Linear transceiver design in uplink coordinated multipoint multiple-input multiple-output systems

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Abstract: The authors investigate linear transceiver design in uplink (UL) coordinated multipoint transmission and reception (CoMP) multiple-input multiple-output (MIMO) systems with joint detections. A two-stage design algorithm is proposed by exploiting the technique of interference alignment, optimising the structure of the effective channel and employing power loading, with the goal to achieve high throughput and convergence performance. In contrast to conventional CoMP transceiver design, which is investigated under a predefined number of data streams transmitted by each user, the authors further investigate the selection of the number of data streams, called the configuration selection, and propose a corresponding low-complexity algorithm. By combining the proposed linear transceiver algorithm and low-complexity configuration selection algorithm, this work presents a new practical framework for linear transceiver design in UL CoMP MIMO systems. The simulation results confirm that the proposed algorithms achieve higher sum-rate performance than prior linear transceivers used in UL CoMP MIMO systems. Furthermore, the proposed transceiver offers comparable performance to existing IA-aided transceivers with significantly faster convergence.

1 Introduction

In wireless cellular communications, the growing demand on data rate and scarcity of spectrum bring great challenges to the research and industrial communities. High spectral efficiency with high throughput for both cell average and edge users is crucial. Techniques of advanced multiple-input multiple-output (MIMO) and spectrum reuse have been substantially studied as a means to improve spectral efficiency and throughput. The challenges remain where the interference among base stations (BSs) restricts the performance of MIMO techniques and efficiency of spectrum reuse. To mitigate the interference among BSs, the coordinated multipoint transmission and reception (CoMP) technique has been introduced and widely investigated [1–3]. In wireless standards such as long-term evolution-advanced (LTE-A) aiming for the fourth generation mobile communications developed by Third Generation Partnership Project (3GPP), CoMP has been officially adopted as a key enabling technique [4–6].

The CoMP technique aims to improve cell average and edge throughput in interference-limited scenarios by exploiting the coordination of BSs. In uplink (UL) CoMP systems, two reception schemes are adopted: joint scheduling and joint detection. Joint scheduling jointly allocates the resources of cooperating BSs to eliminate or reduce interference between cells so as to enhance channel quality; joint detection jointly receives and decodes the data streams from users belonging to the cooperating BSs. Joint detection with multiple antennas on both BSs and users is referred to as the centralised UL CoMP MIMO system. A centralised UL CoMP MIMO system coordinates multiple BSs to alleviate inter-cell interference (ICI), and jointly decodes data symbols by exploiting the spatial degrees-of-freedom (DoFs) and exchanging full channel state information (CSI) and data [7, 8].

The zero-forcing (ZF) and minimum mean square error (MMSE) receivers are among the most commonly used receivers in the centralised UL CoMP MIMO system. Ideally, the centralised UL CoMP MIMO system can be viewed as a large multiple access channel (MAC), where some transceiver design algorithms can be readily applied. In MAC, there are two major criteria for jointly designing transmitters and receivers; one is to maximise sum capacity [9, 10], whereas the other is to minimise the MSE between transmitted symbols and received symbols [11, 12]. In this paper, we focus on transceiver design based on the maximisation of sum capacity. In [9], an algorithm called iterative water-filling to maximise the MAC capacity by jointly designing the transmitters at users is proposed. In [10], the optimality of these transmitters in the case where the number of transmit antennas is larger than the number of receive antennas is investigated. Since the ML and MMSE receivers with successive interference cancellation (MMSE-SIC) are known to be both nearly capacity achieving [13], the sum capacity maximising transceiver can be obtained by combining the transmitters designed by iterative water-filling and ML/MMSE-SIC receivers. Nevertheless, both receivers are nonlinear and challenges still remain. The use of the ML receiver demands high
computational complexity [14], whereas MMSE-SIC receiver causes a large delay and capacity degradation due to SIC processing and error propagation [15]. Moreover, the delay and degradation become more severe as the number of transmitted data streams increases. In a centralised UL CoMP MIMO system, the delay and capacity degradation is particularly hazardous due to the need of handling a large number of data streams.

The linear receiver is advantageous in its low complexity and short delay in decoding the data streams. To design a linear transceiver, conventional linear receivers such as ZF/MMSE receivers can be used with the transmitters designed by iterative water-filling. However, the ZF/MMSE receivers provide limited performance due to the inherent capacity gap [16], even with well-designed transmitters. By exploiting the technique of interference alignment (IA) [17-20], linear transceiver design algorithms are proposed to maximise the achievable sum-rate [21]. These IA-aided algorithms prove effective in enhancing the sum-rate performance, while requiring a large number of iterations to converge to a good solution. The high complexity in the iterations remains to be a critical challenge.

Motivated by the above, we propose a linear transceiver design algorithm for the centralised UL CoMP MIMO system by exploiting the underlying concept of IA. We explore the matrix structure in the system formulation and design a two-stage algorithm that mitigates different types of interference separately to improve the convergence rate. By incorporating appropriate power allocation, the sum-rate performance of the proposed transceiver can be further enhanced. Since the overall performance of the transceiver depends largely on the number of data streams transmitted by each user, we investigate the optimal assignment of data streams to users, called configuration selection, and propose a corresponding low-complexity selection algorithm. The combination of the proposed highly efficient linear transceiver algorithm and low-complexity configuration selection algorithm forms a new framework for linear transceiver design in UL CoMP MIMO systems. The efficacy of the proposed design algorithms is evaluated by computer simulations in two different UL CoMP MIMO systems to confirm its advantages over existing CoMP and IA-aided linear transceivers.

The feasibility of the system model considered in this work has been widely discussed in the industrial standards [6]. As shown in Fig. 1, scenarios in which BSs are connected with dedicated fibre links to exchange CSI and data to form a CoMP system have been considered and defined in the LTE-A standard (CoMP scenarios 3 and 4). For these scenarios, there exists a BS serving as the central unit, with two remote radio heads to assist data reception. Information including data and CSI is collected at the central unit for decoding the desired signals and designing the transceivers. As a final remark, although the proposed algorithm can also be used in a centralised downlink (DL) CoMP system, the requirement of full CSI feedback from all users would limit the feasibility of such implementation.

