Joint Transmit Diversity and Active/Passive Precoding Design for IRS-Aided Multiuser Communication

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***Abstract*—In this paper, we investigate a novel intelligent reflecting surface (IRS)-aided multiuser communication system, where a multi-antenna base station (BS) integrated with an IRS simultaneously serves multiple low-mobility and high-mobility users via transmit diversity and active/passive precoding, respectively. Specifically, we exploit IRS's common phase shift to help achieve transmit diversity for high-mobility users without any channel state information (CSI), while incorporating the active/passive precoding design into the IRS-integrated BS to serve low-mobility users with known CSI. Then, we formulate and solve a new problem to minimize the total transmit power at the BS by jointly optimizing the reflect precoding at the IRS and the transmit precoding at the BS to cope with interference among different users. Simulation results validate the performance superiority of our proposed IRS-aided multiuser communication.**

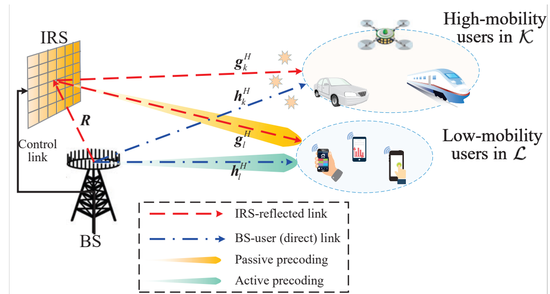

Fig. 1. IRS-aided multiuser downlink communication with high-mobility and low-mobility users.

I. Introduction

IRS, or its equivalents such as reconfigurable intelligent surface (RIS) has drawn considerable attention due to its ability to cost-effectively construct a "smart and reconfigurable radio environment" via real-time passive reflection [1]–[3]. Specifically, IRS is a planar meta-surface comprising a large number of passive elements, each independently adjusting the amplitude and/or phase shifts of the incident signal with ultra-low power consumption. Due to the passive reflection nature of IRS, it does not require any radio-frequency (RF) chain for signal processing/amplification, resulting in even lower power consumption and hardware cost. Not only conceptually appealing, IRS can also be practically fabricated with low profile, lightweight, and conformal geometry, thus facilitating its flexible and dense deployment in future wireless networks to boost both spectral and energy efficiency [1]–[3].

However, prior works on IRS-aided wireless communication have mainly focused on joint active and passive precoding design with the CSI is fully or partially known [1]–[4]. Note that achieving superior passive precoding gains with IRS generally requires more CSI and/or beam training, which, however, can be practically challenging to acquire and maintain. Moreover, since IRS typically consists of a large number of passive elements that lack transmission/processing capabilities, the conventional "all-at-once" IRS channel/beam training schemes (see [4] and the references therein) may incur long training overhead, thus inevitably inducing a considerable delay prior to data transmission. Such training delay is unacceptable for delay-sensitive or high-mobility communication scenarios [5], [6]. In addition, the fast time-varying channel may cause beam misalignment for IRS's passive precoding, severely degrading the performance of IRS in high-mobility and delay-sensitive communication scenarios. As such, besides IRS passive precoding for low-mobility communications, it is essential to develop new and efficient IRS-aided transmission schemes to improve communication reliability for high-mobility communications with very little or even no CSI.

Motivated by the above, we propose a new IRS-aided multiuser downlink communication system as shown in Fig. 1, where an IRS is integrated with a multi-antenna BS to serve multiple high-mobility and low-mobility users simultaneously without and with CSI, respectively. Specifically, we design a new space-time code requiring no CSI to achieve transmit diversity for multiple high-mobility users by exploiting the common phase shift of the IRS. Meanwhile, we design the active/passive precoding at the IRS-integrated BS to serve multiple low-mobility users with their CSI known. Then, we propose an effective algorithm based on alternating optimization (AO) to jointly design the IRS's transmit diversity and passive precoding as well as the BS's active precoding, thus successfully minimizing the total transmit power at the BS subject to individual signal-to-interference-plus-noise ratio (SINR) constraints for all users. Finally, we provide extensive numerical results to validate the performance superiority of our IRS-aided multiuser communication with the proposed co-design of transmit diversity and active/passive precoding.

