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Hybrid SLM-PTS for PAPR Reduction in MIMO-OFDM

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ABSTRACT: Multiple-Input Multiple-Output Orthogonal Frequency Division Multiplexing (MIMO-OFDM) is a promising candidate for 4G broadband wireless communications. However, MIMO-OFDM inherited the problem of high Peak-to-Average Power Ratio (PAPR) from OFDM. Many PAPR reduction techniques were developed in last two decades to reduce the PAPR of OFDM, among them Partial Transmit Sequence (PTS) and Selected Mapping (SLM) show a highly successful PAPR reduction performance. In literature there are three well known approaches for extending SLM and PTS to MIMO-OFDM namely ordinary (oSLM/oPTS), simplified (sSLM/sPTS), and directed (dSLM/dPTS). Hybrid SLM-PTS techniques combine SLM and PTS in four different ways to reduce the required computational complexity lower than both SLM and PTS. Here, we will show the performance of applying ordinary and simplified approaches on the Hybrid SLM-PTS techniques in MIMO-OFDM system. Also, we will investigate the possibility of applying directed approach to Hybrid SLM-PTS techniques by means of proposed approach that combining dSLM and dPTS in one approach.

KEYWORDS: MIMO, OFDM, PAPR, SLM, PTS.

I.INTRODUCTION

The modern day phenomenon of increased thirst for more information and the explosive growth of new multimedia wireless applications have resulted in an increased demand for technologies that support very high speed transmission rates, mobility and efficiently utilize the available spectrum and network resources. Orthogonal Frequency Division Multiplexing (OFDM) is one of the best solutions to achieve this goal and it offers a promising choice for future high speed data rate systems [1]. OFDM, which is one of multi-carrier modulation (MCM) techniques, offers a considerable high spectral efficiency, multipath delay spread tolerance, immunity to the frequency selective fading channels and impulse noise, power efficiency and eliminates the need for equalizers, while efficient hardware implementation can be realized using fast Fourier transform (FFT) techniques[2-3]. Multiple-Input Multiple-Output (MIMO) is known to boost capacity. For high data rate transmission, the multipath characteristic of the environment causes the MIMO channel to be frequency-selective. OFDM can transform such a frequency-selective MIMO channel into a set of parallel frequency-flat MIMO channels, and therefore decrease receiver complexity. The combination of the two powerful techniques, MIMO and OFDM, is very attractive, and has become a most a promising candidate for 4G broadband wireless communications [4]. However, one main disadvantage of MIMO-OFDM is that the signals transmitted on different antennas might have large Peak-to-Average Power Ratio (PAPR), Since MIMO-OFDM system is based on OFDM, it will also suffers from the problem of inherent high PAPR [5]. This phenomenon results from that in the time domain, an OFDM signal is the superposition of many narrowband subcarriers. At certain time instances, the peak amplitude of the signal is large and at the other times is small, that is, the peak power of the signal is substantially larger than the average power of the signal. When a high PAPR OFDM signal passes through a nonlinear device, it may cause in-band distortion and undesired spectral spreading. Thus, handling occasional large peaks leads to low power efficiency and then increases the cost of the RF power amplifier. Therefore, how to find a solution to reduce high PAPR effectively is one of the most important implementation issues in OFDM communications [6-7]. There has been a significant amount of research devoted to the development of PAR reduction algorithms for OFDM. But, in general PAPR reduction techniques achieve PAPR reduction at the expense of transmit signal power increase, bit error rate (BER) increase, data rate loss, computational complexity increase [3]. Selected Mapping (SLM) and Partial

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Transmit Sequence (PTS) schemes are widely studied techniques because they show good PAPR reduction performance without BER degradation, by optimal using of redundancy bits. However, they require many Inverse Fast Fourier Transforms (IFFTs), which cause high computational complexity, and need to transmit the Side Information (SI), delivering which phase rotation vector was used [8]. Recently different hybrid schemes combine PTS and SLM aim to reduce the computational complexity or obtain a better PAPR reduction performance compared with conventional PTS.

In this paper, Hybrid SLM-PTS techniques introduced in a general form which make them capable to be used with any number of sub-blokes, overstep being restricted to two sub-blokes as in [6] and [7]. Extension of Hybrid SLM-PTS techniques to MIMO-OFDM is proposed in this paper, *ordinary* and *simplified* approaches used to implement Hybrid SLM-PTS techniques in MIMO-OFDM systems, done in the same way as it done in [9] for conventional PTS or SLM. whereas *ordinary* approach applies PTS or SLM individually to each antenna in MIMO-OFDM, *simplified* approach on other hand, applies them concurrently. But, neither *ordinary* nor *simplified* indeed use the potential of MIMO transmission for PAR reduction. In MIMO communication, data rate or diversity order can be improved by exploiting the spatial dimension. In the same spirit, treating the parallel transmit signals jointly, PAR reduction can be improved by "reallocate the peak power over the antennas" [10]. With this spirit two approaches presented in [10-11] called *directed* SLM (*d*SLM) and directed PTS (*d*PTS) utilize the potential of MIMO transmission in PAPR reduction.

Also in this paper, a suggested technique for combining *d*SLM and *dPTS* as a *directed* Hybrid SLM-PTS technique is presented here. Whereas Hybrid SLM-PTS techniques are themselves a combination between SLM and PTS.

The rest of this paper is organized as follows: Section II, the related works are presented. Section III provides a brief description of the MIMO-OFDM system model and PAPR equation. Section IV reviews conventional SLM and conventional PTS for single antenna systems. Hybrid SLM-PTS techniques for PAPR reduction are reviewed and generalized in Section V. The extensions of Hybrid SLM-PTS techniques to MIMO-OFDM systems are given in Section VI. In Section VII, simulation results are given and the PAPR reduction performance of the three possible extension approaches per each one of the four Hybrid SLM-PTS techniques to MIMO-OFDM system is compared for different numbers of transmitting antennas. Finally, Section VIII is the conclusion.