This paper is organised as follows: the rationale of adopting linear receivers, model of the centralised UL CoMP MIMO system and formulation of achievable sum-rate are described in Section 2. The adoption of IA to the transceiver design leading to the proposed transceiver design algorithm is elaborated in Section 3. In Section 4, the configuration selection problem is formulated and a near-optimal low-complexity algorithm is provided. Finally, numerical results and conclusions are given in Sections 5 and 6, respectively.

Notations: Boldface capital and lowercase symbols represent matrix and column vectors, respectively. For example, \( \mathbf{X} \in \mathbb{C}^{M \times 1} \) and \( \mathbf{X} \in \mathbb{C}^{M \times N} \) represent an \( M \times 1 \) complex vector and an \( M \times N \) complex matrix, respectively. \( \mathbf{I}_M \) and \( \mathbf{0}_{M \times N} \) represent an \( M \times M \) identity matrix and an \( M \times N \) matrix with all elements equal to zero. \( \mathbf{X} = \text{diag}(X_1, \ldots, X_k) \) is a block diagonal matrix composed of \( X_1, \ldots, X_k \) as the diagonal elements. \( \mathcal{CN}(\mathbf{0}, N_0 \mathbf{I}_M) \) represents the distribution of a circularly symmetric additive white Gaussian noise (AWGN) vector \( \mathbf{x} \in \mathbb{C}^{M \times 1} \) with zero mean and covariance matrix \( N_0 \mathbf{I}_M \). \( \mathbf{X}^T \) and \( \mathbf{X}^{-1} \) represent the transpose, conjugate transpose and inverse of a matrix \( \mathbf{X} \), respectively; \( \mathbf{X}^{(0)}, \mathbf{X}^{(1)}, \{ \mathbf{X}^{(j)} \}, \text{tr}(\mathbf{X}) \) and \( \| \mathbf{X} \|_F \) represent the \( i \)th diagonal element, the \( i \)th column, \( i \)th column to \( j \)th column, trace and the Frobenius norm of a matrix \( \mathbf{X} \), respectively. We denote a scaling function \( S(\mathbf{X}, p) = \sqrt{p} \mathbf{X}/\| \mathbf{X} \|_F \) to scale the Frobenius norm of \( \mathbf{X} \) to \( p \).

## 2 Signal model and performance metric

In this section, we first elaborate the advantages of using a linear receiver over non-linear receivers. Then a centralised UL CoMP MIMO system model with a linear transceiver is described. Finally, the achievable sum-rate used as the design criterion is given.

### 2.1 Rationale of adopting linear receivers

The ML and MMSE-SIC receivers are well known hard-decision non-linear receivers. Under proper conditions, they can be capacity-achieving. However, the ML receiver causes extremely high complexity, while the MMSE-SIC receiver causes a delay which increases linearly with the number of transmitted data streams due to the SIC process.

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**Fig. 1** Illustration of centralised UL CoMP scenario
Even though the MMSE-SIC receiver is nearly capacity-achieving, this would hold only if error propagation can be properly reduced in the SIC process. In both information-theoretical and detection architecture perspectives, errors occur inevitably in a slow fading channel without effective instantaneous rate and power control leading to occurrences of outage [13, 22, 23]. Once error occurs in a stage in SIC, the erroneous cancellation in the subsequent stages leads to error propagation, which hugely degrades the performance. Moreover, it can be observed that error propagation is more severe with a larger number of transmitted data streams [15, 24, 25].

The rationale behind the adoption of linear receivers in this work is to achieve low complexity, low latency and high robustness in the system design. It is noted that the robustness property which prevents error propagation is particularly crucial due to the large system size of the considered centralised UL CoMP MIMO system.

### 2.2 Centralised UL CoMP MIMO system model

As shown in Fig. 2, a centralised UL CoMP MIMO system model includes BSs (M cells), each equipped with \( N_t \) antennas. Each cell is equipped with \( N_t \) antennas, and each cell has \( K \) users for simplicity. This can be easily extended to the case where each cell has a different number of users. The transmitted signal vector described by \( x_l \) belonging to the \( l \)th cell is denoted as \( \{ x_l(1), \ldots, x_l(1+K) \} \in \mathbb{C}^{N_t \times 1} \); the aggregated transmitted signal at the \( l \)th cell is denoted as \( x_l = \left[ (x_l(1-K+1)), \ldots, (x_l(1+K)) \right] ^T \in \mathbb{C}^{MK \times 1} \); and the aggregated number of transmitted data streams at the \( l \)th cell is denoted as \( d_l = \sum_{q=1}^{MK} d_l(q) \). The channel matrix between the \( m \)th BS and all users in the \( l \)th cell is denoted as \( H_{m,l} = [H_{m,l}^1, H_{m,l}^2, \ldots, H_{m,l}^{N_r}] \in \mathbb{C}^{N_r \times N_t} \); the noise vector at the \( m \)th BS is denoted as \( z_m = \mathbb{C}^{N_t \times 1} \) with distribution \( CN(0, N_r I_{N_t}) \). Therefore, the total dimension of the transmission system is \( d_l = \sum_{q=1}^{MK} d_l(q) \). Note that the transmit power for each user is restricted to \( P_{UPW} \), that is

\[
\| V_{(l-1)K+q} x_{(l-1)K+q} \|^2 = P_{UPW}
\]

In a centralised UL CoMP system, the coordination is conducted at a central unit (CU). The CU collects the received signals from all BSs and jointly processes the signals. By stacking the received signals into a vector, the received signal in CU is expressed as

\[
y_{\text{CoMP}} = H V x + z
\]

where \( y_{\text{CoMP}} = [y_1^T, y_2^T, \ldots, y_M^T] \) is the aggregated received signal; \( V = \text{diag}(V_1, V_2, \ldots, V_M) \) is the aggregated precoding matrix; and \( x = [x_1^T, x_2^T, \ldots, x_M^T] \) is the aggregated transmit vector. The aggregated channel is described as

\[
H = \left[ [H_{1,1}, \ldots, H_{1,M}]^T, \ldots, [H_{M,1}, \ldots, H_{M,M}]^T \right]^T
\]

and the aggregated noise vector is \( z = [z_1^T, z_2^T, \ldots, z_M^T] \).

