Interference Mitigation with Block Diagonalization for IRS-aided MU-MIMO Communications

Wilker de O. Feitosa, Igor M. Guerreiro, Fco. Rodrigo P. Cavalcanti, Tarcisio F. Maciel, Maria Clara R. Lobão, Fazal-E-Asim, Behrooz Makki and Gábor Fodor.

***Abstract*— This work investigates interference mitigation techniques in multi-user multiple input multiple output (MU-MIMO) intelligent reflecting surface (IRS)-aided networks, focusing on the base station end. Two methods of precoder design based on block diagonalization are proposed. The first method does not consider the interference caused by the IRS, seeking to mitigate only the multi-user interference. The second method mitigates both the IRS-caused interference and the multi-user interference. A comparison between both methods within a no-IRS MU-MIMO network with strong direct links is provided. The results show that, although in some circumstances IRS interference can be neglected, treating it improves system capacity and provide higher spectral efficiency.**

***Keywords*— MU-MIMO, interference mitigation, block diagonalization, intelligent reflecting surfaces.**

I. INTRODUCTION

The sixth generation (6G) of cellular networks is expected to present significant advances in terms of system capacity, energy efficiency, number of supported users and spectral efficiency (SE) compared to the fifth generation (5G) [1]. To accomplish this goal, new physical layer technologies are investigated to take more advantage of the propagation features of the environment, such as intelligent reflecting surface (IRS) [2], sub-Terahertz bands, distributed multiple input multiple output (MIMO), among others [3].

Multi-user MIMO (MU-MIMO), as a well-established key technology in mobile wireless systems due to its advantages in spatial diversity and multiplexing, will also play an important role in 6G, where its beamforming gains and improvements in SE are desired and enhanced when combined with the aforementioned technologies. One of the challenges in MU-MIMO systems is to deal with multi-user interference. Various methods for mitigating interference on the receiver end, as well as at the transmitter, have been developed over the past years, for instance, the design of robust decoding and precoding filters using methods like zero-forcing (ZF) [4] and block diagonalization (BD) [5].

This work was partially supported by the Ericsson Research, Sweden, and Ericsson Innovation Center, Brazil, under UFC.51, partially supported by the Brazilian National Council for Scientific and Technological Development (CNPq), the Coordenação de Aperfeiçoamento de Pessoal de Nível Superior - Brasil (CAPES) - Finance Code 001, in part by CNPq 306845/2020-2, and Fundação Cearense de Apoio ao Desenvolvimento Científico e Tecnológico (FUNCAP) - Grant FC3-00198-00056.01.00/22.

Wilker de O. Feitosa, Igor M. Guerreiro, Fco. Rodrigo P. Cavalcanti, Tarcisio F. Maciel, Maria Clara R. Lobão and Fazal-E-Asim are with the Wireless Telecommunications Research Group (GTEL), Federal University of Ceará (UFC), Fortaleza, Brazil. E-mails: {wilker, igor, rodrigo, maciel, clara, fazalasim}@gtel.ufc.br. Behrooz Makki and Gábor Fodor are with Ericsson Research. (e-mails: behrooz.makki@ericsson.com, gabor.fodor@ericsson.com).

In 5G and beyond systems, the use of millimeter wave (mmWave) bands is highly desirable due to the large bandwidth available in the spectrum. However, since high frequencies are bound to severe pathloss and penetration loss, communication in those bands is much more susceptible to blockage and poor link conditions without a line of sight (LOS) component. Such effects can even interrupt the connection, thus limiting the capacity of the system. In order to overcome these limitations, the use of mmWave is usually combined with other technologies like massive MIMO (mMIMO) and ultra-dense networks [6]. In these circumstances, the concept of smart radio environment (SRE) can be introduced, which states that the wireless environment can be partially turned into an optimization variable that, jointly with transmitter and receiver properties, can be used to maximize the overall network performance [7].

The concept of IRSs is a candidate enabler for SRE since it can well modify the environment. For instance, IRS can create a virtual LOS component, higher-rank channels, or attenuate undesired signals [8]. IRSs also actuate as antenna arrays, improving signal quality by applying beamforming to the desired signal. Thus, SRE aided by IRSs can be leveraged to diminish the effects of propagation losses, improve the coverage and increase the SE by optimizing the environment between transmitter and receiver to achieve better link conditions [9].

Considering the high pathloss and blockage probability at mmWave bands, the use of IRSs can provide beamforming gains and additional paths to users under poor propagation conditions, thus improving the capacity of MU-MIMO systems. Nonetheless, it is important to notice that even a fully passive IRS, i.e., an IRS without radio frequency (RF) chains, can introduce interference to untargeted users and base stations (BSs), e.g., other nodes nearby the intended user. This raises the question about the need to mitigate such interference and the means to do it.

