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A SELECTIVE LINEAR TRANSCEIVER DESIGN OVER CORRELATED LARGE-MIMO CHANNELS

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Abstract—With tens or an even larger number of antennas utilized, large-MIMO systems have many potential merits. However, there are also some difficulties with its practical realization. For example, the feedback overhead caused by sending back a large precoding matrix is heavy. In this paper, we propose a selective linear transceiver scheme to reduce the overwhelming feedback overhead in correlated large-MIMO systems. In line with the required reduced amount of feedback, antennas which can provide a potentially large diversity gain are firstly chosen independently of the actual channel realization. The transceiver is then designed over correlated MIMO channels in an iterative way to minimize the sum of detection errors under the transmit power constraint. Although optimal solutions for the case of full transceiver have been given under some special scenarios, we modify them to improve the BER performance of systems. Monte-Carlo simulation results verify that the proposed selective linear transceiver is a useful scheme in large-MIMO systems to provide a tradeoff between performance and feedback overhead.

1. INTRODUCTION

It is well known that the performance of MIMO systems can be significantly improved by increasing the number of transmit and/or receive antennas [1]. Using tens or even hundreds of antennas [2], the large-MIMO systems can bear much higher data rate. However, there are also many challenges [2] for practically realizing large-MIMO systems, such as antennas placement [3–5], channel estimation and low-complexity near-optimal detection methods.

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Some low-complexity receivers are proposed in [2, 8]. However, there is only few works on the precoder design for large-MIMO systems [10]. Although traditional joint linear transceiver design methods [6, 9, 11] can be directly utilized, problems with feedback overhead and transmitter/receiver complexity become increasingly prominent in frequency division duplexing (FDD) systems with a large number of antennas.

In this paper, we design a selective linear transceiver over correlated large-MIMO channels. Motivated by reducing the feedback overhead, we constrain each data stream to be transmitted only over the selected antennas. The equalizing matrix can also be similarly constrained out of concern of detection complexity. Independently from the actual channel realization, antenna sets are firstly determined to provide possibly large diversity gain, and the transceiver structure is thus established. In an iterative way, we then design the transceiver under the total transmit power constraint to minimize the sum of detection errors. In the limiting case where both the precoder and the equalizer are full matrices, the proposed method is equivalent to that of [14]. The optimal full transceivers in the sense of minimizing the sum of detection errors are proposed for the special scenarios of correlated transmit antennas only [9] and correlated receive antennas only [14]. We modify both solutions to provide better bit error ratio (BER) performance. The proposed scheme is proven effective in achieving a tradeoff between performance and feedback overhead.

In Section 2, the correlated large-MIMO channel model is introduced and the transceiver design problem is formulated. In Section 3, we firstly introduce the structure of our selective transceiver, and then describe the antenna selection rules and transceiver design algorithm. The full transceiver case is discussed in the last part of Section 3. Numerical results and conclusions are given in Sections 4 and 5 respectively.

Notations: Vectors and matrices are represented by lower and upper case boldface letters. The superscripts $(\cdot)^*$ and $(\cdot)^H$ denote complex conjugate and complex conjugate transpose. $tr(\cdot)$ denotes matrix trace and $\otimes$ represents the Kronecker product. $\mathbf{I}$ represents the identity matrix and $\mathbf{J}_{m,n}^{i,j}$ is a single-entry $m \times n$ matrix with 0 everywhere except that the (i, j) entry is 1. $\mathbf{A}_{S_1,S_2}$ represents a $n_1 \times n_2$ submatrix of $\mathbf{A}$ with rows and columns indexed by the sets $\mathcal{S}_1$ and $\mathcal{S}_2$ respectively, where n_1 and n_2 are the cardinalities of $\mathcal{S}_1$ and $\mathcal{S}_2$.

2. SYSTEM MODEL

2.1. The Kronecker MIMO Channel Model

Due to the large quantity of antennas, correlation among antennas in large-MIMO systems is usually inevitable. We consider a $M_r \times M_t$ MIMO channel matrix with correlation following the widely used Kronecker model [12, 13, 15]

$$\mathbf{H} = \mathbf{H}_1 + \mathbf{C}_r^{1/2}\mathbf{H}_\epsilon \mathbf{C}_t^{T/2} \tag{1}$$

where $\mathbf{H}_1 = \mathbf{C}_r^{1/2}\hat{\mathbf{H}}\mathbf{C}_t^{T/2}$ is the channel estimation; both $\hat{\mathbf{H}}$ and $\mathbf{H}_\epsilon$ have i.i.d. entries distributed as $\mathcal{CN}(0, 1-\sigma_\epsilon^2)$ and $\mathcal{CN}(0, \sigma_\epsilon^2)$ respectively, where σ_ϵ^2 reflects channel estimation accuracy; $\mathbf{C}_t$, $\mathbf{C}_r$ are the covariance matrices at the transmitter and receiver side respectively, and the whole correlation matrix for MIMO systems is their Kronecker product $\mathbf{C} = \mathbf{C}_t \otimes \mathbf{C}_r$.

