

Coordinated Multi-point Transmission with Non-ideal Channel Reciprocity

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Abstract—This paper studies robust transmission strategies for downlink time division duplex coordinated multi-point (CoMP) systems with non-ideal uplink-downlink channel reciprocity due to imperfect antenna calibration. By exploiting the statistics of antenna calibration errors, we first characterize the optimal parametric precoder structure that maximizes the weighted sum rate, based on which a closed-form robust signal-to-leakage-plus-noise ratio (RSLNR) precoder with properly selected parameters is then proposed. Simulation results show that the proposed precoder together with user scheduling provides near-optimal performance and data sharing among coordinated BSs may become detrimental depending on the employed precoders and the accuracy of antenna calibration.

I. INTRODUCTION

Coordinated multi-point (CoMP) transmission is a promising technique to handle inter-cell interference in universal frequency reuse cellular networks [1]. Coherently cooperative transmission from all coordinated base stations (BSs) to all users, or CoMP-JP (joint processing) in the context of Long-Term Evolution-Advanced (LTE-A), is able to fully exploit the potential of CoMP [1, 2], where both data and channel state information (CSI) are shared among the BSs.

Although in principle CoMP-JP is similar to a single-cell multiuser multi-input multi-output (MIMO) system, there are important differences between the two systems, such that the well-explored transmit strategies cannot be extended to CoMP directly. First of all, per-BS power constraints (PBPC) should be considered instead of sum power constraints, which yields more complicated optimization [3]. Second, CoMP channel is a concatenation of multiple single-cell channels. There exist multiplicative noises in the downlink channel which is either led by non-ideal uplink-downlink channel reciprocity in time division duplex (TDD) systems [4], or by practical codebooks in frequency division duplex (FDD) systems [5]. Compared with additive noises such as channel estimation errors, these multiplicative noises are more detrimental because they will hinder the co-phasing of coherent CoMP transmission. On the other hand, the performance gain of CoMP-JP comes at costs of high capacity and low-latency backhaul links as well as

increased signalling overhead. Moreover, even with such an expensive architecture, CoMP-JP cannot achieve the promised performance when imperfect CSI is directly applied [6]. This is because the desired signals from multiple BSs may be added destructively, which in fact becomes interference.

The CSI at the BSs is in demand for all kinds of CoMP systems to gain their benefits. It is widely recognized that TDD is more applicable for CoMP systems than FDD, because FDD needs large feedback overhead for providing CSI to the BSs [5]. In TDD systems, the downlink channel can be obtained by the BSs via estimating the uplink channel exploiting channel reciprocity. However, the uplink and downlink channels are only reciprocal for the propagation channels, which is invalid in practical systems due to the imperfect calibration for analog gains of radio frequency (RF) chains in transmit and receive antennas [4, 7]. In CoMP-JP systems, it has been shown that the non-ideal channel reciprocity will lead to severe performance degradation [4].

In this paper, we design robust multiuser precoder against the calibration errors in CoMP-JP systems. Under the imperfect CSI corrupted by such kind of multiplicative noises, we first establish the equivalence between a weighted sum rate maximization problem and a weighted sum mean square error (MSE) minimization problem both subject to PBPC. Based on this result, the optimal parametric precoder structure for maximizing the weighted sum rate is provided, and a closed-form precoder with properly selected parameters is proposed. Simulation results demonstrate that the proposed closed-form precoder in conjunction with user scheduling achieves near-optimal performance.

Notation: $(\cdot)^T$, $(\cdot)^*$, $(\cdot)^H$, and $(\cdot)^\dagger$ denote the transpose, the complex conjugate, the conjugate transpose, and the Moore-Penrose inverse, respectively. $\text{diag}\{\cdot\}$ denotes a diagonal matrix, $\Re\{\cdot\}$ denotes the real part of a complex number, and $\|\cdot\|$ denotes Euclidian norm. \mathbf{I}_N and $\mathbf{0}_N$ denote $N \times N$ identity and zero matrices, respectively. For a set \mathcal{S} , its elements are $\mathcal{S}(1), \dots, \mathcal{S}(|\mathcal{S}|)$ where $|\mathcal{S}|$ is the cardinality of \mathcal{S} .

II. SYSTEM AND CHANNEL MODEL

A. System Model

Consider a downlink CoMP-JP system where a cluster of N_c coordinated BSs, each equipped with N_t antennas, jointly serve N_u single-antenna users. We assume that the

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data and CSI to multiple users are fully shared among the BSs via noiseless and zero latency backhaul links. For notational simplicity, we denote the u th user and the b th BS as MS_u and BS_b , respectively, for $u \in \mathcal{U} = \{1, \dots, N_u\}$ and $b \in \mathcal{B} = \{1, \dots, N_c\}$. For simplicity, we call CoMP-JP as CoMP for short in the remaining of this paper.