To manage the interference introduced by IRSs, the authors in [10] employ an orthogonalization scheme based on BD in a MU-MIMO scenario. However, their BD approach demands the use of at least one IRS per user. The authors in [11] also address this problem by minimizing the symbol error rate (SER) using 1-bit analog-to-digital converters (ADCs) on the BS side.

In this paper, we study the problem of interference management in IRS-assisted MU-MIMO networks with a single IRS. We propose two precoding methods based on BD. Compared to the solution in [10], our proposed methods mitigate in-

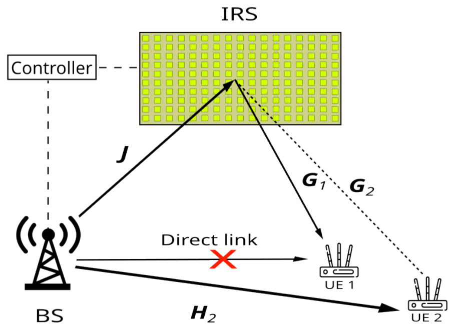

Fig. 1. Multi-user MIMO-IRS assisted systems

terference caused by IRSs not only by being less greedy in terms of computational complexity, but also using fewer RF chains at both the transmitter and the receiver. The first method considers the interference caused by the IRS as negligible and focuses only on multi-user interference mitigation. The second method takes both types of interference into account and also uses part of the IRS signal towards untargeted users as useful signal. The results show that interference can be mitigated with both methods, and in particular, the second method presents the highest SE in comparison with the other method and the state-of-the-art.

II. System model

Consider the downlink of a system composed of a BS with M antennas serving simultaneously two co-channel user equipments (UEs), as shown in Fig. 1. UE1 is equipped with Q antennas and has no direct path to the BS due to, e.g., a strong blockage effect. Therefore, the BS serves UE1 *via* an IRS with N reflecting elements. UE2, with P antennas, has a strong direct link to the BS; thus, the use of the IRS for UE2 is optional, in both cases, $N_s^{(1,2)}$ stands for the number of streams allowed in each user's configuration. The IRS phase-shift controller is assumed to work ideally with the BS. While serving UE1 *via* IRS, UE2 receives from the BS an unintended signal due to beam leakage, whose intensity depends on the propagation conditions and on the BS transmit power.

The received signal model is given as:

$$\begin{bmatrix}\mathbf{y}_1\\ \mathbf{y}_2\end{bmatrix} = \begin{bmatrix}\mathbf{W}_1^{\mathrm{H}} & \mathbf{0}\\ \mathbf{0} & \mathbf{W}_2^{\mathrm{H}}\end{bmatrix}\begin{bmatrix}\bar{\mathbf{H}}_1\\ \bar{\mathbf{H}}_2\end{bmatrix}\begin{bmatrix}\mathbf{F}_1 & \mathbf{F}_2\end{bmatrix}\begin{bmatrix}\mathbf{x}_1\\ \mathbf{x}_2\end{bmatrix} + \begin{bmatrix}\tilde{\mathbf{n}}_1\\ \tilde{\mathbf{n}}_2\end{bmatrix}, \quad (1)$$

where $\mathbf{W}_1 \in \mathbb{C}^{Q\times N_s^{(1)}}$ represents the combiner for UE1, $\mathbf{F}_1 \in \mathbb{C}^{M\times N_s^{(1)}}$ is the digital baseband precoder for UE1, $\mathbf{n}_1 \sim \mathcal{CN}\left(\mathbf{0}, \sigma_1^2\mathbf{I}_Q\right)$ is the circularly symmetric additive white Gaussian noise vector with variance σ_1^2 and $\mathbf{x}_1 \in \mathbb{C}^{N_s^{(1)}\times 1}$ is the data vector for UE1. Similarly, $\mathbf{W}_2 \in \mathbb{C}^{P\times N_s^{(2)}}$ is the combiner for UE2, $\mathbf{F}_2 \in \mathbb{C}^{M\times N_s^{(2)}}$ is the digital baseband precoder for UE2 and $\mathbf{n}_2 \sim \mathcal{CN}\left(\mathbf{0}, \sigma_2^2\mathbf{I}_P\right)$ is the circularly symmetric additive white Gaussian noise vector with variance σ_2^2, and finally, $\mathbf{x}_2 \in \mathbb{C}^{N_s^{(2)}\times 1}$ is the data vector for UE2. At last, $\mathbf{I}_a$ denotes the $a\times a$ identity matrix and $\mathbf{0}$ is a vector of zeros of proper size.