2.2. Linear Transceiver Design Problem Formulation

With the channel model (1), the detected signal $\hat{\mathbf{s}}$ in MIMO systems can be written as

$$\hat{\mathbf{s}} = \mathbf{G}\mathbf{y} = \mathbf{G}(\mathbf{H}\mathbf{F}\mathbf{s} + \mathbf{v}) \tag{2}$$

where $\mathbf{y}$ is the $M_r \times 1$ received symbol vector; $\mathbf{F} \in \mathbb{C}^{M_t \times M_s}$ and $\mathbf{G} \in \mathbb{C}^{M_s \times M_r}$ are the linear precoding and equalizing matrices respectively; $M_s \times 1$ ($M_s \leq \mathrm{rank}(\mathbf{H}) = \min(M_r, M_t)$) transmitted data vector $\mathbf{s}$ is assumed to have the covariance matrix $\mathbf{R}_s = \mathbb{E}[ss^H] = \mathbf{I}$; noise vector $\mathbf{v}$ is zero mean circularly symmetric complex Gaussian distributed with the covariance matrix $\mathbf{R}_v = \mathbb{E}[\mathbf{v}\mathbf{v}^H] = \sigma_v^2\mathbf{I}$, and is uncorrelated with $\mathbf{s}$.

To design the transceiver $\mathbf{F}$ and $\mathbf{G}$ under the criterion of minimizing the sum of mean squared error (MSE) subject to the total transmit power constraint $\mathcal{P}_0$, the following optimization problem can be formulated

$$\begin{aligned} \min_{\mathbf{F},\mathbf{G}} \quad & tr\left\{\mathbb{E}[\mathrm{MSE}(\mathbf{F},\mathbf{G})]\right\} = tr\left\{\mathbb{E}\left[\mathbf{e}\mathbf{e}^H\right]\right\} \\ \text{s.t.} \quad & tr\left(\mathbf{F}\mathbf{F}^H\right) \leq \mathcal{P}_0 \end{aligned} \tag{3}$$

where $\mathbf{e} = \hat{\mathbf{s}} - \mathbf{s}$ is the detection error vector and the MSE matrix thus is

$$\mathrm{MSE}(\mathbf{F},\mathbf{G}) = (\mathbf{G}\mathbf{H}\mathbf{F} - \mathbf{I})(\mathbf{G}\mathbf{H}\mathbf{F} - \mathbf{I})^H + \mathbf{G}\mathbf{R}_v\mathbf{G}^H \tag{4}$$

With the statistical CSI model (1), we can calculate the expected MSE matrix [7] as

$$\mathbb{E}[\mathrm{MSE}(\mathbf{F},\mathbf{G})] = (\mathbf{G}\mathbf{H}_1\mathbf{F} - \mathbf{I})(\mathbf{G}\mathbf{H}_1\mathbf{F} - \mathbf{I})^H + \mathbf{G}\widetilde{\mathbf{R}}_v\mathbf{G}^H \tag{5}$$

where $\widetilde{\mathbf{R}}_v = \mathbf{R}_v + \sigma_\epsilon^{-2} tr(\mathbf{F}\mathbf{F}^H\mathbf{C}_t^T)\mathbf{C}_r$. In the following, $\mathbf{F}$ and $\mathbf{G}$ will be designed at the receiver side based on the knowledge of σ_ϵ^2, $\hat{\mathbf{H}}$, $\mathbf{C}_t$ and $\mathbf{C}_r$, and then $\mathbf{F}$ will be fed back to the transmitter.

3. SELECTIVE LINEAR TRANSCEIVER DESIGN FOR LARGE-MIMO SYSTEMS

3.1. Selective Transceiver Structure

If the precoder $\mathbf{F}$ has $M_t M_s$ nonzero entries, large-MIMO systems may suffer overwhelming feedback overhead. The large dimensions of the matrices $\mathbf{G}$ and $\mathbf{F}$ also make the transceiver complicated. An intuitive approach to deal with these problems is to restrict that only some carefully selected entries of $\mathbf{F}$ are non-zero. It is assumed that the pth data stream is transmitted only over the antenna set $\mathcal{F}(p)$, which has cardinality of $n(\mathcal{F}(p)) = B_1 \leq M_t$. In other words, each data stream has symbols transmitted only on B_1 out of M_t antennas. The equalizer $\mathbf{G}$ can also be similarly constrained over the antenna sets $\mathcal{G}(p)$, of which the cardinality is $n(\mathcal{G}(p)) = B_2 \leq M_r$. The transceiver $\mathbf{F}$ and $\mathbf{G}$ can thus be mathematically expressed as

$$\begin{aligned} \mathbf{F} &= \sum\nolimits_{p=1}^{M_s} \sum\nolimits_{q\in\mathcal{F}(p)} \mathbf{J}_{M_t,M_s}^{q,p} \otimes \mathbf{F}_{p,q} \\ \mathbf{G} &= \sum\nolimits_{p=1}^{M_s} \sum\nolimits_{q\in\mathcal{G}(p)} \mathbf{J}_{M_s,M_r}^{p,q} \otimes \mathbf{G}_{p,q} \end{aligned} \tag{6}$$

where the entry $\mathbf{F}_{p,q}$ generates the transmitted signal component over the qth antenna for transmitting the pth data stream; the $B_1 \times 1$ column vector $\mathbf{F}_p$, formed by stacking the entries $\mathbf{F}_{p,q}$ ($q \in \mathcal{F}(p)$), is the whole precoder for generating the transmitted signal components over the set of antennas $\mathcal{F}(p)$ corresponding to the pth data stream. The $\mathbf{G}_{p,q}$ operates on the received signal of the qth antenna for detecting the pth data stream; the $1 \times B_2$ row vector $\mathbf{G}_p$, formed by stacking the entries $\mathbf{G}_{p,q}$ ($q \in \mathcal{G}(p)$), is the whole equalizer operating on the set of antennas $\mathcal{G}(p)$ to obtain the pth data stream. In short, each data stream is equalized over only B_2 instead of all M_r antennas.