In TDD systems, channel reciprocity allows each BS to obtain its downlink channel to all users by estimating the uplink channel. To highlight the multiplicative noises caused by calibration errors, we assume that the uplink channel estimation is perfect, i.e., we ignore the additive noises. We assume block flat fading channels and denote the downlink channel comprising both propagation channel and analog gains of RF chains from BS_b to MS_u as $\mathbf{h}_{ub,D} \in \mathbb{C}^{N_t \times 1}$, then the downlink CoMP channel from all BSs to MS_u is a concatenation of multiple single-cell channels, which is $\mathbf{h}_{u,D} = [\mathbf{h}_{u1,D}^T, \dots, \mathbf{h}_{uN_c,D}^T]^T$.

When linear precoding is used, the signal received by MS_u can be expressed as

$$y_u = \mathbf{h}_{u,D}^H \mathbf{w}_u x_u + \mathbf{h}_{u,D}^H \sum_{j \neq u} \mathbf{w}_j x_j + z_u, \quad (1)$$

where x_u is the data symbol for MS_u , the data symbols of all users are assumed as independent and identically distributed (i.i.d.) Gaussian random variables with zero mean and unit variance, z_u is the additive white Gaussian noise (AWGN) with zero mean and variance σ_u^2 , and $\mathbf{w}_u \in \mathbb{C}^{N_c N_t \times 1}$ is the precoding vector for MS_u .

We consider PBPC in the optimization. Let \mathbf{B}_b be block-diagonal with block size N_t , where its b th block is \mathbf{I}_{N_t} and all the other blocks are $\mathbf{0}_{N_t}$. Then the power constraints per BS is expressed as

$$\sum_{u=1}^{N_u} \mathbf{w}_u^H \mathbf{B}_b \mathbf{w}_u \leq P \quad \forall b, \quad (2)$$

where $P > 0$ is the maximal transmit power of each BS.

B. Channel Model with Calibration Errors

Based on a popular antenna calibration method, antenna self-calibration, the uplink-downlink channels in CoMP systems are modeled as [4]

$$\mathbf{h}_{u,D} = \mathbf{G}_u \mathbf{h}_{u,U}, \quad (3)$$

where $\mathbf{h}_{u,U} = [\mathbf{h}_{u1,U}^T, \dots, \mathbf{h}_{uN_c,U}^T]^T$ is the uplink CoMP channel, $\mathbf{G}_u = \text{diag}\{\mathbf{g}_u\}$, $\mathbf{g}_u = [\mathbf{g}_{u1}^T, \dots, \mathbf{g}_{uN_c}^T]^T$, and $\mathbf{g}_{ub} = [g_{ub1}, \dots, g_{ubN_t}]^T$ represents the ambiguity factors between the uplink and downlink channels for BS_b and MS_u caused by calibration errors.

As analyzed in [4], self-calibration is readily to implement when the antennas are co-located, but is hard for CoMP systems. On the other hand, individual self-calibration within each BS will lead to different ambiguity factors at multiple coordinated BSs. We refer to the BS-wise ambiguity as ‘‘port error’’ which is denoted as $g_{ub}^{(1)}$. Theoretically, the ambiguity factors of all antennas in the same BS should be identical after the self-calibration, which however are actually time-varying since the analog gains of RF chains vary with temperature,

humidity, etc.. We refer to the time-varying ambiguity as ‘‘residual error’’ and denote it as $g_{ub}^{(2)}$. Then we can express the ambiguity factors between BS_b and MS_u as

$$\mathbf{g}_{ub} = g_{ub}^{(1)} [g_{ub1}^{(2)}, \dots, g_{ubN_t}^{(2)}]^T \triangleq g_{ub}^{(1)} \mathbf{g}_{ub}^{(2)}, \quad (4)$$

where $g_{ub}^{(1)}$ and $\mathbf{g}_{ub}^{(2)}$ are independent from each other, and both are usually modeled as random variables with log-uniformly distributed amplitudes and uniformly distributed phases [8].

Note that although the self-calibration is considered, the model in (3) and (4) is valid when other antenna calibration methods are applied. For example, for over-the-air calibration [7], the multicell channel estimation errors will lead to the port errors and the analog gain drift of RF chains will still lead to the residual errors.

III. OPTIMAL TRANSMISSION STRUCTURE FOR WEIGHTED SUM RATE MAXIMIZATION

In order to alleviate the performance degradation caused by the imperfect channel reciprocity, we resort to robust CoMP precoder design aimed at maximizing the weighted sum rate of multiple users. We assume perfect uplink channel estimation, i.e., $\mathbf{h}_{u,U}$ is known for $u = 1, \dots, N_u$. However, this does not mean a perfect knowledge of the downlink channel $\mathbf{h}_{u,D}$ at the BSs due to the imperfect channel reciprocity. As is commonly assumed for robust optimization, we assume *a priori* knowledge of statistics of the multiplicative ambiguity factors between the uplink and downlink channels.