To further utilise the available degrees of freedom (DoFs) to mitigate interference, the received signal processed through a decoder \( U^H \in \mathbb{C}^{d_l \times N_t} \) yields

\[
y_{\text{LA}} = U^H y_{\text{CoMP}} = U^H H V x + U^H z = H_{\text{eff}} x + \tilde{z}
\]

where \( H_{\text{eff}} = U^H H V \in \mathbb{C}^{d_l \times d_l} \) is the effective channel matrix composed of the decoder, ordinary physical channel

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**Fig. 2** Illustration of centralised UL CoMP system model
and the precoder. The \( \hat{z} = U^H z \) is the effective noise after the decoder.

Finally, the CU processes the received signal as

\[
\hat{x} = F^H y_{IA} = F^H H_{\text{eff}} x + F^H \hat{z}
\]

(5)

where \( \hat{x} \in \mathbb{C}^{d_1 \times 1} \) is the estimated signal by coherent joint reception, and \( F \in \mathbb{C}^{d_1 \times d_1} \) is the equalising matrix for coherent joint reception, which is typically in the form of minimum mean squared error (MMSE) [6–8]

\[
F^H = \left[ H_{\text{eff}}^{-1} C_n^{-1} (H_{\text{eff}} + I) \right]^{-1} H_{\text{eff}}^{-1} C_n^{-1}
\]

(6)

where \( C_n = N_0 U^H U \) is the covariance of noise \( \hat{z} \). With the precoder/decoder designed to mitigate interference, the adoption of the MMSE receiver after the decoder would effectively improve the transceiver performance in the low SNR regime.

### 2.3 Achievable sum-rate

The proposed work aims to jointly design the linear precoding matrices and decoding matrix in the system with the objective of maximising the UL sum capacity. The sum capacity with linear receiver can be explicitly expressed as the achievable sum-rate [16], which in the centralised UL CoMP MIMO system after the decoding matrix can be expressed as

\[
R_{\text{sum}} = \sum_{d=1}^{d_f} \left[ \log_2(1 + \text{SINR}_d) \right]
\]

(7)

where \( \text{SINR}_d \) is the signal-to-noise-plus-interference power ratio of the \( d \)-th data stream after decoding at the receiver, which is given by

\[
\text{SINR}_d = \frac{(F_d^H H_{\text{eff}}^{(d)} H_{\text{eff}}^{(d)} H_d^H)}{(F_d^H H_{\text{eff}}^{(d)} H_{\text{eff}}^{(d)} H_d^H N_0 + U^H U) F_d^H}.
\]

(8)

Note that the difference between achievable sum-rate and sum capacity is that the achievable sum-rate in (7) and (8) incorporates the effect of receiver, which renders performance evaluation more explicit.

### 3 Transceiver design

In this section, we propose a linear transceiver design algorithm aiming to outperform existing linear transceivers in the centralised UL CoMP MIMO system through exploiting the techniques of IA. As IA has not been widely applied to the centralised UL CoMP MIMO system, we first elaborate on the incorporation of IA techniques in the considered setting. Then, a detailed description of the proposed algorithm is given.

#### 3.1 Interference alignment in UL CoMP systems

The basic principle of IA in the UL CoMP system is to suppress interference onto a lower-dimensional subspace, such that each receiver can adopt more effective DoFs to better decode the desired signal. The ideal optimal design criteria of IA in the UL CoMP systems with regards to the \( k \)-th user \( \forall k \in \{1, 2, \ldots, MK\} \) can be described as [17]

\[
\left( \left[ U \right]_{d_k l} \right)^H H_k V_l = 0_{d_k \times d_l}, \quad \forall l \neq k
\]

(9)

where \( H_k = \{ H_k \}^{N_k}_{(k-1)N_k+1} \) and the location index of decoding vector corresponding to the \( d \)-th layer belonging to the \( k \)-th user within \( U \) is denoted as \( d_{k l} = \sum_{i=1}^{k-1} d_i + d_l \).

In the IA-aided UL CoMP systems, the CU attempts to design the decoder and precoder for each user with the goal to conform to constraints in (9), and sends back the corresponding precoder to each user. Since constraints in (9) may not be always feasible, a more plausible approach to solving (9) would be to alternatively maximise SINR of each layer at the output of the decoder. In an IA-aided UL CoMP system, the maximum SINR algorithm in [18] can be reformulated as

\[
(U, V_1, \ldots, V_{d_P}) = \arg \max_{U, V_1, \ldots, V_{d_P}} \text{SINR}_{k,d}.
\]

(10)

The \( \text{SINR}_{k,d} \) is the SINR of the \( d \)-th layer belonging to the \( k \)-th user at the output of the decoder, described as

\[
\text{SINR}_{k,d} = \frac{(U(U_{d_k d})^H H_k V_l (V_l^H H_k^H U_{d_k d}))}{(U(U_{d_k d})^H H_k V_l (V_l^H H_k^H U_{d_k d}))}
\]

(11)

where

\[
B_{k,d} = \sum_{l \neq d} H_l V_l (V_l^H H_l^H + N_0 I_{N_l})
\]

(12)

