Joint Transmitter-Receiver Optimization and Self-interference Suppression in Full-Duplex MIMO Systems

Radwa Sultan[§], *IEEE Member*, Karim G. Seddik[*], *IEEE Senior Member*, Zhu Han[†], *IEEE Fellow Member*, and Behnaam Aazhang[‡], *IEEE Fellow Member*

[§] Electrical and Computer Engineering Department, Manhattan College, Riverdale, NY, USA
[*] Electronics and Communications Engineering Department, American University in Cairo, New Cairo, Egypt
[†] Electrical and Computer Engineering Department, University of Houston, TX, USA
[‡] Department of Electrical and Computer Engineering, Rice University, Houston, USA
rsultan02@manhattan.edu, kseddik@aucegypt.edu, zhan2@uh.edu, aaz@rice.edu

Abstract*—In this paper, we study the effects of joint transmit and receive antennas' selection on full-duplex (FD) multiple-input-multiple-output (MIMO) networks' performance. The antennas' selection problem is, in general, a combinatorial problem whose complexity grows exponentially with the number of antennas. To fully understand the effects of antennas' selection, we study the sum rate maximization problem in a single-cell network with an FD-MIMO base-station (BS). First, we consider a system with a normal-scale MIMO full-duplex BS, i.e., a normal-scale FD MIMO system. The sum rate maximization problem is studied for two different scenarios; in the first scenario, we consider jointly optimizing the transmit and receive antennas' selection with the precoder and the receiver weights. A Generalized Bender's Decomposition based algorithm is proposed to solve the mixed-integer nonlinear sum rate maximization problem. In the second scenario, we consider self-interference cancellation via zero-forcing (ZF) transmission. A heuristic algorithm is proposed to solve the sum rate maximization problem by optimizing the selection of the transmit antennas, receive antennas, and the receive antennas at which self-interference is nulled. Second, in a very-large scale, i.e., massive MIMO system, we derive lower bounds for the uplink and downlink rates with ZF receiver and precoder, respectively. The sum rate is maximized by jointly optimizing the transmit to receive antennas ratio and the ratio of the receive antennas at which self-interference is nulled. Finally, via numerical analysis, we evaluate the performance of the formulated sum rate maximization problems.

Keywords. Antenna Selection, Asymptotic Behavior, Full-Duplex, Generalized Bender's Decomposition, Massive MIMO, Resource Allocation, Transmit Beamforming.

I. Introduction

Full-Duplex (FD) communication and massive multiple-input-multiple-output (MIMO) are envisioned as eminent candidate technologies capable of supplying the ever-growing demand in communication links capacity required in the modern cellular networks. Theoretically speaking, enabling FD communication can double the data rate achieved by half-duplex (HD) communication. In general, self-interference (SI), which is the interference from the FD node transmission on its reception, is considered to be the main obstacle for the FD feasibility until the evolution of SI cancellation techniques [1]–[3]. Additionally, massive MIMO [4]–[6] has proven its ability to increase the spectral efficiency and communication reliability compared to normal-scale MIMO systems, as tremendously increasing the number of antennas helps in expanding the network's capacity. Additionally, with a massive number of antennas, the simplest form of user detection and/or beamforming, such as matched filter (MF) or zero-forcing (ZF), becomes optimal.

However, utilizing FD in MIMO communication networks poses some new challenges [7]–[9]. The first challenge is how to optimize the transmit and receive antenna selections. For a given MIMO base station (BS) operating in half-duplex (HD), all the antennas will be either transmitting or receiving, in the downlink (DL) or the uplink (UL) mode, respectively. However, for a given MIMO-BS operating in FD, we need to choose the number and the indices of transmit antennas and receive antennas. The decision is highly dependent on the channel conditions between the BS antennas and the network users, and this choice can highly affect system performance. Hence, to optimize the network's performance, we need to optimize the transmit and receive antennas' selection. Furthermore, to make FD-MIMO feasible, we need to reduce the SI. However, as the number of antennas increases, the complexity of the analog SI cancellation circuit grows. Accordingly, to prevent radio frequency (RF) saturation of the receiver, SI cancellation via transmit beamforming is needed. As a result, the second challenge is to make the FD-DL precoder's design meet two objectives; the first objective, similar to the DL-HD precoder, is to reduce the effect of the multi-user interference (MUI), and the second objective is to reduce the SI level.

A. Contributions

In this paper, and based on the previously mentioned challenges, we study different scenarios for antenna selection and transmitter-receiver joint optimization to maximize the achievable sum rate in a single cell network with an FD-MIMO base station and multiple users. Our contributions are summarized as follows

1) In a normal-scale FD MIMO setting, we propose a sum rate maximization problem by jointly optimizing the transmit antennas' selection, the receive antennas' selection, the transmit precoder weights, and the receiver weights. As a result of the combinatorial nature of the antennas' selection, the proposed optimization problem is a mixed-integer non-linear

programming (MINLP) problem, which we solve using the *Generalized Bender's Decomposition* (GBD) algorithm.

2) In a normal-scale FD MIMO setting, to investigate the effect of considering SI cancellation, we assume the BS has a transmit ZF precoder and a ZF receiver. In this case, the ZF precoder will help in reducing the SI power on the UL reception. Accordingly, we propose a sum rate maximization problem by jointly optimizing the transmit antennas' selection, the receive antennas' selection, and the SI cancellation. To better optimize the SI cancellation, we need to optimally select the number of receive antennas at which the ZF precoder will cancel the SI. The main idea of SI cancellation via ZF transmit beamforming is to design the DL precoder nulling the SI at the BS receive antennas. However, increasing the number of required interference-nulls decreases the degrees-of-freedom in the DL transmission, and hence, the DL transmission gain, posing a trade-off in the selection of the number of receive antennas at which the ZF precoder will cancel the SI. In this scenario, and to avoid the growing complexity of solving the combinatorial antennas' selection problem, we propose a heuristic, low complexity antennas' selection algorithm.
3) In a massive FD MIMO system, similar to the normal-scale FD-MIMO setting, we consider SI cancellation via ZF. In this case, the ZF precoder will help in reducing the SI power affecting the UL reception. Then, to understand the network's asymptotic behavior, we derive lower bounds for the UL and DL rates. From these bounds, we investigate the trade-off between the DL rate and increasing the number of receive antennas at which the ZF precoder nulls SI. However, under the assumption that the number of transmitting and receiving antennas tend to infinity, all antennas are asymptotically equivalent. Accordingly, it will not be necessary to optimize the indices of the transmitting and receiving antennas, and it will be sufficient to optimize their ratio to the total number of antennas. Finally, we formulate a rate region maximization problem optimizing the transmit to receive antennas ratio and the proportion of the receive antennas at which the ZF precoder nulls SI.

The rest of the paper is organized as follows. In Section II, an overview of some prior work done on resource allocation in FD-MIMO networks is presented. In Section III, the system model is illustrated. In Section IV, the sum rate maximization problem in FD massive MIMO networks is formulated. In Section V, the asymptotic rate behavior is investigated and the rate-region maximization problem is formulated under a massive MIMO setting. Numerical results are presented in Section VI. Finally, the paper is concluded in Section VII.

Notation: Vectors are written in boldface lowercase letters, while matrices are denoted by boldface uppercase letters. The statistical expectation of a random entity z is shown by $\mathbb{E}(z)$. The complex number field is denoted by $\mathbb{C}$. A list of key mathematical symbols is summarized in Table 1.

TABLE I: List of Symbols

Symbol	Definition
N	Number of BS antennas
N_t	Number of transmit antennas
N_r	Number of receive antennas
N_a	Number of receive antennas at which the SI is canceled by transmit-beamforming
K	Number of macro-users
K_d	Number of DL macro-users
K_u	Number of UL macro-users
P_d	DL Transmission Power
P_u	UL Transmission Power
$\mathbf{H}_d$	DL channel matrix
$h_{k,i}$	Channel between the k^{th} DL and i^{th} UL users
$\mathbf{H}_a$	Proportion of the channel between the BS transmit and receive antennas at which the SI will be cancelled
$\mathbf{H}_r$	Proportion of the channel between the BS transmit and receive antennas at which the SI will not be cancelled
$\mathbf{H}_{si}$	SI channel matrix
$\mathbf{H}_t$	Aggregate DL channel matrix
$\mathbf{W}_d$	DL precoding matrix
$\mathbf{F}_{ZF}$	DL-ZF precoding matrix
$\mathbf{W}_{ZF}$	Normalized DL-ZF precoding matrix
$\mathbf{w}_k$	k^{th} DL user's precoding vector
$\mathbf{U}_u$	UL receiver matrix
x_k	k^{th} DL user transmitted signal
$\mathbf{G}$	UL channel matrix
$\mathbf{U}_u$	UL receive matrix
$\mathbf{U}_{ZF}$	ZF-UL receive matrix
$\mathbf{u}_l$	l^{th} UL user's receive vector
C_{BS}	SI cancellation coefficient
$y_{d,k}$	k^{th} DL user received signal
$y_{u,l}$	l^{th} UL user received signal
Γ_{dl}	k^{th} macro DL user received SINR
Γ_{ul}	l^{th} macro UL user received SINR
Γ_{dl}^{ZF}	k^{th} macro DL user received SINR with ZF
Γ_{ul}^{ZF}	l^{th} macro UL user received SINR with ZF
R_d	DL-rate lower bound
R_u	UL-rate lower bound
R_s	sum rate per unit time and bandwidth
R_{ZF}	sum rate per unit time and bandwidth with ZF
$a_{t\|n}$	n^{th} antenna transmission coefficient
$a_{sic\|n}$	n^{th} antenna SI cancellation coefficient
σ^2	AWGN noise variance
α	Proportion of receive antennas on which SI is cancelled
β	Ratio between Tx and Rx antennas

II. Prior Work

With the evolution of SI cancellation techniques, the development of resource allocation schemes for different FD systems has attracted some recent research work [10], [11]. Additionally, authors have shown in [12] that FD can outperform HD in both interference-unaware and interference-aware scenarios. Accordingly, to fully utilize the available FD resources while considering the new challenges arising from deploying FD, efficient and novel resource allocation schemes are strongly needed [13]–[15]. Moreover, studying the potentials of massive MIMO networks and the potential benefits from increasing the number of active antennas are the main interests of recent research work [16]–[19].

As mentioned earlier in Section I, utilizing FD communication in MIMO networks poses the problem of selecting the best combination of transmit antennas and receive antennas at the BS. In [20], a joint sum rate maximization problem that aims at optimizing the half-array antenna mode selection at the base station, with time phases and user assignments, is proposed. Additionally, the authors considered a general max-min rate optimization to maximize the minimum per-user rate while satisfying a given ratio between UL and DL rates. In [21], the authors derived exact closed-form expressions of the outage probability and the symbol error probability for the FD-Spatial Multiplexing (SM)-MIMO system with transmit antenna selection (TAS). Furthermore, via numerical analysis, the authors evaluated the performance of the FD-SM-MIMO system, with and without TAS. In [22], an antenna selection based interference cancellation for an FD-MIMO system is proposed, in which interference cancellation is realized by antenna selection and Eigen beamforming. In [23], the authors examined the outage probability of max-max antenna selection in the FD amplify-and-forward relaying, in which the source, the relay, and the destination are equipped with multiple antennas. In [24], the antenna selection for a two-node FD-MIMO network is studied. It is assumed that each node is equipped with a predefined number of antennas and transmit/receive chains. The proposed selection algorithms are based on magnitude, orthogonality, and determinant criteria. In [25], the authors investigated multiple antenna-selection schemes in bidirectional FD MIMO Systems. After studying the optimal antenna-selection scheme, the authors proposed less complex, near-optimal scheme which utilizes a greedy search method. It was shown that the proposed algorithm has a 15% performance gain over random antenna selection. In [26], two antenna selection schemes, based on maximum sum rate and minimum symbol-error-rate, are proposed. It was shown, via numerical analysis, that these schemes can achieve significant performance gains. However, these selections gains are contingent on the reduction of the SI.

Additionally, SI cancellation in FD-MIMO has attracted more recent research work. In [27], a beam-based adaptive filter structure, with analog least mean square (ALMS) loops, is proposed to reduce the complexity of SI cancellation for FD-MIMO systems. Moreover, results showed that the proposed structure outperforms the ALMS loop employed for an FD single input single output system. In [28], the authors proposed an SI suppression method that adopts null steering with Eigen-beamforming (EB). The proposed SI suppression method relies on manipulating the propagation channel by careful arrangement of the MIMO antennas, which degenerates the rank of the SI channel matrix at the relay station such that the number of eigenmodes is nearly one. In [29], the authors proposed an antenna selection algorithm to reduce the Rician distributed SI. The presented criterion can minimize the effects of residual SI by maximizing the desired signal power to the residual SI and the noise ratio. In [30], two estimation algorithms for a two-stage SI cancellation scheme are presented. In the first stage, at RF SI cancellation, a compressed sensing-based SI channel estimation algorithm is derived. In the second stage, to jointly estimate the residual SI channel, the intended channel and the transmitter nonlinearities, a subspace-based algorithm is proposed; it was shown that the proposed algorithm achieves better performance than the least-square algorithm.

Furthermore, SI cancellation, in FD massive MIMO networks, is considered in [31]. Digital SI cancellation in a single RF chain FD-massive MIMO orthogonal frequency division multiplexing (OFDM) system with phase noise is studied. A weighted linear SI channel estimator is derived to minimize the residual SI power in each OFDM symbol. In [32], the authors studied two methods of partial analog SI cancellation. In the first method, the analog cancelers are assigned to a fixed set of antennas. On the other hand, in the second method, the analog cancelers can be dynamically assigned to any receive antennas based on channel conditions. In [33], assuming that instantaneous channel state information (CSI) of the SI channel is not available, the authors proposed an energy efficiency optimization based self-interference cancellation (SIC) scheme (EE-SIC). The proposed technique can effectively eliminate SI and improve system energy efficiency. In [34], it was shown, in FD massive MIMO networks, that the SI-subtraction outperforms the spatial suppression for SI cancellation in a perfect channel state information case. On the other hand, for imperfect channel estimation, spatial suppression will achieve better UL and total ergodic rates. In [35], the authors proposed an opportunistic user and antenna selection algorithm for FD multi-antenna BS and HD mobile terminal with two active antennas. The algorithm selects UL users' antennas according to their vector channel gain and DL users' antennas based on their signal-to-interference-plus-noise ratio (SINR). In [36], the authors studied a FD-massive MIMO network with a Rician SI-channel. Additonally, the authors utilized a linear ZF precoder with SI-cancellation, and an SI-aware fractional power control mechanism at the UL-FD mobile terminals. It was shown that exploiting massive MIMO can help in overcoming the UL rate bottleneck.

