PAPR REDUCTION IN BLIND MIMO OFDM SYSTEMS BASED ON INDEPENDENT COMPONENT ANALYSIS

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ABSTRACT

We propose three peak-to-average power ratio (PAPR) reduction schemes for multiple-input multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) systems, where the phase shifts and permutation introduced by PAPR reduction are completely resolved by an independent component analysis (ICA) based and precoding aided blind receiver. Therefore, no transmission of side information is needed as required by conventional PAPR reduction schemes, avoiding any spectral overhead. Simulation results show that the proposed schemes reduce the PAPR of the transmit signals considerably.

1. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) and multiple-input multiple-output (MIMO) are two promising candidates for future wireless communication standards. Compared to training based channel estimation and equalization approaches, blind approaches improve the spectral efficiency without extra overhead for training. Independent component analysis (ICA) is an efficient HOS based blind source separation technique, and has been applied to MIMO OFDM systems for blind equalization [1, 2]. However, as a common drawback of blind approaches, ambiguity still exist for the above ICA based approaches. In [3], ambiguity was eliminated by exploiting the correlation introduced by precoding.

One of the main drawbacks of OFDM systems is the high peak-to-average power ratio (PAPR) of OFDM signals. Recently, a variety of PAPR reduction methods have been well studied [4, 5, 6, 7, 8, 9]. Selective mapping (SLM) and partial transmit sequence (PTS) [5] are two flexible probabilistic techniques without signal distortion. MIMO OFDM provides extra spatial freedom which can be exploited to obtain further PAPR reduction. In [6], the SLM and PTS were investigated for MIMO OFDM systems. In [7], Fischer *et al.* proposed a directed SLM method for MIMO OFDM systems by jointly optimizing over all antennas to reduce PAPR of OFDM signal. A scheme of cross-antenna rotation and inversion (CARI) [8] was proposed over space-time block coded (STBC) MIMO OFDM systems. However, the above schemes require some side information to be transmitted to the receiver with high reliability, which reduces the spectral efficiency.

In this paper, we propose three PAPR reduction schemes for MIMO OFDM systems with the ICA based and precoding aided blind receiver. The PAPR of OFDM signals can be reduced by judicious design of precoding, followed by two other proposed PAPR reduction schemes, which are referred to as extended selective mapping (ESLM) and cross-antenna swapping (CAS), respectively. Compared to conventional PAPR reduction schemes [6, 8], our proposed PAPR reduction schemes improve the bandwidth efficiency because no transmission of side information is required. Simulation results show that the proposed blind structure can achieve considerable PAPR reduction without spectral overhead. The impact of the subblock length are also investigated.

The paper is organized as follows. The system model is given in section 2. We briefly review the ICA based blind receiver for MIMO OFDM systems in section 3. In section 4, we present three proposed PAPR reduction schemes. Section 5 gives the simulation results to demonstrate the performance of the proposed approaches. Section 6 is the conclusion.

2. SYSTEM MODEL

Fig. 1 depicts a MIMO OFDM spatial multiplexing system with K subcarriers, N_t transmit antennas and N_r receive antennas. For the purpose of PAPR reduction, each symbol block is partitioned into $M = N_s/P$ subblocks, with N_s and P denoting the block length and the subblock length, respectively. Let $s_n(k,m,i)$ denote the signal on subcarrier k in the i-th symbol of the m-th block transmitted by the n-th antenna. At the transmitter, a non-redundant linear precoding [3] transforms the source data vector $\mathbf{d}(k,m,i) = [d_0(k,m,i),\ldots,d_{N_t-1}(k,m,i)]^T$ of length N_t according to

$$\mathbf{s}(k,m,i) = \frac{1}{\sqrt{1+a^2}}[\mathbf{d}(k,m,i) + a\mathbf{d}_{\text{ref}}(k,i)] \quad (1)$$