Then, the optimal decoding vector \( U(U_{d_k d}) \) maximising \( \text{SINR}_{k,d} \) can be formulated as

\[
U(U_{d_k d}) = (B_{k,d})^{-1} H_k V_l \text{ or } (B_{k,d})^{-1} H_k \left\| V_l \right\|_F.
\]

(13)

To evaluate the precoding vector \( V_l \) for the \( d \)-th layer at the \( k \)-th user, we reformulate \( \text{SINR}_{k,d} \) by the channel reciprocity [17] as follows

\[
\text{SINR}_{k,d} = \frac{(V_l^H H_k^H U_{d_k d} U_{d_k d}^H V_l^H H_k^H U_{d_k d} U_{d_k d}^H V_l)}{(V_l^H H_k^H U_{d_k d} U_{d_k d}^H V_l)}
\]

(14)

where

\[
\tilde{B}_{k,d} = \sum_{l=1, l \neq d_k}^{d_f} H_l^H U_l^H H_k + N_0 I_{N_l}
\]

(15)

The optimal vector \( V_l \) that maximises \( \text{SINR}_{k,d} \) can be formulated as

\[
V_l = \sqrt{P_{\text{UPW}}/d_k} \left( \tilde{B}_{k,d} \right)^{-1} H_k^H U_{d_k d} U_{d_k d}^H V_l \text{ or } \left( \tilde{B}_{k,d} \right)^{-1} H_k^H U_{d_k d} U_{d_k d}^H V_l \right\|_F.
\]

(16)
By using (13) and (16), the precoding and decoding matrices can be evaluated in each iteration. As a final remark, the initial decoding matrix for this iterative algorithm is suggested to be the typical discrete Fourier transform (DFT) matrix $W \in \mathbb{C}^{N_t \times N_K}$. Its element in the $i$th row and the $j$th column is given by

$$w_{ij} = e^{-j2\pi(i-1)(j-1)/N_t \cdot K}.$$

The iterative procedure is summarised in Table 1. The iterative procedure is expected to converge and generate a nearly optimal solution; however, to attain the optimal solution usually requires a tremendous amount of iterations. The algorithm provided in this section is denoted as ‘existing max SINR IA’.

<table>
<thead>
<tr>
<th>Table 1</th>
<th>Procedure for Max SINR IA transceiver design</th>
</tr>
</thead>
<tbody>
<tr>
<td>Initialisation:</td>
<td>Set an initial value for decoding matrix $U_i$; DFT matrix is suggested as initial value for faster convergence</td>
</tr>
<tr>
<td>Step 1:</td>
<td>Compute the precoders $V_i, i = 1, \ldots, \mathcal{M}$ according to (16)</td>
</tr>
<tr>
<td>Step 2:</td>
<td>Compute the decoding $U_i$ according to (13)</td>
</tr>
<tr>
<td>Step 3:</td>
<td>Go back to Step 1 till the constraint on iteration number is achieved or convergence is achieved</td>
</tr>
</tbody>
</table>

3.2 Proposed two-stage transceiver design algorithm

A new algorithm is proposed with the objective to maximise SINR within the small number of algorithmic iterations by exploiting the structure of effective channel. As observed in [26], the sum-rate can be improved through proper power allocation if the effective channel has the nearly diagonal structure leading to less interference. Through preserving available DoFs and conducting power allocation while suppressing interference, a two-stage approach is proposed to solve the max SINR problem as depicted in (10).

At the first stage, our proposed design criterion is to minimise inter-user interference subject to preserving available DoFs. Specifically, to evaluate the first-stage decoder $U_{i,1} \in \mathbb{C}^{N_t \times M \times N_M}$, the interference and noise power at the output of first-stage decoder corresponding to the $i$th user $P_{int,i}$ is defined as

$$P_{int,i} = \text{tr}(U_{i,1}^{H}H_{int,i}^{H}H_{int,i}^{H}U_{i,1}) + \text{tr}(N_0 U_{i,1}^{H}U_{i,1})$$

where $H_{int,i}$ is the associated interference channel for the $i$th user

$$H_{int,i} = [H_iV_1, \ldots, H_{i-1}V_{i-1}, H_{i+1}V_{i+1}, \ldots, H_{MK}V_{MK}].$$

(18)

Then, the design problem for the first stage decoder $U_{i,1}$ can be formulated as

$$\begin{align*}
\text{minimise} & \quad \tilde{P}_{int,i}, \\
\text{subject to} & \quad \|U_{i,1}\|_F^2 = 1, \\
\text{rank}(U_{i,1}) & = N_i M
\end{align*}$$

(19)

where the norm constraint is imposed as a normalisation factor, and the rank constraint ensures preserving DoFs for the decoder. The problem in (19) can be readily solved with a unitary $U_{i,1}$. To conduct power allocation, the first-stage decoder is decomposed as $U_{i,1} = B_{i,1}W_{i,1}$, where $B_{i,1}$ is a unitary basis matrix and $W_{i,1}$ is a diagonal weighting matrix with non-negative entries. Then, $P_{int,i}$ can be expressed as

$$P_{int,i} = \text{tr}(W_{i,1}^{H}B_{i,1}^{H}(T_{int,i}^{H}S_{int,i}^{H}S_{int,i}^{H}T_{int,i}^{H} + N_0 I_{N_M})B_{i,1}W_{i,1})$$

(20)

where $H_{int,i} = T_{int,i}S_{int,i}^H$ is the singular value decomposition (SVD) of $H_{int,i}$. We set $B_{i,1}$ to match singular vectors of interference channel $H_{int,i}$ as $B_{i,1} = T_{int,i}$

(21)

The $P_{int,i}$ can be further expressed as

$$P_{int,i} = \text{tr}(W_{i,1}^{H}(S_{int,i}S_{int,i}^{H} + N_0 I_{N_M})W_{i,1}).$$

(22)

For minimising $P_{int,i}$ and preserving available DoFs, $W_{i,1}$ is set, based on water-filling principle which allocates more power to the channel with less interference and noise power, as

$$W_{i,1} = S\left((S_{int,i}S_{int,i}^{H} + N_0 I_{N_M})^{-1/2}, 1\right).$$

(23)