A. Signal Model

Given propagation scenario depicted in Fig. 1, the channel for UE1 can be defined as $\bar{\mathbf{H}}_1 = \mathbf{G}_1\mathbf{\Omega}\mathbf{J} \in \mathbb{C}^{Q\times M}$, hence, the complete equation of the received signal for UE1 is written as

$$\begin{aligned}\mathbf{y}_1 =& \mathbf{W}_1^{\mathrm{H}}\mathbf{G}_1\mathbf{\Omega}\mathbf{J}\mathbf{F}_1\mathbf{x}_1 + \\ & \underbrace{\mathbf{W}_1^{\mathrm{H}}\mathbf{G}_1\mathbf{\Omega}\mathbf{J}\mathbf{F}_2\mathbf{x}_2}_{\text{interference}} + \underbrace{\mathbf{W}_1^{\mathrm{H}}\mathbf{n}_1}_{\tilde{\mathbf{n}}_1} \in \mathbb{C}^{N_s^{(1)}\times 1},\end{aligned} \quad (2)$$

where $\mathbf{J} \in \mathbb{C}^{N\times M}$ is the channel between the BS and the IRS, $\mathbf{G}_1 \in \mathbb{C}^{Q\times N}$ is the channel between the IRS and the UE1 and $\mathbf{\Omega} = \operatorname{diag}\{\boldsymbol{\omega}\}$, $\boldsymbol{\omega} \in \mathbb{C}^{N\times 1}$, is the IRS phase-shift vector, which, since the IRS is passive, acts like an analog beamforming with constant modulus. Based on (2), the signal-to-interference-plus-noise-ratio (SINR) γ_1 for UE1 is given by

$$\gamma_1 = \operatorname{tr}\left[\mathbf{W}_1^{\mathrm{H}}\mathbf{G}_1\mathbf{\Omega}\mathbf{J}\mathbf{F}_1\mathbf{F}_1^{\mathrm{H}}\mathbf{J}^{\mathrm{H}}\mathbf{\Omega}^{\mathrm{H}}\mathbf{G}_1^{\mathrm{H}}\mathbf{W}_1\mathbf{R}_1^{-1}\right], \quad (3)$$

where $\mathbf{R}_1 = \sigma_1^2\mathbf{I}_{N_s^{(1)}} + \mathbf{W}_1^{\mathrm{H}}\bar{\mathbf{H}}_1\mathbf{F}_2\mathbf{F}_2^{\mathrm{H}}\bar{\mathbf{H}}_1^{\mathrm{H}}\mathbf{W}_1$, and $\operatorname{tr}[\cdot]$ denotes the trace operator.

Likewise, the channel for UE2 can be defined as $\bar{\mathbf{H}}_2 = \mathbf{H}_2 + \mathbf{G}_2\mathbf{\Omega}\mathbf{J} \in \mathbb{C}^{P\times M}$. Therefore, the received signal for UE2 from (1) is given by

$$\begin{aligned}\mathbf{y}_2 =& \mathbf{W}_2^{\mathrm{H}}\mathbf{H}_2\mathbf{F}_2\mathbf{x}_2 + \mathbf{W}_2^{\mathrm{H}}\mathbf{G}_2\mathbf{\Omega}\mathbf{J}\mathbf{F}_2\mathbf{x}_2 + \\ & \underbrace{\mathbf{W}_2^{\mathrm{H}}\mathbf{H}_2\mathbf{F}_1\mathbf{x}_1 + \mathbf{W}_2^{\mathrm{H}}\mathbf{G}_2\mathbf{\Omega}\mathbf{J}\mathbf{F}_1\mathbf{x}_1}_{\text{interference}} + \underbrace{\mathbf{W}_2^{\mathrm{H}}\mathbf{n}_2}_{\tilde{\mathbf{n}}_2} \in \mathbb{C}^{N_s^{(2)}\times 1},\end{aligned} \quad (4)$$

where $\mathbf{H}_2 \in \mathbb{C}^{P\times M}$ is the direct channel between the BS and the UE2, $\mathbf{G}_2 \in \mathbb{C}^{P\times N}$ is the leakage channel between the IRS and UE2. Now let $\mathbf{R}_2 = \sigma_2^2\mathbf{I}_{N_s^{(2)}} + \mathbf{W}_2^{\mathrm{H}}\bar{\mathbf{H}}_2\mathbf{F}_1\mathbf{F}_1^{\mathrm{H}}\bar{\mathbf{H}}_2^{\mathrm{H}}\mathbf{W}_2$. Then, based on (4), the SINR for UE2 is calculated by (5), given on top of the next page.

The key performance indicators used for comparing the techniques are the SE and the sum SE. The SE can be calculated as $\epsilon_j = \log_2\det[\mathbf{I}_{N_s} + \gamma_j], j \in \{1,2\}$, and the sum SE is defined as $\epsilon_{\text{sum}} = \sum_{j\in\{1,2\}}\epsilon_j$.