A. Problem Formulation

The received power at MS_u of the signals from the BSs to MS_j can be obtained from (1) as

$$p_{uj} = |\mathbf{h}_{u,D}^H \mathbf{w}_j|^2. \quad (5)$$

Due to the imperfect channel reciprocity, the BSs do not know the value of p_{uj} . To design a robust precoder against the imperfect CSI, the BSs can estimate its value based on the uplink channel, the statistics of the uplink-downlink channel ambiguity, and the model in (3). The minimum mean square error (MMSE) estimate of p_{uj} can be obtained as

$$\hat{p}_{uj}^{\text{mmse}} = \arg \min_{\hat{p}_{uj}} \mathbb{E}_g \{ |\hat{p}_{uj} - p_{uj}|^2 \} = \mathbf{w}_j^H \mathbf{R}_u \mathbf{w}_j, \quad (6)$$

where $\mathbb{E}_g\{\cdot\}$ denotes the expectation with respect to calibration errors \mathbf{G}_u , $\mathbf{R}_u = \mathbb{E}_g\{\mathbf{G}_u \mathbf{h}_{u,U} \mathbf{h}_{u,U}^H \mathbf{G}_u^H\} = \mathbf{H}_{u,U} \mathbb{E}_g\{\mathbf{g}_u \mathbf{g}_u^H\} \mathbf{H}_{u,U}^H$, and $\mathbf{H}_{u,U} = \text{diag}\{\mathbf{h}_{u,U}\}$.

Based on (6), the signal-to-interference-plus-noise ratio (SINR) and the data rate of MS_u can be estimated as

$$\widehat{\text{SINR}}_u = \frac{\mathbf{w}_u^H \mathbf{R}_u \mathbf{w}_u}{\sum_{j \neq u} \mathbf{w}_j^H \mathbf{R}_u \mathbf{w}_j + \sigma_u^2} \quad \text{and} \quad \hat{R}_u = \log(1 + \widehat{\text{SINR}}_u).$$

The precoder design problem of maximizing the weighted estimated sum rate subject to PBPC can be formulated as

$$\max_{\{\mathbf{w}_u\}} \sum_{u=1}^{N_u} \alpha_u \hat{R}_u \quad (7a)$$

$$\text{s.t.} \quad \sum_{u=1}^{N_u} \mathbf{w}_u^H \mathbf{B}_b \mathbf{w}_u \leq P \quad \forall b, \quad (7b)$$

where the weights α_u , $u = 1, \dots, N_u$ reflect the priorities of different users.

Problem (7) is non-convex and generally NP-hard [6]. It is very difficult to find the globally optimal solution to the problem. In the following subsection, we strive for characterizing the structure of the optimal precoder, which will be exploited to develop efficient suboptimal solutions later.

B. Optimal Precoder Structure

As a first step to find the optimal precoder structure, we prove the equivalence of the weighted estimated sum rate maximization problem and the weighted sum MSE minimization problem, which has been established in [9–11] under the assumption of perfect knowledge of the downlink channel. To show the equivalence of the two problems when CSI is imperfect at the BSs, we start with deriving the estimated MSE of the data streams of the users based on the uplink channel and the statistics of calibration errors.

Denoting the receive filter at MS $_u$ as v_u , the MSE of MS $_u$'s data stream can be obtained from (1) as

$$\begin{aligned} \epsilon_u &\triangleq \mathbb{E}_{x,z} \{|v_u^* y_u - x_u|^2\} \\ &= 1 - 2\Re\{v_u^* \mathbf{h}_{u,D}^H \mathbf{w}_u\} + \left(\sum_{j=1}^{N_u} |\mathbf{h}_{u,D}^H \mathbf{w}_j|^2 + \sigma_u^2\right) |v_u|^2, \end{aligned} \quad (8)$$

where $\mathbb{E}_{x,z}\{\cdot\}$ is the expectation with respect to the data and noises.

With the uplink channel and the statistics of calibration errors, the MMSE estimate of $|\mathbf{h}_{u,D}^H \mathbf{w}_j|^2$ is $\mathbf{w}_j^H \mathbf{R}_u \mathbf{w}_j$, which is given in (6). Thus, we can obtain the estimate of $\mathbf{h}_{u,D}^H \mathbf{w}_u$ as $\sqrt{\mathbf{w}_u^H \mathbf{R}_u \mathbf{w}_u} e^{j\theta_u}$ with a proper phase θ_u . Note that the phase θ_u has no impact on the optimal precoder structure as shown by the following Theorem 2 and Property 1. Then the estimated MSE of MS $_u$'s data stream can be obtained as

$$\begin{aligned} \hat{\epsilon}_u &= 1 - 2\Re\{v_u^* \sqrt{\mathbf{w}_u^H \mathbf{R}_u \mathbf{w}_u} e^{j\theta_u}\} \\ &\quad + \left(\sum_{j=1}^{N_u} \mathbf{w}_j^H \mathbf{R}_u \mathbf{w}_j + \sigma_u^2\right) |v_u|^2. \end{aligned}$$

