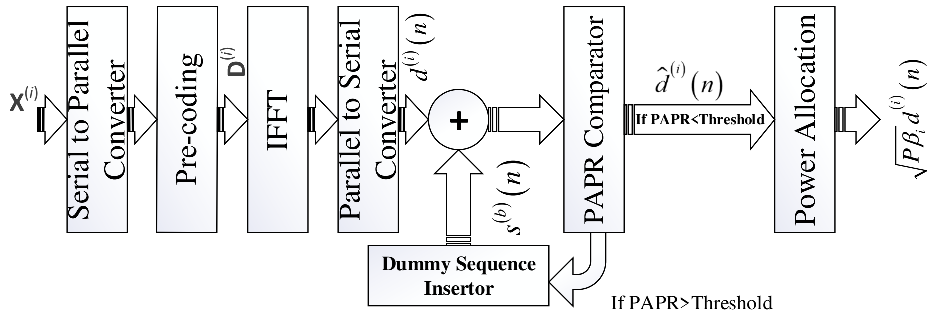


Figure 5: The block diagram of the DSI-Precoding method

as follows:

$$\hat{\mathbf{D}}^{(i)} = \left[D_0^{(i)}, D_1^{(i)}, \ldots, D_{K-1}^{(i)}, W_0^b, W_1^b, \ldots, W_{L-1}^b\right]. \tag{10}$$

Substituting (10) into (5), we can re-write the output signal as the sum of the two separate signals:

$$\begin{aligned}
\hat{d}^{(i)}(n) &= \frac{1}{\sqrt{N}} \sum_{z=0}^{N-1} \hat{D}_z^{(i)} e^{\frac{j2\pi zn}{N}} \\
&= \frac{1}{\sqrt{N}} \Big(X_0^{(i)} e^{\frac{j2\pi 0n}{N}} + X_1^{(i)} e^{\frac{j2\pi 1n}{N}} + \ldots + X_{K-1}^{(i)} e^{\frac{j2\pi(K-1)n}{N}} + W_0^b e^{\frac{j2\pi Kn}{N}} + \ldots \\
&+ W_{L-1}^b e^{\frac{j2\pi(N-1)n}{N}} \Big) = IFFT\left\{\mathbf{D}^{(i)}, \overbrace{0, \ldots, 0}^{L}\right\} + IFFT\left\{\overbrace{0, \ldots, 0}^{K}, \mathbf{W}^b\right\} \\
&= d^{(i)}(n) + s^{(b)}(n).
\end{aligned} \tag{11}$$

where $d^{(i)}(n)$ denotes the time-domain data signal for the i^{th} user and $s^{(b)}(n)$ donates the b^{th} time-domain dummy signal.

The system could pre-calculate and save the time-domain dummy signals. Later, add them to the time-domain data signal on each iteration instead of generating and adding the dummy sequence into the IFFT input. The system checks the PAPR on each iteration to choose the signal with the least PAPR value. This process avoids unnecessary IFFT operations on each loop and makes the system more efficient. Also, the system can utilize the threshold technique to decrease the number of iterations and the overall complexity. In this case, after each iteration, if the PAPR value is less than a predetermined value, the system transmits the signal without checking the other dummy sequences.

Figure 5 illustrates the block diagram of the DSI-Precoding method. First, precoded data passes through the IFFT. Then, the first time-domain dummy signal adds to the time-domain data signal. If the PAPR value is less than the threshold, the signal transmits. Otherwise, another time-domain dummy signal is added to the time-domain data signal, and the system rechecks the PAPR. This process continues until finding an appropriate signal with a PAPR less than the threshold or until checking all candidate signals and selecting one with the least PAPR value.

Note that while the length of the dummy sequence is less than the number of guard subcarriers, the inserted dummy sequence will not reduce the bit rate. Since both of them have no value of information, we could utilize guard subcarriers for placing the dummy sequence [14]. For instance, there are 75 guard subcarriers in the third generation partnership project long term evolution (3GPP LTE) and 55 guard subcarriers in the WiMAX systems with 256 subcarriers.