where $\mathbf{s}(k,m,i) = [s_0(k,m,i),\ldots,s_{N_t-1}(k,m,i)]^T$, and the real number a ($0 \le a \le 1$) is the predefined precoding constant. The PAPR reduction process can be described as $\mathbf{s}'(k,m,i) = \mathbf{Q}(k,m)\mathbf{s}(k,m,i)$, where $\mathbf{Q}(k,m)$ is the subblock based PAPR reduction matrix of size $N_t \times N_t$. Due to the use of cyclic prefix (CP), inter-symbol interference (ISI) is avoided when $L_{cp} \ge (L_c - 1)$, where L_{CP} and L_c denote the length of CP and the maximum channel memory, respectively. The channel is assumed to be quasi-static block fading and remains constant for the duration of an OFDM symbol block. Therefore, the received signal vector $\mathbf{x}(k,m,i)$ of length N_r on subcarrier k within OFDM symbol i of the m-th block can be written as

$$\mathbf{x}(k,m,i) = \mathbf{H}(k)\mathbf{Q}(k,m)\mathbf{s}(k,m,i) + \mathbf{z}(k,m,i) \quad (2)$$

where $\mathbf{H}(k)$ is a channel frequency response matrix of size $N_r \times N_t$ on subcarrier k. $\mathbf{z}(k,i)$ is a additive white Gaus-

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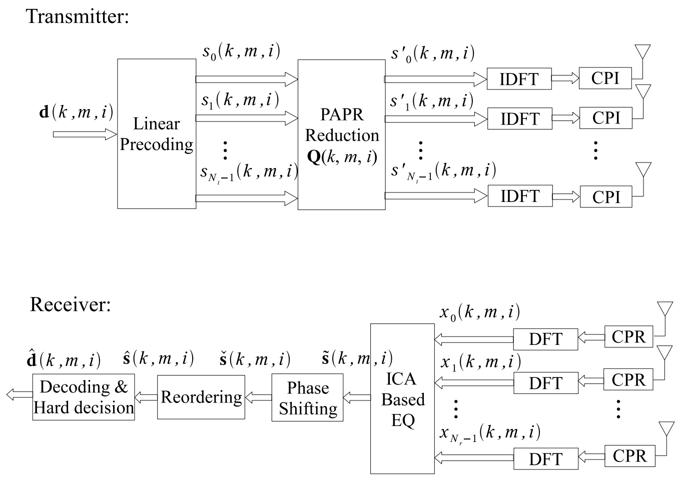

Figure 1: System model of the MIMO OFDM system with ICA based blind receivers

sian noise (AWGN) vector whose entries have zero mean and variance σ_z^2.

3. ICA BASED RECEIVER

ICA [2, 3] is applied to equalize the received signals on each subcarrier k. However, as a common drawback of blind approaches, ambiguity exists in the equalized signal $\tilde{\mathbf{s}}(k,m,i)$

$$\tilde{\mathbf{s}}(k,m,i) = \mathbf{P}_{\text{ICA}}(k,m)\mathbf{G}_{\text{ICA}}(k)\mathbf{Q}(k,m)\mathbf{s}(k,m,i) \quad (3)$$

where the permutation matrix $\mathbf{P}_{\text{ICA}}(k,m)$ accounts for the permutation ambiguity and the diagonal matrix $\mathbf{G}_{\text{ICA}}(k)$ accounts for the phase ambiguity.

The phase ambiguity in the ICA equalized signal $\tilde{\mathbf{s}}(k,m,i)$ can be resolved by the de-rotation operation for each data stream n as $\check{s}_n(k,m,i) = \tilde{s}_n(k,m,i)\left[\alpha_n(k)/|\alpha_n(k)|\right]$, where the factor $\alpha_n(k)$ is $\alpha_n(k) = \left\{\frac{1}{MP}\sum_{m=0}^{M-1}\sum_{i=0}^{P-1}\left[\tilde{s}_n(k,m,i)\right]^4\right\}^{-\frac{1}{4}} e^{j\frac{\pi}{4}}$ for QPSK modulation [2]. Although the phase shifting resolves the phase ambiguity in the equalized signal $\tilde{s}_n(k,m,i)$, it introduces a phase rotation of $b_{\text{ICA},n}(k) = \frac{\pi}{2}l$ ($l \in \{0,1,2,3\}$) in $\check{s}_n(k,m,i)$. The output signal vector after phase shifting under noiseless assumption can be written as

$$\check{\mathbf{s}}(k,m,i) = \mathbf{P}_{\text{ICA}}(k,m)\mathbf{B}_{\text{ICA}}(k,m)\mathbf{Q}(k,m)\mathbf{s}(k,m,i). \quad (4)$$