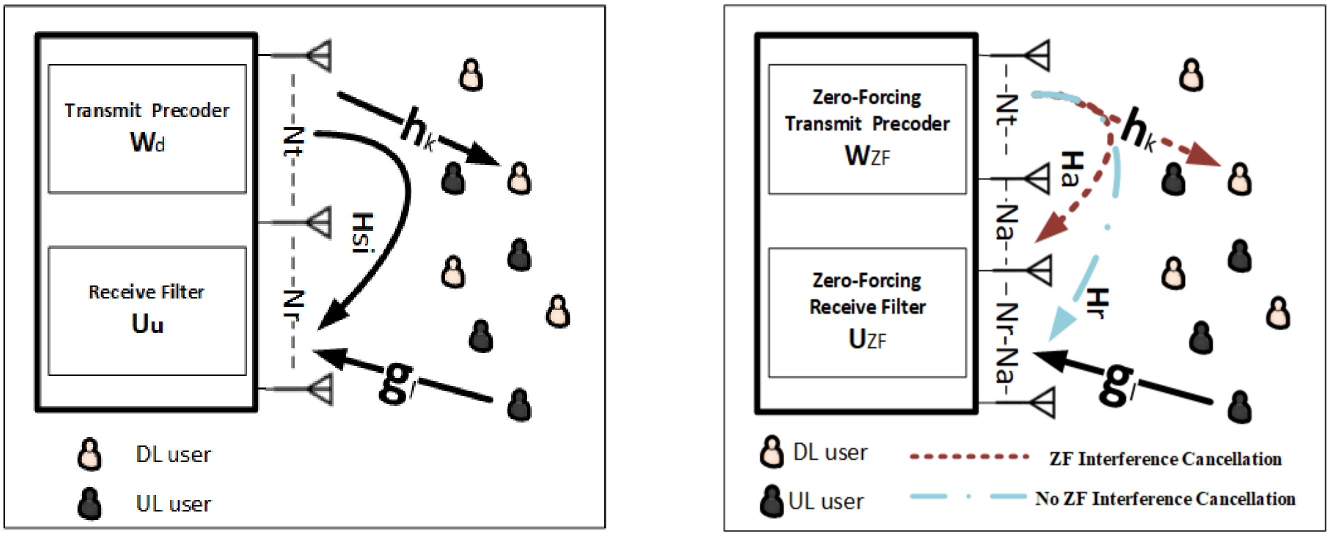
III. System Model

We consider a wireless network with a single N-antenna FD MIMO BS. Under the assumption of FD transmission, the BS is equipped with an FD radio to help suppress the SI. Additionally, the BS antennas are divided into N_t transmit antennas and N_r receive antennas. There are K HD single-antenna users in the network that are divided into K_d users receiving in the DL mode and K_u users transmitting in the UL mode. In this paper, we study two different scenarios. In the first scenario, we assume a general precoder and a general receiver that will be later optimized to maximize the sum rate. In the second scenario, we study the changes in the network after implementing a ZF precoder and a ZF receiver at the BS. Next, we will distinguish between these two scenarios in terms of the received UL signals, the received DL signals, and the receivers' SINRs.

A. Full Duplex MIMO Network

In this case, we assume that the BS has a DL precoder, with precoding matrix $\mathbf{W}_d \in \mathbb{C}^{N_t \times K_d}$, and a UL receiver with a receiving matrix $\mathbf{U}_u \in \mathbb{C}^{K_u \times N_r}$. Additionally, it should be guaranteed that $N_t \geq K_d$ and $N_r \geq K_u$ [1]. The system model is

[1]In the proposed model, it is assumed that K_d and K_u denote the number of the active DL and UL users, respectively, which will be served on the same time-frequency resource block. In other words, the network may have larger numbers of DL and UL users, however, user selection is out of the paper's scope. Accordingly, these conditions are required for both precoding in the DL transmission and the receiver filter in the UL transmission [37], [38].



(a) Full duplex MIMO network

(b) Full duplex MIMO network with ZF

Fig. 1: System model

shown in Fig. 1a. Based on the above assumptions, the received signal at the k^{th} DL user is given by

$$
\begin{aligned}
y_{d,k} = \underbrace{\sqrt{P_d}\mathbf{h}_k\mathbf{w}_k x_k}_{\text{Desired Signal}} + \underbrace{\sum_{\substack{j=1\\ j\neq k}}^{K_d}\sqrt{P_d}\mathbf{h}_k\mathbf{w}_j x_j}_{\text{MUI}} \\
+ \underbrace{\sum_{i=1}^{K_u}\sqrt{P_u}h_{k,i}s_i}_{\text{CCI}} + n_k^d,
\end{aligned}
\tag{1}
$$

where P_d is the DL transmission power, $\mathbf{h}_k \in \mathbb{C}^{1\times N_t}$ is the DL channel vector between the BS transmit antennas and k^{th} HD-DL user. Without loss of generality, all channel coefficients are assumed to be independent and identically distributed (i.i.d.), zero-mean complex Gaussian random variables with unit variance, i.e., we assume a Rayleigh flat fading channel model [39]–[41]. In addition, it is assumed that the channel state information (CSI) is perfectly available at both the BS and users [7] [2]. The vector $\mathbf{w}_k \in \mathbb{C}^{N_t\times 1}$ is the k^{th} column of the precoding matrix $\mathbf{W}_d \in \mathbb{C}^{N_t\times K_d}$ and denotes the precoding vector for the k^{th} DL user's data, x_k is the k^{th} DL user's transmit signal where $\mathbb{E}\{\|x_k\|^2\} = 1$. The second term is the MUI caused by the other DL transmissions in the network. The third term is the co-channel interference (CCI) caused by the UL transmissions from the K_u UL users, where P_u is the UL transmission power. In this paper, we assume that all the K_u UL users have a constant transmit power P_u [42]–[44]. $h_{k,i}$ is the channel coefficient between the i^{th} UL user and the k^{th} DL user, s_i is the i^{th} UL user transmit signal with $\mathbb{E}\{\|s_i\|^2\} = 1$, and n_k^d is the DL additive white complex Gaussian noise (AWGN) term with variance σ^2. From the expression in (1), the DL received SINR at the k^{th} DL user is given by

$$
\Gamma_{dl} = \frac{P_d\|\mathbf{h}_k\mathbf{w}_k\|^2}{\sigma^2 + \sum\limits_{\substack{j=1\\ j\neq k}}^{K_d} P_d\|\mathbf{h}_k\mathbf{w}_j\|^2 + \sum\limits_{i=1}^{K_u} P_u\|h_{k,i}\|^2}. \tag{2}
$$

On the other hand, in the case of the UL transmission, the received signal at the BS from the l^{th} UL user transmission $l \in \{1, 2, \cdots, K_u\}$ is given by

$$
\begin{aligned}
y_{u,l} = \underbrace{\sqrt{P_u}\mathbf{u}_l\mathbf{g}_l s_l}_{\text{Desired Signal}} + \sum_{\substack{i=1\\ i\neq l}}^{K_u}\underbrace{\sqrt{P_u}\mathbf{u}_l\mathbf{g}_i s_i}_{\text{MUI}} \\
+ \underbrace{\sqrt{C_{BS}}\sqrt{P_d}\mathbf{u}_l\mathbf{H}_{si}\mathbf{W}_d\mathbf{x}_d}_{\text{RSI}} + \mathbf{u}_l\mathbf{n}_u,
\end{aligned}
\tag{3}
$$

where $\mathbf{u}_l \in \mathbb{C}^{1\times N_r}$ is the receive vector for the l^{th} UL transmission data, $\mathbf{g}_l \in \mathbb{C}^{N_r\times 1}$ is the UL channel vector between the l^{th} UL transmitter and the BS receive antennas. The second term of the expression in (3) is the MUI on the l^{th} UL user from other UL transmissions in the network; the third term in (3) is the residual self-interference (RSI) signal. The RSI signal is the product of the square-root of the SI cancellation coefficient $0 \leq C_{BS} \leq 1$ which is introduced by the FD radio at the BS, the DL transmission power, the UL receive vector $\mathbf{u}_l$, the self-interference channel matrix $\mathbf{H}_{si} \in \mathbb{C}^{N_r\times N_t}$ between the BS transmit and receive antennas, the DL precoding matrix $\mathbf{W}_d$, and the transmitted DL signal $\mathbf{x}_d \in \mathbb{C}^{K_d\times 1}$ [45]. Dinally, $\mathbf{n}_u \in \mathbb{C}^{N_r\times 1}$ is the AWGN noise vector with variance σ^2. Accordingly, the received SINR for the l^{th} UL user at the BS receive antennas is given by

$$
\Gamma_{ul} = \frac{P_u\|\mathbf{u}_l\mathbf{g}_l\|^2}{\sigma^2\|\mathbf{u}_l\|^2 + \sum\limits_{\substack{i=1\\ i\neq l}}^{K_u} P_u\|\mathbf{u}_l\mathbf{g}_i\|^2 + C_{BS}P_d\|\mathbf{u}_l\mathbf{H}_{si}\mathbf{W}_d\mathbf{x}_\mathbf{d}\|^2}. \tag{4}
$$

From the expressions in (2) and (4), the sum rate per unit time

[2]The results under perfect CSI may act as an upper bound on the sum rate performance for the FD systems.

and unit bandwidth is given by

$$R_s = \sum_{k=1}^{K_d} C_1 \log_2(1+\Gamma_{dl}) + \sum_{l=1}^{K_u} C_2 \log_2(1+\Gamma_{ul}), \quad (5)$$

where C_1 and C_2 are constant scaling factors.

B. Full Duplex MIMO Network with ZF Precoder and ZF Receiver

Usually, in HD-MIMO networks, the ZF precoder and receiver aim to cancel the MUI between the DL transmissions and the UL transmissions, respectively. However, in FD-MIMO networks, besides canceling the DL-MUI, the DL transmit precoder will aim to null the SI on some of the BS receive antennas as well. The SI level is decreased by assuming that this proportion of the BS receive antennas are an additional set of virtual DL users and null interference at them. Since considering the SI cancellation in the precoder design will decrease the DL transmission gain, we propose a variable $0 \le N_a \le \min(N_r, N_t - K_d)$ to be the number of BS receive antennas at which the ZF precoder will cancel SI by transmit-beamforming. When $N_a = 0$, SI will not be canceled at any BS receive antennas by transmit-beamforming, i.e., the ZF precoder will only null the MUI interference between the DL transmissions. On the other hand, when $N_a = N_r$, the precoder will null SI at all the BS receive antennas. Finally, when $1 < N_a < Nr - 1$, the precoder will null SI at only N_a receive antennas. Accordingly, it should be guaranteed that $N_t \ge (K_d + N_a)$ to have enough dimensions to transmit the DL data and to null interference at the intended DL users and BS receive antennas. The system model is shown in Fig. 1b.

Based on the above assumptions, the received signal at the k^{th} DL user is given by

$$y_{d,k|ZF} = \underbrace{\sqrt{P_d}\mathbf{h}_k \mathbf{w}_k^{ZF} x_k}_{\text{Desired Signal}} + \underbrace{\sum_{i=1}^{K_u} \sqrt{P_u} h_{k,i} s_i}_{\text{CCI}} + n_k^d, \quad (6)$$

where $\mathbf{w}_k^{ZF} \in \mathbb{C}^{(N_t)\times 1}$ is the ZF precoding vector for the k^{th} DL user's data. The second term is the CCI caused by the UL transmissions from K_u UL users. The ZF precoding matrix $\mathbf{F}_{ZF} \in \mathbb{C}^{(N_t)\times(N_a+K_d)}$ and the normalized precoding matrix $\mathbf{W}_{ZF} \in \mathbb{C}^{(N_t)\times(N_a+K_d)}$ are given, respectively, by

$$\begin{aligned} \mathbf{F}_{ZF} &= \mathbf{H}_t^H \left(\mathbf{H}_t \mathbf{H}_t^H\right)^{-1}, \\ \mathbf{W}_{ZF} &= \frac{\mathbf{F}_{ZF}}{\|\mathbf{F}_{ZF}\|_F}, \end{aligned} \quad (7)$$

where, $\mathbf{H}_t = \begin{bmatrix}\mathbf{H}_d & \mathbf{H}_a\end{bmatrix}^T \in \mathbb{C}^{(N_a+K_d)\times(N_t)}$ is the aggregated DL channel matrix which is composed of the DL channel matrix $\mathbf{H}_d = [\mathbf{h}_1\ \mathbf{h}_2 \cdots \mathbf{h}_{K_d}]^T \in \mathbb{C}^{K_d \times N_t}$ and $\mathbf{H}_a \in \mathbb{C}^{N_a \times N_t}$ is the channel between the BS transmit and receive antennas at which the SI is nulled. $\|\mathbf{F}_{ZF}\|_F$ denotes the Frobenius norm of the precoding matrix $\mathbf{F}_{ZF}$. When comparing the expressions in (1) and (6), it can be readily noticed that using the ZF transmit precoder cancels the MUI term. From the expression in (6), the DL received SINR at the k^{th} DL user can be given by

$$\Gamma_{dl}^{ZF} = \frac{P_d \|\mathbf{h}_k \mathbf{w}_k^{ZF}\|^2}{\sigma^2 + \sum_{i=1}^{K_u} P_u \|h_{k,i}\|^2}. \quad (8)$$