Theorem 1. Introduce the auxiliary variable $t_u \geq 0$ as a scalar weight for the MSE of MS $_u$'s data stream. The following weighted sum MSE minimization problem

$$\min_{\mathbf{w}, v, t} \sum_{u=1}^{N_u} \alpha_u (t_u \hat{\epsilon}_u - \log t_u) \quad (9a)$$

$$\text{s.t.} \quad \sum_{u=1}^{N_u} \mathbf{w}_u^H \mathbf{B}_b \mathbf{w}_u \leq P \quad \forall b \quad (9b)$$

is equivalent to the weighted estimated sum rate maximization problem (7), in a sense that the two problems have identical global optimal precoders \mathbf{w}^{opt} ¹.

For space limitations, the proof of Theorem 1 is given in [12]. With the equivalence between the two problems, we next examine the structure of the optimal precoder.

Theorem 2. The optimal precoder that maximizes the weighted estimated sum rate subject to PBPC has the following structure

$$\mathbf{w}_u^{\text{opt}} = \sqrt{q_u^{\text{opt}}} \mathbf{f}_u^{\text{opt}} \quad \forall u, \quad (10)$$

¹In (9), the notation \mathbf{w} is short for $\{\mathbf{w}_u\}$, which denotes variables \mathbf{w}_u for all u . The notations $v = \{v_u\}$, $t = \{t_u\}$, $\mathbf{w}^{\text{opt}} = \{\mathbf{w}_u^{\text{opt}}\}$, $v^{\text{opt}} = \{v_u^{\text{opt}}\}$, $t^{\text{opt}} = \{t_u^{\text{opt}}\}$, and $\lambda^{\text{opt}} = \{\lambda_b^{\text{opt}}\}$ are defined similarly in the following.

where $\mathbf{f}_u^{\text{opt}}$ is the eigenvector corresponding to the largest eigenvalue of the following matrix

$$\alpha_u \kappa_u \left(\sum_{j \neq u} \alpha_j \kappa_j \mathbf{R}_j + \sum_{b=1}^{N_c} \nu_b \mathbf{B}_b \right) \dagger \mathbf{R}_u \quad \forall u \quad (11)$$

with parameters $\kappa_u, \nu_b \in [0, 1]$ for all $u \in \mathcal{U}$ and $b \in \mathcal{B}$, and

$$[q_1^{\text{opt}}, \dots, q_{N_u}^{\text{opt}}]^T = \mathbf{\Sigma}^{-1} [\sigma_1^2 d_1, \dots, \sigma_{N_u}^2 d_{N_u}]^T \quad (12)$$

with d_u denoting the largest eigenvalue of the matrix (11) and $\mathbf{\Sigma}$ defined as

$$[\mathbf{\Sigma}]_{uj} = \begin{cases} \mathbf{f}_u^{\text{opt},H} \mathbf{R}_u \mathbf{f}_u^{\text{opt}}, & j = u, \\ -d_u \mathbf{f}_j^{\text{opt},H} \mathbf{R}_u \mathbf{f}_j^{\text{opt}}, & j \neq u. \end{cases} \quad (13)$$

Here, $[\mathbf{\Sigma}]_{uj}$ denotes the element of $\mathbf{\Sigma}$ at the u th row and j th column for $u, j \in \mathcal{U}$.

The proof of Theorem 2 is given in [12]. From the proof, we can obtain the following properties with respect to the parameters κ_u and ν_b , which will be used to develop a closed-form suboptimal precoder in next section.

Property 1. Denote t^{opt} , v^{opt} and λ^{opt} as the optimal solutions and optimal Lagrange multipliers corresponding to PBPC for problem (9). Then the optimal κ_u and ν_b can be expressed as

$$\kappa_u = t_u^{\text{opt}} |v_u^{\text{opt}}|^2 / c \quad \text{and} \quad \nu_b = \lambda_b^{\text{opt}} / c, \quad (14)$$

where t^{opt} and v^{opt} are functions of the optimal precoder \mathbf{w}^{opt} ,

$$t_u^{\text{opt}} = \frac{\sum_{j=1}^{N_u} \mathbf{w}_j^{\text{opt},H} \mathbf{R}_u \mathbf{w}_j^{\text{opt}} + \sigma_u^2}{\sum_{j \neq u} \mathbf{w}_j^{\text{opt},H} \mathbf{R}_u \mathbf{w}_j^{\text{opt}} + \sigma_u^2} \quad \forall u, \quad (15)$$

$$v_u^{\text{opt}} = \frac{\sqrt{\mathbf{w}_u^{\text{opt},H} \mathbf{R}_u \mathbf{w}_u^{\text{opt}} e^{j\theta_u}}}{\sum_{j=1}^{N_u} \mathbf{w}_j^{\text{opt},H} \mathbf{R}_u \mathbf{w}_j^{\text{opt}} + \sigma_u^2} \quad \forall u, \quad (16)$$

and $c = \max(\max_u t_u^{\text{opt}} |v_u^{\text{opt}}|^2, \max_b \lambda_b^{\text{opt}})$. It is worth noting that the scalar c does not change the matrix in (11), which is only for ensuring the values of $\kappa_u, \nu_b \in [0, 1]$.

