Stream-wise Blind Selected Mapping for Single Carrier MU-MIMO

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1 Introduction

Multi-user multiple-input multiple-output (MU-MIMO) is a technique to improve spectrum efficiency (SE) by allowing many users to transmit multiple streams simultaneously [1]. However, the transmit filtering for interference mitigation and equalization increases the peak-to-average power ratio (PAPR) of transmit signal even if single carrier (SC) waveform is used [2], consequently degrades energy efficiency (EE) of battery-powered user equipment (UE).

In this paper, we firstly introduce a principle of blind selected mapping (blind SLM) [3]. Blind SLM applies a phase rotation sequence, which minimizes the PAPR, to the original transmit signal and the phase rotation estimation at the receiver to recover the original transmit signal without side information. Then, an application of blind SLM to MU-MIMO is introduced. Phase rotation sequence is applied to the data streams prior to transmit filtering (called stream-wise blind SLM) for keeping the phase rotation sequence estimation simple.

Here, a singular value decomposition (SVD) based eigenmode transmit filtering and minimum mean-square error (MMSE) based receive filtering are considered as MU-MIMO [4]. Adaptive rank/modulation control (ARMC) is also adopt. Computer simulation assuming SC uplink (UE to base station (BS)) confirms that the blind SLM can lower the PAPR of MU-MIMO without degrading the BER.

2 Principle of Blind SLM

Here, we assume single antenna (SISO) transmission for simple explanation. PAPR of a an oversampled block $\{s(n)\}$ is denoted by PAPR($\{s(n)\}$)

2.1 SLM algorithm

An N_c-length symbols block $\{d(n); n = 0 \sim N_c - 1\}$ is firstly multiplied by a selected phase rotation sequence $\{\Phi_{\hat{m}}(n)\}$, yielding the rotated block $\{d_{\hat{m}}(n)\}$. $\{d_{\hat{m}}(n)\}$ is then transformed to frequency components $\{D_{\hat{m}}(k), k = 0 \sim N_c - 1\}$ by discrete Fourier transform (DFT). After that, transmit filtering $\{W_T(k)\}$ is applied, obtaining frequency-domain signal $\{S_{\hat{m}}(k)\}$. Finally the waveform $\{s(n)\}$ is obtained by applying IDFT to $\{S_{\hat{m}}(k)\}$. $\{\Phi_{\hat{m}}(n)\}$ is selected by

$$\hat{m} = \arg \min_{m=0\sim M-1} (\mathrm{PAPR}(\{s_m(n)\})), \tag{1}$$

where $\{s_m(n)\}$ is a corresponding transmit waveform generated from the phase-rotated block $\{d_m(n) = \Phi_m(n)d(n)\}$. $\{\Phi_m(n); m = 0 \sim M - 1\}$ is the m-th phase rotation sequence in a codebook and is generated randomly as $\Phi_m(n) \in \{1, e^{j2\pi/3}, e^{j4\pi/3}\}$ [3].

2.2 Phase rotation sequence estimation

Phase rotation sequence estimation is based on Euclidean distance calculation between the de-mapped symbols and original constellations. The sequence associated with the de-mapped symbols having the minimum Euclidean distance is selected. Assuming the time domain block before de-mapping is $\{\hat{d}(n); n = 0 \sim N_c - 1\}$, the phase sequence estimation is

$$\tilde{m} = \arg \min_{\substack{m=0\sim M-1 \\ \mathbb{C}\in\Psi_{\mathrm{mod}}}} \left(\epsilon = \sum_{n=0}^{N_c-1} \left| \Phi_m^*(n)\hat{d}(n) - \mathbb{C} \right|^2 \right), \tag{2}$$

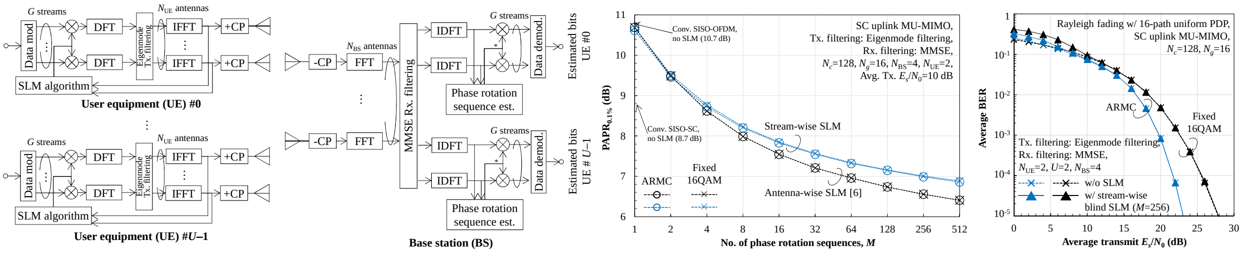
where Ψ_{mod} is the original constellation. (2) can be done based on either maximum-likelihood (ML) or Viterbi algorithm, where the latter one requires less computational complexity. Meanwhile, the use of Viterbi algorithm only remains errors due to frequency-selective fading. A verification and correction based on minimum Hamming distance between the estimated sequence and sequences in the codebook is employed afterward (called 2-step phase rotation estimation [3]).