On the other hand, in the case of the UL transmission, the received signal at the BS from the l^{th} UL transmission $l \in \{1, 2, \cdots, K_u\}$ is given by

$$y_{u,l|ZF} = \underbrace{\sqrt{P_u}\mathbf{u}_l^{ZF}\mathbf{g}_l s_l}_{\text{Desired Signal}} + \underbrace{\sqrt{C_{BS}}\sqrt{P_d}\mathbf{u}_l^{ZF}\mathbf{H}_{si}^{ZF}\mathbf{W}_{ZF}\mathbf{x}_{ZF}}_{\text{RSI}} + \mathbf{u}_l^{ZF}\mathbf{n}_u, \quad (9)$$

where $\mathbf{u}_l^{ZF} \in \mathbb{C}^{1\times N_r}$ is the ZF receive vector for the k^{th} UL transmission data. The ZF UL receiving matrix $\mathbf{U}_{ZF} \in \mathbb{C}^{K_u \times N_r}$ is given by

$$\mathbf{U}_{ZF} = (\mathbf{G}^H\mathbf{G})^{-1}\mathbf{G}^H, \quad (10)$$

where $\mathbf{G} \in \mathbb{C}^{N_r \times K_u} = \begin{bmatrix}\mathbf{g}_1 & \mathbf{g}_2 & \cdots & \mathbf{g}_{K_u}\end{bmatrix}$ is the UL channel matrix between the UL transmitters and the BS receive antennas [3]. The second term in (9) is the RSI power, $\mathbf{H}_{si}^{ZF} = \begin{bmatrix}\mathbf{H}_a & \mathbf{H}_r\end{bmatrix}^T \in \mathbb{C}^{N_r \times N_t}$ is the channel between the BS transmit and receive antennas. In the case of utilizing the ZF precoder, $\mathbf{H}_{si}^{ZF}$ is composed of $\mathbf{H}_a$ and $\mathbf{H}_r \in \mathbb{C}^{N_r - N_a \times (N_t)}$ which is the channel between the BS transmit antennas and receive antennas at which SI will only be canceled through FD radio SI cancellation and not through the transmit precoder. Moreover, $\mathbf{x}_{ZF} = \begin{bmatrix}\mathbf{x}_d & \mathbf{0}_{N_a}\end{bmatrix}^T \in \mathbb{C}^{(K_d+N_a)\times 1}$ is the aggregated transmitted DL signal which is composed of the DL users' signal $\mathbf{x}_d \in \mathbb{C}^{K_d \times 1}$ and the zero signals $\mathbf{0}_{N_a} \in \mathbb{C}^{N_a \times 1}$ which are sent to the N_a BS receive antennas for SI cancellation. Similarly, When comparing expressions (3) and (9), it can be noticed that using the ZF in the receiver cancels the MUI term. Accordingly the received SINR for the l^{th} UL user, at the BS receive antennas, is given by

$$\Gamma_{ul}^{ZF} = \frac{P_u \|\mathbf{u}_l^{ZF}\mathbf{g}_l\|^2}{\sigma^2\|\mathbf{u}_l^{ZF}\|^2 + C_{BS}P_d\|\mathbf{u}_l^{ZF}\mathbf{H}_{si}^{ZF}\mathbf{W}_{ZF}\mathbf{x}_{ZF}\|^2}, \quad (11)$$

From the expressions in (8) and (11), the sum rate per unit time and bandwidth is given by

$$R_{ZF} = \sum_{k=1}^{K_d} C_1 \log_2(1+\Gamma_{dl}^{ZF}) + \sum_{l=1}^{K_u} C_2 \log_2(1+\Gamma_{ul}^{ZF}). \quad (12)$$

where C_1 and C_2 are constant scaling factors.

IV. Sum Rate Maximization in Full-Duplex MIMO Network

In this section, we study the sum rate maximization problem for the models proposed in Section III. For the model in Section III-A, the optimization will be with respect to N_t, N_r, the precoder vectors, and the receiver vectors. For the model in Section III-B, and since ZF precoder and receiver are used, then the optimization will be with respect to N_t, N_r, and N_a. It should be noted that optimizing over N_t, N_r, and N_a depends on the amount of CSI known at the BS, and therefore, we should also consider the antennas' selection problem while optimizing the

[3]The normalization of the DL precoding matrix in (7) is to limit the power transmitted from the BS. However, in the UL transmission, in (10), the normalization of the UL receiving matrix is not required

values of N_t, N_r, and N_a. In other words, we should select the indices of the transmit and receive antennas and not just the number of antennas in each subset. For that purpose, we define the binary variable $a_{t|n} \in \{0,1\}$, $n \in \{1,2,\cdots N\}$. When $a_{t|n}=1$, then the n^{th} antenna is used as a transmit antenna, otherwise, the n^{th} antenna is used as a receive antenna. Accordingly, $N_t = \sum_{n=1}^{N} a_{t|n}$ and $N_r = N - \sum_{n=1}^{N} a_{t|n} = \sum_{n=1}^{N}(1-a_{t|n})$. Moreover, in the case of ZF, we define the binary variable $a_{sic|n} \in \{0,1\}$, $n \in \{1,2,\cdots N\}$. When $a_{sic|n}=1$, then the SI is nulled at the n^{th} antenna by the ZF precoder. Otherwise, the SI is not cancelled at the n^{th} antenna by the ZF precoder. If the n^{th} antenna is configured to be a transmit antenna, i.e., $a_{t|n}=1$, then $a_{sic|n}$ will be automatically set to zero. Next, we will, respectively, formulate and solve the sum rate maximization problems for the two models presented in Section III-A and Section III-B.

A. Joint Transmitter-Receiver Optimization in Full Duplex MIMO Network

In this case, we maximize the sum rate by optimizing N_t, N_r, the precoder weights, and receiver weights. Additionally, we find the indices of the transmit and receive antennas [4]. Accordingly, the sum rate maximization problem can be formulated as

$$\begin{aligned} \max_{\mathbf{a}_t, \mathbf{W}_d, \mathbf{U}_u} \quad & R_s \\ \text{subject to} \quad & \sum_{n=1}^{N} a_{t|n} \geq K_d, \\ & \sum_{n=1}^{N} (1-a_{t|n}) \geq K_u, \\ & P_d \sum_{k=1}^{K_d} \|\mathbf{w}_k\|^2 \leq P_{max}, \\ & a_{t|n} \in \{0,1\}, \ \forall n \in \{1,\cdots,N\}, \end{aligned} \tag{\textbf{P1}}$$

where $\mathbf{a}_t$ is a vector that includes all the $a_{t|n}$ variables. The first constraint guarantees that the number of transmit antennas is larger than the number of DL users. The second constraint guarantees that the number of receive antennas is larger than the number of UL users. The third constraint is to keep the overall DL transmission power bounded to P_{max}, which is the maximum allowable transmission power [5]. The formulation in (**P1**) doesn't impose minimum-rate constraints on the UL and the DL transmissions. This assumption has been adopted in many research work [8], [49], [50]. However, maximizing the sum rate with no minimum-rate constraints may cause some fairness issues among the users. The authors in [48] have addressed this problem by maximizing the logarithmic utility function, which yields a good balance between system throughput and fairness. The optimization problem in (**P1**) is an MINLP problem, which is very hard to solve due to the presence of $a_{t|n}$'s binary variables. Additionally, it should be noticed that the values and the indices of N_t and N_r control $\mathbf{W}_d$ and $\mathbf{U}_u$ matrices' dimensions, respectively, as $\mathbf{W}_d \in \mathbb{C}^{N_t \times K_d}$ and $\mathbf{U}_u \in \mathbb{C}^{K_u \times N_r}$. Therefore, it is not possible to divide the problem in (**P1**) into two separate sub-problems to deal with the integer and continuous variables. Accordingly, to solve the problem in (**P1**), we start by reformulating the problem in terms of $\mathbf{W}_{ex} \in \mathbb{C}^{N \times K_d}$ which is an extended precoder matrix, $\mathbf{U}_{ex} \in \mathbb{C}^{K_u \times N}$ which is an extended receiver matrix and $\mathbf{A}_{t|ex}$ which is an $N \times N$ diagonal matrix with the elements of $\mathbf{a}_t$ on the diagonal. In this case, changing the value of N_t and N_r will not affect the dimensions of $\mathbf{W}_{ex}$ and $\mathbf{U}_{ex}$, respectively. Next, we define the sum rate in terms of $\mathbf{W}_{ex}$, $\mathbf{U}_{ex}$, and $\mathbf{A}_{t|ex}$ as

$$\begin{aligned} R_{ex} = & \\ & \sum_{k=1}^{K_d} \log_2 \left(1 + \frac{P_d \|\mathbf{h}_{k|ex} \mathbf{A}_{t|ex} \mathbf{w}_{k|ex}\|^2}{\sigma^2 + \sum\limits_{\substack{j=1 \\ j \neq k}}^{K_d} P_d \|\mathbf{h}_{k|ex} \mathbf{A}_{t|ex} \mathbf{w}_{j|ex}\|^2 + \sum\limits_{i=1}^{K_u} P_u \|h_{k,i}\|^2} \right) \\ & + \sum_{l=1}^{K_u} \log_2 \left(1 + \frac{P_u \|\mathbf{u}_{l|ex} (\mathbf{I}_N - \mathbf{A}_{t|ex}) \mathbf{g}_{l|ex}\|^2}{\sigma^2 \|\mathbf{u}_{l|ex}(\mathbf{I}_N - \mathbf{A}_{t|ex})\|^2 + CCI_{ex} + RSI_{ex}} \right), \end{aligned} \tag{13}$$

where

$$CCI_{ex} = \sum_{\substack{i=1 \\ i \neq l}}^{K_u} P_u \|\mathbf{u}_{l|ex}(\mathbf{I}_N - \mathbf{A}_{t|ex}) \mathbf{g}_{i|ex}\|^2, \tag{14}$$

$$RSI_{ex} = C_{BS} P_d \|\mathbf{u}_{l|ex}(\mathbf{I}_N - \mathbf{A}_{t|ex}) \mathbf{H}_{si|ex} \mathbf{A}_{t|ex} \mathbf{W}_{ex} \mathbf{x}_{ex}\|^2, \tag{15}$$

$\mathbf{h}_{k|ex} \in \mathbb{C}^{1 \times N}$ is the extended DL channel between the k^{th} DL user and all the BS antennas, $\mathbf{w}_{k|ex} \in \mathbb{C}^{N \times 1}$ is the k^{th} column of the extended precoder matrix $\mathbf{W}_{ex}$, $\mathbf{u}_{l|ex} \in \mathbb{C}^{1 \times N}$ is the l^{th} row of the extended transmitter matrix $\mathbf{U}_{ex}$, $\mathbf{g}_{l|ex} \in \mathbb{C}^{N \times 1}$ is the extended UL channel between the l^{th} UL user and all the BS antennas and $\mathbf{H}_{si|ex} \in \mathbb{C}^{N \times N}$ is the channel matrix between the BS antennas.

Therefore, the sum rate maximization problem in (**P1***) can be reformulated as

$$\begin{aligned} \max_{\mathbf{A}_{t|ex}, \mathbf{W}_{ex}, \mathbf{U}_{ex}} \quad & R_{ex} \\ \text{subject to} \quad & \mathrm{Tr}(\mathbf{A}_{t|ex}) \geq K_d, \\ & N - \mathrm{Tr}(\mathbf{A}_{t|ex}) \geq K_u, \\ & P_d \|\mathbf{W}_{k|ex}\|^2 \leq P_{max}, \\ & \mathbf{A}_{t|ex}(n,n) \in \{0,1\} \ \forall n \in \{1,\cdots,N\}, \end{aligned} \tag{\textbf{P1*}}$$

where the first constraint ensures that $N_t = \mathrm{Tr}(\mathbf{A}_{t|ex})$, which is the sum of the diagonal elements of $\mathbf{A}_{t|ex}$, exceeds K_d, and the second constraint guarantees that $N_r = N - N_t = N - \mathrm{Tr}(\mathbf{A}_{t|ex})$ exceeds K_u. To solve the problem in (**P1***), which is an MINLP problem, we adopt the GBD algorithm. Next, we will given an overview on the GBD algorithm, and how it could be used to solve the problem in (**P1***).

[4]It should be noted the sum rate maximization formulation suffers from a fairness problem as user with bad channel conditions are normally not assigned any resources. To address this issue some papers, in other contexts, has added a per user minimum rate constraint, e.g., [46]. Some other papers have adopted a different formulation, e.g., the max-min formulation where the minimum data rate is maximized [47]. In [48], authors have addressed this problem by maximizing the logarithmic utility function, which yields a good balance between system throughput and fairness. However, in our work, we focus on the sum rate maximization formulation similar to [8], [49], [50]. We leave addressing the fairness issue as a direction for future work.

[5]It should be noticed that the receiver weights' power is not a function of the UL transmission powers from the UL users. Accordingly, we assume that the UL transmission power P_u is constant, and in that case, in (**P1**), we will not need a constraint to limit the UL transmission power.

Data: $\gamma^\nu \in (0,1]$, $\mathbf{x}^0 \in \mathcal{X}$; set $\nu = 0$.

(S.1) If x^ν is a stationary point of ($\mathcal{P}$): STOP;
(S.2) Compute $\hat{\mathbf{x}}(\mathbf{x}^\nu)$, the solution of ($\mathcal{P}_{x^\nu}$);
(S.3) Set $\mathbf{x}^{\nu+1} = \gamma^\nu(\hat{\mathbf{x}}(\mathbf{x}^\nu) - \mathbf{x}^\nu)$;
(S.4) $\nu \leftarrow \nu + 1$ and go to step (S.1);

Algorithm 1: NOVA Algorithm for ($\mathcal{P}$)

1) GBD Algorithm for Joint Transmitter-Receiver Optimization in FD-MIMO Network : GBD is a procedure to solve MINLP problems that have both continuous and integer variables [51]. The main idea of GBD is to divide the MINLP problem into two sub-problems; the primal problem and the master problem. The primal problem, which is related to the continuous variables, is obtained by fixing the values of $\mathbf{A}_{t|ex}$ variables to a certain feasible point $\overline{\mathbf{A}}_{t|ex}$ and rewrite the problem in terms of $\mathbf{W}_{ex}$ and $\mathbf{U}_{ex}$. The second sub-problem is called the master problem which is only related to the integer variables. These two problems are solved iteratively until their solutions converge. At each iteration, and based on the primal problem feasibility, either an optimality cut or a feasibility cut is added to the master problem.