It is also important to notice that mmWave and Terahertz systems tend to have fewer RF chains in their configurations, due to their massive number of antennas. In this paper, all hybrid beamforming is done considering that the number of RF chains is smaller than the number of antennas [12].

B. Propagation Model

Considering the scenario illustrated in Fig. 1, our adopted channel model [13] is given by

$$\begin{aligned}\mathbf{H}_{r,t} =& \sqrt{\frac{K}{K+1}}A_0\mathbf{a}_r\left(\theta_{r,0},\phi_{r,0}\right)\mathbf{a}_t^{\mathrm{T}}\left(\theta_{t,0},\phi_{t,0}\right) + \\ & \sqrt{\frac{1}{K+1}}\left(\frac{1}{\sqrt{S}}\sum_{s=1}^{S}A_s\mathbf{a}_r\left(\theta_{r,s},\phi_{r,s}\right)\mathbf{a}_t^{\mathrm{T}}\left(\theta_{t,s},\phi_{t,s}\right)\right),\end{aligned} \quad (6)$$

in which K is the Rician K-factor; $\mathbf{a}_r\left(\theta_{r,s},\phi_{r,s}\right)$ and $\mathbf{a}_t\left(\theta_{t,s},\phi_{t,s}\right)$ represent the steering vectors at the receiver r and the transmitter t, respectively, for the s-th ray, with $s = 0,\ldots,S$. The index $s=0$ indicates the LOS component of the channel. The angles θ and ϕ correspond to the horizontal and

$$\gamma_2 = \mathrm{tr}\left[(\mathbf{W}_2^{\mathrm{H}}\mathbf{H}_2\mathbf{F}_2\mathbf{F}_2^{\mathrm{H}}\mathbf{H}_2^{\mathrm{H}}\mathbf{W}_2 + \mathbf{W}_2^{\mathrm{H}}\mathbf{G}_2\boldsymbol{\Omega}\mathbf{J}\mathbf{F}_2\mathbf{F}_2^{\mathrm{H}}\mathbf{J}^{\mathrm{H}}\boldsymbol{\Omega}^{\mathrm{H}}\mathbf{G}_2^{\mathrm{H}}\mathbf{W}_2)\mathbf{R}_2^{-1}\right] \tag{5}$$

vertical directions, respectively. The term A_s is the channel coefficient that contains the pathloss, shadowing and fast-fading. The channels between the receiver r and transmitter t, $\mathbf{H}_{r,t}$, are defined as $\mathbf{H}_{\text{IRS,BS}} = \mathbf{J}$, $\mathbf{H}_{\text{UE2,BS}} = \mathbf{H}_2$, $\mathbf{H}_{\text{UE1,IRS}} = \mathbf{G}_1$ and $\mathbf{H}_{\text{UE2,IRS}} = \mathbf{G}_2$.

For the modeling of the steering vectors, the IRS is designed as a uniform rectangular array (URA). The UEs and the BS are considered to be equipped with horizontal uniform linear arrays (ULAs), for which the angle ϕ is disregarded.

Due to the use of mmWave bands, the pathloss, shadowing and Rician K-factor for the considered system are modeled according to [14] considering the urban macro (UMa) scenario:

$$\mathrm{PL} = 28 + 22\log_{10}(d_{3D}) + 20\log_{10}(f_c), \tag{7}$$

in which d_{3D} is the absolute distance between the transmitter and the receiver and f_c is the carrier frequency. The shadow fading is modeled according to a log-normal distribution with standard deviation $\sigma = 4$ dB. The Rician K-factor also follows a log-normal distribution, with $K \sim \mathcal{N}(9, 3.5)$ dB [14], for all the links in the Fig. 1.

The scattering in a UMa scenario is considered to be rich, i.e., the channel has a large number of multi-paths. Therefore, all channels in the studied scenario are considered to have a full rank.

C. IRS Phase Shift Setting

Seeing that the main use of the IRS in this work is to provide an alternative path for users under blockage, for the design of the IRS phase-shift, a singular value decomposition (SVD) is performed on the channels $\mathbf{J}$ and $\mathbf{G}_1$, which can be individually estimated by techniques like [16], [17]. The phase-shift vector $\boldsymbol{\omega}$ is generated through the Hadamard product of the left singular vector of $\mathbf{J}$ associated with its highest singular value, $\mathbf{u}_{\mathbf{J}}$, and the right singular vector of $\mathbf{G}_1$ associated with its highest singular value, $\mathbf{v}_{\mathbf{G}_1}^*$:

$$\boldsymbol{\omega} = -\angle(\mathbf{u}_{\mathbf{J}} \odot \mathbf{v}_{\mathbf{G}_1}^*) \in \mathbb{C}^{N\times 1}, \tag{8}$$

where $\odot$ denotes Hadamard product, and $*$ is the conjugate of a vector.