II. RELATED WORK

Many PAPR reduction techniques have been proposed in the literature. These techniques can be broadly classified into three main categories: Signal distortion techniques, multiple signalling and probabilistic techniques, and coding techniques [1]. Clipping and filtering [12-13], windowing [14], peak cancellation [15], tone reservation (TR) [16] and Companding [17], are all belong to signal distortion techniques, where the PAPR reduced by distorting the transmitted OFDM signal before it passes through the power amplifier, this will brought errors to the system. On other hand, coding schemes such as block coding [18], LDPC coding [19] or turbo coding [20], whose always use to correct errors in the communication systems, are also have the capability to reduce the PAPR. Multiple signal representation and probabilistic techniques, include Selective mapping (SLM) [21], partial transmit sequence (PTS) [22], interleaving [23], tone injection (TI) [16], Dummy Sequence Insertion (DSI) [24], and active constellation extension (ACE) [25], in which several candidate signals are generated and the one with the minimum PAPR is selected for transmission.

Among these techniques PTS and SLM are highly successful PAPR reduction techniques. However, the highly computational complexities of both techniques limit their PAPR reduction capability. To reduce the required number of IFFT_s (computational complexity) and obtain a significant PAPR reduction performance in OFDM systems, a Hybrid SLM-PTS algorithm combining SLM and PTS was firstly given in [26] known as Conventional Hybrid (CH). Other Hybrid methods such as Additional Hybrid (AH), Switching Hybrid (SH) were introduced in [6], a Modified Hybrid algorithm (MH) combining AH with SH schemes is also proposed in [6]. Moreover one of these techniques (MH) combines with Dummy Sequence Insertion (DSI) in [7] to produce (DH) technique.

In MIMO-OFDM systems, a straightforward way for PAPR reduction is to apply existing algorithms separately on each transmit antenna. It is effective to reduce PAPR, but requires high complexity and large amount SI [5]. A first extension of SLM and PTS to MIMO-OFDM was given in [9]. It applies SLM or PTS to each antenna in MIMO-OFDM individually, this procedure called *ordinary* SLM (*o*SLM) or *ordinary* PTS (*o*PTS) respectively. Another approach proposed by Baek *et al.* in [9] aims to reduce number of SI bits called *simplified* SLM (*s*SLM) or *simplified* PTS (*s*PTS).

But, neither *ordinary* nor *simplified* indeed use the potential of MIMO transmission for PAR reduction. In [10] *directed* SLM (*d*SLM) was the first approach utilizes the potential of MIMO transmission in reducing the PAPR, after that "*directed*" approach was applied to PTS in [11] leaded directed PTS (*d*PTS).

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III. SYSTEM MODEL AND PAPR DESCRIPTION

As usual in OFDM, the information carrying symbols $S_{(\mu,k)}$ (QAM symbol) of the μ th transmit antenna are specified in frequency domain (carrier$k$) and are combined into the vector $\mathbf{S}_\mu = [S_{(\mu,1)} \quad S_{(\mu,2)} \quad \cdots \quad S_{(\mu,N-1)}]$ of length N (number of subcarriers). This vector is transformed into the time-domain vector x_μ (OFDM frame) via an IFFT, written as $\mathbf{x}_\mu = \mathbf{IFFT}(\mathbf{S}_\mu)$, with components

$$x_{(\mu,n)} = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_{(\mu,k)}\, e^{j\frac{2\pi nk}{N}} \quad ,0 \le n \le N-1 \tag{1}$$

Assuming statistically independence of the frequency-domain symbols $S_{(\mu,k)}$ and sufficiently large N, due to the central limit theorem, the resulting time-domain samples $x_{(\mu,n)}$ are approximately Gaussian distributed which leads to a high PAPR. If multiple transmit antennas are present, we consider the worst-case peak power over all transmit antennas being crucial. Define the PAPR of one OFDM frame as

$$\text{PAPR} = \max_{\mu=1,2,\dots,N_T} \text{PAPR}_\mu = \frac{\text{Max}_{n\in[0,N]} \left|x_{(\mu,n)}\right|^2}{E\left\{\left|x_{(\mu,n)}\right|^2\right\}} \tag{2}$$

where $E\{\cdot\}$ denotes the expectation operator. Note that the maximization is carried out over all time-domain samples within one OFDM frame and over all transmit antennas. As common in literature, we consider the PAPR of the discrete time signal [11]. Then, the Complementary Cumulative Distribution Function (CCDF), which is the probability that the PAPR of an OFDM symbol exceeds the given threshold $PAPR_o$, can be expressed as

$$\text{CCDF} = \text{prob}\{\text{PAPR} > PAPR_o\} \tag{3}$$

noteworthy, for conventional (single-antenna) OFDM without any PAPR reduction technique the CCDF of the OFDM signals is written as

$$\text{prob}\{\text{PAPR} > PAPR_o\} = 1 - (1 - e^{-PAPR_o})^N \tag{4}$$

in MIMO OFDM, the probability that the PAPR of a randomly generated N_T OFDM symbol over all N_T transmit antennas exceeds $PAPR_o$, is given by [5];

$$\text{prob}\{\text{PAPR} > PAPR_o\} = 1 - (1 - e^{-PAPR_o})^{N_T N} \tag{5}$$

IV. CONVENTIONAL SLM AND PTS FOR SINGLE ANTENNA SYSTEMS

A. Conventional SLM

The conventional SLM scheme generates U alternative OFDM signal sequences $\mathbf{x}^u$, $0 \le u \le U-1$ for the same input symbol sequence $\mathbf{S}$ (the antenna index μ is suppressed in this section). To generate U alternative OFDM signal sequences, U distinct phase rotation vectors $\mathbf{P}^u$ known to both transmitter and receiver are used, Where $\mathbf{P}^u = \left[P^u_{(0)} \;, P^u_{(1)} \;, \dots , P^u_{(N-1)}\right]$ With $P^u_{(k)} \;,= e^{j\,\phi^u_{(k)}}$, $\phi^u_{(k)} \in [0, 2\pi)$ in general $P^u_{(k)}$ are unit magnitude complex number selected from a binary or quaternary elements sets that is $\{\pm 1\}$ or $\{\pm 1 , \pm j\}$. $\mathbf{P}^0$ is the all-one vector for generating the original OFDM signal sequence and thus $\mathbf{x}^0 = \mathbf{x}$. An input symbol sequence S is multiplied by each phase rotation vector $\mathbf{P}^u$ element by element. Then an input symbol sequence S is represented by U different alternative input symbol sequences $\mathbf{S}^u$, where $S^u_{(k)} = S_{(k)}\, P^u_{(k)}$, $0 \le u \le U-1$.