To solve the MINLP in (**P1***) using the GBD, the first step is to formulate the primal problem, after setting $\mathbf{A}_{t|ex} = \overline{\mathbf{A}}_{t|ex}$, as follows [6]

$$\begin{aligned} \min_{\mathbf{W}_{ex},\mathbf{U}_{ex}} \quad & -R_{ex|\overline{\mathbf{A}}_{t|ex}} \\ \text{subject to} \quad & P_d\|\mathbf{W}_{ex}\|^2 \le P_{max}. \end{aligned} \tag{\textbf{P1.1*}}$$

If the primal problem is feasible, we update the solution upper bound (UBD), then we will need to calculate the optimality cut of this feasible iteration, which will be later used in formulating the master problem. The optimality cut for each $j_1 = 1, \cdots J_1$ of feasible iterations is given by

$$\mathcal{L}(\mathbf{A}_{t|ex}, \mu^{j_1}) = \min_{\mathbf{W}_{ex},\mathbf{U}_{ex}} (-R_{ex} + \mu^T G(\mathbf{A}_{t|ex}, \mathbf{W}_{ex}, \mathbf{U}_{ex})). \tag{16}$$

However, if the primal problem is infeasible then we will need to calculate the feasibility cut of this infeasible iteration, which will be similarly used in formulating the master problem. The feasibility cut, for each $j_2 = 1, \cdots J_2$ of infeasible iterations, is given by

$$\mathcal{L}_2(\mathbf{A}_{t|ex}, \lambda^{(j_2)}) = \min_{\mathbf{W}_{ex},\mathbf{U}_{ex}} (\lambda^T G(\mathbf{A}_{t|ex}, \mathbf{W}_{ex}, \mathbf{U}_{ex})), \tag{17}$$

where the feasibility cut is obtained from solving the feasibility problem. The feasibility problem is given by

$$\max_{\lambda} \min_{\mathbf{W}_{ex},\mathbf{U}_{ex}} \lambda^T G(\overline{\mathbf{A}}_{t|ex}, \mathbf{W}_{ex}, \mathbf{U}_{ex}), \tag{\textbf{Q}}$$

where $G(\mathbf{A}_{t|ex}, \mathbf{W}_{ex}, \mathbf{U}_{ex}))$ denotes the constraint functions defined in (**P1**). The next step is to formulate the master problem as follows,

$$\begin{aligned} \min_{\mathbf{A}_{t|ex},\gamma} \quad & \gamma \\ \text{subject to} \quad & \gamma \ge \mathcal{L}(\mathbf{A}_{t|ex}, \mu^{j_1}),\ j_1 = 1, \cdots, J_1, \\ & 0 \ge \mathcal{L}_2(\mathbf{A}_{t|ex}, \lambda^{j_2}),\ j_2 = 1, \cdots, J_2. \end{aligned} \tag{\textbf{P1.2*}}$$

After solving the master problem, we update the solution's lower bound (LBD). As previously mentioned, both the primal and the master problems are solved iteratively till their solutions, i.e., the UBD and the LBD, respectively, converge.

To solve the primal problem in (**P1.1****), we adopt the iNner cOnVex Approximation (NOVA) algorithm proposed in [52], [53]. The idea of the NOVA algorithm is to solve a sequence of strongly convex inner approximations of (**P1.1****). Next, we will give a brief overview of the NOVA algorithm, and discuss how it could be used to solve the primal problem in (**P1.1****).

2) An Overview of the NOVA Algorithm for Constrained Non-convex Optimization: Assume a minimization problem of a non-convex objective function $Q : \mathcal{K} \to \mathbb{R}$ with a convex set $\mathcal{K}$ subject to some convex constraints and some non-convex constraints $z_j(x) \le 0$, with $z_j : \mathcal{K} \to \mathbb{R}$,

$$\begin{aligned} \min_{\mathbf{x}} \ & Q(\mathbf{x}) \\ \text{subject to} \quad & z_j(\mathbf{x}) \le 0,\ j = 1, \cdots, m, \\ & x \in \mathcal{K}, \end{aligned} \tag{$\mathcal{P}$}$$

where the feasible region of ($\mathcal{P}$) is denoted by $\mathcal{X}$. The main goal of the NOVA algorithm is to efficiently compute local optimal solutions of ($\mathcal{P}$), while preserving the feasibility of the iterations [7]. Given $x^\nu \in \mathcal{X}$, the approach solves a sequence of strongly convex inner approximations of ($\mathcal{P}$) of the form:

$$\begin{aligned} \min_{\mathbf{x}} \ & \tilde{Q}(\mathbf{x}; \mathbf{x}^\nu) \\ \text{subject to} \quad & \tilde{z}_j(\mathbf{x}, \mathbf{x}^\nu) \le 0,\ j = 1, \cdots, m, \\ & x \in \mathcal{K}, \end{aligned} \tag{$\mathcal{P}_{x^\nu}$}$$

where $\tilde{Q}(\mathbf{x}; \mathbf{x}^\nu)$ and $\tilde{z}_j(\mathbf{x}, \mathbf{x}^\nu)$ represent approximations of $Q(\mathbf{x})$ and $z_j(\mathbf{x})$ at the current iteration $\mathbf{x}^\nu$, respectively, and with $\mathcal{X}(\mathbf{x}^\nu)$ as the feasible region of ($\mathcal{P}_{x^\nu}$). $\tilde{Q}(\mathbf{x}; \mathbf{x}^\nu)$ is a strong convex function on $\mathcal{K}$, and $\tilde{z}_j(\cdot, \mathbf{y})$ is convex over $\mathcal{K}$ for all $\mathbf{y} \in \mathcal{X}$. It can be shown that each sub-problem ($\mathcal{P}_{x^\nu}$) is strongly convex, and thus has a unique solution which is denoted by $\hat{\mathbf{x}}(\mathbf{x}^\nu)$:

$$\hat{\mathbf{x}}(\mathbf{x}^\nu) \triangleq \operatorname*{\mathbf{argmin}}_{\mathbf{x} \in \mathcal{X}(\mathbf{x}^\nu)} \tilde{Q}(x; x^\nu). \tag{18}$$

Afterwards, Algorithm 1 is utilized to obtain ($\mathcal{P}_{x^\nu}$) solution. Starting from an initial feasible point $\mathbf{x}^0$, the proposed method iteratively computes $\hat{\mathbf{x}}(\mathbf{x}^\nu)$, the solution to the problem in (18), and then taking a step from $\mathbf{x}^\nu$ towards $\hat{\mathbf{x}}(\mathbf{x}^\nu)$. In Step S.1, it is required to check the stationarity of the current solution. A suitable termination check in Step S.1 is $\|\hat{\mathbf{x}}(\mathbf{x}^\nu) - \mathbf{x}^\nu\| \le \epsilon_n$, where ϵ_n is the desired accuracy. Additionally, we use a diminishing step size rule to update γ^ν:

$$\gamma^\nu = \gamma^{\nu-1}(1 - \varepsilon\gamma^{\nu-1}),\ \nu \ge 1, \tag{19}$$

[6]GBD requires that the initial value $\overline{\mathbf{A}}_{t|ex}$ should be any feasible point in the optimization problem set that satisfies the first two constraints in (**P1***).

[7]For the interested reader, the convergence requirements of NOVA algorithm are stated in [52].

Data: all CSI information, ϵ^0

1) Initialize the upper bound UBD and the lower bound LBD
2) **if** $UBD \leq LBD + \epsilon^0$ **then**
   $(\mathbf{A}^*_{t|ex}, \mathbf{W}^*_{ex}, \mathbf{U}^*_{ex})$ is the problem solution;
   else
   Solve the Primal Problem
   if *the primal problem (**P1.1****) is feasible* **then**
   Solve the primal problem and obtain the solution and multipliers $(\mathbf{W}^*_{ex}, \mathbf{U}^*_{ex}, \mu^*)$;
   Determine $\mathcal{L}(\mathbf{A}_{t|ex}, \mu^{K_1})$;
   If $-R_{ex}(\overline{\mathbf{A}}_{t|ex}, \mathbf{W}^*_{ex}, \mathbf{U}^*_{ex}) < UBD$, set $UBD = -R_{ex}(\overline{\mathbf{A}}_{t|ex}, \mathbf{W}^*_{ex}, \mathbf{U}^*_{ex})$;
   else
   Solve the feasibility Problem in $\mathbf{Q}$ and Obtain its solution $(\mathbf{W}^*_{ex}, \mathbf{U}^*_{ex}, \lambda^*)$;
   Determine $\mathcal{L}_2(\mathbf{A}_{t|ex}, \lambda^{(J_2)})$;
   end
   end
3) Solve the master problem in (**P1.2***) and obtain its solution $(\gamma^*, \mathbf{A}^*_{t|ex})$;
4) Set the $LBD = \gamma^*$;
5) Set $\overline{\mathbf{A}}_{t|ex} = \mathbf{A}^*_{t|ex}$;
6) Return to Step 2;

Algorithm 2: Joint Transmitter-Receiver Optimization in Full Duplex MIMO Network

where $\gamma^0 \in (0, 1]$ and $\varepsilon \in (0, 1)$. For more details on the NOVA algorithm and the choice of γ, ϵ, and ε, please refer to [52], [53]. In the proposed primal problem in (**P1.1****), we only need to derive an approximation $\tilde{R}_{ex}$ for the objective function R_{ex}. A valid choice of $\tilde{R}_{ex}$, as suggested by [52], is the first order approximation of R_{ex}, that is, $\tilde{R}_{ex} = \sum_{i=1}^{I} \tilde{R}_{ex}(\mathbf{x}_i, \mathbf{x}^\nu)$, with each

$$\tilde{R}_{ex}(\mathbf{x}_i, \mathbf{x}^\nu) \triangleq \nabla_{\mathbf{x}_i} R_{ex}(\mathbf{x}^\nu)^T(\mathbf{x_i} - \mathbf{x}_i^\nu) + \frac{\tau_i}{2}\|\mathbf{x}_i - \mathbf{x}_i^\nu\|^2 \quad (20)$$

where $\mathbf{x}_i$ is the i^{th} block of variables. In our problem, $\{\mathbf{x}_1, \cdots \mathbf{x}_{K_d}, \mathbf{x}^T_{K_d+1} \cdots, \mathbf{x}^T_{K_d+K_u}\} = \{\mathbf{w}_{1|ex}, \cdots, \mathbf{w}_{K_d|ex}, \mathbf{u}^T_{1|ex}, \cdots, \mathbf{u}^T_{K_u|ex}\}$. Accordingly, the primal problem approximation is given by,

$$\begin{aligned} \min_{\mathbf{W}_{ex}, \mathbf{U}_{ex}} \quad & -\tilde{R}_{ex|\overline{\mathbf{A}}_{t|ex}} \\ \text{subject to} \quad & P_d\|\mathbf{W}_{ex}\|^2 \leq P_{max}. \end{aligned} \qquad \textbf{(P1.1**)}$$

Afterwards, we can adopt Algorithm 1 to solve the primal problem approximation in (**P1.1****). The complete solution algorithm for (**P1***) is presented in Algorithm 2. After initializing the UBD, and the LBD, we solve the primal problem with the NOVA algorithm, i.e., Algorithm 1. Afterwards, based on the primal problem feasibility, we will add either an optimality cut or a feasibility cut to the master problem. The next step is to solve the master problem and update the LBD. The algorithm will keep on iterating between the primal and the master problems till their solutions converge, as indicated in the second step in Algorithm 2.

B. Joint Transmit Antennas, Receive Antennas, and SI Cancellation Optimization in Full Duplex MIMO Network with Zero-Forcing Precoder and Zero-Forcing Receiver

In this case, we try to find the optimal values for N_t, N_r, and N_a that maximize the network sum rate. Additionally, we need to find the indices of the transmit antennas, receive antennas, and the receive antennas at which the SI will be nulled. Accordingly, the sum rate maximization problem can be written as

$$\begin{aligned} \max_{N_t, N_r, N_a, \mathbf{a}_t^{ZF}, \mathbf{a}_{sic}} \quad & R_s \\ \text{subject to} \quad & N_t \geq K_d + N_a, \\ & N_r \geq K_u, \\ & 0 \leq N_a \leq \min(N_r, N_t - K_d), \\ & \sum_{n=1}^{N} a_{t|n} = N_t, \\ & \sum_{n=1}^{N} (1 - a_{t|n}) = N_r, \\ & \sum_{n=1}^{N} (1 - a_{t|n})(a_{sic|n}) = N_a, \end{aligned} \qquad \textbf{(P2)}$$

where $\mathbf{a}_t^{ZF}$ is a vector including all the $a_{t|n}$ coefficients and $\mathbf{a}_{sic}$ is a vector including all the $a_{sic|n}$ coefficients. The optimal solution of (**P2**) can be obtained by brute-force exhaustive searching among $\sum_{N_t=K_d}^{N-K_u} \binom{N}{N_t} \Big[\binom{N_r}{1} + \binom{N_r}{2} + \cdots + \binom{N_r}{\min(N_r, N_t - K_d)} \Big]$ cases. Therefore, the exhaustive search is impractical to implement for a relatively large number of antennas. Accordingly, we propose a heuristic two-step antenna selection algorithm; in the first step, we determine the indices of the transmit and receive antennas. In the second step, we determine the indices of the receive antennas at which SI will be cancelled by transmit precoder.

1) Transmit and Receive antennas' selection in FD-MIMO Networks: In this section, we explain the proposed iterative algorithm for selecting the transmit and receive antennas for the model described in Section III-B. The algorithm steps are illustrated in Algorithm 3. Initially, it is assumed that the antennas are randomly selected to be either transmit or receive antennas while satisfying the first two constraints stated in (**P2**). From the CSI information at the BS, the BS antennas are sorted in descending order based on the sum of the UL and DL channel gains to all the UL and DL users, respectively. The first step is to start with the antenna with the largest sum of UL and DL channel gains and checking whether fixing this antenna to be a transmit or receive antenna will be more beneficial for the sum rate. At this step, we do not consider SI cancellation via ZF transmit beamforming. After we determine the mode of the first antenna, we move to the antenna with the second-largest sum of UL and DL channel gains. Similarly, and by knowing the first antenna's mode, the second antenna's mode is chosen to achieve a higher gain in the sum rate. Subsequently, the algorithm decides the mode of all the other remaining antennas. It should be noted that in each step, the algorithm checks that setting the antenna's mode will not violate the constraints on the number of antennas. After

Data: all CSI information, $\mathbf{A}_t^0$, ϵ, $R_S(\mathbf{A}_t^0)$;
Result: Find $\mathbf{A}_t^* = [a_{t|1}, a_{t|2}, \cdots, a_{t|N}]^T$;

Initially:
$\mathbf{A}_t = \mathbf{A}_t^0$, Iteration $j = 1$;

1: Sorting

(a) Calculate $\lambda_n = \sum_{k=1}^{K_d} \|h_k^n\|^2 + \sum_{l=1}^{K_u} \|g_l^n\|^2 \ \forall \ n \in \{1..N\}$;

(b) Form $\mathbf{\Lambda}$ such that $\mathbf{\Lambda}(1) = n(\max(\lambda_n))$, $\mathbf{\Lambda}(N) = n(\min(\lambda_n))$;

2: Sequential Transmit/Receive Antenna Selection
For each antenna: $(a_{t|\Lambda(i)} = 1)$ **if** $R_{ZF}|(a_{t|\Lambda(i)} = 1) \geq R_{ZF}|(a_{t|\Lambda(i)} = 0)$ **else**$(a_{t|\Lambda(i)} = 0)$

3: Check Convergence
(a) **if** $R_{ZF}(\mathbf{A}_t^j)$-$R_{ZF}(\mathbf{A}_t^{j-1}) \leq \epsilon_1$ **then**
  $\mathbf{A}_t^* = \mathbf{A}_t^j$;
else
  j=j+1;
  Return to Step. 2;
end

Algorithm 3: Transmit and Receive Antenna Selection is FD-MIMO Networks

selecting the modes for all the antennas, the algorithm reiterates the selection process starting from the first antenna until the change in the sum rate is within a certain tolerance. It should be noticed that the antenna selection process is performed at the BS. Additionally, the complexity of Algorithm 3 can be obtained adding the complexity of each operation in the algorithm and choose the worst-case term, i.e., the highest order term, which dictates the algorithm limiting behaviour. Accordingly, Algorithm 3 has a computational complexity of $\mathcal{O}(N^2)$, i.e., polynomial complexity.