4.1. A Novel Dummy Sequence Generation Method

To improve the PAPR reduction property of the DSI-Precoding method, we exploit a special set of dummy sequences. These sequences generate in the following steps:

1. Construct J Golay complementary pairs [32], with the same alphabet of the system modulation, where J is the total number of iterations to find an appropriate signal with the least PAPR value. Assume sequence $\mathbf{q}$ and sequence $\mathbf{p}$ are Golay complementary pairs. In this case, the inserted dummy sequence is constructed as follows:

$$\bar{\mathbf{W}}^b = \left[\underbrace{q_0, \ldots, q_{\frac{L}{2}-1}, p_0, \ldots, p_{\frac{L}{2}-1}}_{L} \right], \quad 1 \leq b \leq J \; . \tag{12}$$

2. Precode each dummy sequence with the same precoding method used for the input data, i.e.,

$$\mathbf{W}^b = \bar{\mathbf{P}}.\bar{\mathbf{W}}^b = \left[W_0^b, W_1^b, \ldots, W_{L-1}^b \right], \quad 1 \leq b \leq J \; , \tag{13}$$

where $\bar{\mathbf{P}}$ is a $L \times L$ precoding matrix.

3. Calculate the 4N-point IFFT of each precoded dummy sequence by placing K zeros at the beginning of them:

$$s^{(b)}(n) = IFFT \left\{ \overbrace{0, \ldots, 0}^{K}, W^b \right\}, \quad 1 \leq b \leq J \; . \tag{14}$$

4. Save J time-domain dummy signals to check the PAPR value in each iteration.

In [33], the authors proved that the upper bound of the PAPR for any Golay complementary pair is 2. Hence, they are a proper candidate to be the dummy sequence that might reduce the signal PARP. Let $\mathbf{q}$ and $\mathbf{p}$ be complex sequences with equal length of N. $\mathbf{q}$ and $\mathbf{p}$ are Golay complementary pairs if they satisfy the following condition:

$$C_{\mathbf{p},\mathbf{p}}(\tau) + C_{\mathbf{q},\mathbf{q}}(\tau) = \begin{cases} 0 & , \tau \neq 0 \\ 2N, & \tau = 0 \end{cases} , \tag{15}$$

where $C_{\mathbf{q},\mathbf{q}}(\tau)$ and $C_{\mathbf{p},\mathbf{p}}(\tau)$ are the aperiodic autocorrelation functions of sequence $\mathbf{q}$ and $\mathbf{p}$. The aperiodic autocorrelation functions of a sequence like $\mathbf{u}$ with a length of N is defined as follows:

$$C_{\mathbf{u},\mathbf{u}}(\tau) = \begin{cases} \sum\limits_{k=0}^{N-1-\tau} u_k u_{k+\tau}^* & , 0 \leq \tau \leq N-1 \\ \sum\limits_{k=0}^{N-1+\tau} u_{k-\tau} u_k^* & , 1-N \leq \tau < 0 \\ 0, otherwise \end{cases} , \tag{16}$$

where $\mathbf{u}^*$ is the complex-conjugate of $\mathbf{u}$.

The detection of the inserted dummy sequence is not necessary for the SIC at the receiver, and the wrong detection of it will not hurt the data (proof in Appendix A).

5. Computational Complexity

Computational complexity is one of the important factors in wireless communication systems, especially in OFDM-NOMA systems. In the C-PTS and Gray-PF-PTS C methods with 4-times oversampling, Adjacent or Pseudo-Random partitioning scheme, and Cooley-Tukey IFFT algorithm, the system needs [34], [10]:

$$\begin{aligned}
&C_{Mult.}^{C-PTS} = M\left(2N\log_2(4N)\right) + 4V^{M-1}K(M+1),\\
&C_{Add.}^{C-PTS} = M\left(4N\log_2(4N)\right) + 4V^{M-1}K(M-1),\\
&,\\
&C_{Mult.}^{\mathrm{Gray-PF-PTS\ C}} = M\left(2N\log_2(4N)\right) + \left[\left[12K \times 2^{M-1}\left(\frac{M}{2}+3\right)\right] - 28K\right],\\
&C_{Add.}^{\mathrm{Gray-PF-PTS\ C}} = M\left(4N\log_2(4N)\right) + \left[\left[12KM \times 2^{M-1}\right] - 16K\right],
\end{aligned} \tag{17}$$

complex operations, where $C_{Mult.}$ and $C_{Add.}$ are the number of complex multiplication and addition operations. Due to the inserted dummy sequence, the DSI-PTS method needs fewer subblocks than the C-PTS method to have the same PAPR reduction [14]. Also, by using the suggested dummy sequence insertion technique, it will need $4JN$ more additions to add J dummy sequences and check their PAPR values.