Property 2. The estimated SINR of MS $_u$ with the optimal precoders \mathbf{w}^{opt} equals to the largest eigenvalue of the matrix shown in (11). Thus, it is easy to see that the estimated SINR is a decreasing function of ν_b (or λ_b^{opt} considering (14)) $\forall b$ and an increasing function of κ_u (or $t_u^{\text{opt}} |v_u^{\text{opt}}|^2$).

Property 3. The optimal Lagrange multipliers λ^{opt} satisfy

$$\sum_{b=1}^{N_c} \lambda_b^{\text{opt}} \leq \frac{\sum_{u=1}^{N_u} \alpha_u}{P}, \quad (17)$$

which is proved in [12].

A similar optimal precoder structure governed by $N_c + N_u$ parameters has recently been derived in [6], which is however very different from ours. On one hand, perfect CSI at the BSs is considered in [6], while we consider imperfect CSI with multiplicative noises. On the other hand, [6] derives the results based on a framework of uplink-downlink duality, while we exploit the equivalence between the weighted estimated sum rate maximization problem and the weighted sum MSE minimization problem, although the proposed optimal precoder structure also has $N_c + N_u$ parameters.

Remark: The proposed method to obtain the optimal precoder structure is applicable for the scenarios with partial data

sharing among coordinated BSs. To this end, we can first express the precoding vector as $\mathbf{w}_u = [\mathbf{w}_{u1}^T, \dots, \mathbf{w}_{uN_c}^T]^T$ with \mathbf{w}_{ub} denoting the precoder at BS_{*b*} for MS_{*u*}, then optimize the weighted sum MSE minimization problem (9) with additional constraints $\mathbf{w}_{ub} = \mathbf{0}$ if BS_{*b*} does not have the data of MS_{*u*}. Further details can be found in [12].

IV. CLOSED-FORM MULTICELL PRECODER

Since the weighted estimated sum rate maximization problem with imperfect CSI under PBPC is NP-hard, it is not feasible in practice to optimally select the parameters involved in the structured precoder. In this section, we will propose a closed-form precoder with properly selected parameters according to the properties developed in Section III-B.

According to Property 1, the scalar normalization factor c does not affect the optimal precoder. Thus, we can obtain the parameters κ_u and ν_b by equivalently finding $t_u^{\text{opt}}|v_u^{\text{opt}}|^2$ and λ_b^{opt} as shown in (14).

We begin with examining the selection of $t_u^{\text{opt}}|v_u^{\text{opt}}|^2$, which can be expressed based on (15) and (16) as

$$t_u^{\text{opt}}|v_u^{\text{opt}}|^2 = \frac{1}{\underbrace{\sum_{j \neq u} \mathbf{w}_j^{\text{opt},H} \mathbf{R}_u \mathbf{w}_j^{\text{opt}} + \sigma_u^2}_{\text{Signal power}} + \underbrace{\mathbf{w}_u^{\text{opt},H} \mathbf{R}_u \mathbf{w}_u^{\text{opt}} + \sum_{j \neq u} \mathbf{w}_j^{\text{opt},H} \mathbf{R}_u \mathbf{w}_j^{\text{opt}} + \sigma_u^2}_{\text{Interference power}}}. \quad (18)$$

The coherent CoMP transmission can enhance the signal strength and eliminate the inter-cell and inter-user interference simultaneously. This indicates that the optimal precoder can provide high signal power and low interference power, which makes the following approximation of $t_u^{\text{opt}}|v_u^{\text{opt}}|^2$ reasonable

$$t_u^{\text{opt}}|v_u^{\text{opt}}|^2 \approx \frac{1}{\sigma_u^2} \quad \forall u. \quad (19)$$

Such an approximated value of $t_u^{\text{opt}}|v_u^{\text{opt}}|^2$ decreases with the noise variance σ_u^2 . It is consistent with Property 2, which tells us that the estimated SINR increases with the growth of $t_u^{\text{opt}}|v_u^{\text{opt}}|^2$.