2) Antenna Selection for SI-Nulling is FD-MIMO Networks: In this section, we explain the proposed iterative algorithm for selecting the receive antennas at which the SI is nulled by ZF transmit beamforming. Initially, we consider no SI-nulling at any receive antenna. By using a similar approach to the algorithm in Section IV-B1, each receive antenna will be ranked according to the gain in sum rate achieved when SI-nulling is done exclusively at this antenna. Then, the algorithm checks if nulling the SI on the highest-gain antenna will increase or decrease the sum rate. If the SI-nulling on this antenna increases the sum rate, then the value of $a_{sic|n} = 1$, otherwise, $a_{sic|n} = 0$. Afterward, the algorithm proceeds to the antenna with the second-highest gain and based on the decision taken on the first antenna, the algorithm decides whether or not to null the SI on the second antenna. This process repeats for each receive antenna. Note that, in each step, the algorithm checks that the constraints on the number of antennas ae not violated. Subsequently, the algorithm reiterates the selection process until the change in the sum rate is within a certain tolerance. The complete procedure is illustrated in Algorithm 4.

Assuming that there is no SI-nulling on any of the receive antenna, the algorithm calculates the sum rate $R_{sn}^{'}$ when SI is canceled on the n^{th} antenna exclusively. Afterward, the antenna indices are sorted in the vector $\mathbf{C}$ in a descending order based on the highest sum rate gain, i.e., $\mathbf{C}(1) = n(\max(R_{sn}^{'}))$, $\mathbf{C}(N) = n(\min(R_{sn}^{'}))$. The next step is to check the effect of canceling the SI at more receive antennas on the sum rate. To do this, we first assume, while satisfying the constraints on the number of antennas, that we cancel the SI on the first receive antenna index in $\mathbf{C}(1)$, i.e., $a_{sic|\mathbf{C}(i)} = 1$. Afterward, with SI cancellation at $\mathbf{C}(1)$, the algorithm checks if additionally canceling SI at $\mathbf{C}(2)$ will increase the sum rate. If this additional SI cancellation increases the sum rate then we set $a_{sic|\mathbf{C}(2)} = 1$; otherwise $a_{sic|\mathbf{C}(i)} = 0$. This process repeats for all receive antennas. Afterward, the algorithm iterates until a convergence, within a certain tolerance, ϵ_2, is reached. Note that, The antenna selection process for SI nulling is performed at the BS. Finally, by adopting the same calculation procedure used for Algorithm 3, we can find that Algorithm 4 has a computational complexity of $\mathcal{O}(N^2)$, i.e., polynomial complexity.

V. Asymptotic Sum Rate Behavior or Full-Duplex MIMO Networks with Zero-Forcing

In this section, we study the asymptotic rate behavior when $N = N_t + N_r \to \infty$. We assume that the BS has a ZF precoder and a ZF receiver. In this case, we can assume that $N_a = \alpha N_r$. The first step in studying the asymptotic behavior is to evaluate how the DL and the UL rates are affected when the number of antennas tends to infinity. First, in the following proposition, we start by studying the behavior of the DL transmission.

Proposition 1. *A lower bound for the total FD DL rate R_d is given by*

$$R_d = K_d \log_2 \left(1 + \frac{N_t - (K_d + N_a)}{K_d + N_a} \left(\frac{P_d}{\sigma^2 + P_u K_u} \right) \right). \quad (21)$$

Proof: From (8), the total DL rate is the sum of the DL rates from the DL-HD users and the FD nodes. Accordingly, the lower bound of the total DL rate can be derived as

$$\begin{aligned} R_d &= K_d \mathbb{E} \left[\log_2 \left(1 + \Gamma_{dl}^{ZF} \right) \right], \\ &\overset{(a)}{\geq} K_d \log_2 \left\{ 1 + \left[\mathbb{E} \left(\frac{1}{\Gamma_{dl}^{ZF}} \right) \right]^{-1} \right\} \\ &\overset{(b)}{=} K_d \log_2 \left\{ 1 + \left[\mathbb{E} \left(\frac{\sigma^2 + \sum_{i=1}^{Ku} P_u \|h_{k,i}\|^2}{P_d / \|\mathbf{F}_{ZF}\|_F^2} \right) \right]^{-1} \right\} \\ &\overset{(c)}{=} K_d \log_2 \left\{ 1 + P_d \frac{N_t - (K_d + N_a)}{(K_d + N_a)\left(\sigma^2 + \mathbb{E}\{\sum_{i=1}^{K_u} P_u \|h_{k,i}\|^2\}\right)} \right\} \\ &\overset{(d)}{=} K_d \log_2 \left\{ 1 + \frac{N_t - (K_d + N_a)}{K_d + N_a} \left(\frac{P_d}{\sigma^2 + P_u K_u} \right) \right\}, \end{aligned} \quad (22)$$

where (a) results from applying Jensen inequality for the convex function $\log_2(1 + 1/x)$ and assuming that, in massive MIMO systems, each HD-DL user from the K_d users will "asymptotically" experience the same receive SINR as a result of channel hardening [54], [55] . (b) follows from the ZF array gain properties, in which $\mathbf{F}_{ZF}\mathbf{H}_t = \mathbf{I}_{K_d+N_a}/\|\mathbf{F}_{ZF}\|_F$. Therefore,

Data: all CSI information, $\mathbf{A}_t$, ϵ_2, $R_S(\mathbf{A}_t)$
Result: Find $\mathbf{A}^*_{sic}$

Initially:
$a_{sic|n} = 0$, $\forall\ n \in \{1, 2, \cdots, N_r\}$, Iteration $j = 1$;

1. **for** $n = 1 : N_r$ **do**
   (a) Set $a_{sic|m} = 0\ \forall m \neq n \in \{1, 2, \cdots, N_r\}$;
   Calculate $R^{'}_{sn}$;
end
2. Form a vector $\mathbf{C}$ such that $\mathbf{C}(1) = n(\max(R^{'}_{sn}))$, $\mathbf{C}(N) = n(\min(R^{'}_{sn}))$;
3. For each antenna, **if** $R_{ZF}|(a_{sic|\mathbf{C}(i)} = 1) \geq R_{ZF}|(a_{sic|\mathbf{C}(i)} = 0)$ **then** $a_{sic|\mathbf{C}(i)} = 1$
4. **if** $R_{ZF}(\mathbf{A}^j_{sic})$-$R_{ZF}(\mathbf{A}^{j-1}_{sic}) \leq \epsilon_2$ **then**$\mathbf{A}^*_{sic}$ = $\mathbf{A}^j_{sic}$ **else** j=j+1; Return to Step 3;

Algorithm 4: Antenna Selection for SI-Nulling

$\mathbf{h}_k\mathbf{w}_i = \delta_{ki}/\|\mathbf{F}_{ZF}\|_F$, *where* $\delta_{ki} = 1$ *when* $k = i$ *and* $\delta_{ki} = 0$ *otherwise. Additionally, (c) follows by substituting the value of* $\|\mathbf{F}_{ZF}\|_F$ *calculated in [56]. Finally, (d) is obtained by knowing that* $\sum_{i=1}^{K_u} \|h_{k,i}\|^2 \sim E(K_u, 1)$ *which is the Erlang distribution with* K_u *shape parameter and unity rate parameter, and similarly,* $\sum_{\substack{i=1\\ i\neq m}}^{K_u} \|h_{k,i}\|^2 \sim E(K_u - 1, 1)$ *which is the Erlang distribution with* $(K_u - 1)$ *shape parameter and unity rate parameter.*

In the following proposition we provide a lower bound for the UL rate when $N \to \infty$.

Proposition 2. *A lower bound for the FD UL rate* R_u *is given by*

$$R_u = K_u \log_2\left(1 + \frac{P_u(N_r - K_u)}{\sigma^2 + C_{BS}P_d(1-\alpha)}\right). \tag{23}$$

The expression in (23) can be derived following the derivation in [56, Proposition 2].

In order to clarify the trade-off between the DL rate and SI cancellation by transmit beamforming, the lower bounds of the DL and UL rates, given in (21) and (23), respectively, in terms of $\beta = N_t/N_r$ and $\alpha = N_a/N_r$ are written as follows

$$\begin{aligned} R_d(\beta,\alpha) &= K_d \log_2\left(1 + \left(\frac{P_d}{\sigma^2 + P_uK_u}\right)\left(\frac{\beta}{\frac{K_d(\beta+1)}{N} + \alpha} - 1\right)\right) \\ R_u(\beta,\alpha) &= K_u \log_2\left(1 + \frac{P_u(\frac{N}{\beta+1} - K_u)}{\sigma^2 + C_{BS}P_d(1-\alpha)}\right). \end{aligned} \tag{24}$$

It can be readily verified, based on the expressions derived in (24), that increasing α leads to better SI cancellation and higher UL rate; however, it causes a decrease in the DL rate. Therefore, it is required to optimize β, as well as α, to maximize (enlarge) the network's rate region.

A. Antenna Ratio and Self-Interference Cancellation Optimization for Rate Region Maximization

In this section, we start by deriving a feasible rate region which is obtained by calculating the maximum value of $R_d(\beta,\alpha)$ [8], then for each possible value of $R_d(\beta,\alpha) = R$, we optimize β and α to maximize $R_u(\beta,\alpha)$ [9]. By setting the DL rate to a constant value R, the relation between β and α is given by

$$\beta\left(1 - \Psi\frac{K_d}{N}\right) = \Psi\left(\frac{K_d}{N} + \alpha\right), \tag{25}$$

where $\Psi = 1 + \frac{(\sigma^2+P_uK_u)(2^{R/K_d}-1)}{P_d}$. Since $\Psi > 1$, the relation in (25) satisfies that $\beta > (\alpha + K_d/N)/(1 - K_d/N)$. Therefore, the feasible rate region maximization problem can be given by

$$\begin{aligned} \max_{\beta,\alpha} \quad & R_u(\beta,\alpha) \\ \text{subject to} \quad & 0 \leq \alpha \leq 1, \\ & \beta > \frac{\alpha + \frac{Kd}{N}}{1 - \frac{Kd}{N}}, \\ & \beta\left(1 - \Psi\frac{K_d}{N}\right) = \Psi\left(\frac{K_d}{N} + \alpha\right). \end{aligned} \tag{26}$$

To find the solution of rate region maximization problem in (26), we will first study the convexity of the objective function in the following proposition.

Proposition 3. *The total UL rate* $R_u(\beta,\alpha)$*, defined in (24), is a convex function in both* β *and* α *given that* $N_r \geq \frac{4}{3}K_u$.

The proof for **Proposition 3** is presented in [56, Appendix B].

Therefore, we can reformulate the rate maximization in terms of β or α without affecting the convexity of the problem. After substituting the value of β in terms of α in (26), the rate maximization problem is given by

$$\begin{aligned} \max_{\alpha} \quad & R_u(\alpha) \\ \text{subject to} \quad & 0 \leq \alpha \leq 1. \end{aligned} \tag{\textbf{P3}}$$

Since the objective function is convex [10], the maximization problem's solution lies on the boundaries of the feasible set defined by the constraints in (**P3**). Therefore, after some simple mathematical manipulation, it can be readily shown that the values of α and β maximizing R_u for given R_d are given by

$$\alpha^* = 1, \text{ and } \beta^* = \frac{\Psi(1 + \frac{K_d}{N})}{1 - \Psi(\frac{K_d}{N})}. \tag{27}$$

On the other hand, the rate maximization problem can also be derived by setting the value of $R_u(\beta,\alpha)$ to a certain constant

[8] The maximum DL rate is achieved when the UL transmission is inactive

[9] The rate region can also be obtained by calculating the maximum value of $R_u(\beta,\alpha)$, i.e., when the DL transmission is inactive. Then for each possible value of $R_u(\beta,\alpha)$, we optimize β and α to maximize $R_d(\beta,\alpha)$.

[10] It should be noticed that the UL rate is monotonically increasing with α. However, the rate maximized in (**P3**) is the UL rate at a constant DL rate, which is a convex function in α. This is due to the fact that, for small values of α, increasing α will increase β to keep R_d constant, which will decrease the UL rate. However, at large values of α, i.e., more SI cancellation, the effect of increasing β will be less significant on the UL rate and accordingly increasing α will increase the UL rate. Accordingly, the UL rate, at a constant DL rate, will be convex function in α

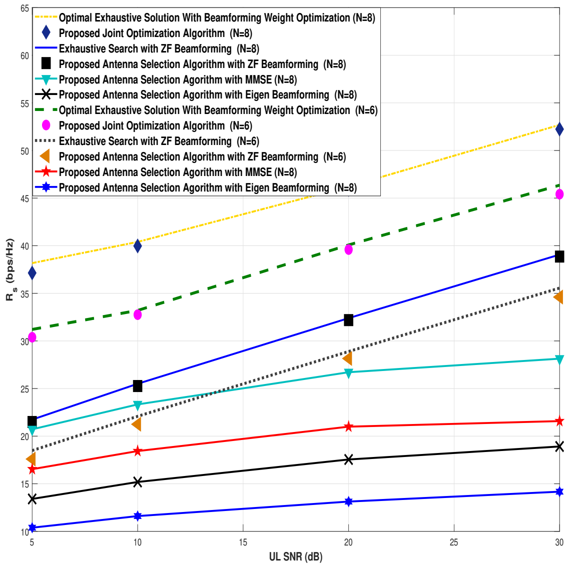


Fig. 2: Variation of the sum rate with the UL SNR for $N = 6$, $K_d = 3$, and $K_u = 2$ and $N = 8$, $K_d = 3$, and $K_u = 2$.