For multiplying a $K \times K$ precoding matrix by a $1 \times K$ input vector, we need $O(K^2)$ operations. But fortunately, the fast implementation needs only $\frac{K}{2}\log_2 K$ additions or subtractions and $K\log_2 K$ multiplications [29]. For the Cooley-Tukey IFFT algorithm with 4-times oversampling, the system needs $2N\log_2(4N)$ complex multiplication and $4N\log_2(4N)$ complex addition operations [35]. Furthermore, Adding J dummy signals to the time-domain data signal and selecting one with the least PAPR needs $4JN$ complex addition operations. Consequently, the DSI-Precoding method needs:

$$\begin{aligned}
C_{Mult.}^{DSI-Precoding} &= K\log_2(K) + 2N\log_2(4N),\\
C_{Add.}^{DSI-Precoding} &= \frac{k}{2}\log_2(K) + 4N\log_2(4N) + 4JN,
\end{aligned} \tag{18}$$

complex multiplication and addition operations.

Table 2 demonstrates the number of complex operations for six systems with 128, 256, and 512 subcarriers. The length of input data (K) is 96, 192, and 384 for the systems with 128, 256, and 512 subcarriers. The first and third system exploits the C-PTS and Gray-PF-PTS C methods for PAPR reduction in which $M = 4$ and $V = 4$. The second system uses the DSI-PTS method with hundred iterations ($J = 100$) in which $M = 3$ and $V = 4$. The fourth system exploits precoding methods for PAPR reduction. The fifth and sixth system employs the DSI-Precoding technique with $J = 10$ and $J = 100$ iterations, respectively. The table indicates that the precoding methods with fast implementation have the least computational complexity compared to the others. However, they have a lower

Table 2: Number of Complex Operation Needed for Each Method

	128		256		512	
	$C_{Add.}$	$C_{Mult.}$	$C_{Add.}$	$C_{Mult.}$	$C_{Add.}$	$C_{Mult.}$
C-PTS	92,160	132,092	188,416	266,240	383,488	536,576
DSI-PTS with 100 iterations	81,408	39,680	165,888	80,896	337,920	164,864
Gray-PF-PTS C	53,760	52,608	111,616	107,264	231,424	218,624
Precoding	5,056	3,200	11,264	7,168	24,832	15,872
DSI-Precoding with 10 iterations	10,000	2,936	21,208	6,576	44,656	14,560
DSI-Precoding with 100 iterations	56,124	2,936	113,369	6,576	228,976	14,560

PAPR reduction than the other methods. The proposed method has better PAPR reduction compared to the precoding methods at the cost of some extra complex additions for inserting the dummy sequence.

6. Numerical Results and Discussions

In this section, first, we will study the effect of the proposed method using the Aperiodic-Autocorrelation Function (ACF). Later, we will present Power Spectral Density (PDS) analysis of the proposed method. Finally, we will analyze the PAPR reduction and BER performance of the system via computer simulations.

6.1. The Effect of Precoding and Adding Dummy Sequence

To study the effect of joint precoding and dummy sequence insertion, we utilize the ACF. Low autocorrelation in the IFFT input leads to a lower risk of in-phase addition at the IFFT [22], [28]. Low in-phase addition at the IFFT means low PAPR. Figure 6 depicts the ACF for three sequences of size 64 taken from the QPSK constellation. The first sequence uses no precoding method, the second sequence uses the DSMT precoding, and the third one uses

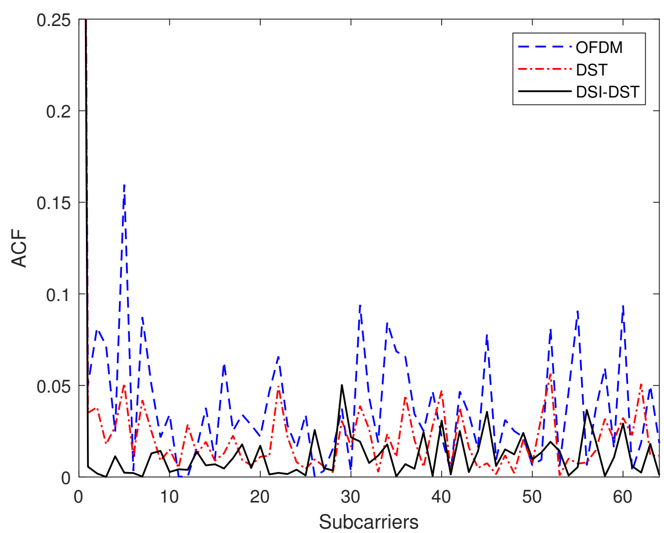


Figure 6: ACF characteristics of three sequences with different PAPR reduction methods