Given B_1 and B_2, we need to choose the elements of the sets $\mathcal{F}(p)$ and $\mathcal{G}(p)$, which are the antennas utilized for transmission and equalization of the symbol of the pth data stream. The optimal antenna sets $\mathcal{F}$ and $\mathcal{G}$ can be found by exhaustive search over all possible sets and comparing the resultant objective function value of (3) with the knowledge of the channel $\mathbf{H}$. However, it not only causes very high complexity but also increases the feedback overhead since the transmitter needs to be informed about $\mathcal{F}(p)$.

As a result, we assume that the selected antenna sets do not change with the channel realization $\mathbf{H}$. We select those antennas that can provide statistically large transmit and receive diversity. Since the correlation between antennas usually decreases with antenna distance, the diversity provided by choosing neighboring antennas is limited due to their strong correlations. To obtain as a large diversity as possible, we choose the antennas based on the following two rules: (a) the minimal distance between each pair of selected antennas should be as large as possible; (b) the antennas should be selected with an equal probability.

For clarity, we show an exemplary $\mathbf{F}$ for a 32×16 MIMO system in Figure 1 when each of 16 data streams is transmitted over 4 or 16 antennas. The black squares represent the selected antennas for transmission. For each data stream, the minimal distance between different selected antenna pairs are all equal. Besides, the transmitted symbol over each antenna is formed by an equal number of data streams. Compared with a full precoding matrix, the feedback overhead of our transceiver when $B_1 = 4$ and 16 can be reduced by 87.5% and 50% respectively.

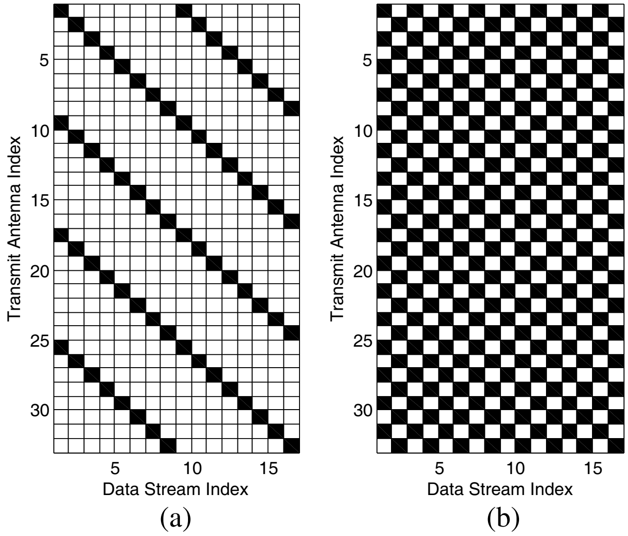


Figure 1. Selected transmit antennas when (a) $B_1 = 4$ and (b) $B_1 = 16$.

3.2. Lagrange Multiplier Method Based Solution

The objective function in (3) is non-convex even for both full $\mathbf{F}$ and $\mathbf{G}$. However, it has been pointed out in [14] that the global optimal solution satisfies the first-order Karush-Kuhn-Tucker (KKT) conditions. Although the transceiver structure is constrained as shown in (6), we can still find the solution according to the first-order KKT conditions. The Lagrange function of the optimization problem shown in (3) can be expressed as

$$\begin{aligned} L(\mathbf{G};\mathbf{F};\lambda) \;=\; & tr\left[(\mathbf{G}\mathbf{H}_1\mathbf{F}-\mathbf{I})\,(\mathbf{G}\mathbf{H}_1\mathbf{F}-\mathbf{I})^H+\mathbf{G}\widetilde{\mathbf{R}}_v\mathbf{G}^H\right] \\ & +\lambda\times\left[tr\left(\mathbf{F}\mathbf{F}^H\right)-\mathcal{P}_0\right] \end{aligned} \tag{7}$$

where λ is a nonnegative Lagrange multiplier.

The first-order partial derivative [16] of L with respect to $\mathbf{G}^*_{p,q}$ and $\mathbf{F}^*_{p,q}$ can be obtained as

$$\frac{\partial L}{\partial \mathbf{F}^*_{p,q}} \;=\; \Phi_{q,\mathcal{F}(p)}\mathbf{F}_p-\Gamma_{q,p}+\lambda\mathbf{F}_{q,p} \tag{8a}$$

$$\frac{\partial L}{\partial \mathbf{G}^*_{p,q}} \;=\; \mathbf{G}_p\Xi_{\mathcal{G}(p),q}-\Delta_{p,q} \tag{8b}$$

where

$$\begin{aligned} &\Gamma=\mathbf{H}_1^H\mathbf{G}^H \\ &\Phi=\Gamma\Gamma^H+\sigma_\epsilon^2 tr(\mathbf{G}\mathbf{C}_r\mathbf{G}^H)\mathbf{C}_t^T \\ &\Delta=\mathbf{F}^H\mathbf{H}_1^H \\ &\Xi=\Delta^H\Delta+\widetilde{\mathbf{R}}_v \end{aligned} \tag{9}$$