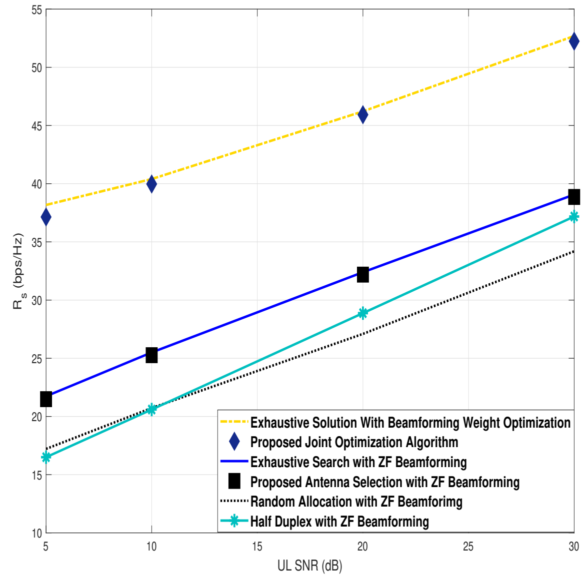


Fig. 3: Variation of the sum rate with the UL SNR for different antennas' selection algorithms in a FD network.

value $R^{'}$. Then, for each possible value of $R^{'}$, we optimize β and α to maximize $R_d(\beta, \alpha)$. Therefore, for a constant $R_u(\beta, \alpha) = R^{'}$, the relation between β and α is given by

$$\beta = \frac{NP_u}{P_u K_u + (2^{R^{'}/K_u} - 1)(\sigma^2 + CP_d(1-\alpha))} - 1. \tag{28}$$

It should be noted that in order to guarantee that $\beta > (\alpha + (K_d/N))/(1 - (Kd/N))$, then the FD radio should satisfy the following condition

$$C_{BS} \leq \frac{1}{Pd}\left(\frac{P_u(N-K_u)}{2^{R^{'}/K_u} - 1} - \sigma^2\right). \tag{29}$$

Accordingly, the rate maximization problem is given by

$$\begin{aligned} \max_{\alpha} \quad & R_d(\alpha) \\ \text{subject to} \quad & 0 \leq \alpha \leq 1. \end{aligned} \tag{\textbf{P4}}$$

In that case, the optimization problem defined in (**P4**) is not convex. Therefore, we will derive the Lagrangian function and apply the Karush-Kuhn-Tucker (KKT) conditions [57], [58] to solve the dual problem of (**P4**). The dual problem of (**P4**) is given by

$$\begin{aligned} \max_{\alpha, \lambda_1, \lambda_2} \quad & \mathfrak{L}(\alpha, \lambda_1, \lambda_2) = R_d - \lambda_1(\alpha) + \lambda_2(1-\alpha) \\ \text{subject to} \quad & 0 \leq \alpha \leq 1. \end{aligned} \tag{30}$$

The solution of the dual problem is given in the following proposition.

Proposition 4. *The solution of the dual problem using the KKT conditions is given by*

$$\alpha = 0, \lambda_1 = \lambda_2 = 0. \tag{31}$$

Proof: The system of equations obtained from applying the KKT conditions is given by

$$\begin{aligned} & \frac{\partial \mathfrak{L}(\alpha, \lambda_1, \lambda_2)}{\partial \alpha} = 0 \\ & \lambda_1 \times \alpha = 0 \\ & \lambda_2 \times (1-\alpha) = 0. \end{aligned} \tag{32}$$

After some simple mathematical steps, it can be shown that the solution of the dual problem in (30) satisfying (32) is given by $\alpha = 0, \lambda_1 = \lambda_2 = 0$.

Note that obtaining the feasible rate region by solving (**P3**) is more straightforward than solving (**P4**), due to the convexity of the optimization problem function in (**P3**) with respect to α.

VI. Numerical Analysis

In this section, we simulate a square grid network with the BS in its center and the users are randomly located within the grid. or a normal-scale FD MIMO setting, we evaluate the performance of the joint optimization problem proposed in Section IV-A and compare it to the joint antenna selection and SI cancellation with ZF receiver and precoder proposed in Section IV-B. Additionally, since the problems proposed in Section IV-A and Section IV-B are MINLP and integer programming (IP) problems, respectively, their performance should be validated by comparing it to the optimal solution. For the problem in Section IV-A, the optimal solution is obtained by exhaustive search, i.e., by enumerating all possible combinations of transmit and receive antennas' selections. Then, for each combination, we optimize over the precoder and the receiver weights. Finally, we choose the combination with the highest sum rate. Similary, for the problem in Section IV-B, the optimal solution is obtained by exhaustive search. Finally, we compare the performance of the joint antenna selection and SI cancellation to that of the ZF receiver and precoder proposed in

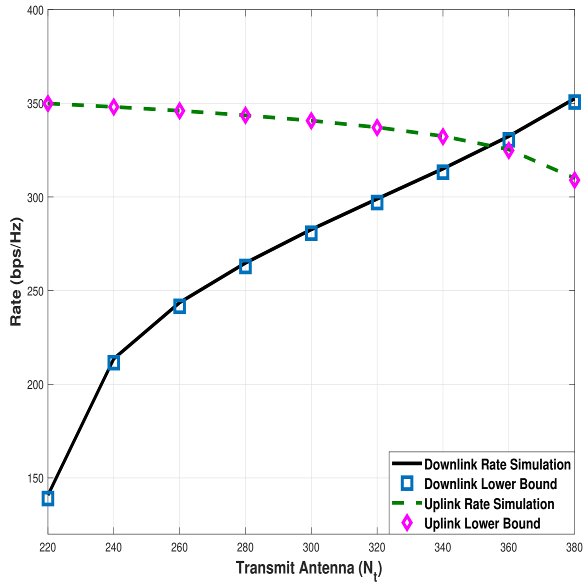


Fig. 4: validating derived downlink and uplink bounds for different transmit antennas. Parameters used to generate this figure: $N = 400$, $K_d = 30$, $K_u = 10$, $\alpha = 1$, $\sigma^2 = -110dB$, $C = 80dB$, $P_d = 10W$ and $P_u = 2mW$.

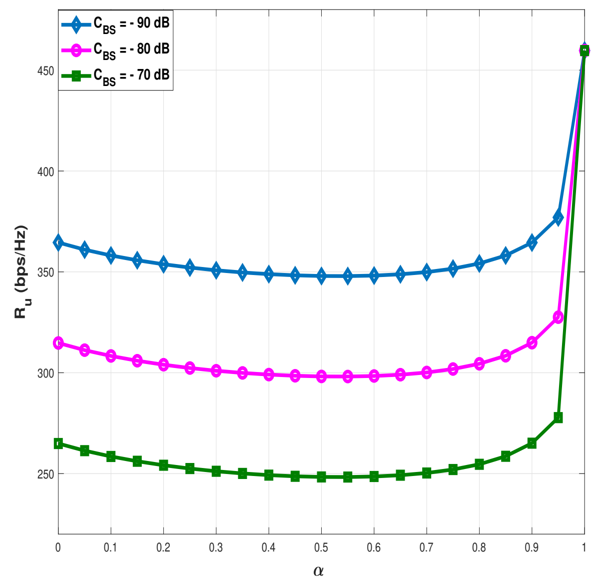


Fig. 5: Validating rate maximization problem solution when fixing $R_d = R$ and maximizing R_u. Parameters used to generate this figure: $N = 200$, $K_d = 20$, $K_u = 15$, $\sigma^2 = -110dB$, $R_d = R = 200bps/Hz$, $P_d = 10W$ and $P_u = 2mW$.

Section IV-B with the rate region maximization derived in Section V.

For a normal-scale FD-MIMO setting, Fig. 2 shows the variation of the sum rate with the uplink signal-to-noise (SNR) defined as P_u/σ^2 and compare the performance of the algorithms proposed in Section IV-A and Section IV-B to the optimal exhaustive search based solution. First, the network is simulated for $N = 6$, $K_d = 3$, and $K_u = 2$. From the results in Fig. 2, it is clear that increasing the UL SNR under constant UL transmission power, i.e., decreasing the noise variance σ^2, increases the total sum rate of the network. Additionally, when comparing the joint sum rate maximization scheme proposed in Section IV-A to the optimal exhaustive search, we can validate that the proposed algorithm can achieve near-optimal performance. Similarly, when considering SI cancellation via ZF, we can verify that the proposed antenna selection algorithm in Section IV-B can achieve near-optimal performance. However, when comparing the achievable sum rate of both algorithms, it is clear that the first algorithm achieves better performance. As mentioned before, increasing the number of required interference-nulls decreases the degrees-of-freedom in the DL transmission, and hence, the DL transmission gain, and accordingly the sum rate. Then, we can conclude that optimizing the precoder and transmitter weights with the antenna selection achieves better performance than optimizing the antenna selection under the restriction of ZF precoder and ZF transmitter. For further validation, the network is simulated for $N = 8$, $K_d = 3$, and $K_u = 2$. Similarly, increasing the UL SNR by decreasing the noise variance increases the sum rate, R_s. Additionally, the results verify that the proposed algorithms can approximate the optimal solution at a much lower complexity. Finally, to study the effect of SI cancellation via ZF, the network is simulated using minimum mean-squared-error (MMSE) beamforing and EB. From the results, we can see the effectiveness of SI cancellation via ZF as it outperforms both MMSE and EB. Additionally, we can notice that the performance gap between ZF and MMSE, or EB, increases with the UL SNR. This result is because increasing the UL SNR by decreasing σ^2 will increase the SI in the UL transmission. Accordingly, both the MMSE and the EB sum rates will be much lower than the "ZF with SI cancellation" sum rate.

Fig. 3 compares the performance of the proposed algorithms to the performance of random antenna selection and half-duplex transmission for a network with $N = 8$, $K_d = 3$, and $K_u = 2$, with ZF precoder and receiver. In the random selection scheme, the antennas are randomly selected to be either transmit or receive antennas while satisfying the constraints that $N_t \geq K_d$ and $N_r \geq K_u$. We study the effect of optimizing the antenna selection process with ZF precoder and receiver, as proposed in Section IV-B. As expected, selecting transmit and receive antennas randomly degrades the FD performance, and accordingly, both algorithms in Section IV-A and Section IV-B outperform random antenna selection. This result validates that the random selection of transmitting and receiving antennas might miss better selection chances, which can potentially achieve a higher sum rate. Finally, when compared to half-duplex communication, the proposed algorithm can effectively improve the network's sum rate. However, the results show that increasing the UL SNR narrows the performance-gap between FD and HD sum rates. This result is because increasing the UL SNR, by decreasing σ^2, will increase the CCI in the DL transmission and the SI in the UL transmission. Accordingly, the improvement in the FD sum rate will be slower than in the HD sum rate. Hence, the performance-gap will narrow.

For a massive MIMO network, we start by validating the bounds derived in (21) and (23). The system is simulated with

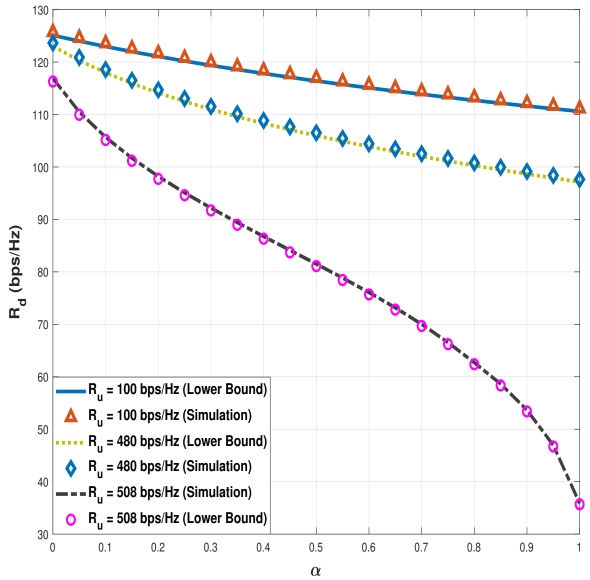


Fig. 6: Validating rate maximization problem solution when fixing R_u and maximizing R_d. Parameters used to generate This figure: $N = 200$, $K_d = 10$, $K_u = 15$, $\sigma^2 = -110dB$, $P_d = 10W$ and $P_u = 2mW$,for $R_u = 100, 480, 508bps/Hz$.

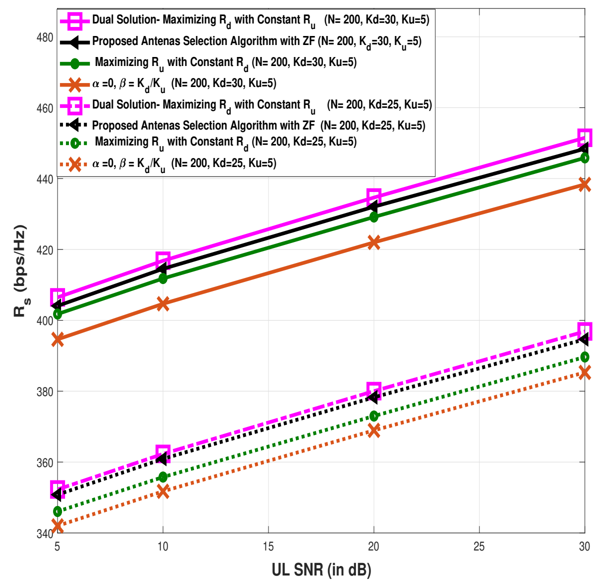


Fig. 7: Variation of the sum rate with The UL-SNR in a massive MIMO network.