Then we adopt the same approach to obtain the first-stage precoder $V_{i,1} \in \mathbb{C}^{N_t \times N_i}$ by reciprocity property. The interference and noise power at the output of first-stage virtual decoder (as the first-stage precoder in forward link) belonging to the $i$th user, denoted as $P_{int,i}$, is defined as

$$P_{int,i} = \text{tr}(V_{i,1}^{H}\tilde{H}_{int,i}^{H}H_{int,i}V_{i,1}) + \text{tr}(N_0 V_{i,1}^{H}V_{i,1})$$

(24)

where $\tilde{H}_{int,i}$ is the interference channel for the $i$th user, expressed as

$$H_{int,i} = [H_{i}^{H}U_{1,1}, \ldots, H_{i}^{H}U_{1,i-1}, H_{i}^{H}U_{1,i+1}, \ldots, H_{i}^{H}U_{1,MK}].$$

(25)

Then, the optimisation design problem for first-stage precoder $V_{i,1}$ corresponding to the $i$th user can be formulated as

$$\begin{align*}
\text{minimise} & \quad \tilde{P}_{int,i}, \\
\text{subject to} & \quad \|V_{i,1}\|_F^2 = \bar{P}_{UPW}, \\
\text{rank}(V_{i,1}) & = N_i
\end{align*}$$

(26)

Afterwards, the first-stage precoder is expressed as $V_{i,1} = \tilde{B}_{i,1}W_{i,1}$ where $\tilde{B}_{i,1}$ and $W_{i,1}$ are

$$\tilde{B}_{i,1} = \bar{T}_{i,1}; \tilde{W}_{i,1} = S\left((S_{int,i}S_{int,i}^{H} + N_0 I_{N_M})^{-1/2}, \bar{P}_{UPW}\right)$$

(27)

and $\tilde{H}_{int,i}$ is the SVD of $\tilde{H}_{int,i}$. The above procedure completes the first stage.

At the second stage, the design criterion is to decouple the desired signals to eliminate intra-user interference. The effective channel for the $i$th user after the first stage is given...
by
\[ \hat{H}_i = U_{i,1}^H H_i V_{i,1} \]  \hspace{1cm} (28)
and the SVD of the effective channel with singular values arranged in descending order is given by
\[ \hat{H}_i = \tilde{\mathbf{T}}_i \tilde{\mathbf{S}}_i \tilde{\mathbf{Q}}_i^H. \]  \hspace{1cm} (29)
To resolve the intra-user interference, the second-stage decoder \( U_{2,i} \) and precoder \( V_{2,i} \) of the \( i \)th user are chosen as
\[ U_{2,i} = \{ \tilde{\mathbf{T}}_i \}_{1}^{d_i}, \quad V_{2,i} = \{ \tilde{\mathbf{Q}}_i \}_{1}^{d_i}. \]  \hspace{1cm} (30)
By the linearity of the two stages, the decoder can be obtained as
\[ U = S(U_1 U_2, 1) \]  \hspace{1cm} (31)
where \( U_1 = [U_{1,1}, U_{1,2}, \ldots, U_{1,M_K}] \) and \( U_2 = \text{diag}(U_{2,1}, U_{2,2}, \ldots, U_{2,M_K}) \) are the first and second-stage decoders, respectively. Similarly, the precoder \( V_i \) for the \( i \)th user can be expressed as
\[ V_i = S(V_1 V_2, P_{\text{UPW}}). \]  \hspace{1cm} (32)
From the above, the decoders and precoders of the first stage and second stage are first constructed. Then the combined decoder and precoders are obtained after multiple iterations until the convergence condition is reached. It should be noted that the initial precoding matrix of the \( i \)th user for the proposed two-stage algorithm is given by
\[ V_i = \{ \mathbf{W} \}_{1}^{d_i}. \]

The complete iterative procedure is summarised in Table 2. As will be observed in the simulation section, the above developed transceiver design offers comparable performance to existing Max SINR IA transceivers with a much higher convergence rate. It should be noted that, in order to notify the users of their corresponding precoders designed at CU, a feedback channel is required in the downlink to send the precoding matrices to users from the BSs. In this work, we assume that there exists an ideal channel for such precoder feedback. Since this feedback is carried out in the DL, the UL sum-rate performance metric as considered in this paper will not be affected.

**Table 2**  Procedure for two-stage transceiver design

<table>
<thead>
<tr>
<th>Step</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Compute first-stage decoder ( U_{1,i} ) according to (21) and (23)</td>
</tr>
<tr>
<td>2</td>
<td>Compute first-stage precoder ( V_{1,i} ) according to (27)</td>
</tr>
<tr>
<td>3</td>
<td>Compute second-stage precoder ( V_{2,i} ) and decoder ( U_{2,i} ) according to (30)</td>
</tr>
<tr>
<td>4</td>
<td>Obtain final decoder ( U^f ) and precoders ( V_i ) according to (21) and (32)</td>
</tr>
<tr>
<td>5</td>
<td>Go back to Step 1 till the constraint on iteration number is achieved or convergence</td>
</tr>
</tbody>
</table>

4  Configuration selection for transceiver and a low-complexity algorithm

In the previous sections, we have considered transceiver design under a given number of data streams transmitted by each user; a question naturally arises as to what assignment of data streams to the users can provide the best performance under the proposed transceiver design framework. This is referred to as the configuration selection problem. In this section, we formulate the problem and propose a low-complexity algorithm to find the solution.

4.1  Problem formulation of configuration selection

Since the performance of the transceiver is hugely affected by the number of data streams to be transmitted and the characteristics of channel, the solution to configuration selection may not be trivial. If more data streams are transmitted, the transceiver has to utilise more DoFs to align the interference, which would reduce the number of DoFs for effective data streams. On the other hand, transmitting more data streams provides potentially higher rates if the interference can be properly mitigated. Thus, there is an inherent trade-off between using DoFs for interference mitigation and data stream transmission.