The PAPR reduction function $\mathbf{Q}(k,m)$ may introduce phase shifts or permutation ambiguity, which can be resolved by reordering, along with the ICA related quadrant and permutation ambiguity given by $\mathbf{B}_{\text{ICA}}(k,m)$ and $\mathbf{P}_{\text{ICA}}(k,m)$, respectively.

Due to the precoding, the transmitted symbols are correlated with the reference symbols. Define $\rho_{\pi_n(k,m),n}$ to be the cross correlation between the detected stream $\pi_n(k,m)$ and the reference signal stream n on subcarrier k in subblock m,

$$\rho_{\pi_n(k,m),n} = \frac{1}{P}\sum_{i=0}^{P-1}\check{s}_{\pi_n(k,m)}(k,m,i)\left[d_{\text{ref},n}(k,i)\right]^*. \quad (5)$$

Based on [3], the permutation $\pi(k,m)$ corresponding to the largest $|\rho_{\pi_n(k,m),n}|$ has the highest probability of being the correct permutation, *i.e.*,

$$\pi^{\text{cor}}(k,m) = \underset{\pi(k,m)=\left[\pi_0(k,m),\ldots,\pi_{N_t-1}(k,m)\right]}{\arg\max}\sum_{n=0}^{N_t-1}|\rho_{\pi_n(k,m),n}|. \quad (6)$$

The estimated data $\hat{\mathbf{s}}(k,m,i)$ after reordering is obtained as

$$\hat{\mathbf{s}}(k,m,i) = \left[\mathbf{B}(k,m)\right]^{-1}\cdot \left[\check{s}_{\pi_0^{\text{cor}}(k,m)}(k,m,i),\ldots,\check{s}_{\pi_{N_t-1}^{\text{cor}}(k,m)}(k,m,i)\right]^T \quad (7)$$

where the entry $b_n(k,m)$ of the diagonal matrix $\mathbf{B}(k,m)$ is given by

$$b_n(k,m) = \left[je^{-j\frac{\pi}{4}}\,\text{sign}\left(\frac{\rho_{\pi_n(k,m),n}}{|\rho_{\pi_n(k,m),n}|}e^{j\frac{\pi}{4}}\right)\right]^{-1} \quad (8)$$

for QPSK modulation, with sign$(\cdot)$ denoting the sign function. Next,we decode and make hard decision $\hat{\mathbf{d}}(k,m,i)$ for the source data streams $\mathbf{d}(k,m,i)$.

4. PAPR REDUCTION

The PAPR reduction of OFDM signals is achieved by the design of reference data for precoding, followed by two other proposed block based PAPR reduction schemes, which are referred to as ESLM and CAS. ESLM modifies the phases of a subblock of OFDM symbols simultaneously and employs the minimum maximum criterion [8] to make the selection. CAS utilizes the spatial freedom of MIMO systems, where data blocks on an arbitrary subcarrier for different transmit antennas are swapped to lower the maximum PAPR. Note that the phase shifts and permutation of the transmit data streams introduced by PAPR reduction can be resolved blindly by reordering as described in Section 3, along with the quadrant and permutation ambiguity due to ICA equalization. Therefore, no transmission of side information is needed by the proposed PAPR reduction schemes, which saves precious bandwidth over conventional PAPR reduction schemes [4, 5, 6]. Moreover, our proposed PAPR reduction schemes do not introduce any signal distortion, and therefore have no effect on the BER performance.