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These U alternative input symbol sequences are transformed by U IFFTs to generate U alternative OFDM signal sequences $\mathbf{x}^{u} = \boldsymbol{IFFT}\,(\boldsymbol{S}^{u})$, and the PAPR values of them are calculated. Finally, the alternative OFDM signal sequence $\mathbf{x}^{\hat{u}}$ having the minimum PAPR, among number of alternative signal representations $I_{SLM} = U$, is selected for transmission [8] as

$$\hat{u} = \underset{0 \le u \le U-1}{\operatorname{argmin}} \left(\frac{\mathrm{Max}_{n \in [0,N]} \left| x_{(n)}^{u} \right|^{2}}{E\left\{ \left| x_{(n)}^{u} \right|^{2} \right\}} \right) \tag{6}$$

Number of side information of SLM equal to [10]

$$SI = \lceil \log_2(U) \rceil \qquad \text{(bits)} \tag{7}$$

B. Conventional PTS

The main idea is that the input symbol $\mathbf{S} = [S_{(1)} \quad S_{(2)} \quad \cdots \quad S_{(N-1)}]$ is partitioned into $\mathbf{V}$ disjoint subblocks $\mathbf{S}^{v} = \left[S_{(0)}^{v}\ , S_{(1)}^{v}\ , \dots , S_{(N-1)}^{v}\right], 1 \le v \le \mathbf{V}$. All the sub-carriers positions which are presented in other subblocks must be zero, so that the sum of all the subblocks constitutes the original signal such that $\mathbf{S} = \sum_{v=1}^{\mathbf{V}} \mathbf{S}^{v}$. Then, subblocks are combined with rotational factors b^{V} $(|b^{V}| = 1, b^{V} = e^{j\phi^{V}})$ where $\phi^{V} \in [0, 2\pi)$, to minimize the PAPR. The subblocks may be transformed by $\mathbf{V}$ separate and parallel IFFTs. Mathematically, this operation can be described as:

$$\tilde{\mathbf{x}} = \boldsymbol{IFFT}\left(\sum_{v=1}^{\mathbf{V}} (b^{v} \boldsymbol{S}^{v}) \right) = \sum_{v=1}^{\mathbf{V}} b^{v} \boldsymbol{IFFT}(\boldsymbol{S}^{v}) = \sum_{v=1}^{\mathbf{V}} b^{v} \mathbf{x}^{v} \tag{8}$$

where $\mathbf{x}^{v}$ is N-point IFFT of each sub-block. Ideally, the optimized rotation parameter set reads

$$[\hat{b}^{1} \quad \cdots \quad \hat{b}^{V}] = \underset{[b^{1} \ \cdots \ b^{V}]}{\operatorname{argmin}} \left(\frac{\mathrm{Max}_{n \in [0,N]} \left| \sum_{v=1}^{V} b^{v} x_{(n)}^{v} \right|^{2}}{E\left\{ \left| \sum_{v=1}^{V} b^{v} x_{(n)}^{v} \right|^{2} \right\}} \right) \tag{9}$$

resulting in the optimum transmit sequence

$$\hat{\mathbf{x}} = \sum_{v=1}^{\mathbf{V}} \hat{b}^{v} \mathbf{x}^{v} \tag{10}$$

That has the lowest PAPR of all alternative transmit sequences that can be generated by this method. Rotational factors are determined optimally and iteratively for the minimum PAPR. In Exhaustive search PTS, the phase factors are restricted to a finite set of values:

$$\phi^{v} \in \{\, (2\pi l/W), \text{where } l = 0,1, \dots , W-1\} \qquad , 1 \le v \le \mathbf{V} \tag{11}$$

Thus W determine the number of allowed phase factors, $\hat{b}^{v}$ preferably chosen from the set $\{\pm 1, \pm j\}$ i.e.$W = 4$. The transmitter tests all possible rotational factors. Therefore, the total number of alternative signal representations will be ($\phi^{0} = 0$ without any performance loss) $I_{PTS_{MAX}} = W^{(\mathbf{V}-1)}$ .This value increases exponentially with the number of subblocks $\mathbf{V}$. Therefore, Optimum PTS (OPTS) may not be feasible for a large number $\mathbf{V}$ [26].

Restricting the search space to a given number of $I_{PTS} \le I_{PTS_{MAX}}$ different, arbitrary chosen combinations (vectors $\mathbf{b}^{i} = [b^{1} \quad \cdots \quad b^{v}]; \boldsymbol{i} = 1, \dots , I_{PTS}$) is also possible. Thereby the complexity of the PAR reduction – given by the number $I = I_{PTS}$ of superpositions (candidates) which have to be evaluated (calculating their PAPR) – can be

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controlled. In addition, independent of the number of examined superpositions, $\boldsymbol{V}$ IFFTs have to be calculated to obtain the partial transmit sequences $\mathbf{x}^{\mathrm{v}}$.
In order to recover the transmitted signal correctly, for coherent reception, the receiver must be aware of the actually used weighting vector $\mathbf{b}^{\hat{i}} = [\hat{b}^1 \ \cdots \ \hat{b}^{\mathrm{V}}]$. Thus, transmission of side information SI is necessary. Assuming a codebook of all I_{PTS} possible combinations code vectors $\mathbf{b}^{i} = [b^1 \ \cdots \ b^{\mathrm{v}}]; \boldsymbol{i} = 1, \ldots, I_{PTS}$ is available jointly to transmitter and receiver, it is sufficient to transmit the index $\hat{\boldsymbol{\imath}}$ of the optimum combination. This index can be represented by [11];

$$\mathrm{SI} = \lceil \log_2(I_{PTS}) \rceil \qquad \text{(bits)} \tag{12}$$

V. HYBRID SLM-PTS TECHNIQUES FOR SINGLE ANTENNA SYSTEMS

A. Conventional Hybrid (CH)

Conventional hybrid (CH) method was the first hybrid technique that combines PTS and SLM, introduced by Pushkarev in [26] it reduce the required computational complexity less than that required by the conventional PTS for the same amount of redundancy .The original OFDM symbol $\boldsymbol{S}$ (the antenna index μ is suppressed in this section) is multiplied with the U phase rotation sequences, and then each of the new OFDM symbols are partitioned into $\boldsymbol{V}$ pairwise disjoints sub-blocks. Those OFDM sub-block values are calculated by each optimization block in each PTS blocks. This can be written as