Stacking the results of (8), we can also obtain the following first-order partial derivatives of L with respect to $\mathbf{F}^*_p$ and $\mathbf{G}^*_p$

$$\frac{\partial L}{\partial \mathbf{F}^*_p} \;=\; \Phi_{\mathcal{F}(p),\mathcal{F}(p)}\mathbf{F}_p-\Gamma_{\mathcal{F}(p),p}+\lambda\mathbf{F}_p \tag{10a}$$

$$\frac{\partial L}{\partial \mathbf{G}^*_p} \;=\; \mathbf{G}_p\Xi_{\mathcal{G}(p),\mathcal{G}(p)}-\Delta_{p,\mathcal{G}(p)} \tag{10b}$$

The optimal solution of (3) must satisfy the first-order KKT conditions, which means that both $\partial L/\partial\mathbf{F}^*_p$ in (10a) and $\partial L/\partial\mathbf{G}^*_p$ in (10b) are zero matrices. Consequently,

$$\mathbf{F}_p \;=\; \left(\Phi_{\mathcal{F}(p),\mathcal{F}(p)}+\lambda\mathbf{I}_{B_1}\right)^{-1}\Gamma_{\mathcal{F}(p),p} \tag{11a}$$

$$\mathbf{G}_p \;=\; \Delta_{p,\mathcal{G}(p)}\Xi^{-1}_{\mathcal{G}(p),\mathcal{G}(p)} \tag{11b}$$

where λ is numerically found as shown in the Appendix so that the total transmit power constraint can be satisfied. Using the equations in (11a) and (11b), we propose the following iterative algorithm to design the linear transceiver. In the extreme case where both $\mathbf{G}$ and $\mathbf{F}$ are full matrices ($B_1 = M_t$, $B_2 = M_r$), our solution is identical to that of [14, Table 1] and λ can be directly calculated by $\lambda = tr(\mathbf{G}\widetilde{\mathbf{R}}_v\mathbf{G}^H)/\mathcal{P}_0$ [14, Lemma1].

3.3. Full Transceiver Scenario

When both the precoding and equalizing matrices are full, the optimization problem (3) has a closed-form solution if either ρ_r or ρ_t is zero. We restate the solutions here for further analysis.

Algorithm 1 Iterative Transceiver Design Method for Large-MIMO Systems

1. Randomly generate a precoder $\mathbf{F}^{(0)}$ satisfying the total power constraint and then calculate the corresponding equalizing matrix $\mathbf{G}$ via (11b). Evaluate the objective function $f(0) = tr\{\mathbb{E}[\mathrm{MSE}(\mathbf{G}^{(0)}, \mathbf{F}^{(0)})]\}$ via (5).
2. For the jth ($1 \leq j \leq J$) iteration, update the transceiver $\mathbf{F}^{(j)}$ and $\mathbf{G}^{(j)}$ via (11), and calculate $f(j) = tr\{\mathbb{E}[\mathrm{MSE}(\mathbf{G}^{(j)}, \mathbf{F}^{(j)})]\}$ via (5).
3. If the convergence threshold is satisfied $\frac{|f(j)-f(j-1)|}{f(j-1)} \leq \epsilon$, stop; otherwise, repeat step 2.

3.3.1. Solution for the Special Case $\rho_t \neq 0$, $\rho_r = 0$ [14]

The optimal solution of (3) when $\rho_r = 0$ is proposed in [14, Section III]. Define the eigenvalue decomposition (EVD)

$$\begin{aligned} &\left(\sigma_v^2\mathbf{I} + \sigma_\epsilon^2\mathcal{P}_0\mathbf{C}_t^T\right)^{-1/2}\mathbf{H}_1^H\mathbf{H}_1\left(\sigma_v^2 + \sigma_\epsilon^2\mathcal{P}_0\mathbf{C}_t^T\right)^{-1/2} \\ &= \begin{pmatrix} \mathbf{V} & \widetilde{\mathbf{V}} \end{pmatrix} \begin{pmatrix} \Lambda & 0 \\ 0 & \widetilde{\Lambda} \end{pmatrix} \begin{pmatrix} \mathbf{V} & \widetilde{\mathbf{V}} \end{pmatrix}^H \end{aligned} \tag{12}$$

where Λ contains all positive eigenvalues arranged in a decreasing order, $\widetilde{\Lambda}$ contains zero eigenvalues, $\mathbf{V}$ and $\widetilde{\mathbf{V}}$ contain the corresponding eigenvectors. The optimal solution of (3) is

$$\begin{aligned} \mathbf{F}_{opt} &= \left(\sigma_v^2\mathbf{I} + \sigma_\epsilon^2\mathcal{P}_0\mathbf{C}_t^T\right)^{-1/2}\mathbf{V}\Lambda_{\mathbf{F}_{opt}} \\ \mathbf{G}_{opt} &= \Lambda_{\mathbf{G}_{opt}}\mathbf{V}^H\left(\sigma_v^2\mathbf{I} + \sigma_\epsilon^2\mathcal{P}_0\mathbf{C}_t^T\right)^{-1/2}\mathbf{H}_1^H \end{aligned} \tag{13}$$