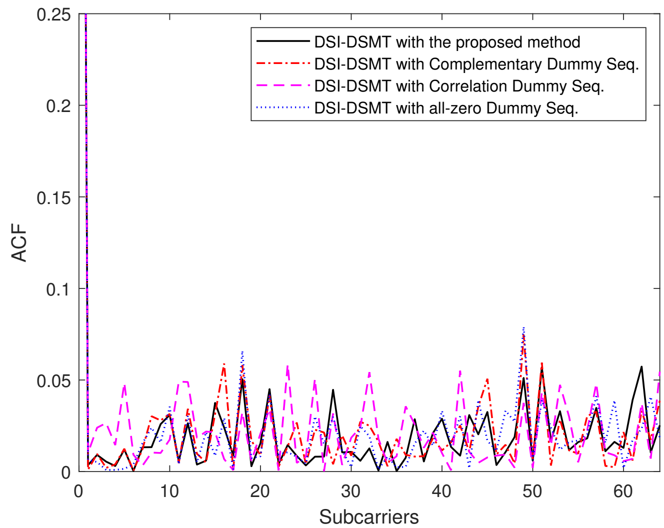


Figure 7: ACF characteristics of DSI-DSMT signals with different dummy sequence generation techniques

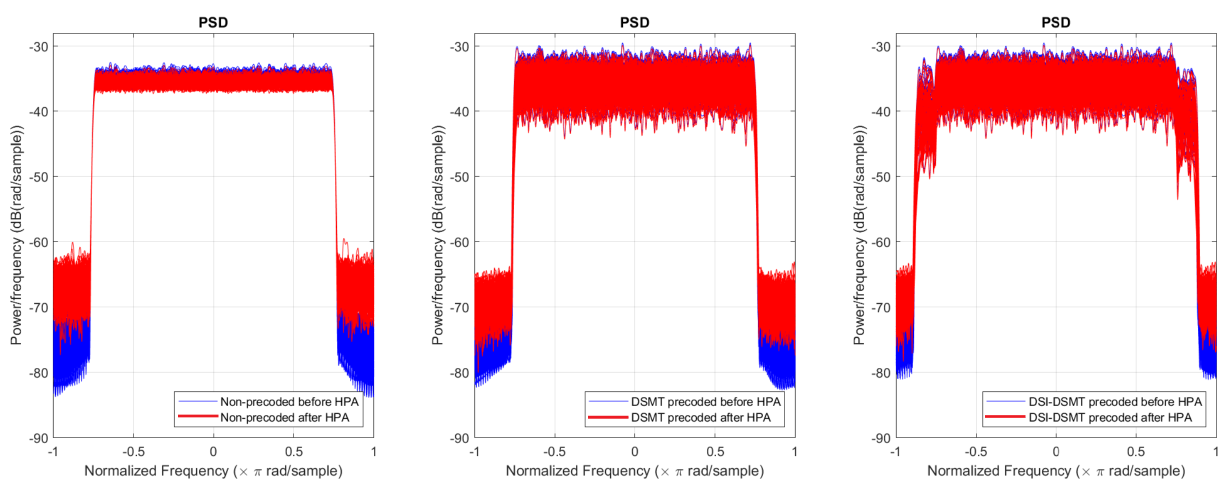


Figure 8: Power Spectral Density of Non-Precoded, DSMT precoded, and DSI-DSMT precoded signals before and after the HPA

the DSI-DSMT method. As it can be seen, the Non-Precoded sequence and the sequence generated by the proposed method have the highest and the least side-lobes, respectively. A higher side-lobe indicates higher autocorrelation among the input data and verse versa. For the DSMT precoded sequence, the side-lobe is between the first and third sequences. Figure 7 shows the ACF for DSI-DSMT precoded signals with the proposed and conventional dummy sequence generation techniques. As it can be seen, the DSI-DSMT signal with the proposed dummy sequence generation algorithm has the least side-lobes, which means a lower chance of in-phase addition.

6.2. Power Spectral Density (PSD) Analysis

In this subsection, we will analyze the PSD performance of the proposed method. Figure 8 shows PSD for non-precoded, DSMT precoded, and DSI-DSMT precoded signals before