We consider all possible combinations of numbers of data streams to be transmitted for each user as the feasible configuration set, where a single combination is denoted as a configuration. In this setting, the problem of finding the configuration providing the largest achievable sum-rate in the feasible set can be formulated into a search problem as
\[ \theta = \arg \max_{\theta_i} C_{\theta_i} = \sum_{k=1}^{M_K} d_k^0 \log \left( 1 + \text{SINR}_{\theta_k} \right) \]  \hspace{1cm} (33)
where \( \Theta \) is the feasible set containing all possible configurations, \( \theta_i \) is the element in \( \Theta \), \( d_k^0 \) is the number of transmitted data streams of the \( k \)th user for \( \theta_i \) and SINR\(_k^{(\theta)}\) is the SINR of the \( j \)th data stream in the \( k \)th user for \( \theta_i \).

Note that due to the limited number of transmit antennas, \( \Theta \) is a finite set containing \((N_T)^{\text{MK}}\) elements. An intuitive approach to solving (33) is to exhaustively search for all the possible configurations and calculate the achievable sum-rates for each configuration. Then the one with the largest achievable sum-rate is selected. However, the exhaustive search induces high complexity due to both the computations in designing the transceiver and the large amount of configurations to be searched in the feasible set. A low-complexity configuration selection algorithm that can provide a near-optimal solution is thus desired.

4.2  Simplified metric for SINR prediction

We first derive a model to represent the SINR of the data streams in the proposed transceiver. A simple metric to predict the SINR of different configurations is then derived from the model.

For high SNR cases, the MMSE decoder can be approximated as
\[ F^H = (H_{\text{eff}}^H C_n H_{\text{eff}})^{-1} H_{\text{eff}}^H v_n^{-1}. \]  \hspace{1cm} (34)
Substituting (34) into (5), the signal detection vector can be
expressed as
\[ \hat{x} = x + \left( H_{\text{eff}}^H C_n^{-1} H_{\text{eff}} \right)^{-1} H_{\text{eff}}^H C_n^{-1} U^H z. \] (35)

From (35), the covariance matrix of noise \( \tilde{z} \) is given by
\[ C_{\tilde{z}} = \left( H_{\text{eff}}^H C_n^{-1} H_{\text{eff}} \right)^{-1}. \] (36)

The \( H_{\text{eff}} \) and \( C_n^{-1} \) are approximately diagonal matrices if the interference is well aligned, that is, the transceiver satisfies the IA feasibility condition. Thus, the covariance matrix can also be approximated as a diagonal matrix. As a result, SINR also be approximated as a diagonal matrix. As a result, the covariance matrix can be written as
\[ C_{\tilde{z}} = \left( H_{\text{eff}}^H C_n^{-1} H_{\text{eff}} \right)^{-1}. \] (36)

The \( H_{\text{eff}} \) and \( C_n^{-1} \) are approximately diagonal matrices if the interference is well aligned, that is, the transceiver satisfies the IA feasibility condition. Thus, the covariance matrix can also be approximated as a diagonal matrix. As a result, SINR can be expressed as
\[ \text{SINR}_d = \sigma_n^2 \left| \mathbf{H}_{\text{eff}}^{(dd)} \right|^2 \] (37)

where SINR\(_d\) is the \( d \)th streams among all the data streams corresponding to \( d \)th element of \( x \). Without any pre-defined preference in detecting data streams, the diagonal terms of \( C_n^{-1} \) are expected to be nearly identical. Therefore, (37) can be written as
\[ \text{SINR}_d = \sigma_n^2 \left| \mathbf{H}_{\text{eff}}^{(dd)} \right|^2 \] (38)

where \( \sigma_n^2 \) is the expected diagonal element in \( C_n \). To further analyse (38), we express \( \mathbf{H}_{\text{eff}}^{(dd)} \) as
\[ (\mathbf{H}_{\text{eff}}^{(dd)}) = (\mathbf{U}^{(d)})^H (\mathbf{H} V)^{j(d)}. \] (39)

The \( d \)th column of \( \mathbf{H} V \) can be decomposed as
\[ (\mathbf{H} V)^{j(d)} = \sum_{i=1}^{N_r M} a_{di} b_i \] (40)

which is the linear combination of columns in a basis matrix \( \mathbf{B} \) containing \( N_r M \) orthonormal vectors. Since there are \( d_T - 1 \) interferers for a data stream, the dimension of subspace containing the interference is \( d_T - 1 \) in the worst case, that is, the interference vectors are linearly independent. Therefore, the interference for a data stream can be represented by using the first \( d_T - 1 \) vectors of \( \mathbf{B} \) without loss of generality. With well-aligned interference, the best decoder for the \( d \)th data stream in the worst-case scenario is given by
\[ \mathbf{U}_W^{(d)} = \sum_{i=d_T}^{N_r M} c_{di} b_i \] (41)

where the coefficients are chosen to match the coefficients in (40), that is
\[ c_{di} = \frac{d_{T, i}}{|a_{di}|} \times \sqrt{MN_T - d_T + 1}. \]

In order to take the gain of IA into consideration, that is, when the dimension of subspace containing the interference is smaller than \( d_T - 1 \), a calibration term \( C_d \) is added. The predicted receiver is then expressed as
\[ \mathbf{U}^{(d)} = C_d \sum_{i=d_T-1}^{MN_T} c_{di} b_i. \] (42)

Since the exact \( \mathbf{H} \) is unavailable before computing the exact transceiver, we consider the coefficients \( \{c_{di}\} \) in (40) as random variables with identical distributions and variance \( \sigma_n^2 \). By substituting (39), (40) and (42) into (38), and taking expectation, the model of the predicted SINR for the \( d \)th data stream is established by the expected SINR, and expressed as
\[ \text{SINR}_d = E \left\{ \sigma_n^2 \left| \mathbf{H}_{\text{eff}}^{(dd)} \right|^2 \right\} = (N_r M - d_T + 1)C_d \sigma_n^2 \] (43)

The \( \sigma_n^2 \) implicitly represents the channel gain and transmit power of \( d \)th user. If the channel gain and transmit power are large, the variance will be large and vice versa. The \( C_d \) here accounts for the calibration of the gain due to interference alignment, and would be larger if the interference is better aligned than the worst case. It is important to note that the calibration terms and variances are assumed to be the same for all data streams transmitted from the same user, and the power allocation of the data streams within the same user is assumed to be uniform.