$$[\hat{b}_u^1 \ \cdots \ \hat{b}_u^V] = \underset{[b_u^1 \ \cdots \ b_u^V]}{\operatorname{argmin}} \left(\frac{\mathrm{Max}_{n \in [0,N]} \left| \sum_{\mathrm{v}=1}^{\mathrm{V}} b_u^{\mathrm{v}} \, x_{(n)}^{(u,\mathrm{v})} \right|^2}{\mathrm{E}\left\{ \left| \sum_{\mathrm{v}=1}^{\mathrm{V}} b_u^{\mathrm{v}} \, x_{(n)}^{(u,\mathrm{v})} \right|^2 \right\}} \right) \tag{13}$$

where $x_{(n)}^{(u,\mathrm{v})}$ is the subcarrier n in v sub-block in the u PTS block, b_u^{v} is a rotational factor that will be applied to the v sub-block in the u PTS block, $1 \le u \le U, 1 \le v \le \boldsymbol{V}$.Each PTS blocks produce an output signal $\hat{\mathbf{x}}^{\boldsymbol{u}}$,

$$\hat{\mathbf{x}}^{\boldsymbol{u}} = \sum_{\mathrm{v}=1}^{\mathrm{V}} \hat{b}_u^{\mathrm{v}} \, \mathbf{x}^{(u,\mathrm{v})} \tag{14}$$

Where $1 \le u \le U$, $\mathbf{x}^{(u,\mathrm{v})}$ is time domain version of the v sub-block in the u PTS block [6-7] . The one with the lowest PAPR among the U signals will be selected by a selection block. This can be written as

$$\hat{u} = \underset{0 \le u \le U-1}{\operatorname{argmin}} \left(\frac{\mathrm{Max}_{n \in [0,N]} \left| \hat{x}_{(n)}^{u} \right|^2}{\mathrm{E}\left\{ \left| \hat{x}_{(n)}^{u} \right|^2 \right\}} \right) \tag{15}$$

Thus, number of alternatives that can be produced by CH I_{CH}, equal to number of PTS Blocks $U_{Total} = U$ multiplied by the given number $I = I_{PTS} \le I_{\mathrm{PTS}_{MAX}}$ of superpositions (candidates) in each PTS Block $I_{CH} = U_{Total} \, I_{PTS} \le UW^{(V-1)}$. Whereas the number of required side information bits can be written as

$$SI = \lceil \log_2(U) + \log_2(I_{\mathrm{PTS}}) \rceil \qquad \text{(bits)} \tag{16}$$

that includes the side information of PTS part and SLM part.

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B. Additional Hybrid (AH)

PAPR reduction is improved in CH scheme, by generating a large number of alternative OFDM signal sequences without increasing the number of IFFTs to avoid high computational complexity [7]. AH scheme introduced by Hong Chou *et al.* In [6] combines CH scheme [26] with modified SLM scheme [27] to produce number of alternative signal representations more than CH using the same number of IFFTs , the excess alternative OFDM signal sequences are generated by the linear combination of the sub-block signals from different PTS blocks after IFFTs operations. Using the linear property of Fourier transform, the linear combination of these sequences can be obtained by:

$$\mathbf{x}^{(u,\mathrm{v})} = \mathrm{C}^{\mathrm{i}}\mathbf{x}^{(\mathrm{i},\mathrm{v})} + \mathrm{C}^{\mathrm{k}}\mathbf{x}^{(\mathrm{k},\mathrm{v})} \tag{17}$$

Where $U+1 \le u \le U^2, 1 \le \mathrm{i,k} \le U \;, 1 \le \mathrm{v} \le \boldsymbol{V}$, and $\mathrm{C}^{\mathrm{i}}, \mathrm{C}^{\mathrm{k}}$ are some coefficients to be chosen later.
To ensure $\mathbf{x}^{(u,\mathrm{v})}$ have an average power equal to the half of the sum of the average power of $\mathbf{x}^{(\mathrm{i},\mathrm{v})}$ and $\mathbf{x}^{(\mathrm{k},\mathrm{v})}$, two conditions must be satisfied $\mathrm{C}^{\mathrm{i}} = \frac{1}{\sqrt{2}}, \mathrm{C}^{\mathrm{k}} = \pm j\frac{1}{\sqrt{2}}$ and $\mathrm{P}^{u}_{(\mathrm{k})} \in \{\pm 1\}$. From U binary phase rotation sequences, we can obtain $2 * \binom{U}{2}$ excessive $\boldsymbol{V}$ sub-blocks sequences with $\binom{U}{2} = U(U-1)/2$ and, thus, there are total $U_{Total} = U^2$ PTS Blocks (each with $\boldsymbol{V}$ subblock) for AH scheme [6].
Assume an$[(U^2 - U) \times \mathbf{2}]$ indices book matrix $\boldsymbol{Set_{AH}}$ contains $(U^2 - U)$ indices sets of all possible linear combinations between the given U PTS Block. For example if we have three PTS blocks ($U = 3$) each with 2 subblocks ($\boldsymbol{V} = 2$), then $\boldsymbol{Set_{AH}} = [1\;\;2\,;1\;\;2\,;1\;\;3;1\;\;3;2\;\;3;2\;\;3]$. Note that, each two consecutive sets have the same indices; first set for addition and the other for subtraction.
The optimization of the first U PTS blocks is the same as CH. Whereas the latest $(\mathrm{U}^2 - \mathrm{U})$ PTS blocks optimized as follows

$$[\hat{\mathrm{b}}_u^1 \;\cdots\; \hat{\mathrm{b}}_u^{\boldsymbol{V}}] = \underset{[\mathrm{b}_u^1 \;\cdots\; \mathrm{b}_u^{\mathrm{V}}]}{\mathrm{argmin}} \left(\frac{\max\limits_{\mathrm{n}\in[0,\mathrm{N}]} \left| \sum_{v=1}^{\mathrm{V}} \left(\mathrm{b}_v^u \left(\frac{1}{\sqrt{2}} \mathrm{x}_{(\mathrm{n})}^{(\boldsymbol{Set_{AH}}(u,1),\mathrm{v})} + (-1)^u \frac{1}{\sqrt{2}} \mathrm{x}_{(\mathrm{n})}^{(\boldsymbol{Set_{AH}}(u,2),\mathrm{v})} \right) \right) \right|^2}{\mathrm{E}\left\{ \left| \sum_{\mathrm{v}=1}^{\mathrm{V}} \left(\mathrm{b}_v^u \left(\frac{1}{\sqrt{2}} \mathrm{x}_{(\mathrm{n})}^{(\boldsymbol{Set_{AH}}(u,1),\mathrm{v})} + (-1)^u \frac{1}{\sqrt{2}} \mathrm{x}_{(\mathrm{n})}^{(\boldsymbol{Set_{AH}}(u,2),\mathrm{v})} \right) \right) \right|^2 \right\}} \right) \tag{18}$$