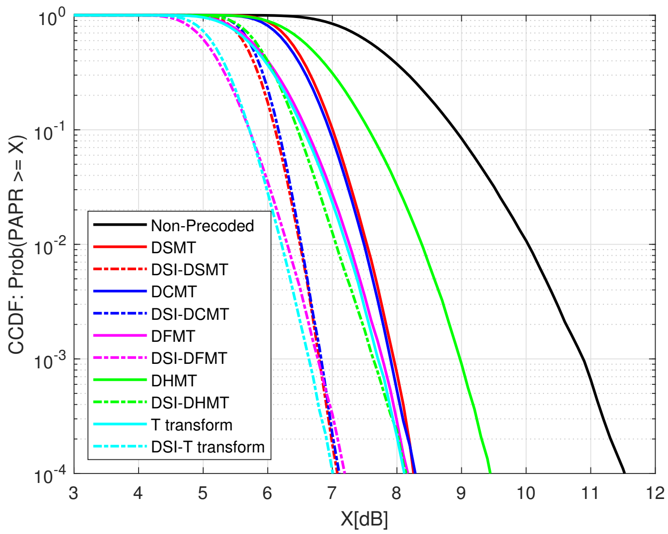


Figure 9: PAPR CCDF curves for Non-Precoded and DSI-Precoding signals with DSMT, DCMT, DFMT, DHMT, and T transform precoding matrices and ten iterations

and after the HPA. All of the systems have 256 subcarriers with 64 guard subcarriers. In the DSI-DSMT method, the length of the dummy sequence is 32, and the number of iterations for finding a proper dummy sequence is 100. The results indicate that there is a lower distortion caused by the HPA in the signals with the DSMT and DSI-DSMT PAPR reduction methods compared to the non-precoded one. In the DSI-DSMT signal, there is an extra distortion in the guard interval because of the inserted dummy sequence in the guard subcarriers. However, that part will interfere with the next system's guard interval, and it will not distort the data.

6.3. Simulation Results

In this part, we will evaluate the BER and PAPR reduction performance of the proposed method via computer simulations. MATLAB has been used as simulation software. The system used in this simulation has modeled in section 2. For simplicity, we assumed there are only two users in the network, which are transmitting their messages over the same time-frequency resources via power-domain NOMA. The power allocation coefficients for the first and second user is $\beta_1 = 0.8$ and $\beta_2 = 0.2$. The number of subcarriers is 256 with 64 guard subcarriers. So, the length of the dummy sequence in the DSI and DSI-Precoding methods is 64. The digital signal modulation scheme used in this simulation is quadrature phase-shift keying (QPSK). We utilized the method described in [36] to generate complementary pairs for the proposed dummy sequence generation algorithm. Finally, the additive white Gaussian noise (AWGN) channel is used in the simulations.

Figure 9 shows PAPR Complementary Cumulative Distribution Function (CCDF) curves for Non-Precoded, precoded, and DSI-Precoded signals with DSMT, DCMT, DFMT, DHMT, and T transform precoding matrices and ten iterations. DHMT has the worst PAPR reduction performance among the compared precoding matrices. Likewise, DSI-DHMT precoded signal has the worst performance between the DSI-Precoding methods. The precoded

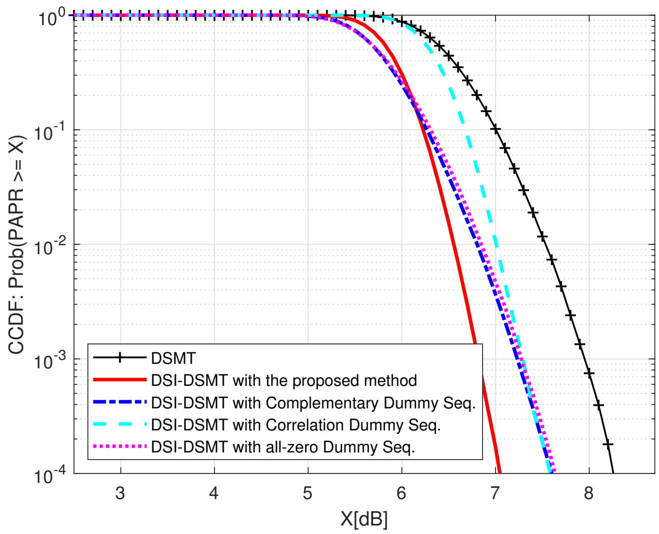


Figure 10: PAPR CCDF curves for comparing the proposed dummy sequence generation method with the conventional ones with ten iterations for finding a proper dummy sequence

signals with DSMT, DCMT, DFMT, and T-transform matrices have almost similar PAPR reduction performances. Similarly, DSI-Precoded signals with DSTM, DCMT, DFMT, and T-transform matrices have almost similar PAPR performances. As it can be seen, the proposed DSI-Precoding technique does well with different precoding matrices. It can reduce the PAPR value regardless of the used PAPR reduction precoding matrix. Therefore, in this simulation, we will use DSMT as the precoding matrix for evaluating the BER and PAPR reduction performance of the proposed technique. Other precoding matrices with the same PAPR reduction performance will have similar results.