4.1 PAPR Reduction by Precoding

In the paper, an underlined symbol is used to denote the IDFT output signal. According to (1), the transmitted signals after IDFT are the superimposition of the source signals and the reference signals, which are $\underline{s}_n(t,m,i) = \frac{1}{\sqrt{1+a^2}}[\underline{d}_n(t,m,i) + a\underline{d}_{\text{ref},n}(t,i)]$. The precoding does not increase the average signal power, *i.e.*, $\underset{t}{\text{E}}\left\{|\underline{s}_n(t,m,i)|^2\right\} = \underset{t}{\text{E}}\left\{|\underline{d}_n(t,m,i)|^2\right\} = 1$. Therefore, we consider only the peak power of signals for PAPR reduction via precoding. Let $|\underline{d}_{n,\max}(t,m,i)|^2 = \arg\max_{0\leq t<T}|\underline{d}_n(t,m,i)|^2$ denote the peak power of the source signal without PAPR reduction. The difference between the power of the precoded signal and the peak power of the

source signal is given by

$$\begin{aligned}\beta_n(t,m,i) &= |\underline{s}_n(t,m,i)|^2 - |\underline{d}_{n,\max}(t,m,i)|^2 \\ &= \frac{1}{1+a^2}\left[|\underline{d}_n(t,m,i)|^2 + a^2|\underline{d}_{\mathrm{ref},n}(t,i)|^2\right. \\ &\quad +2a\mathrm{Re}\{\underline{d}_n(t,m,i)\underline{d}^*_{\mathrm{ref},n}(t,i)\} \\ &\quad \left.-(1+a^2)|\underline{d}_{n,\max}(t,m,i)|^2\right]. \end{aligned} \quad (9)$$

Assuming $|\underline{d}_{\mathrm{ref},n}(t,i)|^2 \leq \xi$ ($\xi \geq 1$) and using $|\underline{d}_n(t,m,i)|^2 \leq |\underline{d}_{n,\max}(t,m,i)|^2$, (9) becomes

$$\begin{aligned}\beta_n(t,m,i) \leq\ & \frac{1}{1+1/a^2}\left[\xi - |\underline{d}_{n,\max}(t,m,i)|^2\right] \\ &+\frac{2a\mathrm{Re}\{\underline{d}_n(t,m,i)\underline{d}^*_{\mathrm{ref},n}(t,i)\}}{1+a^2}. \end{aligned} \quad (10)$$

Since the mean of $\underline{d}_{\mathrm{ref},n}(t,i)$ is $\mathop{\mathrm{E}}_t\{\underline{d}_{\mathrm{ref},n}(t,i)\} = 0$, the mean of $\beta_n(t,m,i)$ satisfies

$$\mu_\beta \leq \frac{1}{1+1/a^2}\left[\xi - |\underline{d}_{n,\max}(t,m,i)|^2\right]. \quad (11)$$

By minimizing ξ so that $\xi \ll |\underline{d}_{n,\max}(t,m,i)|^2$, we can guarantee $\mu_\beta \ll 0$. With careful design of the reference data, the probability that the proposed precoding results in a reduced PAPR can approach 1. It can also be deduced from (11) that the larger the precoding constant a, the smaller the value of μ_β, *i.e.*, the larger the impact of precoding on PAPR reduction. Therefore, a design criterion of the reference data to minimize the PAPR is given by:

$$\min \arg \max_t \left\{|\underline{d}_{\mathrm{ref},n}(t,i)|^2\right\} \quad (12)$$

From [3], the reference data is $d_{\mathrm{ref},n}(k,i) = c_k M_P(n,i)$, with $M_P(n,i)$ denoting the entry at row n and column i in Hadamard matrix $\mathbf{M}_P$. Since $M_P(n,i)$ is independent of subcarriers, we have