where $U+1 \le u \le U^2, 1 \le \mathrm{v} \le \boldsymbol{V}$
we have to select and transmit the resulting OFDM signal sequence $\bar{\boldsymbol{x}}$, which has the minimum PAPR among the whole OFDM signal sequences of overall lowest PAPR sequences $\hat{\mathbf{x}}^{\boldsymbol{u}}$ as the following

$$\hat{u} = \underset{0\le \mathrm{u} \le \mathrm{U}^2}{\mathrm{argmin}} \left(\frac{\mathrm{Max}_{\mathrm{n}\in[0,\mathrm{N}]} \left|\hat{\mathrm{x}}_{(\mathrm{n})}^{u}\right|^2}{\mathrm{E}\left\{\left|\hat{\mathrm{x}}_{(\mathrm{n})}^{u}\right|^2\right\}} \right) \tag{19}$$

This analysis can be applied to any number of subblocks $\boldsymbol{V}$, not to $\boldsymbol{V} = 2$ only as in [6].
The number of side information bits , the transmitter have to send to inform the receiver which alternative signal representation from $I_{AH} = U_{Total}\, I_{PTS} \le U^2\, W^{(V-1)}$ have been selected as the best one , can be written as

$$SI = \lceil \log_2(\mathrm{U}^2) + \log_2(I_{PTS}) \rceil \qquad \text{(bits)} \tag{20}$$

C. Switching Hybrid (SH)

Instead of generating alternative OFDM sequences with linear combination as in AH, Switching hybrid (SH) scheme introduced in [6] combines the switching technique with the CH scheme to generate a large number of alternatives OFDM signal sequences without increasing the number of IFFTs. Here we extend analysis to any number of subblocks without restricting $\boldsymbol{V} = 2$ as in [6]. Switching block uses the original U sets of $\boldsymbol{V}$ subblocks $[\mathbf{x}^{(u,1)} \;\; \cdots \;\; \mathbf{x}^{(u,\boldsymbol{V})}]$,

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where $u = 1, \dots, U$ to generate excessive $(U^V - U)$ OFDM sequences without increasing number of IFFTs more than the number required for generating U OFDM sequences. Thus, the total number of alternative signal representations generated by SH is $I_{SH} \leq U^V\, I_{PTS} = U^V\, W^{(V-1)}$.
Assume an $[(U^V - U) \times \boldsymbol{V}]$indices book matrix $\boldsymbol{Set_{SH}}$ contains $(U^V - U)$ indices sets of all possible switching between the given U PTS Block. Excluding sets that have all its indices belong to the same PTS Block. For example if we have two PTS blocks $(U = 2)$ each one of them partioned into three subblocks $(\boldsymbol{V} = 3)$, then $\boldsymbol{Set_{SH}} = [1\ \ 1\ \ 2\,;1\ \ 2\ \ 1\,;1\ \ 2\ \ 2\,;2\ \ 2\ \ 1\,;2\ \ 1\ \ 2\,;2\ \ 1\ \ 1]$, excluding two sets $[1\ \ 1\ \ 1\,;2\ \ 2\ \ 2]$ as they represent the first two PTS Blocks of the CH.
The optimization of the first U PTS Blocks is the same as CH. Whereas the latest $(U^V - U)$ PTS blocks optimized as follows

$$[\hat{b}_u^1 \quad \cdots \quad \hat{b}_u^V] = \underset{[b_u^1 \quad \cdots \quad b_u^V]}{\operatorname{argmin}} \left(\frac{\operatorname{Max}_{n\,\in[0,N]} \left| \sum_{v=1}^{V} b_u^v\, x_{(n)}^{(\boldsymbol{Set_{SH}}(u,v),v)} \right|^2}{E\left\{ \left| \sum_{v=1}^{V} b_u^v\, x_{(n)}^{(\boldsymbol{Set_{SH}}(u,v),v)} \right|^2 \right\}} \right) \tag{21}$$

where $U + 1 \leq u \leq U^V, 1 \leq v \leq \boldsymbol{V}$
After optimization blocks, the OFDM sequence with the lowest PAPR sequence among the outputs of the total number $U_{Total} = U^V$ of PTS Block is selected for transmission. Required number of side information bits of SH will be,

$$SI = \lceil \log_2(U^V) + \log_2(I_{PTS}) \rceil \qquad \text{(bits)} \tag{22}$$

D. Modified Hybrid (MH)

Modified hybrid (MH) algorithm introduced also in [6], generates more and more alternative OFDM sequences by combining AH and SH schemes. This scheme produces number of alternative signal representations equal to $I_{MH} = U_{Total}\, I_{PTS} \leq (U^V + U^2 - U)\, W^{(V-1)}$, where the total number of PTS Block U_{Total} equal to $(U^V + U^2 - U)$. This number of alternatives requires number of side information bits equal to

$$SI = \lceil \log_2(U^V + U^2 - U) + \log_2(I_{PTS}) \rceil \qquad \text{(bits)} \tag{23}$$

The optimization of the first U PTS Blocks is the same as CH, while the optimization of the $(U^2 - U)$ PTS Blocks that belong to AH part follows (18), while the others $(U^V - U)$ PTS Blocks that belong to SH part can be optimized according to (21).