Figure 10 compares the PAPR CCDF curves of the DSMT and DSI-DSMT precoded signals with different dummy sequence generation techniques and ten iterations for finding a proper dummy sequence. The first DSI-DSMT system utilizes the proposed dummy sequence generation technique while the others exploit complementary sequence, correlation sequence, and initially all-zero sequence methods to form the dummy sequence. In the complementary dummy sequence method, a complementary sequence is inserted as the dummy sequence, and on each iteration, the system replaces the inserted complementary sequence with another one. This method is the most effective and commonly used dummy sequence generation technique in the previous works. In the correlation sequence method, the input data is divided into several subblocks, and a correlation sequence inversely corresponding to the first symbols of each partitioned subblocks is found and inserted as the dummy sequence. Later, on each iteration, the system uses the flipping technique on the correlation sequence to form new dummy sequences [15]. In the initially all-zero sequence method, first, an all-zero sequence is inserted as the dummy sequence. Then, on each iteration, those zeros sequentially flipped to create new dummy sequences. As it can be seen, our proposed dummy sequence generation approach has the best PAPR reduction performance compared to the others. Complementary dummy sequence and initially all-zero sequence methods have almost the same PAPR reduction. The DSI-DSMT system with the correlation sequence method has

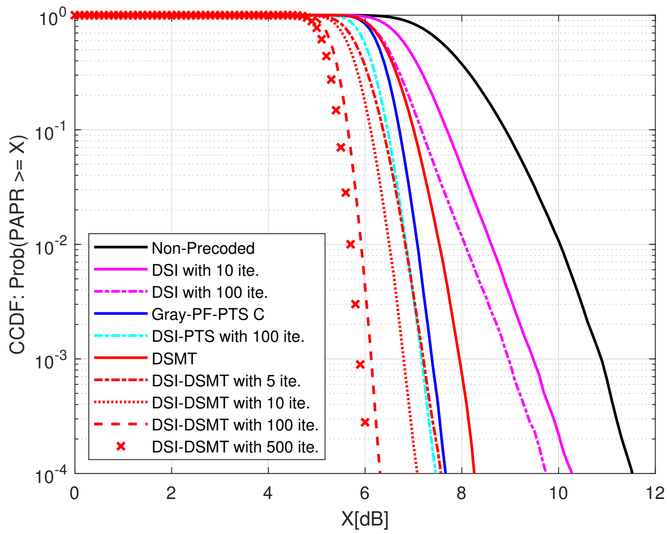
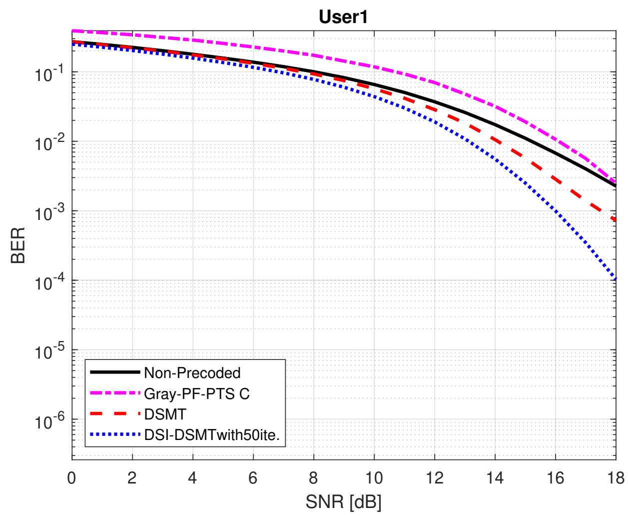


Figure 11: PAPR CCDF curves for Non-Precoded, DSI, DSI-PTS, Gray-PF-PTS C, DSMT, and DSI-DSMT precoded signals with different iterations in DSI-based techniques

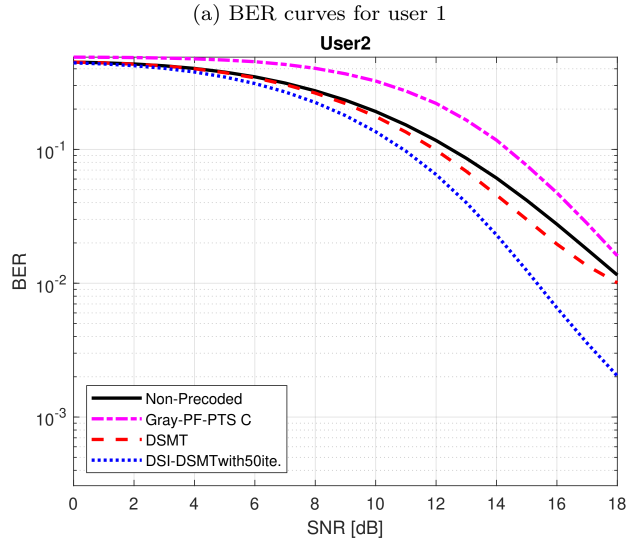
the worst PAPR reduction performance among the compared ones.