One shortcoming of this approximation is that the values of $t_u^{\text{opt}}|v_u^{\text{opt}}|^2$ are strictly positive for all users, while we can predict that the users with poor channel quality will not be served with the optimal precoder, i.e., the corresponding $t_u^{\text{opt}}|v_u^{\text{opt}}|^2$ become zeros. This can be circumvented by employing user schedulers together with the proposed precoder, i.e., the parameters $t_u^{\text{opt}}|v_u^{\text{opt}}|^2$ are selected as (19) for the scheduled users, and as zeros for the other users. Let $\mathcal{S} \subseteq \mathcal{U}$ denote the set of scheduled users.

We then consider the selection of λ_b^{opt} . Property 3 gives an upper bound of $\sum_{b=1}^{N_c} \lambda_b^{\text{opt}}$ that is $\sum_{j \in \mathcal{S}} \alpha_j / P$, based on which we can select its value as²

$$\lambda_b^{\text{opt}} = \frac{\sum_{j \in \mathcal{S}} \alpha_j}{N_c P} \quad \forall b, \quad (20)$$

²Herein, since only users in \mathcal{S} are served, the term $\sum_{u=1}^{N_u} \alpha_u$ in Property 3 is accordingly replaced by $\sum_{j \in \mathcal{S}} \alpha_j$.

which is a decreasing function of the transmit power P . This is consistent with Property 2 because the estimated SINR decreases with the growth of λ_b^{opt} .

With the selected parameters, the matrix in (11) can be rewritten as

$$\alpha_u \left(\sum_{j \neq u} \alpha_j \frac{\sigma_u^2}{\sigma_j^2} \mathbf{R}_j + \frac{\sigma_u^2 \sum_{j \in \mathcal{S}} \alpha_j}{N_c P} \mathbf{I}_{N_c N_t} \right)^\dagger \mathbf{R}_u \quad \forall u \in \mathcal{S}. \quad (21)$$

Therefore, the precoding vector of MS_{*u*} can be expressed as

$$\mathbf{w}_u = \sqrt{q_u} \mathbf{f}_u, \quad (22)$$

where the beamforming vector \mathbf{f}_u and the allocated power q_u can be computed according to Theorem 2.

Since the parameters are selected in a suboptimal way, the obtained precoders may not satisfy PBPC. We can ensure PBPC by scaling the precoding vectors of all users as

$$\mathbf{w}_u^{\text{RSLNR}} = \sqrt{\rho} \mathbf{w}_u \quad (23)$$

with $\rho = P / \max_b \sum_{u \in \mathcal{S}} \mathbf{w}_u^H \mathbf{B}_b \mathbf{w}_u$.

When all the users have the same weights of priority and noise variances, i.e., $\alpha_{\mathcal{S}(1)} = \dots = \alpha_{\mathcal{S}(|\mathcal{S}|)}$ and $\sigma_{\mathcal{S}(1)}^2 = \dots = \sigma_{\mathcal{S}(|\mathcal{S}|)}^2$, the matrix in (21) can be simplified as

$$\left(\sum_{j \neq u} \mathbf{R}_j + \frac{\sigma_u^2 |\mathcal{S}|}{N_c P} \mathbf{I}_{N_c N_t} \right)^\dagger \mathbf{R}_u \quad \forall u \in \mathcal{S}. \quad (24)$$

In this case, to find the beamforming vector \mathbf{f}_u of the proposed closed-form precoder (i.e., the eigenvector corresponding to the largest eigenvalue of the matrix in (24)), is equivalent to solve the following Rayleigh quotient maximization problem

$$\max_{\mathbf{f}_u} \frac{\mathbf{f}_u^H \mathbf{R}_u \mathbf{f}_u}{\mathbf{f}_u^H \sum_{j \neq u} \mathbf{R}_j \mathbf{f}_u + \frac{\sigma_u^2 |\mathcal{S}|}{N_c P}}, \quad (25)$$

where the objective can be regarded as the signal-to-leakage-plus-noise ratio (SLNR) of MS_{*u*}.

Therefore, we call the proposed closed-form precoder robust SLNR (RSLNR) precoder.

V. SIMULATION RESULTS

In this section, we evaluate the performance of the proposed RSLNR precoder.

The calibration errors consist of port errors and residual errors. In simulations, we put emphasis on analyzing the impact of port phase errors on the performance of CoMP systems. This is because the residual errors are generally much smaller than the port errors, and the port phase errors will prevent the co-phasing of coherent CoMP transmission and affect the performance of CoMP systems severely. To this end, we introduce a parameter θ and model the phases of the port errors as uniformly distributed random variables within $[-\theta, \theta]$. The phases of the residual errors are modeled as uniformly distributed random variables within $[-10^\circ, 10^\circ]$ [8], and the amplitudes of the port errors and the residual errors are modeled as log-uniformly distributed random variables within $[-3, 3]$ dB and $[-1, 1]$ dB [8], respectively. In addition, it is assumed that the residual errors are independent among