VI. MIMO EXTENSION OF HYBRID SLM-PTS TECHNIQUES

A. Ordinary and Simplified Hybrid SLM-PTS Techniques

It is natural to individually apply any one of the Hybrid SLM-PTS techniques to each transmitting antenna independently, this procedure called *ordinary* CH (*o*CH), *ordinary* AH (*o*AH), *ordinary* SH (*o*SH), or *ordinary* MH (*o*MH) depending on which Hybrid technique is used. In each transmitting antenna the alternative signal representation with the lowest PAPR among I alternatives is independently selected for transmission. Now the number of IFFTs required for reducing the PAPR of the MIMO-OFDM system will be N_T multiple of that required by single antenna OFDM system, i.e. number of IFFTs will be $N_T\boldsymbol{UV}$ that is right for all of the four Hybrid SLM-PTS techniques. Also these four *ordinary* techniques will require side information bits, N_T multiple of (16), (20), (22) and (23) respectively (i.e. $N_T * SI\ bits$ will be required, where $SI\ bits$ will be transmitted from each antenna).
In order to reduce signalling overhead, *simplified* approach that was first introduced in [9] can be applied here to any one of Hybrid SLM-PTS techniques. Of course these approaches will be called *simplified* CH (*s*CH), *simplified* AH (*s*AH), *simplified* SH (*s*SH), or *simplified* MH (*s*MH).
In these *simplified* approaches all the N_T OFDM frames are simultaneously multiplied with the same phase rotation vector $\mathbf{P}^u$, before being divided into sub-blocks and then combined after being multiplied simultaneously with the same set of rotational factors $\mathbf{b}^{\boldsymbol{i}}$. The alternative signal representation with the lowest average PAPR over all N_T transmit

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antennas is selected. Consequently, number of required side information bits, for these four *simplified* approaches will the same as (16), (20), (22) and (23) respectively (SI *bits* will be transmitted from one transmitting antenna, while each receiving antenna will receive the same information bits by means of receiver diversity [9]). However, no complexity reduction is achieved as still $N_T U\boldsymbol{V}$ IFFTs operations have to be calculated.

B. Directed Hybrid SLM-PTS Techniques

Both *ordinary* and *simplified* are just a simple application of single antenna PAPR reduction techniques at all $\mathrm{N_T}$ antennas of the transmitter [11]. Neither *ordinary* nor *simplified* indeed use the potential of MIMO transmission for PAPR reduction, *d*SLM was the first scheme that developed to treat the parallel transmits signals jointly for improving the PAPR reduction performance. Main idea of *d*SLM is to invest complexity only where PAPR reduction is really needed. Instead of performing U trials for each of the $\mathrm{N_T}$ transmitters, the budget of $\mathrm{N_T}U$ IFFTs is used to successively improve the highest PAPR over the antennas. For that, in the first step the PAPR of the $\mathrm{N_T}$ initial (original) OFDM symbols is calculated. Then, in each successive step, the OFDM frame with currently highest PARP is considered and using a next phase rotation vector $\mathbf{P}^{u}$, a reduction of PAPR is tried. This procedure is continued $\mathrm{N_T}(U-1)$ times, leading to the same complexity as *ordinary* or *simplified* SLM. Since, including the initial step, at maximum $\mathrm{N_T}(U-1)+1$ trials may be performed for one antenna, $\mathrm{N_T}\lceil \log_2(\mathrm{N_T}(U-1)+1)\rceil$ bits side information are required [10]. However, *d*SLM needs more side information bits than *o*SLM and *s*SLM, but it gives a better performance than both of them.

In [11] *directed* approach was applied to PTS. The idea of *d*PTS is to increase the number of possible alternative signal representations (by increasing the combinations of the weighting vectors I), but to keep the complexity (i.e., the amount of IFFT computations $\mathrm{N_T}\boldsymbol{V}$ and superpositions I) the same compared to *o*PTS or *s*PTS. As in *d*SLM, not all possible candidates are evaluated for each transmit antenna, but this method always considers that antenna which currently exhibits the highest PAPR and tries to reduce it.

For that, in the first the partial sequences of all antennas are determined. Then, in each successive step, the antenna with the highest PAPR is considered and another signal representation is tested. At the first $\mathrm{N_T}$ superpositions the PAPR of all $\mathrm{N_T}$ transmit antennas is calculated. The remaining budget of $\mathrm{N_T}(I-1)$ superpositions is successively spent on that antenna exhibiting the worst PAPR.

The number of alternative signal representations, which should be evaluated, must be restricted to $I \le ((I_{PTS}-1)/\mathrm{N_T}+1)$. If always one certain antenna exhibits the currently worst PAPR $\mathrm{N_T}(I-1)+1$ candidates are assessed. This number I, of course, has to be smaller than the maximum possible number of candidates I_{PTS} for each antenna. Compared to oPTS/sPTS the average number of superpositions is given by I and the number of side information bits is $\mathrm{N_T}\lceil \log_2(\mathrm{N_T}(I-1)+1)\rceil$ [11].

Hybrid SLM-PTS techniques can exploit the potential of MIMO transmission by using the Principle of *directed* approach mentioned in [10] and [11], to make the maximum exploitation of the available number of PTS Blocks U_{Total}^{MIMO} , that depend on the used Hybrid SLM-PTS technique , as illustrated in (24) below,

$$U_{Total}^{MIMO} = \begin{cases} (N_T U) & CH \\ (N_T U)^V & AH \\ (N_T U)^2 & SH \\ (N_T U)^V + (N_T U)^2 & MH \end{cases} \tag{24}$$

Different thoughts of how to use *directed* approach with Hybrid SLM-PTS techniques can be suggested, one of this suggestions, is to utilize *d*SLM by using the available U_{Total}^{MIMO} PTS Blocks, and invest all possible I superpositions in each PTS Block on the antenna which exhibiting the worst PAPR in each cycle of the available U_{Total}^{MIMO} cycles. Another suggestion is to perform *d*PTS by directing the total $\mathrm{N_T}I$ superpositions to follow the antenna which exhibits the highest PAPR among the $\mathrm{N_T}$ antennas, and repeat this *d*PTS for each one of the available U_{Total} PTS Blocks. But, unfortunately both of the two preceding suggestions perform worse than *ordinary* approach although they have the same complexity. So that they are not presented here and their results omitted for conciseness.

Other suggestion presented in this paper, is to combine *d*SLM and *d*PTS in adequate method, whereas Hybrid SLM-PTS techniques themselves composed of SLM and PTS. Proposed *directed* Hybrid SLM-PTS approaches will be denoted as *directed* CH (*d*CH), *directed* AH (*d*AH), *directed* SH (*d*SH), or *directed* MH (*d*MH) depending on the

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Hybrid SLM-PTS technique that will be used to reduce the PAPR in MIMO-OFDM system (i.e. CH, AH, SH, or MH respectively).

The idea of this approach is to invest complexity only where PAPR reduction is really needed, as in conventional *d*SLM and *d*PTS. Instead of using all the U_{Total} PTS Blocks of each transmitting antenna independently. Also, instead of independently evaluate all the I possible candidates of each U_{Total} PTS Block of each transmitting antenna. The budget of $\mathrm{N_T}\, U_{Total}\, \boldsymbol{V}$ IFFTs is used to successively improve the currently highest PAPR over the antennas. Also the $U_{Total}^{MIMO} I$ superpositions must be successively spent on that antenna exhibiting the worst PAPR.