$$\underline{d}_{\mathrm{ref},n}(t,i) = M_P(n,i)\underline{c}(t) \quad (13)$$

where $\underline{c}(t)$ is the continuous signal corresponding to $\underline{\mathbf{c}} = [\underline{c}_0,\ldots,\underline{c}_{K-1}]^T$, which is the IDFT of vector $\mathbf{c} = [c_0,\ldots,c_{K-1}]^T$. In order to design $\mathbf{c}$, we are inspired by the pseudonoise number (PN) sequence. Assuming $\underline{\mathbf{c}}_{\mathrm{PN}} = [\underline{c}_{\mathrm{PN},0},\underline{c}_{\mathrm{PN},1},\ldots,\underline{c}_{\mathrm{PN},\nu-1}]^T$ is the IDFT of a PN sequence $\mathbf{c}_{\mathrm{PN}}$ of length ν, the power of $\underline{\mathbf{c}}_{\mathrm{PN}}$ can be obtained as

$$|\underline{c}_{\mathrm{PN},i}|^2 = \begin{cases} \frac{1}{\nu} & i=0 \\ 1+\frac{1}{\nu} & i=1,\ldots,\nu-1 \end{cases}. \quad (14)$$

Obviously, the PAPR of $\underline{\mathbf{c}}_{\mathrm{PN}}$ is close to 0 dB when ν is large. A heuristic design method for vector $\mathbf{c}$ is as follows:

1. Determine the length ν of the PN sequence, which is $\nu = 2^{\lfloor \log_2 K \rfloor} - 1$
2. Generate a PN sequence $\mathbf{c}_{\mathrm{PN}} = [c_{\mathrm{PN},0}, c_{\mathrm{PN},1},\ldots,c_{\mathrm{PN},\nu-1}]^T$
3. Pad a length $(K-\nu)$ random sequence $\mathbf{c}_{\mathrm{pad}} = [c_{\mathrm{pad},0},\ldots,c_{\mathrm{pad},K-\nu-1}]^T$ with $c_{\mathrm{pad},k} \in \{-1,1\}$ ($k = 0,\ldots,K-\nu-1$) to $\mathbf{c}_{\mathrm{PN}}$, resulting in $\mathbf{c} = [\mathbf{c}^T_{\mathrm{PN}}, \mathbf{c}^T_{\mathrm{pad}}]^T$. This step can be repeated with different random sequences $\mathbf{c}_{\mathrm{pad}}$. The vector $\mathbf{c}$ is selected, whose IDFT, denoted by $\underline{\mathbf{c}}$, has the lowest peak power.

Even though the padding process in step 3 slightly increases the peak power of vector $\underline{\mathbf{c}}$, which comprise K samples of $\underline{c}(t)$ during an OFDM symbol period, the increase is insignificant as $(K-\nu)$ is small. Using (13), we obtain $|\underline{d}_{\mathrm{ref},n}(t,i)|^2 \approx 1$ ($0 \leq t < T$) with a large ν, *i.e.*, the PAPR of the reference signal $\underline{d}_{\mathrm{ref},n}(t,i)$ is close to 0 dB. Thus, the threshold for the peak power of $\underline{d}_{\mathrm{ref},n}(t,i)$ is $\xi \approx 1$, which results in a reduced PAPR of the precoded signal according to (11).

In this case, the PAPR reduction matrix $\mathbf{Q}(k,m,i)$ in (2) is an identity matrix, *i.e.*,

$$\mathbf{Q}_{\mathrm{precoding}}(k,m) = \mathbf{I}_{N_t}. \quad (15)$$

4.2 PAPR Reduction by Extended Selective Mapping

In this subsection, we extend the SLM scheme in [6] to further reduce the PAPR of the precoded OFDM signals. The drawback of SLM in [6] is that critical side information must be transmitted for each OFDM symbol to inform the receiver of the selected candidate index, which reduces the spectral efficiency. With our proposed ESLM, this spectral overhead can be avoided by reordering as described in Section 3.