Figure 11 compares the PAPR CCDF curves of DSI-DSMT signals with the PAPR CCDF curves of DSMT, DSI, DSI-PTS, and Gray-PF-PTS C signals for various iterations. In the Gray-PF-PTS C and DSI-PTS, the input data divide into four and three disjoint subblocks, respectively. There are four possible phase rotation factors in the PTS-based techniques. As it can be seen, the suggested method with five iterations has almost the same PAPR reduction performance compared to the PTS-based schemes. In ten iterations, the proposed method outperforms the others. It has a PAPR greater than 6.8dB with the probability of 10^{-3} while the Gray-PF-PTS C and DSI-PTS with hundred iterations have a PAPR greater than 7.8dB and 7.6dB. The DSMT precoding method has a PAPR greater than 8.2dB with the same probability. The DSI method has the worst performance with a PAPR value greater than 8.9dB and 9.5dB for 100 and 10 iterations. Increasing the number of iterations in the DSI-DSMT method increases the PAPR reduction performance of the system. However, there is a little difference in the PAPR reduction between 100 and 500 iterations (about 0.2dB). The proposed technique with 100 iterations has a PAPR greater than 6.1dB with a probability of 10^{-3}.

Figure 12 shows the BER curves for the first and second user in the described OFDM-NOMA system with different PAPR reduction methods. Rapp's model [37] is considered for modeling the effect of the HPA. The Gray-PF-PTS C method has the worst BER performance because it needs to transmit the selected phase rotation factors. However, in the realm of high signal-to-noise ratio (SNR), it outperforms the Non-Precoded system because of the lower PAPR. The system with the DSMT precoding method has better BER performance due to the lower PAPR compared to the Non-Precoded signal and no need for transmitting side-information. The DSI-DSMT method with 50 iterations has the best BER performance. It needs no side information to be sent and has the least PAPR value, which results in lower BER. The second user has better channel condition compared to the first user and hence,



(a) BER curves for user 1



(b) BER curves for user 2

Figure 12: BER vs SNR for Non-Precoded, DSMT precoded and DSI-DSMT OFDM-NOMA systems

higher SNR. To ensure user fairness, they should have the same BER performance. As it can be seen from figures, for the BER of 2.5×10^{-3}, the first user needs an SNR of 15dB. Whereas, at the same BER, the second user needs an SNR of 18dB.

7. Conclusion

In this paper, we proposed a novel PAPR reduction method based on the DSI and precoding techniques. The suggested method benefits from a new dummy sequence generation algorithm. Simulation results prove that the proposed algorithm outperforms the conventional precoding, DSI, and PTS methods in PAPR reduction and BER performance. Moreover, it is computationally less complex than the compared PTS-based methods and does not transmit side information. Considering the results, DSI-Precoding is one of the most worthwhile PAPR reduction techniques to be used in future OFDM-based systems, especially OFDM-NOMA ones.

Appendix A.

Consider a base station exploits power-domain OFDM-NOMA to broadcast two users' data. Using (1), (2) and (11), the transmitted base-band OFDM-NOMA signal for the first and second users will be as follows:

$$\begin{aligned}
x\left(n\right) &= \frac{\sqrt{P\beta_1}}{\sqrt{N}}\Big(X_0^{(1)}e^{\frac{j2\pi 0n}{N}}+X_1^{(1)}e^{\frac{j2\pi 1n}{N}}+\ldots+X_{K-1}^{(1)}e^{\frac{j2\pi(K-1)n}{N}}+W_0^{b_1}e^{\frac{j2\pi Kn}{N}}+\ldots\\
&+W_{L-1}^{b_1}e^{\frac{j2\pi(N-1)n}{N}}\Big)+\frac{\sqrt{P\beta_2}}{\sqrt{N}}\Big(X_0^{(2)}e^{\frac{j2\pi 0n}{N}}+X_1^{(2)}e^{\frac{j2\pi 1n}{N}}+\ldots+X_{K-1}^{(2)}e^{\frac{j2\pi(K-1)n}{N}}\\
&+W_0^{b_2}e^{\frac{j2\pi Kn}{N}}+\ldots+W_{L-1}^{b_2}e^{\frac{j2\pi(N-1)n}{N}}\Big)=\frac{1}{\sqrt{N}}\Big(\big(\sqrt{P\beta_1}X_0^{(1)}+\sqrt{P\beta_2}X_0^{(2)}\big)e^{\frac{j2\pi 0n}{N}}\\
&+\big(\sqrt{P\beta_1}X_1^{(1)}+\sqrt{P\beta_2}X_1^{(2)}\big)e^{\frac{j2\pi 1n}{N}}+\ldots+\big(\sqrt{P\beta_1}X_{K-1}^{(1)}+\sqrt{P\beta_2}X_{K-1}^{(2)}\big)e^{\frac{j2\pi(K-1)n}{N}}\\
&+\big(\sqrt{P\beta_1}W_0^{b_1}+\sqrt{P\beta_2}W_0^{b_2}\big)e^{\frac{j2\pi Kn}{N}}+\ldots+\big(\sqrt{P\beta_1}W_{L-1}^{b_1}+\sqrt{P\beta_2}W_{L-1}^{b_2}\big)e^{\frac{j2\pi(N-1)n}{N}}\Big)\\
&;0\le n\le N-1,
\end{aligned} \tag{A.1}$$

where P is the total transmit power, β_i is the fraction of total power assigned to the i^{th} user, $\mathbf{X}^{(i)}=\left[X_0^{(i)},X_1^{(i)},...,X_{N-1}^{(i)}\right]$ is the input data vector for the the i^{th} user, and $\mathbf{W}^{(i)}=\left[W_0^{b_i},W_1^{b_i},...,W_{N-1}^{b_i}\right]$ is the selected dummy sequence for the the i^{th} user.

The second user is performing SIC. It should detect the first one's data and subtract it from the received signal to extract its data. Assume $\tilde{\mathbf{X}}^{(i)}$ is the first user's detected symbol vector, and the second user have perfectly decoded the first user's data ($\tilde{\mathbf{X}}^{(i)}=\mathbf{X}^{(i)}$). The second user does not know the inserted dummy sequence. The time-domain signal of the second user after SIC will be as follows:

$$\begin{aligned}
\tilde{x}^{(2)}\left(n\right)&=x\left(n\right)-\frac{\sqrt{P\beta_1}}{\sqrt{N}}\left(\tilde{X}_0^{(1)}e^{\frac{j2\pi 0n}{N}}+\tilde{X}_1^{(1)}e^{\frac{j2\pi 1n}{N}}+\ldots+\tilde{X}_{K-1}^{(1)}e^{\frac{j2\pi(K-1)n}{N}}\right)\\
&=\frac{\sqrt{P\beta_2}}{\sqrt{N}}\Bigg(\left(X_0^{(2)}\right)e^{\frac{j2\pi 0n}{N}}+\left(X_1^{(2)}\right)e^{\frac{j2\pi 1n}{N}}+\ldots+\left(X_{K-1}^{(2)}\right)e^{\frac{j2\pi(K-1)n}{N}}\\
&+\left(\frac{\sqrt{P\beta_1}}{\sqrt{P\beta_2}}W_0^{b_1}+W_0^{b_2}\right)e^{\frac{j2\pi Kn}{N}}+\ldots+\left(\frac{\sqrt{P\beta_1}}{\sqrt{P\beta_2}}W_{L-1}^{b_1}+W_{L-1}^{b_2}\right)e^{\frac{j2\pi(N-1)n}{N}}\Bigg)\\
&;0\le n\le N-1.
\end{aligned} \tag{A.2}$$

The detected data vector for the second user will be:

$$\begin{aligned}
\mathbf{X}^{(2)}&=\frac{1}{\sqrt{P\beta_2}}\mathrm{FFT}\left(\tilde{x}^{(2)}\left(n\right)\right)\\
&=\left[X_0^{(2)},X_1^{(2)},...,X_{K-1}^{(2)},\frac{\sqrt{P\beta_1}}{\sqrt{P\beta_2}}W_0^{b_1}+W_0^{b_2},...,\frac{\sqrt{P\beta_1}}{\sqrt{P\beta_2}}W_{L-1}^{b_1}+W_{L-1}^{b_2}\right].
\end{aligned} \tag{A.3}$$

As it can be seen, the wrong detection of the inserted dummy sequence in the SIC does not hurt the data. Note that the inserted dummy sequence has no value of information.

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