The proposed approach begins with directing the first PTS Block to the antenna which exhibits the worst PAPR among the $\mathrm{N_T}$ antennas. But the total $\mathrm{N_T}\, I$ superpositions of that PTS Block will not be spent completely on this antenna, if its PAPR gets lower than another antenna, the rest of the $\mathrm{N_T}\, I$ superpositions will be directed to the another antenna which exhibiting the worst PAPR, and so on. These procedures will continue for each one of the U_{Total}^{MIMO} PTS Blocks.

As this approach is mainly concerned with preserving number of IFFTs the same as *ordinary* and *simplified*, if one PTS Block used to reduce the PAPR of two antennas this will reduce the remained number of PTS Blocks by one , and so on. As a result of that, some of the latest $(U_{Total}^{MIMO} - \mathrm{N_T} U)$ PTS Block in *d*AH, *d*SH, and *d*MH which are composed of combinations between any two PTS Blocks from the first $\mathrm{N_T} U$ PTS Blocks, or composed of switchings among the first $\mathrm{N_T} U$ PTS Blocks, may be unusable, as some or all of their underlying PTS subblocks were not generated for that antenna. This will reduce the total number of alternative signal representations; consequently this may degrade the performance of this approach than *ordinary* approach. However, the proposed *directed* Hybrid SLM-PTS approaches may increase number of alternative signal representations than *ordinary* and *simplified* approaches in case of all the $\mathrm{N_T} U$ PTS Blocks are used. In turn, better performance can be achieved at the price of additional complexity. In Comparison to *ordinary* and *simplified* approaches number of side information bits will be,

$$SI = \left\lceil \log_2(U_{Total}^{MIMO}) + \log_2(N_T\, I) \right\rceil \qquad \text{(bits)} \qquad (25)$$

In the next section we will investigate which one of the *directed* SLM-PTS approaches *d*CH, *dAH*, *d*SH, or *d*MH which will have a better performance than *ordinary* approach due to the excess in the number of alternative signal representations, and which will perform worse than *ordinary* approach due to the losses in the number of PTS Blocks.

VII. SIMULATION AND RESULTS

In this section we compare the PAPR reduction performance of the three different approaches *ordinary*, *simplified* and *directed* per each one of the Hybrid SLM-PTS techniques in a MIMO-OFDM system for different numbers of antennas. Simulation parameters are listed in Table 1.

Table 1: Simulation parameters

Simulation Parameters	Specifications	Simulation Parameters	Specifications
Number of OFDM symbols	50000	Number of allowed rotation factors	$W = 4$
Number of subcarriers	64	Number of superpositions	$I = 8$
Number of transmit antenna	$\mathrm{N_T} = 2, 4, 8$	Phase rotation factors , for PTS	$\mathrm{b^V} \in \{\pm 1, \pm j\}$
Number of sub-blocks	$\boldsymbol{V} = 4$	Phase rotation vectors , for SLM	$\mathrm{P}_{(k)}^{u} \in \{\pm 1\}$
Number of phase rotation vectors	$U = 4$	Modulation scheme	QPSK

In Fig. 1(a), we compare the CCDF in case of no PAPR reduction with that of *ordinary*, *simplified*, and *directed* CH. The plot shows the behavior for a different number of transmit antennas ($N_T = 2, 4, 8$). Also as a reference the results for a single antenna system are also given (gray solid for no PAR reduction and gray dotted for CH with $I = 8$). Compared to the situation with no PAR reduction, the three CH approaches are able to reduce the PAPR significantly. Evidently, *simplified* approach *s*CH performs worse than *ordinary* approach *o*CH as less combination of the weighting factors are utilized. However, both reduction schemes perform worse than CH in the single antenna case and for an increasing number of transmit antennas N_T the results get even worse. This reflects the fact that *simplified* and *ordinary* approaches are just a simple application of single antenna PAPR reduction techniques to a Multiantenna transmitter.

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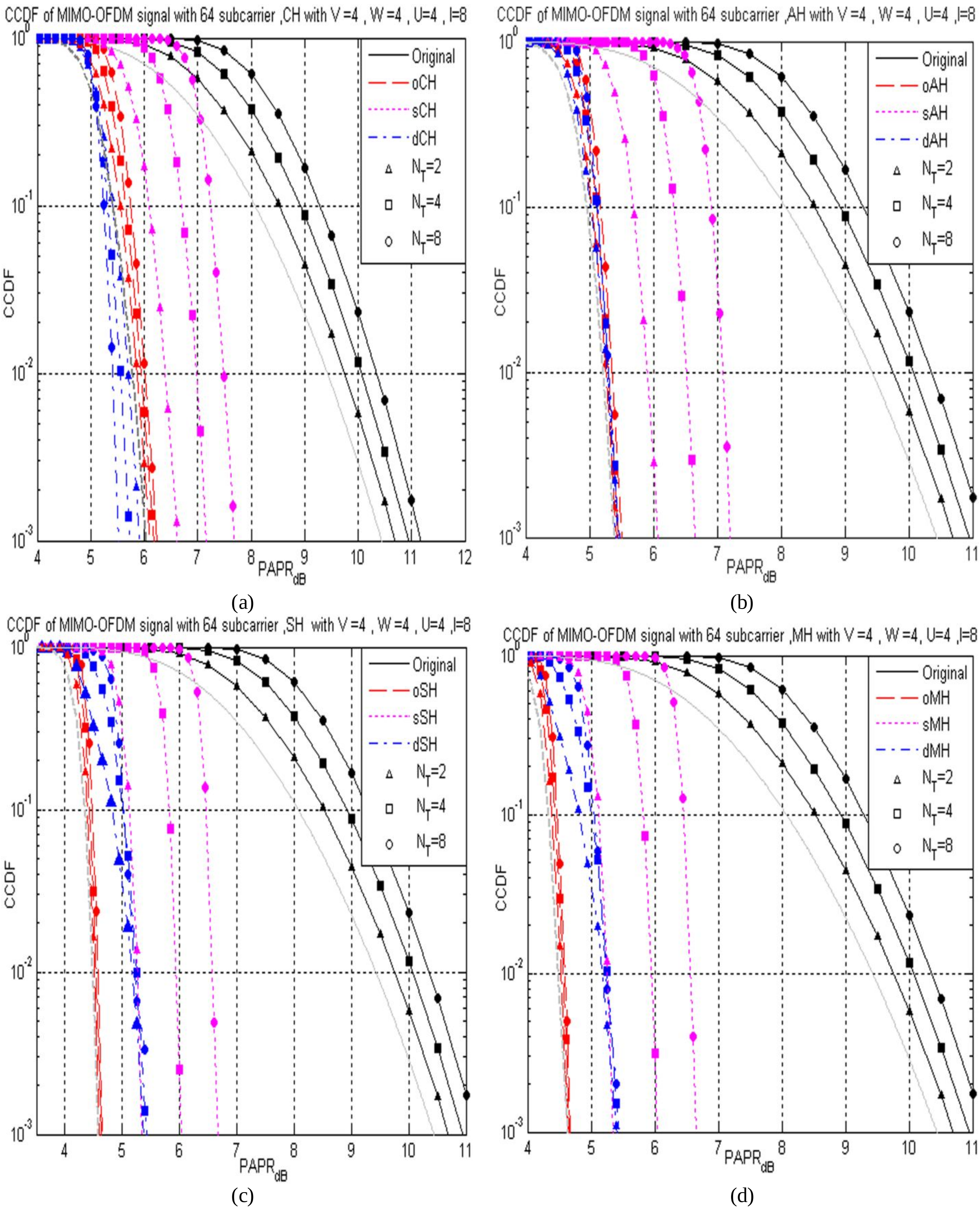