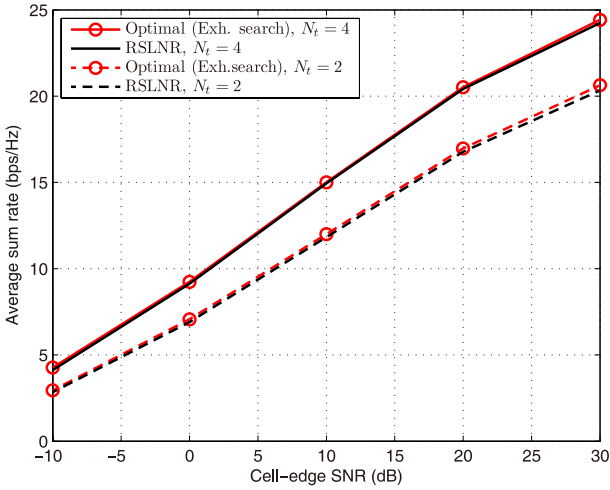


Fig. 1. Average sum rate with RSLNR precoder and the optimal precoder through exhaustive searching versus cell-edge SNR. $N_c = 2$ and $N_u = 2$.

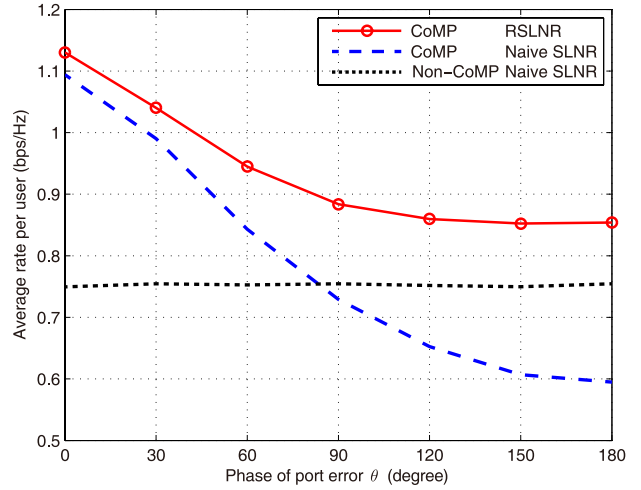


Fig. 2. Average data rate per user with the three considered strategies versus θ . $N_c = 3$ and $N_u = 30$.

all antennas in all coordinated BSs and the port errors are independent among different BSs.

We consider the network layout consisting of three multiple-antenna BSs, which cooperate coherently to serve multiple users. The cell radius r is set to 250 m, and the interference from non-cooperative cells is modeled as white noise. Denoting the average receive SNR of users located at the cell boundary as SNR_{edge} , then the average receive SNR of a user from a BS with distance d is computed as $\text{SNR}_{\text{edge}} + 37.6 \log_{10}(\frac{d}{r}) + \chi$, where χ represents log-normal shadowing with the standard deviation of 8 dB, and $d > 50$ m. The i.i.d. small-scale Rayleigh flat fading downlink channels and perfect uplink channel estimation are considered. The weights α_u are set equal for all users.

First, we evaluate the performance of the proposed RSLNR precoder, where the widely applied greedy user scheduler (GUS) [13] is employed to achieve multiuser diversity gain. For comparison, the performance of the “optimal” precoder is shown, which is obtained through exhaustive searching over $N_c + N_u$ parameters to maximize the weighted estimated sum rate.³ Due to the high complexity of exhaustive searching, we consider the case with $N_c = 2$ and $N_u = 2$. In Fig. 1, the average sum rate of the RSLNR precoder with GUS and the optimal precoder versus cell-edge SNR is depicted with $\theta = 120^\circ$ for $N_t = 2, 4$. As seen in the figure, the RSLNR precoder with GUS performs very close to the optimal precoder in all considered scenarios, and has no sum rate floor in high SNR region despite that the imperfect CSI at the BSs is considered.

Next, we simulate a scenario where 30 users are uniformly distributed in the three cells, to show the performance gain

³Finding the exact optimal precoder by exhaustive searching is impractical, since the parameters are in $[0, 1]$ whose number is infinite. In the simulations, the exhaustive search is performed over 100 points within properly refined search interval for each parameter. The difference between the results of exhaustive search and the real optimal precoder is hard to quantify since the optimization problem is NP-hard.

of the proposed precoder. Except for the proposed RSLNR precoder, an SLNR precoder [14] that simply regards the uplink channel as the downlink channel without considering the imperfect channel reciprocity is also simulated for both CoMP systems and Non-CoMP systems,⁴ which is denoted as the naive SLNR precoder. To ensure fairness among users which is critical for CoMP systems, we apply the GUS in a Round-Robin (RR) fashion similar to [15], which is used together with all the precoders. Since each user is only served once during a scheduling period, the obtained user data rate is normalized by the RR scheduling period.