3 Stream-wise blind SLM for MU-MIMO

A conventional SLM for MIMO was studied in [5], which employing phase rotation individually at each transmit antenna (called antenna-wise SLM). However, The SLM techniques in [3] and [5] cannot be used straightforwardly in MU-MIMO with transmit/receive filtering since it is difficult to realize blind phase sequence estimation (i.e., it needs to consider all possible filter coefficients). To realize low-PAPR MU-MIMO with a simple phase estimation, a stream-wise blind SLM is introduced. Here, we consider an MU-MIMO consisting of U UEs, each equipped with N_{UE} antennas and sending $G(u) \leq N_{\mathrm{UE}}$ streams, and a single BS equipped with N_{BS} antennas (shown in Fig. 1(a)).

3.1 Transmitter (UE)

At the u-th UE, information sequence is modulated into $G(u)$ streams of N_c-length block $\{\mathbf{d}_u(n)\}$ with $\mathbf{d}_u(n) = [d_{u,0}(n), ..., d_{u,g}(n), ..., d_{u,G(u)-1}(n)]^T$. $G(u)$ and the corresponding modulation level, $N_{\mathrm{mod}}(u, g)$, are determined by ARMC so as to minimize the theoretical BER under the data rate constraint η bps/Hz/UE [4]. Each stream is multiplied by the selected phase sequence $\{\Phi_{m(u)}(n)\}$ and then is transformed into frequency-domain components block by DFT, yielding $\{\mathbf{D}_u(k)\}$ with $\mathbf{D}_u(k) =$



(a) SC uplink MU-MIMO with stream-wise SLM　(b) $\mathrm{PAPR}_{0.1\%}$ versus M　(c) BER performance

Fig. 1　Stream-wise blind SLM for SC uplink MU-MIMO.

$[D_{u,0}(k), ..., D_{u,g}(k), ..., D_{u,G(u)-1}(k)]^T$. Then, an $N_{\mathrm{UE}} \times G$ SVD based eigenmode transmit filtering matrix, $\mathbf{W}_{T,u}(k)$, is multiplied to $\mathbf{D}_u(k)$, yielding the frequency-domain transmit signal, $\mathbf{S}_u(k) = [S_{u,0}(k), ..., S_{u,n_{\mathrm{UE}}}(k), ..., S_{u,N_{\mathrm{UE}}-1}(k)]^T$, as

$$\mathbf{S}_u(k) = \sqrt{2E_s/T_s}\mathbf{W}_{T,u}(k)\mathbf{D}_u(k), \tag{3}$$

where E_s and T_s represents symbol energy and symbol duration. $\{\mathbf{S}_u(k)\}$ is then transformed back into time domain by IDFT, yielding the waveforms $\{\mathbf{s}_u(n)\}$ with $\mathbf{s}_u(n) = [s_{u,0}(n), ..., s_{u,n_{\mathrm{UE}}}(n), ..., s_{u,N_{\mathrm{UE}}-1}(n)]^T$.

A phase rotation sequence selection is based on minimizing the maximum PAPR among N_{UE} antennas. $\{\Phi_{\hat{m}(u)}(n)\}$ is selected for the u-th user as

$$\hat{m}(u)=\arg\min_{m=0\sim M-1}\left(\max_{n_{\mathrm{UE}}=0\sim N_{\mathrm{UE}}-1} \mathrm{PAPR}\left(\{s^m_{u,n_{\mathrm{UE}}}(n)\}\right)\right). \tag{4}$$

Here, $\{\mathbf{s}^m_{u,n_{\mathrm{UE}}}(n)\}$ is the waveform at the n_{UE}-th antenna and is obtained from $\{\Phi_m(n)\mathbf{d}_u(n)\}$.