Instead of adjusting the phase of each OFDM symbol separately as in [6], we perform the phase adjustment based on subblocks, as shown in Fig. 2. For each subblock m ($m = 0,\ldots,M-1$), the PAPR reduction matrix $\mathbf{Q}(k,m,i)$ in (2) is a diagonal matrix, given by

$$\mathbf{Q}_{\mathrm{ESLM}}(k,m) = \mathrm{diag}\{[q_0(k,m), q_1(k,m),\ldots,q_{N_t-1}(k,m)]^T\}.$$

It was shown in [6] that the phase factor set $\{\pm 1, \pm j\}$, which comprises the possible values of the phase adjustment factor $q_n(k,m)$ ($n = 0,\ldots,N_t-1$), can obtain the near optimal performance. In this case, the diagonal matrix $\mathbf{B}(k,m)$ in (7) can be expressed as $\mathbf{B}(k,m) = \mathbf{B}_{\mathrm{ICA}}(k)\mathbf{Q}_{\mathrm{ESLM}}(k,m)$, which accounts for the phase shifts due to both ICA equalization and the ESLM PAPR reduction. The combined phase shifts can be eliminated by reordering as described in Section 3, which implies that no transmission of side information is needed.

Since we adjust P OFDM symbols in a subblock simultaneously, there are P resulting PAPR values for each subblock. We employ the minimum maximum criterion in [8] for selection, which is to select the candidate with the smallest maximum PAPR value. The ESLM algorithm can be described by the following steps for each subblock m:

1. Randomly generate an adjustment sequence $\mathbf{q}_n(m) = [q_n(0,m),\ldots,q_n(K-1,m)]^T$ of length K for each stream n ($n = 0,\ldots,N_t-1$)
2. Adjust the phases using (16), and calculate the P resulting PAPR values
3. Find the minimum PAPR value
4. Repeat step 1 to step 3 until the maximum number of iterations U is reached
5. Find the optimal candidate based on the minimum maximum criterion

To improve the PAPR reduction performance, a shorter subblock length P is desirable so that phases can be adjusted with more flexibility. However, a too small valued P will cause a significant degradation of BER performance due to the residual ambiguity [3]. Therefore, the subblock length P should be chosen carefully to reduce the PAPR significantly while maintaining a near-optimal BER performance.

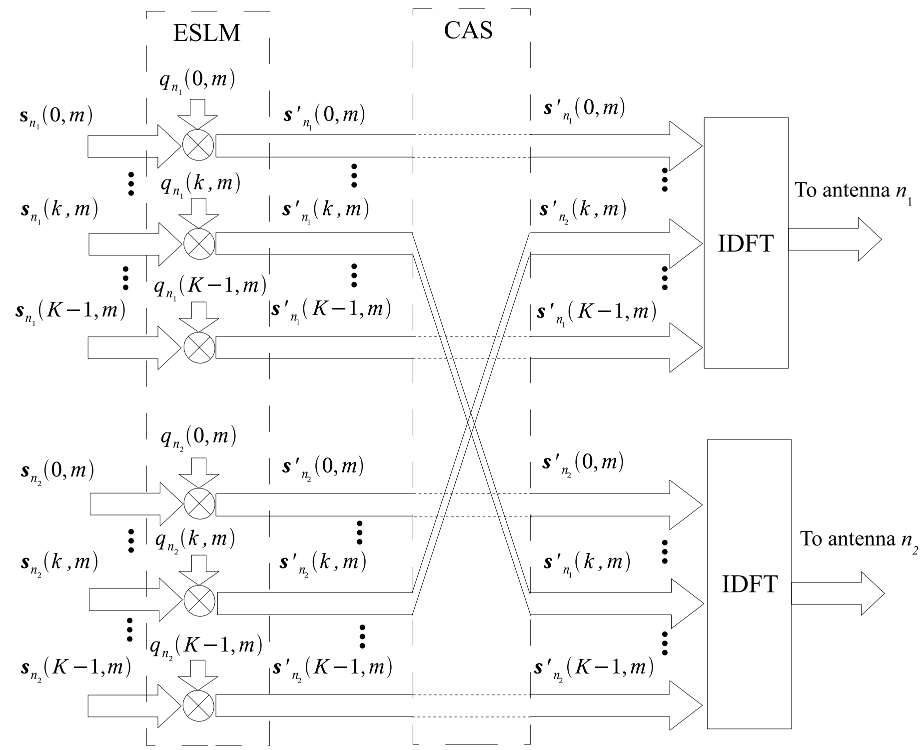

Figure 2: Combined extended selective mapping (ESLM) and cross antenna swapping (CAS) on subblock m. ESLM modifies the phase of each stream in parallel, while CAS swaps data stream n_1 and data stream n_2 on subcarrier k