In Fig. 2, the average data rate per user with the three considered strategies is shown versus the statistics of phases of the port error θ . Compared with Non-CoMP systems, CoMP systems do not always perform better, which depends on the extent of the imperfect reciprocity between uplink and downlink channels. With accurate antenna calibration (i.e., for small θ), the cooperation of multiple BSs can significantly improve the system performance. Yet with large antenna calibration errors and the naive SLNR precoder, the performance of CoMP systems is even inferior to that of Non-CoMP systems. By using the proposed RSLNR precoder, an evident performance gain can be observed over the naive SLNR precoder for both CoMP and Non-CoMP systems.

Finally, we analyze the impact of data sharing on the performance of CoMP systems when the calibration errors are different. In practice, the amount of data shared among the coordinated BSs depends on the implementation costs and the achievable performance gain. In the simulations, we consider a simple data sharing strategy, with which the data of a user are shared among the BSs who have large average channel gains to the user. Specifically, for MS_u we denote the average channel gain from its master BS as α_{u0} in dB; then BS_b will

⁴In Non-CoMP systems, the user is only served by its master BS (the BS that provides the maximum receive power), and the inter-cell interference exists.

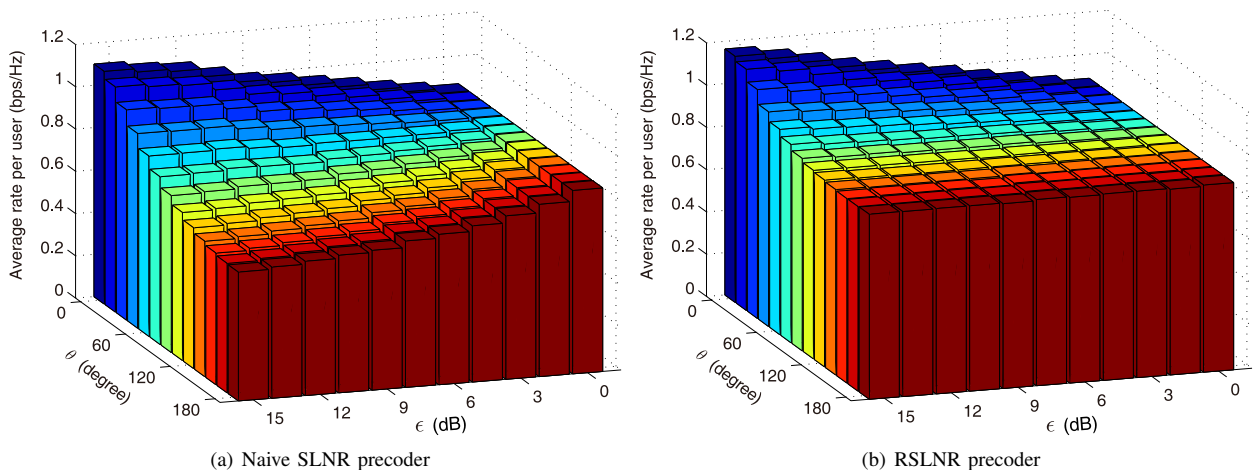


Fig. 3. Average data rate per user with the RSLNR precoder and the naive SLNR precoder as a function of data sharing threshold ϵ and port phase errors θ . $N_c = 3$ and $N_u = 30$.

be shared with the data of MS_u if the average channel gain α_{ub} from BS_b satisfies $\alpha_{u0} - \alpha_{ub} \leq \epsilon$, where ϵ is a pre-determined threshold. It is easy to see that $\epsilon = 0$ means no data sharing (then the CoMP is coordinated beamforming), while $\epsilon = +\infty$ means full data sharing. Fig. 3(a) plots the average data rate per user with the naive SLNR precoder as a function of data sharing threshold ϵ and port phase errors θ , where the cell-edge SNR is 10 dB and $N_t = 4$. It is shown that the performance does not necessarily increase when more data are shared (i.e., when ϵ grows from 0 dB to 15 dB). For large θ , the performance degrades with the growth of ϵ . This is because the imperfect CSI turns the desired signals from the coordinated BSs to each user into the inter-cell inference. When the RSLNR precoder is applied, as shown in Fig. 3(b), more data sharing is always beneficial but the performance gain decreases when the calibration errors increase.

VI. CONCLUSIONS

In this paper, we developed precoders to alleviate the performance degradation caused by imperfect antenna calibration in TDD CoMP systems. With the knowledge of the statistics of antenna calibration errors, we derived the optimal precoder structure by establishing the equivalence between the weighted estimated sum rate maximization problem and the weighted sum MSE minimization problem. By properly selecting the involved parameters, we further proposed a closed-form robust linear precoder, which provides near-optimal performance via applied together with user schedulers as shown by simulations. With non-ideal reciprocity, CoMP does not always outperform Non-CoMP and data sharing among the coordinated BSs is not always beneficial, depending on the employed precoders and the accuracy of antenna calibration.

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