3.2 Receiver (BS)

The received signal blocks through N_{BS} antennas are transformed into frequency domain by DFT to obtain a frequency-domain received signal vector at the k-th subcarrier $\mathbf{R}(k)$. Then, a $(U \cdot N_{\mathrm{UE}}) \times N_{\mathrm{BS}}$ MMSE based receive filtering matrix $\mathbf{W}_R(k)$ is multiplied to $\mathbf{R}(k)$ to obtain the frequency-domain output blocks $\{\hat{\mathbf{D}}(k)\}$ with $\hat{\mathbf{D}}(k) = [\hat{\mathbf{D}}_0(k), ..., \hat{\mathbf{D}}_u(k), ..., \hat{\mathbf{D}}_{U-1}(k)]^T$ and $\hat{\mathbf{D}}_u(k) = [\hat{D}_{u,0}(k), ..., \hat{D}_{u,g}(k), ..., \hat{D}_{u,G(u)-1}(k)]^T$. After that, $\{\hat{\mathbf{D}}(k)\}$ is transformed back into time domain by IDFT to acquire $\{\hat{\mathbf{d}}(n)\}$, with $\hat{\mathbf{d}}(n) = [\hat{\mathbf{d}}_0(n), ..., \hat{\mathbf{d}}_u(n), ..., \hat{\mathbf{d}}_{U-1}(n)]^T$ and $\hat{\mathbf{d}}_u(n) = [\hat{d}_{u,0}(n), ..., \hat{d}_{u,g}(n), ..., \hat{d}_{u,G(u)-1}(n)]^T$.

The phase rotation sequence estimation is carried out for each UE's streams. The time-domain received candidate corresponding to the v-th de-mapping sequence is $\{\hat{\mathbf{d}}^v_u(n) = \Phi^*_v(n)\hat{\mathbf{d}}_u(n)\}$, with $\hat{\mathbf{d}}^v_u(n) = [\hat{d}^v_{u,0}(n), ..., \hat{d}^v_{u,g}(n), ..., \hat{d}^v_{u,G(u)-1}(n)]^T$. The estimated phase rotation sequence of the u-th UE, $\{\Phi_{\tilde{m}(u)}(n)\}$, is

$$\tilde{m}(u)=\arg\min_{\substack{v=0\sim M-1\\ \mathbb{C}\in\Psi_{\mathrm{mod}}(u,g)}}\left(\sum_{g=0}^{G(u)-1}\sum_{n=0}^{N_c-1}\left|\hat{d}^v_{u,g}(n)-\mathbb{C}\right|^2\right). \tag{5}$$

$\Psi_{\mathrm{mod}}(u,g)$ is the original constellation of the g-th stream of the u-th user. Finally, the streams before de-modulation is given by $\tilde{\mathbf{d}}_u(n) = \Phi^*_{\tilde{m}(u)}(n)\hat{\mathbf{d}}_u(n)$.

4 Performance evaluation

PAPR at the complimentary cumulative distribution function (CCDF) equals 10^{-3}, called $\mathrm{PAPR}_{0.1\%}$, and uncoded BER are examined. We set U=2, N_{UE}=2, N_{BS}=4, N_c=128 and η=8 bps/Hz/UE. Ideal channel state information (CSI) estimation is assumed. In Fig. 1(b), when M=1 (no SLM), the use of transmit filtering increases the PAPR of SC waveform when comparing with SISO case. PAPR can be reduced by increasing M. Assuming M=256, the stream-wise blind SLM can lower the $\mathrm{PAPR}_{0.1\%}$ by 3.7 dB (0.5 dB higher than [5]).

Fig. 1(c) shows the BER at M=256. There is no major difference among BER when the average transmit E_s/N_0 is high. Moreover, the use of ARMC also improves the BER performance of stream-wise blind SLM at low E_s/N_0 region compared to that of 16QAM modulation. This is because the ARMC allocates most of the information bits to the stream with high channel gain and avoid using the stream with low channel gain, which improves the phase rotation estimation error.

5 Conclusion

In this paper, we firstly introduced the principle of stream-wise blind SLM. Then, the stream-wise blind SLM for MU-MIMO was introduced. The stream-wise blind SLM applies phase rotation sequence to each of the transmit data streams prior to transmit filtering. Simulation results for SC uplink MU-MIMO with ARMC confirmed that the stream-wise blind SLM can reduce the PAPR by about 3.7 dB without significant degradation in the uncoded BER performance.

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