III. Proposed Precoding Design

Our proposed IRS-aided network operates in two stages: i) There is the IRS phase-shift design, also known as IRS passive beamforming, and then ii) the digital precoder and combiner are computed.

To support multi-user communication, we propose two BD-based precoding schemes implemented at the BS. On the UEs side, a traditional ZF combiner matched to the intended channel is employed. In this study, all channel responses are assumed known at the BS by relying on the fact that practice they can be estimated, as, e.g., demonstrated in [15]–[17].

A. General BD Framework

When it comes to MU-MIMO interference mitigation techniques, BD is well-known for its efficiency in maximizing either the throughput or the fairness[1] of the system [5]. BD-based precoders have the potential to cancel interference toward non-intended users by introducing the following constraint:

$$\tilde{\mathbf{H}}_u\mathbf{F}_k = \mathbf{0}_{N_{M,u}\times N_{M,u}}, \quad \forall u \neq k, \tag{9}$$

where $u, k = 1, \dots, L$, represent UE indexes, L is the total number of UEs in the system, and $\tilde{\mathbf{H}}_l$ is the complementary channel of the l-th UE, defined as:

$$\tilde{\mathbf{H}}_l = \begin{bmatrix}\mathbf{H}_1^{\mathrm{T}} & \dots & \mathbf{H}_{l-1}^{\mathrm{T}} & \mathbf{H}_{l+1}^{\mathrm{T}} & \dots & \mathbf{H}_L^{\mathrm{T}}\end{bmatrix}^{\mathrm{T}}. \tag{10}$$

In order to achieve the result in (9) and mix the signal for the intended users coherently, it is necessary for the precoder $\mathbf{F}_k$ to lie in the null-space of $\tilde{\mathbf{H}}_l$ and on the signal-space of $\mathbf{H}_k$. Both of which can be obtained by the SVD of $\tilde{\mathbf{H}}_l$ and $\mathbf{H}_k$, respectively. This technique must obey the restriction that the number of transmitting antennas should be greater than or equal to the total number of receiving antennas.

The precoder $\mathbf{F}_k$ is constructed as follows:

$$\begin{aligned}
\tilde{\mathbf{H}}_l &= \tilde{\mathbf{U}}_l\tilde{\boldsymbol{\Lambda}}_l\begin{bmatrix}\tilde{\mathbf{V}}_l^{\text{non-zero}} & \tilde{\mathbf{V}}_l^{\text{zero}}\end{bmatrix}^{\mathrm{H}}, \\
\mathbf{H}_k\tilde{\mathbf{V}}_l^{\text{zero}} &= \mathbf{U}_k\boldsymbol{\Lambda}_k\begin{bmatrix}\mathbf{V}_k^{\text{non-zero}} & \mathbf{V}_k^{\text{zero}}\end{bmatrix}^{\mathrm{H}}, \\
\mathbf{F}_k &= \tilde{\mathbf{V}}_l^{\text{zero}}\mathbf{V}_k^{\text{non-zero}}.
\end{aligned} \tag{11}$$

As shown in [10], the classic BD technique is not feasible for IRS-aided scenarios, given that the UEs channels and the IRS phase-shift vectors are coupled, which leads to the right singular vectors of $\tilde{\mathbf{H}}_l$ and $\mathbf{H}_k$ not being disjoint. In this context, two techniques are proposed in the sequel to overcome this issue.

B. Partial IRS BD (PIB) Precoder Design

In this first approach, the IRS precise beamforming [18] is taken into account and it is assumed that the beam leakage is minimal. Therefore, for the system presented in Fig. 1, the IRS leakage channel $\mathbf{G}_2\boldsymbol{\Omega}\mathbf{J}$ is neglected in the precoder design and $\bar{\mathbf{H}}_1 = \mathbf{G}_1\boldsymbol{\Omega}\mathbf{J}$ and $\mathbf{H}_2$ are the only contemplated channels, which can be classically block diagonalized by considering the complementary channels as follows:

$$\begin{aligned}
\tilde{\mathbf{H}}_1 &= \mathbf{H}_2, \\
\tilde{\mathbf{H}}_2 &= \mathbf{G}_1\boldsymbol{\Omega}\mathbf{J}.
\end{aligned} \tag{12}$$

Given the complementary channels in (12), the precoder can be design using (11).

The assumption that $\mathbf{G}_2\boldsymbol{\Omega}\mathbf{J}$ is negligible compared to $\mathbf{H}_2$ makes this method suitable for interference mitigation in the considered scenario. The veracity of this assumption will be further analyzed in Section IV.