$N = 400$ antennas, $K_d = 30$ users, $K_u = 10$ users, and $\alpha = 1$ to show the variation of the DL and UL rates with different numbers of transmit antennas N_t. The results in Fig. 4 verify that the derived DL and UL bounds are tight bounds, as the derived bounds and the rates obtained from simulations show a good match for different values of N_t. Additionally, it can be noticed that increasing N_t remarkably affects the DL rate. As shown in Fig. 4, with full SI cancellation via ZF transmit beamforming, i.e., $\alpha = 1$, if the number of transmit antennas is small, the expected DL rate is low. However, as the number of antennas increases, the effective DL antenna array gain increases, and therefore, the DL rate increases. On the other hand, for $\alpha = 1$, i.e., full SI-nulling on the receive antennas, the UL rate is high. Accordingly, the decrease in the UL SINR caused by increasing the number of transmit antenna will slightly affect the UL rate.

In Fig. 5, we verify the solution of the problem (**P3**) and the value of α derived in (27), that maximizes the UL rate at a given DL rate, in a massive MIMO setting. The system is simulated with $N = 200$ antennas, $K_d = 20$ users, $K_u = 15$ users, and $R_d = 200bps/Hz$. The results show the variation of R_u with α, for different values of the cancellation parameter C_{BS}. From the results in Fig. 5, we can verify that R_u is a convex function of α. Additionally, the maximum of the uplink rate R_u is always at $\alpha = 1$, which matches the rate region maximization problem solution in (27). A final note is that the maximum UL rate achieved at $\alpha = 1$ is the same for all values of SI cancellation parameter C_{BS}, which is a very intuitive result because at $\alpha = 1$, i.e., at full SI cancellation, we have $RSI = 0$ and therefore, the value of C_{BS} will not affect the UL performance. Note that the corresponding value of β that maximizes the UL rate at a given DL rate, is directly calculated from the relation given in (27).

In Fig. 6, we validate the solution of the dual problem derived in (31), for different values of $R_u = R^{'}$, in a system with $N = 200$ antennas, $K_d = 10$ users, and $K_u = 15$ users. The variation of $R_d(\alpha)$ with α validates that $R_d(\alpha)$, for a constant $R_u = R^{'}$, is neither a convex nor a concave function of α. Furthermore, the results in Fig. 6 validate the dual problem solution in (31), as for different values of $R^{'}$, $R_d(\alpha)$ is always maximized at $\alpha = 0$. Moreover, note that for $R_u = R^{'} = 100bps/Hz$ and $R_u = R^{'} = 408bps/Hz$, $R_d(\alpha)$ is slowly varying with α. However, for $R_u = R^{'} = 508bps/Hz$, $R_d(\alpha)$ is rapidly decreasing with α. The DL rate behavior can be explained by the relation between β and R_u, given in (28), which shows that the value of β decreases with $2^{R^{'}/K_u}$. Accordingly, increasing the uplink rate to $R_u = 508bps/Hz$ will cause a large drop in the value of β. When β is small, i.e., small number of the transmit antenna, SI cancellation with beamforming will greatly degrade the DL rate.

Next, we compare the proposed rate maximization algorithm derived in Section IV-B with the asymptotic rate region maximization solution in Section V-A. Fig. 7 shows the variation of the sum rate with the UL SNR= P_u/σ^2. First, the algorithms are compared for a network with $N = 200$ antennas, $K_d = 30$ users and $K_u = 5$ users. The results in Fig. 7 show that the proposed rate maximization algorithm presented in Section IV-B almost achieves the same sum rate achieved by maximizing the feasible rate region in (27). Additionally, it can be noticed that the dual problem solution in (31) results in an upper bound with a very small duality gap. When comparing the proposed scheme to the ZF-scheme which sets $\beta = K_d/K_u$ and $\alpha = 0$, as suggested in [59], it can be seen that the derived solutions achieve better sum rate which validates the necessity of optimizing both the antennas ratio and the SI cancellation. For further validation, the network is simulated with $N = 200$ antennas, $K_d = 30$ users, and $K_u = 5$. Similarly, from the results, we can verify the importance of jointly optimizing both the antennas ratio and the SI cancellation. These results show that for a massive MIMO setting, we just need to

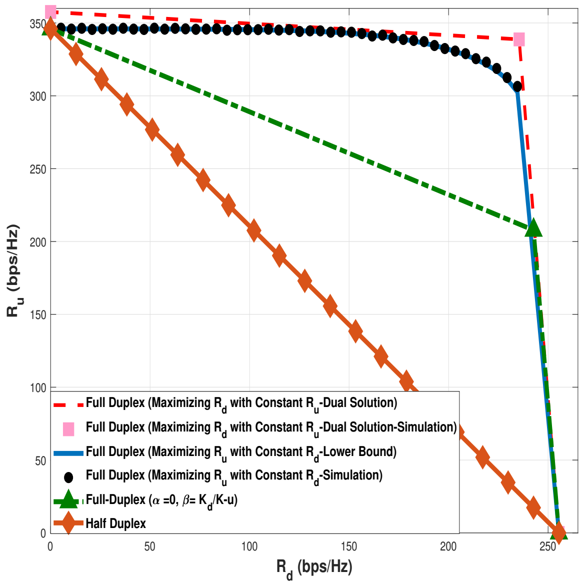


Fig. 8: Feasible rate region for different full duplex and half duplex systems. Parameters used to generate this figure: $N = 300$, $K_d = 20$, $K_u = 10$, $\sigma^2 = -110dB$, $P_d = 10W$ and $P_u = 2mW$.

optimize the antenna ratio. In other words, when the number of antennas grows, it is sufficient to find the transmit to receive antennas ratio without the need to select their indices as in the normal-scale MIMO system case.

Finally, in Fig. 8, the system is simulated with $N = 300$ antennas, $K_d = 20$ users, and $K_u = 10$ to show our derived feasible rate regions for different FD and HD systems. The results compare the derived bounds with the rates obtained from numerical simulations. Additionally, it compares the rate region obtained from solving the optimization problems in (27) and (31) and compares them with the rate region obtained by setting $\beta = K_d/K_u$ and $\alpha = 0$ as suggested in [59], and with the HD rate region. Similarly, as shown in the results of Fig. 7, it can be noticed that solving the dual problem of (31) results in an upper bound with a very small duality gap. Additionally, it can be seen the derived solutions achieve better rate regions than that achieved by the HD system and that achieved when setting $\beta = K_d/K_u$ and $\alpha = 0$ which again validates the importance of jointly optimizing the antennas ratio and the SI cancellation for the rate region maximization problem.

VII. Conclusion

In this paper, we study the challenges of full-duplex MIMO communications. It is shown that, for a normal-scale FD-MIMO setting, and in order to maximize the network sum rate, it is required to jointly optimize the antenna selection with transmitter and receiver weights. Additionally, under the assumption of zero-forcing transmitter and receiver, it is shown that to maximize the network sum rate, it is required to jointly optimize selecting transmit antennas, receive antennas, and receive antennas at which self-interference is nulled via transmit beamforming. From numerical results, it is shown that, in a normal-scale FD-MIMO setting, jointly optimizing the antenna selection, the precoder weights, and the receiver weights achieves a better performance that optimizing the antenna selection under the restriction of zero-forcing precoder and receiver. Furthermore, the asymptotic behavior of full-duplex massive MIMO systems is presented. We derive lower bounds for the downlink and uplink capacities and we show the trade-off between the downlink rate and reducing the self-interference using transmit beamforming. Accordingly, a rate region maximization problem optimizing the transmit and receive antennas ratio and the proportion of self-interference cancellation is formulated. From the numerical results, in the case of self-interference cancellation using zero-forcing transmit beamforming, we show that in a massive MIMO setting we just need to optimize the ratio of receive antennas at which interference is nulled; there is no need for specifying the antenna indices in this case as all antennas are asymptotically equivalent in our massive MIMO setting. Additionally, the derived bounds and the rate maximization solution are verified. It is shown that the proposed scheme achieves better rate region than a previously proposed scheme which sets antennas ratio to the ratio between downlink and uplink users.

References

[1] M. Duarte and A. Sabharwal, “Full-duplex wireless communications using off-the-shelf radios: Feasibility and first results,” in *Conference Record of the Forty Fourth Asilomar Conference on Signals, Systems and Computers (ASILOMAR)*, Pacific Grove, CA, Nov. 2010.

[2] D. Bharadia, E. McMilin, and S. Katti, “Full duplex radios,” in *Proceedings of the ACM SIGCOMM*, Hong Kong, China, Aug. 2013.

[3] M. Jainy, J. I. Choiy, T. M. Kim, D. Bharadia, S. Seth, K. Srinivasan, P. Levis, S. Katti, and P. Sinha, “Practical, real-time, full duplex wireless,” in *the 17th Annual International Conference on Mobile Computing and Networking (MobiCom)*, Las Vegas, NV, Sep. 2011.

[4] F. Rusek, D. Persson, B. K. Lau, E. G. Larsson, T. L. Marzetta, O. Edfors, and F. Tufvesson, “Scaling up MIMO: Opportunities and challenges with very large arrays,” *IEEE Signal Processing Magazine*, vol. 30, no. 1, pp. 40 – 60, Jan. 2013.

[5] T. L. Marzetta, “Noncooperative cellular wireless with unlimited numbers of base station antennas,” *IEEE transactions on wireless communications*, vol. 9, no. 11, pp. 3590–3600, Nov. 2010.

[6] J. Hoydis, K. Hosseini, S. t. Brink, and M. Debbah, “Making smart use of excess antennas: Massive MIMO, small cells, and TDD,” *Bell Labs Technical Journal*, vol. 18, no. 2, pp. 5–21, Aug. 2013.

[7] V.-D. Nguyen, H. V. Nguyen, C. T. Nguyen, and O.-S. Shin, “Spectral efficiency of full-duplex multi-user system: Beamforming design, user grouping, and time allocation,” *IEEE Access*, vol. 5, pp. 5785–5797, Mar. 2017.

[8] S. Huberman and T. Le-Ngoc, “MIMO full-duplex precoding: A joint beamforming and self-interference cancellation structure,” *IEEE Transactions on Wireless Communications*, vol. 14, no. 4, pp. 2205–2217, Apr. 2015.

[9] D. Bharadia and S. Katti, “Full duplex MIMO radios,” in *11th USENIX Symposium on Networked Systems Design and Implementation (NSDI)*, Seattle, WA, Apr. 2014, pp. 359–372.

[10] L. Song, R. Wichman, Y. Li, and Z. Han, *Full-duplex communications and networks*. Cambridge University Press, UK, 2017.

[11] L. Song, Y. Li, and Z. Han, “Resource allocation in full-duplex communications for future wireless networks,” *IEEE Wireless Communications*, vol. 22, no. 4, pp. 88–96, Aug. 2015.

[12] N. H. Mahmood, G. Berardinelli, F. M. Tavares, and P. Mogensen, “On the potential of full duplex communication in 5G small cell networks,” in *IEEE Vehicular Technology Conference (VTC Spring), 81st*, Glasgow, UK, May. 2015.

[13] R. Sultan, L. Song, K. G. Seddik, and Z. Han, “Joint resource management with distributed uplink power control in full-duplex OFDMA networks,” *IEEE Transactions on Vehicular Technology*, Jan. 2020.

[14] R. Sultan, L. Song, K. Seddik, and Z. Han, “Full-duplex meets multiuser MIMO: Comparisons and analysis,” *IEEE Transactions on Vehicular Technology*, vol. 66, no. 1, pp. 455 – 467, Jan. 2017.

[15] R. Sultan, L. Song, and Z. Han, "Impact of full duplex on resource allocation for small cell networks," in *IEEE Signal and Information Processing (GlobalSIP), Global Conference on*, Atlanta, GA, Dec. 2014, pp. 1257–1261.

[16] R. Sultan, L. Song, K. G. Seddik, and Z. Han, "Users association in small cell networks with massive MIMO," in *International Conference on Computing, Networking and Communications (ICNC)*, Kauai, HI, Feb. 2016.

[17] E. G. Larsson, O. Edfors, F. Tufvesson, and T. L. Marzetta, "Massive MIMO for next generation wireless systems," *IEEE communications magazine*, vol. 52, no. 2, pp. 186–195, Feb. 2014.

[18] L. Lu, G. Y. Li, A. L. Swindlehurst, A. Ashikhmin, and R. Zhang, "An overview of massive MIMO: Benefits and challenges," *IEEE journal of selected topics in signal processing*, vol. 8, no. 5, pp. 742–758, Oct. 2014.

[19] R. Sultan, L. Song, K. G. Seddik, and Z. Han, "Full duplex in massive MIMO systems: Analysis and feasibility," in *IEEE Globecom Workshops (GC Wkshps)*, Washington, DC, Dec. 2016.

[20] H. V. Nguyen, V.-D. Nguyen, O. A. Dobre, Y. Wu, and O.-S. Shin, "Joint antenna array mode selection and user assignment for full-duplex MU-MISO systems," *IEEE Transactions on Wireless Communications*, vol. 18, no. 6, pp. 2946–2963, Jun. 2019.

[21] L. V. Nguyen, B. C. Nguyen, X. N. Tran, and L. T. Dung, "Transmit antenna selection for full-duplex spatial modulation multiple-input multiple-output system," *IEEE Systems Journal*, pp. 1–9, Jan. 2020.

[22] N. Takemura, N. Honma, and A. Kawagoe, "An interference cancellation using antenna selection of inter-user for full-duplex MIMO system," in *IEEE International Symposium on Antennas and Propagation (ISAP)*, Xi'an, China, Oct. 2019.

[23] B. Demirkol, M. Toka, and O. Kucur, "Outage probability of max-max antenna selection in dual-hop full duplex amplify-and-forward relaying network over nakagami-m fading channel," in *27th IEEE Signal Processing and Communications Applications Conference (SIU)*, Sivas, Turkey, Aug. 2019.

[24] D. G. Wilson-Nunn, A. Chaaban, A. Sezgin, and M.-S. Alouini, "Antenna selection for full-duplex MIMO two-way communication systems," *IEEE Communications Letters*, vol. 21, no. 6, pp. 1373–1376, Jun. 2017.

[25] S. Jang, M. Ahn, H. Lee, and I. Lee, "Antenna selection schemes in bidirectional full-duplex MIMO systems," *IEEE Transactions on Vehicular Technology*, vol. 65, no. 12, pp. 10 097–0100, Dec. 2016.

[26] M. Zhou, H. Cui, L. Song, and B. Jiao, "Transmit-receive antenna pair selection in full duplex systems," *IEEE Wireless Communications Letters*, vol. 3, no. 1, pp. 34–37, Feb. 20142013.