The diagonal $M_s \times M_s$ matrices $\Lambda_{\mathbf{F}_{opt}}$ and $\Lambda_{\mathbf{G}_{opt}}$ are given by

$$\begin{aligned}\Lambda_{\mathbf{F}_{opt}} &= \left(\tau_1^{1/2}\mu_1^{-1/2}\sigma_v\Lambda^{-1/2} - \tau_1\Lambda^{-1}\right)_+^{1/2} \\ \Lambda_{\mathbf{G}_{opt}} &= \left(\tau_1^{-1/2}\mu_1^{1/2}\sigma_v^{-1}\Lambda^{-1/2} - \mu_1/\sigma_v^2\Lambda^{-1}\right)_+^{1/2}\Lambda^{-1/2}\end{aligned} \tag{14}$$

where

$$\begin{aligned}\tau_1 &= \frac{a_2\mathcal{P}_0}{a_3\mathcal{P}_0 + a_1a_3 - a_2a_4} \\ \mu_1 &= \frac{a_2\sigma_v^2(a_3\mathcal{P}_0 + a_1a_3 - a_2a_4)}{(\mathcal{P}_0 + a_1)^2\mathcal{P}_0}\end{aligned} \tag{15}$$

Denote $M_s' \leq M_s$ as the number of nonzero diagonal entries of $\Lambda_{\mathbf{F}_{opt}}$, the coefficients $a_1 \sim a_4$ are traces of the $M_s' \times M_s'$ top-left submatrices of Λ^{-1}, $\Lambda^{-1/2}$, $\Lambda^{-1/2}\mathbf{V}^H(\sigma_v^2\mathbf{I} + \sigma_\epsilon^2\mathcal{P}_0\mathbf{C}_t^T)^{-1}\mathbf{V}$ and $\Lambda^{-1}\mathbf{V}^H(\sigma_v^2\mathbf{I} + \sigma_\epsilon^2\mathcal{P}_0\mathbf{C}_t^T)^{-1}\mathbf{V}$, respectively. The value of M_s' can be found by an iterative method shown in [14, Appendix E].

3.3.2. Solution for the Special Case $\rho_t = 0$, $\rho_r \neq 0$ [9]

The optimal solution of (3) in the case of $\rho_t = 0$ has been given in [9, *Theorem 1*], which consists of eigenvectors of the matrix $\mathbf{H}_1^H(\sigma_v^2\mathbf{I} + \sigma_\epsilon^2\mathcal{P}_0\mathbf{C}_r)^{-1}\mathbf{H}_1$ and a power allocation strategy.

3.3.3. Modified Solutions

The solutions given in [14] and [9] can achieve the minimal MSE trace and the resultant MSE matrices (5) are diagonal. However, the MSEs for different data streams (i.e., the diagonal entries of MSE matrices) are not equal. If the same modulation is used on each data stream, such MSEs variation will degrade the BER performance especially in the high SNR regime, where the overall BER is dominated by the data stream with the largest MSE.

The average BER can be utilized as the criterion for transceiver design as shown in [6, Section V(C)]. It has been proved in [6] that the BER is a Schur-convex function in the low BER regime. However, even when different data streams employ the same constellation, the optimal power allocation problem is still hard to be solved [6, Equation (51)].

Since the solutions given in [14] and [9] are both unique up to a unitary transform, we can multiply the precoding matrix with a unitary matrix and change the equalizing matrix accordingly. Although the resultant new MSE matrices are not diagonal anymore, the diagonal entries of the new MSE matrices are majorized by those of the original ones. Such unitary matrix can be a Hadamard or a standard DFT matrix, which endows the new MSE matrix with identical diagonal entries [6, Theorem 1] while maintaining the trace unvaried. Although this is not the optimal solution for minimizing the overall system BER, it achieves better BER performance via an identical MSE for different data streams. In the case of $\rho_r = 0$ [14], the optimal solution (13) is modified as:

$$\begin{aligned} \mathbf{F}'_{opt} &= \mathbf{F}_{opt}\mathbf{F}_{dft} \\ \mathbf{G}'_{opt} &= \mathbf{F}^H_{dft}\mathbf{G}_{opt} \end{aligned} \tag{16}$$

where $\mathbf{F}_{dft}$ is the standard M_s-point DFT matrix. When $\rho_t = 0$ [9], we can similarly right multiply the optimal precoding matrix with $\mathbf{F}_{dft}$ and then change the MMSE equalizer accordingly.

4. NUMERICAL RESULTS AND DISCUSSION

We define the transmit signal to noise ratio as $\mathrm{SNR}_{\mathrm{dB}} \triangleq 10\log_{10}\mathcal{P}_0/\sigma_v^2$. The large-MIMO system is equipped with $M_t = 32$ transmit and $M_r = 16$ receive antennas, where $M_s = 16$ QPSK modulated data streams are transmitted. The correlation matrices are set as $[\mathbf{C}_t]_{i_1,j_1} = \rho_t^{|i_1-j_1|}$ for $i_1, j_1 \in \{1, 2, \ldots, M_t\}$ and $[\mathbf{C}_r]_{i_2,j_2} = \rho_r^{|i_2-j_2|}$ for $i_2, j_2 \in \{1, 2, \ldots, M_r\}$, where $0 \leq \rho_t,\ \rho_r \leq 1$.

Each data stream is assumed to be transmitted over 4, 16 or 32 antennas, where the detailed antenna selection for $B_1 = 4$ and 16 are shown in Figure 1. Besides, the equalizing matrix is assumed to be full ($B_2 = M_r$) for better BER performance. In the following, different transceivers are compared under three different scenarios: correlated transmit antennas only, correlated receive antennas only, and both correlated transmit/receive antennas.