[1]In the sense of guaranteeing a minimum Quality of Service (QoS) for all UEs.

C. Full IRS BD (FIB) Precoder Design

Unlike the previous method, this second technique considers in its design the contributions of $\mathbf{G}_2\mathbf{\Omega}\mathbf{J}$ on both useful signal and interference components of (4). To do that, instead of considering the channel $\bar{\mathbf{H}}_2 = \mathbf{H}_2 + \mathbf{G}_2\mathbf{\Omega}\mathbf{J}$ of (1), these components are treated as two separate independent channels of UE2. Based on this consideration, the complementary channels become

$$\begin{aligned} \tilde{\mathbf{H}}_1 &= \left[\mathbf{H}_2^{\mathrm{T}} \quad (\mathbf{G}_2\mathbf{\Omega}\mathbf{J})^{\mathrm{T}}\right]^{\mathrm{T}}, \\ \tilde{\mathbf{H}}_2 &= \mathbf{G}_1\mathbf{\Omega}\mathbf{J}. \end{aligned} \tag{13}$$

It is important to notice that, because the IRS channels are coupled, i.e., they depend on the IRS phase-shift vector, the BD technique is not able to fully cancel all the interfering signals [10]. Thus, in the configuration of (13), the signals traversing these coupled channels can be completely canceled after applying the precoder designed with (11) at the cost of receiving some residual interference from the direct channel $\mathbf{H}_2$.

D. Combiner Design

For both precoding techniques, it is considered that the UEs are employing traditional ZF combining on the strongest channel, which can be expressed as:

$$\begin{aligned} \mathbf{W}_1 &= \left[\mathbf{G}_1\mathbf{\Omega}\mathbf{J}\mathbf{F}_1\right]^{\dagger}, \\ \mathbf{W}_2 &= \left[\mathbf{H}_2\mathbf{F}_2\right]^{\dagger}, \end{aligned} \tag{14}$$

where $(\cdot)^{\dagger}$ represents the Moore-Penrose pseudo-inverse.

IV. Simulations Results

In this section, the performance of each of the proposed methods will be evaluated in terms of SE and compared with i) the solution presented in [10], and with ii) a block-diagonalized MU-MIMO system without IRS that considers a (hypothetical) strong direct path between BS and UE1, then acting as a benchmark.

The scenario illustrated in Fig. 1 was simulated considering $M = 32$ antennas at the BS, $N = 64$ reflecting elements at the IRS, $P = Q = 8$ antennas at the UEs, and the number of RF chains in both UEs was set to 2, which leads to $N_s = N_s^{(1)} = N_s^{(2)} = 2$ streams. The other parameters considered to simulate the studied scenario are exposed in Table I, where d_{2D} represents the horizontal distance.

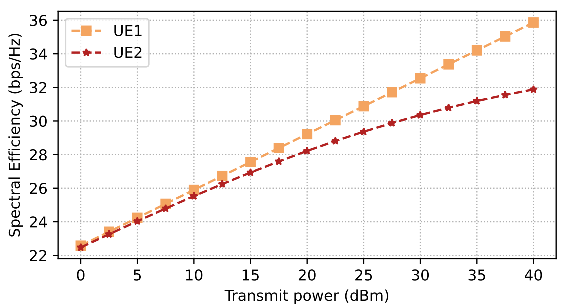

Fig. 2. SE vs BS total transmit power for PIB precoder design.

TABLE I

Simulation parameters.

Parameter	Value
Carrier Frequency	f_c = 28 GHz
Antenna Spacing	$d = \lambda/2$ m
UE height	1.5m
IRS height	8m
BS height	25m
d_{2D} between BS and IRS	100m
d_{2D} between BS and UEs	100m
d_{2D} between IRS and UE1	51.7m
d_{2D} between IRS and UE2	100m
Number of channel realizations	10000
Noise power	$\sigma_1^2 = \sigma_2^2 = -80$ dBm

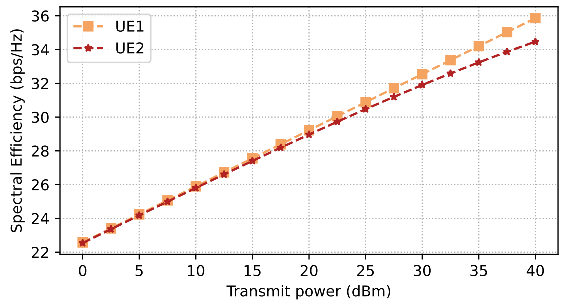

Fig. 3. SE vs BS total transmit power for FIB precoder design.