4.3 PAPR Reduction by Cross-Antenna Swapping

The spatial freedom of MIMO OFDM systems can be exploited to reduce the PAPR [8]. We propose a CAS scheme, which swaps the modulated signal subblocks for different antennas on an arbitrary subcarrier. In this case, the PAPR reduction matrix $\mathbf{Q}_{\text{CAS}}(k,m)$ represents the permutation ambiguity due to the cross-antenna rotation, which can be eliminated by reordering, along with the ICA related permutation ambiguity expressed by $\mathbf{P}_{\text{ICA}}(k,m)$ in (4).

Finding the optimal swapping sequence is a hard combinatorial optimization problem, where there are $((N_t-1)!)^K$ swapping combinations for an OFDM system with N_t transmit antennas and K subcarriers. Therefore, a practical suboptimal stop-and-go algorithm is motivated. The proposed CAS algorithm, as shown in Fig. 2, is as follows.

1. Calculate P PAPR values, and stream $\mathbf{s}_{n_1}(k,m)$ ($n_1 = 0,\ldots,N_t-1$) with the highest PAPR value is selected
2. Swap stream $\mathbf{s}_{n_1}(k,m)$ with a randomly selected stream $\mathbf{s}_{n_2}(k,m)$ ($n_2 = 0,\ldots,N_t-1$ and $n_2 \neq n_1$) on an arbitrary subcarrier k
3. Calculate P the PAPR values and get the highest PAPR value. If it has been reduced after the swapping operation, save the result and go back to step 1. Otherwise, retrieve the result before swapping and then go back to step 1. Cease the algorithm if the highest PAPR value has not been reduced after U_{th} steps, where U_{th} denotes the threshold of the number of iterations.

Moreover, we can combine ESLM and CAS schemes to further reduce the PAPR of OFDM signals. After phase adjustment of the transmitted signals by ESLM, CAS between different antennas can be employed, as illustrated in Fig. 2.

5. SIMULATION RESULTS

We use simulation results to demonstrate the performance of the proposed PAPR reduction schemes for MIMO OFDM systems with $K = 64$ subcarriers, $N_t = 4$ transmit antennas and $N_r = 4$ receive antennas. The data rate is 16 Mbps with QPSK modulation. A CP of length $L_{\text{cp}} = 15$ is used. The Clarke's block fading channel model is employed, which remains constant during a block of $N_s = 256$ OFDM symbols and follows the exponential power delay profile with a root mean square (RMS) delay spread of $T_{\text{RMS}} = 1.4$ normalized to the sampling time interval. Given a subblock length P, the value of the precoding constant a is chosen to provide a good BER performance over a wide range of SNRs [3]. An oversampling factor of $L = 4$ is used to get the accurate PAPR value. The number of iterations for ESLM is $U = 16$, and the threshold for the number of iterations in CAS is $U_{\text{th}} = 6$, so that ESLM and CAS have a similar complexity for fair comparison.

Fig. 3 shows the performance of the proposed PAPR reduction schemes with a subblock length of $P = 64$, in terms of the CCDF of PAPR, which is the probability that the PAPR exceeds a certain value γ, denoted by $\Pr(\text{PAPR} > \gamma)$. At $\text{CCDF} = 10^{-4}$, ESLM provides nearly the best and same performance as the combined ESLM and CAS scheme, with a gain of around 2 dB over the case without PAPR reduction. ESLM also outperforms CAS by 1 dB at $\text{CCDF} = 10^{-4}$ with a similar complexity. The additional design criterion in (12) for the reference data yields a PAPR reduction of around 0.3 dB at $\text{CCDF} = 10^{-4}$ without extra complexity required, as the reference data are generated offline, following the procedure described in Subsection 4.1.