4.1. Correlated Transmit Antennas Case ($\rho_t \neq 0$, $\rho_r = 0$)

In Figure 2, we firstly compare the BER performance of full transceivers. The optimal solution of (3) given in [14] and the proposed modified version (16) are compared under three σ_ϵ^2 values. With more accurate channel estimation (smaller σ_ϵ^2), the BER performance of both transceivers is improved. However, there is always performance advantage of our modified scheme over that of the transceiver [14]. The MSE matrix has identical diagonal entries via (16) while its trace remains the same. In other words, the resultant MSEs on different data streams are identical, which provides better BER performance when the same modulation is adopted on each stream. The SNR gain at

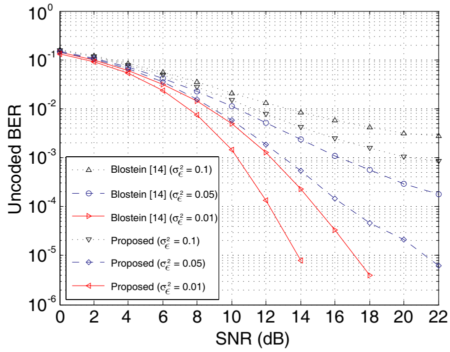
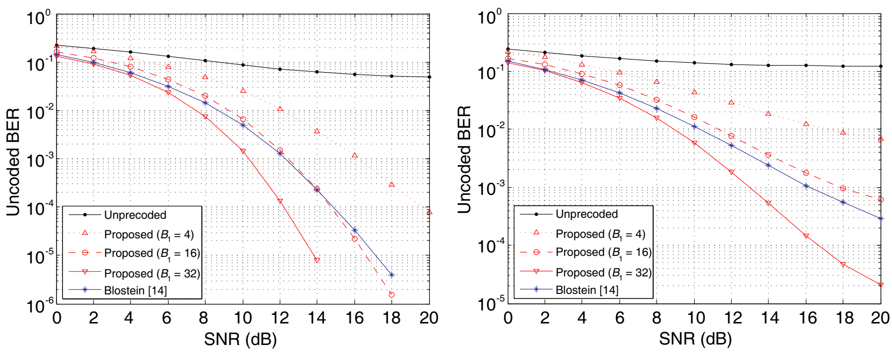


Figure 2. Uncoded BER vs. SNR of the proposed transceiver ($B_1 = M_t$, $B_2 = M_r$) and the scheme in [14] when $\rho_t = 0.5$ and $\rho_r = 0$.



(a) (b)

Figure 3. Uncoded BER vs. SNR of the unprecoded scheme, the proposed transceiver ($B_2 = M_r$), and the scheme in [14] when $\rho_t = 0.5$, $\rho_r = 0$ and (a) $\sigma_\epsilon^2 = 0.01$ or (b) $\sigma_\epsilon^2 = 0.05$.

10^{-4} brought about by such modification is about 3 dB for $\sigma_\epsilon^2 = 0.01$ and larger than 5 dB for $\sigma_\epsilon^2 = 0.05$.

Next, we evaluate the BER performance of the proposed selective transceiver with different B_1 values in Figure 3. The unprecoded scheme is evaluated for comparison, where the data streams are transmitted over randomly selected antennas under the same power constraint. Although the unprecoded scheme has performance improved with decreasing σ_ϵ^2, it always suffers a severe error floor. Even with only 4 antennas selected to transmit each data stream, the proposed transceiver performs much better than the unprecoded

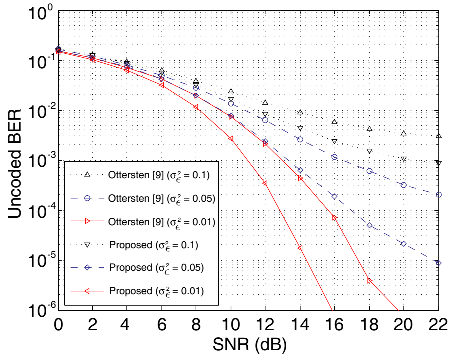


Figure 4. Uncoded BER vs. SNR of the proposed transceiver ($B_1 = M_t$, $B_2 = M_r$) and the scheme in [9] when $\rho_t = 0$ and $\rho_r = 0.5$.

scheme in terms of the disappearance of the error floor and a larger diversity order. For example, the BER at SNR of 20 dB is decreased from 0.05 to 8×10^{-5} when $\sigma_\epsilon^2 = 0.01$. The proposed scheme performs close to or even better than the transceiver of [14] when B_1 increases to 16. We see that there is a SNR gap between our transceiver and the scheme in [14] of only about 1 dB when $\sigma_\epsilon^2 = 0.05$. Under the condition of $\sigma_\epsilon^2 = 0.01$, our scheme has almost the same performance as that of [14] and even a little larger diversity order.