Fig. 2 presents the SE for each UE as a function of the transmit power employed at the BS when the PIB method is applied in the system. It can be noticed that the SE for UE1 grows linearly, which is expected given that both UEs are in the high signal-to-noise-ratio (SNR) regime and because this technique is able to fully cancel signals flowing through $\mathbf{G}_1\mathbf{\Omega}\mathbf{J}\mathbf{F}_2$. In contrast to it, the UE2 presents two behaviors: until 15 dBm, the SE of UE2 grows linearly, similarly to UE1, and after that it presents a still increasing behavior but with a lower slope. This is also expected since this technique does not cancel the interference arriving from channel $\mathbf{G}_2\mathbf{\Omega}\mathbf{J}$ and as the power of the BS increases so does the interfering signal coming through $\mathbf{G}_2\mathbf{\Omega}\mathbf{J}$.

Considering the assumptions made in Section III-B and based on this result, we can conclude that if the interfering channel has low gain, the beam leakage can be considered minimal and be neglected in the precoder design.

Fig. 3 exhibits the SE for each UE as a function of the transmit power employed at the BS when the FIB method is applied on the system. As expected, the FIB technique outperforms the PIB, since it is more capable of canceling interference arriving through $\mathbf{G}_2\mathbf{\Omega}\mathbf{J}$ and even thought the UE2 has some residual interference arriving through $\mathbf{H}_2$, it does not impact significantly on its performance.

In the following, a comparison between the no-IRS MU-MIMO and the IRS-aided MU-MIMO will be provided. In the no-IRS MU-MIMO, all system parameters are kept the same and the only modification occurs on UE1, whose IRS is replaced by a direct channel with strong LOS similarly to UE2. In this scenario the BS precoder is designed following the classical BD method.

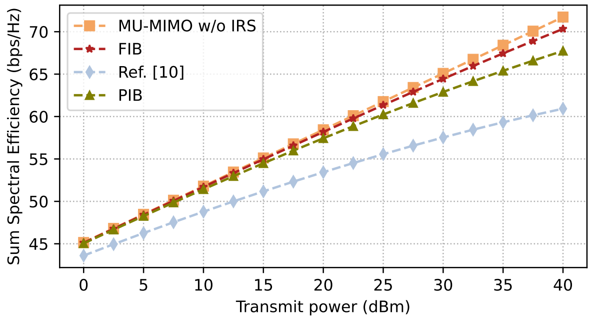

Fig. 4. SE vs BS transmit power for all compered precoder design.

Fig. 4 shows the sum SE as a function of the transmit power at the BS for each method presented, as well as for an adapted version of the precoder from [10]. Therein, only one user performs interference cancellation, since its proposal needs one IRS per user. An additional curve for the MU-MIMO without IRS is also provided, in which the IRS link is replaced by a direct link with the same propagation properties of $\mathbf{H}_2$ for comparison purposes.

It can be observed that the no-IRS setup presents the highest SE. This is due the considered direct link for UE1. However, in the cases where that link is not available, as defined in Section II, the IRS MU-MIMO with the FIB method presents the best results, provided that it is the most robust method. However, the overall performance of the three scenarios does not significantly differs from each other. In contrast, the method in [10] has the worst performance, since its solution need one IRS per user. Hence, one user can fully cancel the interference and the other has unmitigated interference arriving from the IRS.

As for computational complexity, the precoder design method of [10] presents the highest complexity, since it cancels the interference on BS-IRS link instead of BS-IRS-UE link, requiring an SVD of higher dimension for the precoder design. It is followed by the FIB method, which uses the direct link and the BS-IRS-UE link. The PIB is the least complex method herein, since it uses only the BS-IRS-UE link in its formulations.

It is worth mentioning that, in an IRS-aided network, as the PIB method performs similarly to the FIB one and outperforms the method of [10], but with reduced complexity. The PIB may be useful in scenarios in which high spatial multiplexing gains are achievable, directional antennas are employed at the receiver, the interfering channel suffers from blockage, and in many other cases where the interfering channel is negligible compared to the direct one.

V. Conclusion

In this paper, the impact of multi-user interference in IRS-aided networks was studied. Two precoding methods based on BD were proposed to spatially orthogonalize users' signals. It was observed that both methods perform well when the interfering channels created by the IRS has low gain, especially PIB, which has the same limitations of the original BD technique. The FIB technique, even though being more demanding in computational and antenna resources, grants better interference cancellation and performs closely to the no-IRS MU-MIMO case, since the former not only better orthogonalizes the users but also takes advantage of the additional path provided by the IRS for the non-intended users. Future works include the study of multi-cell and multi-IRS scenarios, other mitigation interference methods, joint IRS phase shift and precoder optimization and IRS splitting for serving multiple users.