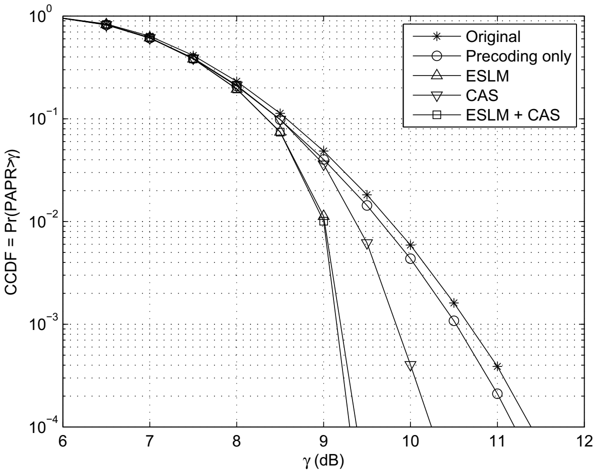

Figure 3: PAPR CCDFs for different schemes with subblock length $P = 16$

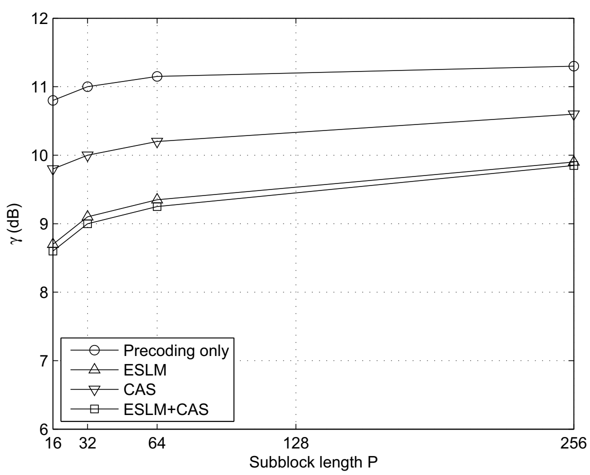

Figure 4: Impact of subblock length P on PAPR reduction with CCDF $= 10^{-4}$

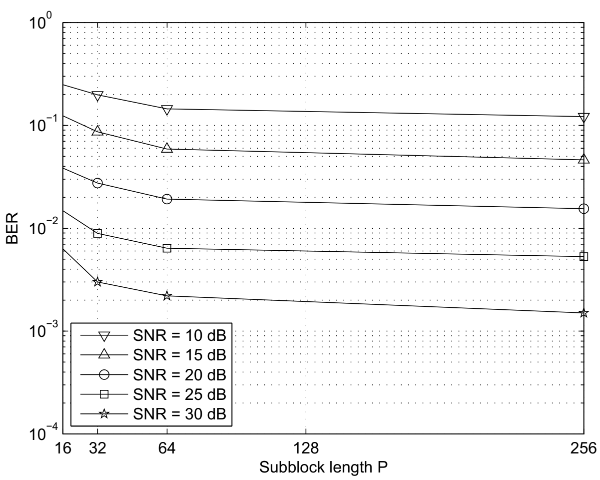

Figure 5: Impact of subblock length P on BER performance

The impact of the subblock length P on the performance of PAPR reduction is shown in Fig. 4 at CCDF $= 10^{-4}$. The effect of block length P on BER performance is shown in Fig. 5. From these two figures, we can observe that over the wide range of $P = 32 - 256$, the BER performance is close to the optimal case while PAPR performance improves with the decrease of P. Therefore, a medium to large subblock length P is preferable, which results in significant PAPR reduction while maintaining a near-optimal BER performance.

6. CONCLUSIONS

We have proposed three PAPR reduction schemes for ICA based blind MIMO OFDM systems. The proposed PAPR reduction schemes outperform significantly the case without precoding, at no cost of spectral overhead. In particular, ESLM demonstrates the best performance on PAPR reduction. A medium to large subblock length is preferable, which results in significant PAPR reduction while maintaining a near-optimal BER performance.

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