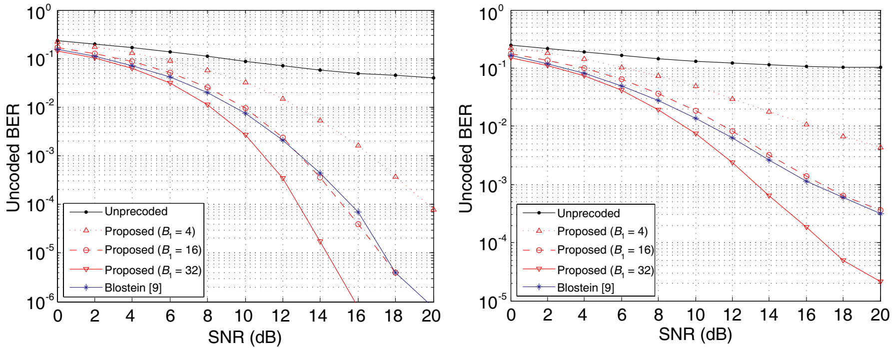
4.2. Correlated Receive Antennas Case ($\rho_t = 0$, $\rho_r \neq 0$)

For the case of $\rho_t = 0$, the BER performance of the full transceiver proposed in [9, Section 6] and its modified version described in the paragraph following (16) is compared in Figure 4. Due to the identical diagonal entries of the transformed MSE matrix of the modified transceiver, the modified transceiver always outperform the scheme in [9] under different channel accuracy scenarios.

In Figure 5, we evaluate the BER performance of the proposed transceiver with different B_1 values when $\rho_t = 0$. It can be observed that the proposed transceiver significantly outperforms the unprecoded scheme even with such small B_1 as 4, and exhibits about the same BER performance with that of the scheme in [9] when B_1 increases to 16.

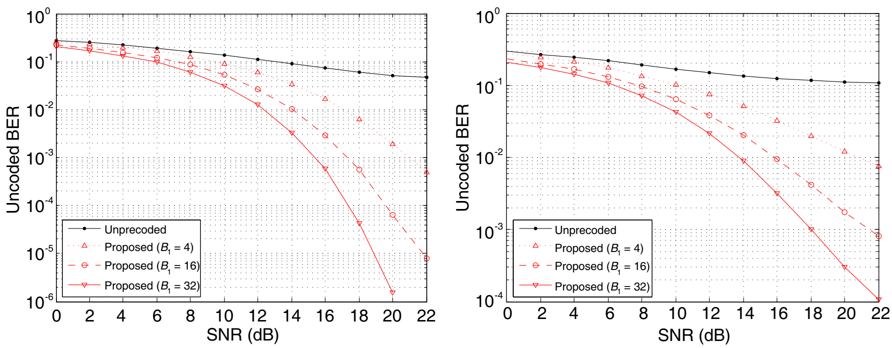
4.3. Correlated Transmit and Receive Antennas Case ($\rho_t \neq 0$, $\rho_r \neq 0$)

In Figure 6, we evaluate the proposed transceiver in the case of both correlated transmit and receive antennas. So far, there is no closed-



(a) (b)

Figure 5. Uncoded BER vs. SNR of the unprecoded scheme, the proposed transceiver ($B_2 = M_r$) and the scheme in [9] when $\rho_t = 0$, $\rho_r = 0.5$ and (a) $\sigma^2 = 0.01$ or (b) $\sigma_\epsilon^2 = 0.05$.



(a) (b)

Figure 6. Uncoded BER vs. SNR of the unprecoded scheme, the proposed transceiver ($B_2 = M_r$) when $\rho_t = 0.2$, $\rho_r = 0.8$ and (a) $\sigma_\epsilon^2 = 0.01$ or (b) $\sigma_\epsilon^2 = 0.05$.

form solution for such case. With only 4 transmit antennas selected, the proposed transceiver performs much better than the unprecoded scheme, which suffers from severe error floors of 0.05 and 0.1 when $\sigma_\epsilon^2 = 0.01$ and 0.05 respectively. In the left of Figure 6, we can observe that there is a SNR gain of about $2 \sim 3$ dB at BER of 10^{-4} when the proposed transceiver increases the parameter B_1 from 16 to 32. However, the proposed transceiver with $B_1 = 16$ shows about the same diversity order as that of $B_1 = 32$. Similar relationship can also be observed when σ_ϵ^2 increases to 0.05. As a result, the proposed transceiver with $B_1 = 16$ can acquire a large portion of performance improvement brought by a full precoding matrix.

5. CONCLUSION

A selective linear transceiver, where data streams are precoded over deliberately selected antennas, is proposed for correlated large-MIMO systems in this paper. We firstly select antennas to provide potentially large diversity without the knowledge of actual channel realization. Correspondingly, the structure of the precoding and equalizing matrices are established. Based on the Lagrange multipliers method, the transceiver is then designed over Kronecker correlated MIMO channels to minimize the sum of detection errors under the total transmit power constraint.

With part of transmit antennas selected, the feedback overhead of the precoding matrix can be significantly reduced. The numerical results show that good error performance can be achieved by transmitting data only over the selected small portion of transmit antennas. When the transceiver consists of full precoding and equalizing matrices, optimal solutions with closed-form expression are provided in previous works for the scenario of either uncorrelated transmit or uncorrelated receive antennas. In this paper, such solutions are modified so that the resultant MSE matrix has identical diagonal entries with its trace remaining unaltered. Simulations show that the modified transceivers can improve the BER performance of systems. Our transceiver possesses the features of both the transmit antenna selection scheme and traditional linear precoding scheme. The tradeoff between performance and feedback overhead can thus be achieved with a large degree of freedom.