Towards this end, a model for the SINR of data streams is established. As \( C_d, \sigma_d^2 \), and \( \sigma_n^2 \) are unknown in the model, a reference configuration is used to derive SINR without \( C_d, \sigma_d^2 \) and \( \sigma_n^2 \). We suggest that the reference configuration be the configuration with each user transmitting a single data stream. The SINR for each data stream of the reference configuration can be obtained by the complete design of the transceiver. Using (43) and the computed SINR of the reference configuration, the effects of \( C_d, \sigma_d^2 \) and \( \sigma_n^2 \) can be computed. With the assumption of uniform power allocation, the SINR for each data stream of a given configuration can be predicted. The SINR of the \( j \)th data stream in the \( k \)th user for \( \theta_q \) is then given by
\[ \text{SINR}_{j,k}^{(q)} = \text{SINR}_{k}^{\text{ref}} \frac{1}{d_q^{(q)}} \frac{N_r M - d_q^{(q)} + 1}{N_r M - K + 1} \] (44)

where SINR\(_{k}^{\text{ref}}\) is the SINR of the \( k \)th user of the reference configuration; \( d_q^{(q)} \) is the number of data streams transmitted by the \( k \)th user in the \( q \)th configuration; \( d_q^{(q)} \) is the total number of data streams transmitted in the \( q \)th configuration. By using (44), (7) and the SINR of the reference configuration, the predicted achievable sum-rates of all the configurations can be computed. It is noted that even if the derivation of the metric in (44) is not for low SNR cases, the prediction precision is reasonably good for the low SNR regime as can be observed in the numerical results.

### 4.3 Reduced-complexity search strategy

In the previous section, a simplified metric to reduce the complexity in computing the achievable sum-rate is proposed. In this section, we propose to reduce the high complexity arising from the exponential increase and polynomial increase in the number of configurations with respect to the number of users in a CoMP system and number of transmit antennas, respectively. A new search strategy is proposed to reduce the size of the search space.
In the proposed strategy, the reference configuration is set as the initial point in the search space. Then the number of data streams transmitted by the first user is altered to generate the achievable sum-rate of the new configuration. The achievable sum-rate is computed for every feasible configuration using transceiver design algorithm in Section 3. The metric in Section 4.2 converts the complicated computation in the exact transceiver design into a simple metric computation; the proposed search strategy in this section renders the number of configurations to increase only linearly, instead of being exponential or polynomial. A combination of these two findings gives a procedure that is expected to be of very low complexity compared with exhaustive search. The complete procedure is summarised in Table 3.

5 Numerical results

In this section, we provide numerical results to demonstrate the performance of the proposed algorithms. The algorithms are evaluated in a typical CoMP scenario in 3GPP recommendation [5], and in another scenario with a larger cooperation size. The typical CoMP MIMO system consists of three BSs in the cooperation group. Each BS has a single user in its coverage participating in CoMP, and is equipped with four antennas. In the system with a larger CoMP size, two users are in the coverage of each BS, and each BS is equipped with eight antennas. For both systems, each user is equipped with four antennas. The channel matrices for the direct links and interference links are set as described in the system model in Section 2. The parameter $\epsilon$ as mentioned in Section 2.2 is set to be 0.4 for all cases, except for Fig. 4, to evaluate the typical scenario in which users are around the half-radius concentric region in the cell, that is, users are in the middle between the BS and the cell boundary. In Fig. 4, $\epsilon = 0.9$ is chosen to assess the case where users are located near the cell edge. The achievable sum-rate is calculated by (7). The SNR is defined as the transmit power per user over the average noise power.

In Figs. 3 and 4, the proposed transceiver design algorithm is compared with conventional CoMP adopting the joint detection receiver without any precoders (CoMP), and the Max SINR IA algorithm described in Section 3.1. The dash lines ‘-’ represent the achievable sum-rates of the typical CoMP system with different transceiver design algorithms and number of iterations. The solid lines ‘-’ represent the performance with a larger CoMP size. Each user in both systems transmits three fixed data streams. From Fig. 3, the proposed transceiver design algorithm is observed to outperform the conventional CoMP transceiver in both scenarios. Meanwhile, the proposed transceiver design algorithm provides comparable performance as Max SINR algorithm while requiring a much smaller number of iterations, or much lower complexity. Furthermore, the Max SINR algorithm provides almost no performance improvement if the number of iterations is not sufficiently high. In Fig. 4, similar results to those in Fig. 3 can be observed. The achievable sum-rate of the proposed transceiver at high SNR regime in Fig. 4 is larger compared

<table>
<thead>
<tr>
<th>Table 3 Procedure for configuration selection</th>
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<tr>
<td><strong>Initial stage:</strong></td>
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<tr>
<td>Step 1: Set reference configuration $\theta_i$</td>
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<tr>
<td>Step 2: Compute SINR, and sum capacity for reference configuration using transceiver design algorithm in Section 3</td>
</tr>
<tr>
<td>Step 3: Set the selected configuration $\theta_i$</td>
</tr>
<tr>
<td><strong>Search stage:</strong></td>
</tr>
<tr>
<td>Step 1: Set $n = 1$. Set $\theta_i$ according to selected $\theta_i$</td>
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<tr>
<td>Step 2: Calculate SINR and achievable sum rate for $d_n = 1, 2, \ldots$, $N$, according to (44) and (7)</td>
</tr>
<tr>
<td>Step 3: Substitute $d_n$ in $\theta_i$ with one that has largest sum capacity. Set $n = n + 1$</td>
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<tr>
<td>Step 4: Go back to step 2 until $n = MK$</td>
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<tr>
<td>Step 5: Output selected configuration</td>
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Fig. 3 Achievable sum rate performance of transceiver design algorithms in a typical CoMP system (dash lines) and larger CoMP system (solid lines) with $\epsilon = 0.4$