References

[1] Z. Zhang, Y. Xiao, Z. Ma, *et al.*, "6G Wireless Networks: Vision, Requirements, Architecture, and Key Technologies," *IEEE Veh. Technol. Mag.*, vol. 14, no. 3, pp. 28–41, 2019.

[2] S. Hassouna, M. A. Jamshed, J. Rains, *et al.*, "A survey on reconfigurable intelligent surfaces: Wireless communication perspective," *IET Communications*, vol. 17, no. 5, pp. 497–537, 2023.

[3] N. Rajatheva, I. Atzeni, S. Bicais, *et al.*, *Scoring the terabit/s goal:broadband connectivity in 6g*, 2021. arXiv: `2008.07220 [eess.SP]`.

[4] M. Joham, W. Utschick, and J. Nossek, "Linear transmit processing in MIMO communications systems," *IEEE Trans. Signal Process.*, vol. 53, no. 8, pp. 2700–2712, 2005.

[5] Q. Spencer, A. Swindlehurst, and M. Haardt, "Zero-forcing methods for downlink spatial multiplexing in multiuser MIMO channels," *IEEE Trans. Signal Process.*, vol. 52, no. 2, pp. 461–471, 2004.

[6] S. A. Busari, K. M. S. Huq, S. Mumtaz, L. Dai, and J. Rodriguez, "Millimeter-wave massive mimo communication for future wireless systems: A survey," *IEEE Commun. Surveys Tuts.*, vol. 20, no. 2, pp. 836–869, 2018.

[7] M. Di Renzo, A. Zappone, M. Debbah, *et al.*, "Smart Radio Environments Empowered by Reconfigurable Intelligent Surfaces: How It Works, State of Research, and The Road Ahead," *IEEE J. Sel. Areas Commun.*, vol. 38, no. 11, pp. 2450–2525, 2020.

[8] Q. Wu, S. Zhang, B. Zheng, C. You, and R. Zhang, "Intelligent Reflecting Surface-Aided Wireless Communications: A Tutorial," *IEEE Trans. Commun.*, vol. 69, no. 5, pp. 3313–3351, 2021.

[9] S. Gong, X. Lu, D. T. Hoang, *et al.*, "Toward Smart Wireless Communications via Intelligent Reflecting Surfaces: A Contemporary Survey," *IEEE Commun. Surveys Tuts.*, vol. 22, no. 4, pp. 2283–2314, 2020.

[10] B. Ning, P. Wang, L. Li, Z. Chen, and J. Fang, "Multi-IRS-Aided Multi-User MIMO in mmWave/THz Communications: A Space-Orthogonal Scheme," *IEEE Trans. Commun.*, vol. 70, no. 12, pp. 8138–8152, 2022.

[11] Y. Zheng, T. Lin, and Y. Zhu, "Passive Beamforming for IRS-Assisted MU-MIMO Systems With One-Bit ADCs: An SER Minimization Design Approach," *IEEE Commun. Lett.*, vol. 26, no. 5, pp. 1101–1105, 2022.

[12] J. Zhang, X. Yu, and K. B. Letaief, "Hybrid beamforming for 5g and beyond millimeter-wave systems: A holistic view," *IEEE Open J. Commun. Soc.*, vol. 1, pp. 77–91, 2020.

[13] R. W. Heath Jr. and A. Lozano, *Foundations of MIMO Communication*. Cambridge University Press, 2018.

[14] 3GPP, "Study on channel model for frequencies from 0.5 to 100 GHz," 3rd Generation Partnership Project (3GPP), TR 38.901, Sep. 2017, v.14.2.0. [Online]. Available: `http://www.3gpp.org/DynaReport/38901.htm` (visited on 09/26/2017).

[15] Fazal-E-Asim, A. L. F. de Almeida, B. Sokal, B. Makki, and G. Fodor, *Two-dimensional channel parameter estimation for irs-assisted networks*, 2023. arXiv: `2305.04393 [eess.SP]`.

[16] G. T. de Araújo, A. L. F. de Almeida, R. Boyer, and G. Fodor, "Semi-Blind Joint Channel and Symbol Estimation for IRS-Assisted MIMO Systems," *IEEE Trans. Signal Process.*, vol. 71, pp. 1184–1199, 2023.

[17] P. R. B. Gomes, G. T. d. Araújo, B. Sokal, A. L. F. d. Almeida, B. Makki, and G. Fodor, "Channel estimation in ris-assisted mimo systems operating under imperfections," *IEEE Transactions on Vehicular Technology*, pp. 1–14, 2023.

[18] A. Zappone, M. Di Renzo, F. Shams, X. Qian, and M. Debbah, "Overhead-Aware Design of Reconfigurable Intelligent Surfaces in Smart Radio Environments," *IEEE Trans. Wireless Commun.*, vol. 20, no. 1, pp. 126–141, 2021.