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APPENDIX A. NUMERICAL METHODS TO OBTAIN $\boldsymbol{\lambda}$

Similar to that of [17, Section III], we denote the eigenvalue decomposition of $\Phi_{\mathcal{F}(p),\mathcal{F}(p)} = \mathbf{U}_p\mathbf{D}_p\mathbf{U}_p^H$ with $\mathbf{U}_p$ and $\mathbf{D}_p$ contain eigenvectors and nonnegative eigenvalues respectively. The precoder in (11a) can be rewritten as $\mathbf{F}_p = \mathbf{U}_p(\mathbf{D}_p + \lambda\mathbf{I})^{-1}\mathbf{U}_p^H\Gamma_{\mathcal{F}(p),p}$. Define the Hermitian matrix $\mathbf{Q}_p = \mathbf{U}_p^H \Gamma_{\mathcal{F}(p),p} \Gamma_{\mathcal{F}(p),p}^H \mathbf{U}_p$, we have

$$tr\left(\mathbf{F}_p \mathbf{F}_p^H\right) = tr\left\{[\mathbf{D}_p + \lambda \mathbf{I}]^{-2} \mathbf{Q}_p\right\} = \sum\nolimits_{i=1}^{2B_1+1} \frac{\mathbf{Q}_p(i,i)}{[\mathbf{D}_p(i,i)+\lambda]^2} \quad \text{(A1)}$$

and

$$\begin{aligned} tr\left(\mathbf{F}\mathbf{F}^H\right) &= \sum\nolimits_{p=0}^{M_s-1} tr(\mathbf{F}_p \mathbf{F}_p^H) \\ &= \sum\nolimits_{p=0}^{M_s-1} \sum\nolimits_{i=1}^{2B_1+1} \frac{\mathbf{Q}_p(i,i)}{[\mathbf{D}_p(i,i)+\lambda]^2} \leq \mathcal{P}_0 \end{aligned} \quad \text{(A2)}$$

where λ can be numerically found to satisfy (A2).

REFERENCES

1. Foschini, G. J. and M. J. Gans, "On limits of wireless communications in a fading environment when using multiple antennas," *Wirel. Pers. Commun.*, Vol. 6, Mar. 1998.
2. Vardhan, K. V., S. K. Mohammed, A. Chockalingam, and B. S. Rajan, "A low-complexity detector for large MIMO systems and multicarrier CDMA systems," *IEEE J. Sel. Areas Commun.*, Vol. 26, 473–485, 2008.
3. Elnour, B. and D. Erricolo, "A novel colocated cross-polarized two-loop PCB antenna in the ISM 2.4-GHz band," *IEEE Antennas Wirel. Propag. Lett.*, Vol. 9, 1237–1240, 2010.
4. Krairiksh, M., P. Keowsawat, C. Phongcharoenpanich, and S. Kosulvit, "Two-probe excited circular ring antenna for MIMO application," *Progress In Electromagnetics Research*, Vol. 97, 417–431, 2009.
5. Chung, J.-Y., T. Yang, and J. Lee, "Low correlation MIMO antennas with negative group delay," *Progress In Electromagnetics Research C*, Vol. 22, 151–163, 2011.
6. Palomar, D. P., J. M. Cioffi, and M. A. Lagunas, "Joint Tx-Rx beamforming design for multicarrier MIMO channels: A unified framework for convex optimization," *IEEE Trans. Signal Process.*, Vol. 51, 2381–2401, 2003.
7. Gupta, A. and D. Nagar, *Matrix Variate Distributions*, Chapman & Hall/CRC, London, UK, 2000.
8. Srinidhi, N., T. Datta, A. Chockalingam, and B. S. Rajan, "Layered tabu search algorithm for large-MIMO detection and a lower bound on ML performance," *IEEE Trans. Commun.*, Vol. 59, 2955–2963, 2011.

9. Zhang, X., D. P. Palomar, and B. Ottersten, "Statistically robust design of linear MIMO transceivers," *IEEE Trans. Signal Process.*, Vol. 56, 3678–3689, 2008.
10. Artigue, C. and P. Loubaton, "On the precoder design of flat fading MIMO systems equipped with MMSE receivers: A large-system approach," *IEEE Trans. Inf. Theory*, Vol. 57, 4138–4155, 2011.
11. Scaglione, A., P. Stoica, S. Barbarossa, G. B. Giannakis, and H. Sampath, "Optimal designs for space-time linear precoders and decoders," *IEEE Trans. Signal Process.*, Vol. 50, 1051–1064, 2002.
12. Are, H. and G. David, "Precoded orthogonal space-time block codes over correlated ricean MIMO channels," *IEEE Trans. Signal Process.*, Vol. 55, 779–783, 2007.
13. Ding, M. H. and S. D. Blostein, "Maximum mutual information design for MIMO systems with imperfect channel knowledge," *IEEE Trans. Inf. Theory*, Vol. 56, 4793–4801, Oct. 2010.
14. Ding, M. H. and S. D. Blostein, "MIMO minimum total MSE transceiver design with imperfect CSI at both ends," *IEEE Trans. Signal Process.*, Vol. 57, 1141–1150, Mar. 2009.
15. Li, X., S. Jin, X. Q. Gao, and K. K. Wong, "Near-optimal power allocation for MIMO channels with mean or covariance feedback," *IEEE Trans. Commun.*, Vol. 58, 289–300, 2010.
16. Hjørungnes, A., *Complex Valued Matrix Derivatives — With Applications in Signal Processing and Communications*, Cambridge University Press, 2011.
17. Shen, H., B. Li, M. Tao, and X. Wang, "MSE-based transceiver designs for the MIMO interference channel," *IEEE Trans. Wireless Commun.*, Vol. 9, 3480–3489, 2010.