Fig. 1. Comparison of the CCDF of original with that of *ordinary*, *simplified*, and *directed* (a) CH, (b) AH, (c) SH, and (d) MH.MIMO-OFDM system with N_T = (Δ) 2 , (□) 4 , and (O) 8 transmit antennas. As reference the single antenna case is plotted in gray with no PAR reduction (dotted) and CH (solid).

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In contrast to that, the *directed* approach is able to exploit the multiple transmit antennas; *d*CH always outperforms single antenna CH and the performance gets even better for increasing N_T .this is obvious from Fig. 1(a),where *d*CH reduce the PAPR by 0.2 dB , 0.52 dB ,0.70 dB lower than *o*CH in case of 2,4,8 antennas respectively at 0.1% CCDF

In Fig. 1(b), Fig. 1(c), and Fig. 1(d) we do the same thing as in Fig. 1(a), but with AH, SH, and MH respectively. In Fig. 1(b), all of the three AH approaches are able to reduce the PAPR significantly. However, no one of them outperforms single antenna AH. In contrast to *d*CH, PAPR reduction performance of *d*AH gets worse for an increasing number of transmit antennas N_T, due to the losses in the total number of PTS Blocks. In addition to that, *d*AH introduces PAPR reduction performance comparable to that of *o*AH or with non noticeable improvement. However, number of required SI bits will be larger than *o*AH. Thus, in MIMO-OFDM systems that will use AH technique, *o*AH will be the best choice.

In Fig. 1(c), PAPR reduction performance of *d*SH gets worse for an increasing number of transmit antennas N_T, also it is worse than *o*SH approach for any number of antennas N_T, due to the losses in the total number of PTS Blocks. Thus, *o*SH approach will be the best choice for SH technique. The same thing can be noted for MH technique, shown in Fig. 1(d), *o*MH approach is the best choice.

Finally, from previous simulations and analysis, presented in this paper, we can say that, *directed* approach needed to be applied to Hybrid SLM-PTS techniques for exploiting the potential of MIMO transmission as in *d*PTS and *d*SLM. Applying *d*SLM approach only or *d*PTS approach only to Hybrid SLM-PTS techniques; degrade the PAPR reduction performance of Hybrid SLM-PTS techniques worse than *ordinary* approaches if they were used. Combining *d*PTS and *d*SLM in one approach is proposed in this paper. Results showed that, among the four *directed* approaches of the Hybrid SLM-PTS techniques (*d*CH, *d*AH, *d*SH, and *d*MH), only *d*CH approach which perform better than *ordinary* approach *o*CH. Others *directed* Hybrid SLM-PTS approaches (*i.e.* *d*AH, *d*SH, and *d*MH) have a performance worse than or at best equal to *ordinary* approaches (*i.e.* *o*AH, *o*SH, and *o*MH), with larger number of SI bits than *ordinary* approaches, of course.

VIII. CONCLUSION

Hybrid SLM-PTS techniques used to provide the same (or better) PAPR reduction performance as PTS or SLM with less number of IFFTs. *ordinary*, *simplified*, and *directed* are three different approaches for using PTS or SLM in reducing the PAPR of MIMO-OFDM systems. Likewise, this approaches used to apply Hybrid SLM-PTS techniques to MIMO-OFDM systems, in this paper. Appling *ordinary* and *simplified* is a straightforward procedure similar to that in PTS or SLM. However, neither *ordinary* nor *simplified* indeed use the potential of MIMO transmission for PAPR reduction, only *directed* approach which can exploit the potential of MIMO transmission. Take in consideration that, any Hybrid SLM-PTS technique composed of PTS and SLM. *directed* approach combining *d*SLM and *d*PTS in one adequate approach is proposed in this paper. But, unfortunately only *d*CH which performs better than *o*CH, while the others *directed* approaches (i.e. *d*AH, *d*SH, and *d*MH) perform worse than *ordinary* approaches (i.e. *o*AH, *o*SH, and *o*MH).

This is illustrated in Fig. 1. Fig. 1(a) shows that *d*CH is better than *o*CH, *s*CH and also single antenna CH, in addition of enhancing PAPR reduction performance with the increase in the number of antennas. On other hand, Fig. 1(b), show that single antenna AH is better than *o*AH, *s*AH and *d*AH, although *d*AH uses the potential of MIMO, it only better than *s*AH, but has a performance similar to *o*AH. While *ordinary* approach shows a PAPR reduction performance better than *simplified* and *directed* approaches in SH and MH techniques, this is shown in Fig. 1(c) and Fig. 1(d) respectively whose also show that *directed* approach is unfortunately worse than *ordinary* approach and single antenna in SH and MH techniques.

To sum up, proposed *directed* approach is suggested to be used only with CH, while *ordinary* approach is the best choice for the others Hybrid SLM-PTS techniques. However, if we are concerned with reducing the required number of SI bits, *Simplified* approach will be the best choice for all the Hybrid SLM-PTS techniques.

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