[27] A. T. Le, L. C. Tran, X. Huang, and Y. J. Guo, "Beam-based analog self-interference cancellation in full-duplex MIMO systems," *IEEE Transactions on Wireless Communications*, pp. 1–1, Jan. 2020.

[28] M. Tsunezawa, K. Takahashi, N. Honma, K. Murata, and K. Nishimori, "Antenna arrangement suitable for full-duplex MIMO," *IEEE Transactions on Antennas and Propagation*, vol. 65, no. 6, pp. 2966–2974, Jun. 2017.

[29] S. Narieda, "Antenna selection for reduction of rician distributed self-interference effects in full duplex radio," *Electronics Letters*, vol. 52, no. 3, pp. 234–235, Feb. 2016.

[30] A. Masmoudi and T. Le-Ngoc, "Self-interference cancellation for full-duplex MIMO transceivers," in *IEEE Wireless Communications and Networking Conference (WCNC)*, New Orleans, LA, Mar. 2015.

[31] M. He and C. Huang, "Self-interference cancellation for full-duplex massive MIMO OFDM with single rf chain," *IEEE Wireless Communications Letters*, vol. 9, no. 1, pp. 26–29, Jan. 2020.

[32] N. M. Gowda and A. Sabharwal, "Jointnull: Combining partial analog cancellation with transmit beamforming for large-antenna full-duplex wireless systems," *IEEE Transactions on Wireless Communications*, vol. 17, no. 3, pp. 2094–2108, Mar. 2018.

[33] D. Chen, F. Liu, C. Feng, J. Shi, and X. Han, "Energy efficiency optimization based self-interference cancellation in massive MIMO full-duplex system," in *IEEE/CIC International Conference on Communications in China (ICCC)*, Qingdao, China, Oct. 2017.

[34] Y.-G. Lim, D. Hong, and C.-B. Chae, "Performance analysis of self-interference cancellation in full-duplex large-scale MIMO systems," in *IEEE Global Communications Conference (GLOBECOM)*, Washington, DC, Dec. 2016.

[35] F. Maciel-Barboza, J. Sánchez-García, S. Armas-Jiménez, and L. Soriano-Equigua, "Uplink and downlink user and antenna selection for mmWave full duplex multiuser systems," in *2nd IEEE International Conference on Frontiers of Signal Processing (ICFSP)*, Warsaw, Poland, Oct. 2016, pp. 142–146.

[36] A. Shojaeifard, K.-K. Wong, M. Di Renzo, G. Zheng, K. A. Hamdi, and J. Tang, "Massive MIMO-enabled full-duplex cellular networks," *IEEE Transactions on Communications*, vol. 65, no. 11, pp. 4734–4750, Nov. 2017.

[37] C. Zhang, W. Xu, and M. Chen, "Hybrid zero-forcing beamforming/orthogonal beamforming with user selection for MIMO broadcast channels," *IEEE Communications Letters*, vol. 13, no. 1, pp. 10–12, Jan. 2009.

[38] H. Yang and T. L. Marzetta, "Performance of conjugate and zero-forcing beamforming in large-scale antenna systems," *IEEE Journal on Selected Areas in Communications*, vol. 31, no. 2, pp. 172–179, Feb. 2013.

[39] J. Koh, Y.-G. Lim, C.-B. Chae, and J. Kang, "On the feasibility of full-duplex large-scale MIMO cellular systems," *IEEE Transactions on Wireless Communications*, vol. 17, no. 9, pp. 6231–6250, Sep. 2018.

[40] X. Wang, J. Liu, C. Zhai, P. Xing, and L. Zheng, "Optimization of energy efficiency for the full-duplex massive MIMO systems," *Journal of Systems Engineering and Electronics*, vol. 27, no. 2, pp. 326–332, Apr. 2016.

[41] B. Yin, M. Wu, C. Studer, J. R. Cavallaro, and J. Lilleberg, "Full-duplex in large-scale wireless systems," in *IEEE Asilomar Conference on Signals, Systems and Computers*, Pacific Grove, CA, Nov. 2013.

[42] K. Singh, S. Biswas, T. Ratnarajah, and F. A. Khan, "Transceiver design and power allocation for full-duplex MIMO communication systems with spectrum sharing radar," *IEEE Transactions on Cognitive Communications and Networking*, vol. 4, no. 3, pp. 556–566, Sep. 2018.

[43] I. Atzeni and M. Kountouris, "Full-duplex MIMO small-cell networks with interference cancellation," *IEEE Transactions on Wireless Communications*, vol. 16, no. 12, pp. 8362–8376, Dec. 2017.

[44] R. K. Mungara, I. Thibault, and A. Lozano, "Full-duplex MIMO in cellular networks: System-level performance," *IEEE transactions on wireless communications*, vol. 16, no. 5, pp. 3124–3137, May 2017.

[45] X. Wang, D. Zhang, K. Xu, and C. Yuan, "On the sum rate of multi-user full-duplex massive MIMO systems," in *IEEE International Conference on Communication Systems (ICCS)*, Shenzhen, China, Dec. 2016.

[46] Q. Li, S. X. Wu, S. Wang, and J. Lin, "Joint uplink/downlink discrete sum rate maximization for full-duplex multicell heterogeneous networks," *IEEE Transactions on Vehicular Technology*, vol. 69, no. 3, pp. 2758–2770, Mar. 2020.

[47] A. C. Cirik, M. J. Rahman, and L. Lampe, "Robust fairness transceiver design for a full-duplex MIMO multi-cell system," *IEEE Transactions on Communications*, vol. 66, no. 3, pp. 1027–1041, Mar. 2018.

[48] A. C. Cirik, "Fairness considerations for full duplex multi-user MIMO systems," *IEEE Wireless Communications Letters*, vol. 4, no. 4, pp. 361–364, Aug. 2015.

[49] P. Aquilina, A. C. Cirik, and T. Ratnarajah, "Weighted sum rate maximization in full-duplex multi-user multi-cell MIMO networks," *IEEE Transactions on Communications*, vol. 65, no. 4, pp. 1590–1608, Apr. 2017.

[50] A. C. Cirik, R. Wang, Y. Hua, and M. Latva-aho, "Weighted sum-rate maximization for full-duplex MIMO interference channels," *IEEE Transactions on communications*, vol. 63, no. 3, pp. 801–815, Mar. 2015.

[51] A. M. Geoffrion, "Generalized benders decomposition," *Springer Journal of optimization theory and applications*, vol. 10, no. 4, pp. 237–260, 1972.

[52] G. Scutari, F. Facchinei, and L. Lampariello, "Parallel and distributed methods for constrained nonconvex optimization-part I: Theory," *IEEE Transactions on Signal Processing*, vol. 65, no. 8, pp. 1929–1944, Apr. 2017.

[53] G. Scutari, F. Facchinei, L. Lampariello, S. Sardellitti, and P. Song, "Parallel and distributed methods for constrained nonconvex optimization-part II: Applications in communications and machine learning," *IEEE Transactions on Signal Processing*, vol. 65, no. 8, pp. 1945–1960, Apr. 2017.

[54] S. Wagner, R. Couillet, M. Debbah, and D. T. Slock, "Large system analysis of linear precoding in correlated MISO broadcast channels under limited feedback," *IEEE transactions on information theory*, vol. 58, no. 7, pp. 4509–4537, Jul. 2012.

[55] Y. Jang, K. Min, S. Park, and S. Choi, "Spatial resource utilization to maximize uplink spectral efficiency in full-duplex massive MIMO," in *IEEE International Conference on Communications (ICC)*, London, UK, Jun. 2015.

[56] R. Sultan, K. G. Seddik, Z. Han, and B. Aazhang, "Asymptotic behavior analysis and performance optimization in full duplex massive MIMO," in *IEEE Global Communications Conference*, Singapore, Dec. 2017.

[57] S. Boyd, S. P. Boyd, and L. Vandenberghe, *Convex optimization*. Cambridge university press, 2004.

[58] A. Ben-Tal and A. Nemirovski, *Lectures on modern convex optimization: analysis, algorithms, and engineering applications*. SIAM, 2001.

[59] K. Min, Y. Jang, S. Park, and S. Choi, "Antenna ratio for sum-rate maximization in MU-MIMO with full-duplex large array BS," in *IEEE International Conference on Communications (ICC)*, London, UK, Jun. 2015, pp. 1589–1594.

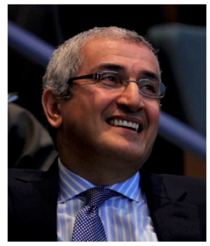
Radwa Sultan (S'05-M'17) received her B.Sc (with highest honors) and the M.S.degrees in electrical engineering from Alexandria University, Alexandria, Egypt, in 2009 and 2013, respectively, and her PhD degree in Electrical and Computer Engineering Department at the University of Houston, Texas, in 2017. Currently, she is an assistant professor in the Electrical and Computer Engineering Department in Manhattan College, New York. She has served on the technical program committees of several IEEE journals and conferences in the areas of wireless communication. Her primary research interests are intelligent reflecting surface, Internet-of-Things, software defined radio Systems, mmWave networks, small cell Networks, full-duplex communication, massive MIMO, signal processing, wireless resource allocation, and big data analysis in wireless communication.

Karim G. Seddik (S'04-M'08-SM'14) Karim Seddik received the BSc with an honor and MSc degrees in electrical engineering from Alexandria University, Alexandria, Egypt, in 2001 and 2004, respectively, and the Ph.D. degree from the University of Maryland, College Park, MD, USA, in 2008. He is currently a Professor in the electronics and communications engineering department, at the American University in Cairo (AUC) and the Associate Dean for Graduate Studies and Research, School of Sciences and Engineering (SSE) at the AUC. Before joining AUC, he was an assistant professor at Alexandria University. His research interests include applications of machine learning in communication networks, intelligent reflecting surfaces, age of information, cognitive radio communications, and layered channel coding. Seddik has served on the technical program committees of numerous IEEE conferences in the areas of wireless networks and mobile computing.

He is a recipient of the State Encouragement Award in 2016 and the State Medal of Excellence in 2017. He is a recipient of the certificate of honor from the Egyptian President for being ranked first among all departments in the College of Engineering, Alexandria University in 2002. He received the Graduate School Fellowship from the University of Maryland in 2004 and 2005, and the Future Faculty Program Fellowship from the University of Maryland in 2007. He also co-authored a conference paper that received the best conference paper award from the IEEE communication society technical committee on green communications and computing in 2019.

Zhu Han (S'01-M'04-SM'09-F'14) received the B.S. degree in electronic engineering from Tsinghua University, in 1997, and the M.S. and Ph.D. degrees in electrical and computer engineering from the University of Maryland, College Park, in 1999 and 2003, respectively.

From 2000 to 2002, he was an R&D Engineer of JDSU, Germantown, Maryland. From 2003 to 2006, he was a Research Associate at the University of Maryland. From 2006 to 2008, he was an assistant professor at Boise State University, Idaho. Currently, he is a John and Rebecca Moores Professor in the Electrical and Computer Engineering Department as well as in the Computer Science Department at the University of Houston, Texas. His research interests include wireless resource allocation and management, wireless communications and networking, game theory, big data analysis, security, and smart grid. Dr. Han received an NSF Career Award in 2010, the Fred W. Ellersick Prize of the IEEE Communication Society in 2011, the EURASIP Best Paper Award for the Journal on Advances in Signal Processing in 2015, IEEE Leonard G. Abraham Prize in the field of Communications Systems (best paper award in IEEE JSAC) in 2016, and several best paper awards in IEEE conferences. Dr. Han was an IEEE Communications Society Distinguished Lecturer from 2015-2018, AAAS fellow since 2019 and ACM distinguished Member since 2019. Dr. Han is 1% highly cited researcher since 2017 according to Web of Science. Dr. Han is also the winner of 2021 IEEE Kiyo Tomiyasu Award, for outstanding early to mid-career contributions to technologies holding the promise of innovative applications, with the following citation: "for contributions to game theory and distributed management of autonomous communication networks."



Behnaam Aazhang (Fellow, IEEE) Behnaam Aazhang received his B.S. (with highest honors), M.S., and Ph.D. degrees in Electrical and Computer Engineering from University of Illinois at Urbana-Champaign in 1981, 1983, and 1986, respectively. From 1981 to 1985, he was a Research Assistant in the Coordinated Science Laboratory, University of Illinois. In August 1985, he joined the faculty of Rice University, Houston, Texas, where he is now the J.S. Abercrombie Professor in the Department of Electrical and Computer Engineering Professor and Director of Rice‚Äôs Neuroengineering Initiative. From 2006 till 2014, he held an Academy of Finland Distinguished Visiting Professorship appointment (FiDiPro) at the University of Oulu, Oulu, Finland. Dr. Aazhang is a Fellow of IEEE and AAAS, a distinguished lecturer of IEEE Communication Society. He received an Honorary Doctorate degree from the University of Oulu, Finland (the highest honor that the university can bestow) in 2017. He is also the recipient of the IEEE ComSoc CTTC Outstanding Service Award "For innovative leadership that elevated the success of the Communication Theory Workshop" in 2016 and Outstanding Technical Achievement Award "For consistent, fundamental contributions to multiuser communication theory for wireless networks" in 2017. He is a recipient of 2019 SIGMOBILE Test of Time (ToT) award recognizing outstanding papers that have had a lasting impact on the field of mobile computing for the paper titled "Design of WARP: a wireless open-access research platform" and 2004 IEEE Communication Society's Stephen O. Rice best paper award for a paper with A. Sendonaris and E. Erkip. In addition, Sendonaris, Erkip, and Aazhang received IEEE Communication Society's 2013 Advances in Communication Award for the same paper. He has been listed in the Thomson-ISI Highly Cited Researchers and has been keynote and plenary speaker of several conferences.

His research interests are signal and data processing, information theory, dynamical systems, and their applications to neuroengineering with focus areas in (i) understanding neuronal circuits connectivity and the impact of learning on connectivity (ii) developing minimally invasive and non-invasive real-time closed-loop stimulation of neuronal systems to mitigate disorders such as epilepsy, Parkinson, depression, obesity, and mild traumatic brain injury (iii) developing a patient-specific multisite wireless monitoring and pacing system with temporal and spatial precision to restore the healthy function of a diseased heart, (iv) developing algorithms to detect, predict, and prevent security breaches in cloud computing and storage systems.