Fig. 4 Achievable sum rate performance of transceiver design algorithms in a typical CoMP system (dash lines) and larger CoMP system (solid lines) with $\epsilon = 0.9$
to those in Fig. 3. This is because the proposed linear transceiver can effectively exploit the stronger coordinating links to enhance system performance thanks to IA. For the remaining simulations, the results with \( \varepsilon = 0.9 \) are omitted for brevity.

To observe more results on the number of iterations required for the proposed algorithm and Max SINR algorithm, Figs. 5 and 6 show the convergence behaviours of these two algorithms in the typical CoMP system, and the system with a larger CoMP size, respectively, with different SNR values. The ‘Sum-Rate-Ratio’ is defined as the achievable sum-rate of the proposed linear transceivers at 40,000 iterations divided by the achievable sum-rate of the transceivers at the number of iterations indicated by the horizontal axis, that is, achievable sum-rates of different schemes at different iterations are normalised by the achievable sum-rate of the proposed linear transceiver at 40,000 iterations. With such definition, the curves converging to one faster exhibit better convergence behaviour. Besides, the convergence can be seen as the curves of sum-rate ratios flatten. Observing from Figs. 5 and 6, the proposed algorithm converges faster and has better performance than the Max SINR algorithm in both systems. In summary, it is confirmed that the proposed transceiver design algorithm achieves better performance and requires significantly lower complexity compared with the Max SINR IA algorithm.

Next, the proposed configuration selection algorithms are evaluated in Figs. 7 and 8 for the typical and larger CoMP systems, respectively. The simulations are conducted for the following combinations: ‘exhaustive search + conventional metric’ (original approach described in Section 4.1), ‘exhaustive search + proposed metric’ and ‘proposed search + proposed metric’. The case of randomly selecting the configuration is also simulated as the trivial benchmark for comparison. As observed in Figs. 7 and 8, the proposed simplified metric gives nearly the same achievable sum-rate as the conventional metric, and the proposed search strategy gives nearly the same achievable sum-rate as the exhaustive search strategy. Thus, the proposed search strategy combined with the simplified metric can indeed achieve nearly the same performance as fully exhaustive search with significantly lower complexity. As expected, the advantage of configuration selection over random selection decreases at low SNR. The simulation results reveal that the proposed algorithm remains effective even if there are up to hundreds of candidate configurations to be selected.

Finally, the performance of the proposed algorithms is compared with transceivers composed of iterative water-filling designed transmitters and linear ZF/MMSE receivers. Since the iterative water-filling solutions implicitly assign the number of transmitted data streams for each user through the rank of precoding matrices, the transceivers designed by iterative water-filling are compared with the proposed transceiver combined with the configuration selection algorithm. The results are shown in Figs. 9 and 10 with the same environment settings as in Figs. 7 and 8, respectively. The achievable sum-rate of the transceiver using the ideal MMSE-SIC receiver (i.e. without error propagation) is also provided as the upper bound benchmark. As observed from Figs. 9 and 10, the
performance of transceivers using linear receivers degrades substantially from the upper bound, especially for high SNR. This is because the linear receivers cannot effectively mitigate the interference. In contrast, the proposed transceiver exhibits less degradation at high SNR thanks to IA. While there is a certain performance gap between the idealised and the proposed transceivers, the MMSE-SIC receiver could potentially suffer from performance degradation due to error propagations. It is noted that although the achievable sum-rate of MMSE-SIC with error propagation may not be mathematically tractable, the resulting performance degradation is expected to be significant as reported in the literatures [15, 24, 25].

6 Conclusions

In this work, the motivation and system framework of adopting linear receivers in the centralised UL CoMP MIMO system is first elaborated. Then a new transceiver design algorithm is developed by exploiting the technique of IA, optimising the structure of the effective channel and employing power loading. The transceiver is further cascaded with an MMSE receiver to achieve better noise suppression in the low SNR regime. Compared with the existing ZF/MMSE-based linear CoMP transceivers, the proposed transceiver offers superior achievable sum-rate performance. Compared with existing IA-aided transceivers, the proposed transceiver offers comparable sum-rate performance with a much higher convergence rate. In contrast to conventional CoMP transceiver design which is investigated under a predefined number of data streams transmitted by each user, the proposed method further investigates configuration selection to optimally assign data streams to different users. A simplified SINR metric and an efficient search strategy are proposed, leading to a very low-complexity algorithm for configuration selection compared with exhaustive search. The combination of the proposed linear transceiver and low-complexity configuration selection constructs a new practical framework for linear transceiver design in UL CoMP MIMO systems. The proposed design method is evaluated through computer simulations and compared with the conventional CoMP system with IA-aided and water-filling-based transceivers. The results demonstrate that the proposed transceiver offers (1) the advantageous performance of IA-aided transceivers with a significantly improved convergence behaviour; and (2) the complexity of conventional linear transceivers with a significantly better sum-rate performance. Although the water-filling-based transceiver with a non-linear receiver seems to outperform the proposed transceiver in the high SNR regime, the potential drawbacks of error propagation and long processing latency can be avoided in our proposed approach to facilitate applications in a large CoMP system.

7 Acknowledgments

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8 References

6 3GPP TR 36.819, ‘Coordinated multi-point operation for LTE physical layer aspects (Release 11)’, December 2011