II. System Model

As shown in Fig. 1, we consider an IRS-aided multiuser downlink communication system, where an IRS comprising N passive reflecting

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elements is co-located and integrated with a BS which is deployed with M antennas to enhance its communication with multiple single-antenna users over a given frequency band. Notably, we consider a new transmission architecture, referred to as the "IRS-integrated BS" in this paper, by taking the IRS as an integral part of the BS, thus enabling the BS to replace the conventional IRS controller for directly tuning the reflection of IRS in real-time.

In addition, we classify users into two groups according to their mobility, i.e. low-mobility and high-mobility users, represented by the sets $\mathcal{L} = \{1, \ldots, L\}$ and $\mathcal{K} = \{L+1, \ldots, L+K\}$, respectively, where the corresponding set sizes, L and K, denote the number of users within each group. Attentive to the different communication requirements and/or channel conditions of these two groups, we consider two basic communication models under a practical transmission protocol: 1) downlink multicasting (MC), where the BS sends common information in short packets to high-mobility users in $\mathcal{K}$; 2) downlink broadcasting (BC), where the BS sends independent information in long packets to different low-mobility users in $\mathcal{L}$, where each transmission packet constitutes one pilot sequence followed by one long data frame.

Let $\boldsymbol{R} \in \mathbb{C}^{N \times M}$, $\boldsymbol{h}_j^H \in \mathbb{C}^{1 \times M}$, and $\boldsymbol{g}_j^H \in \mathbb{C}^{1 \times N}$ denote the baseband equivalent channels for the BS$\to$IRS, BS$\to$user j, and IRS$\to$user j links, respectively, with $j \in \mathcal{L} \cup \mathcal{K}$. For high-mobility users, considering the rich scattering environment and long distance between the BS/IRS and users (see Fig. 1), we assume that the BS$\to$user k channel $\boldsymbol{h}_k^H$ and IRS$\to$user k channel $\boldsymbol{g}_k^H$ follow the Rayleigh fading channel model, i.e.,

$$\boldsymbol{h}_k \sim \mathcal{N}_c\left(\boldsymbol{0}, \frac{\beta}{d_k^{\alpha}} \boldsymbol{I}_M\right), \boldsymbol{g}_k \sim \mathcal{N}_c\left(\boldsymbol{0}, \frac{2\beta}{d_k^{\alpha}} \boldsymbol{I}_N\right), \forall k \in \mathcal{K} \tag{1}$$

where β denotes the reference path gain at the distance of 1 meter (m), the factor of 2 in the latter accounts for the half-space reflection of each IRS element, α denotes the path loss exponent between the IRS-integrated BS and user k, and d_k denotes the propagation distance from the IRS-integrated BS to user k. While the BS$\to$IRS channel $\boldsymbol{R}$ and other channels associated with low-mobility users in $\mathcal{L}$, i.e., $\left\{\boldsymbol{g}_l^H, \boldsymbol{h}_l^H\right\}, \forall l \in \mathcal{L}$, can be completely arbitrary without assuming any specific channel model in this paper.

Furthermore, we let $\boldsymbol{\theta} \triangleq [\theta_1, \theta_2, \ldots, \theta_N]^T$ denote the IRS reflection vector, where the reflection amplitudes of all passive reflecting elements are set as one or the maximum value, i.e., $|\theta_n| = 1, \forall n = 1, \ldots, N$. Thus, under the unit-modulus constraint, any IRS reflection vector $\boldsymbol{\theta}$ can be expressed as

$$\boldsymbol{\theta} = e^{j\varphi} \bar{\boldsymbol{\theta}} \quad \Leftrightarrow \quad \theta_n = e^{j\varphi} \bar{\theta}_n, \qquad \forall n = 1, \ldots, N \tag{2}$$

where $\bar{\boldsymbol{\theta}} \triangleq [\bar{\theta}_1, \bar{\theta}_2, \ldots, \bar{\theta}_N]^T$ denotes the (passive) precoding vector with $|\bar{\theta}_n| = 1, \forall n = 1, \ldots, N$, and φ denotes the IRS's common phase shift applied to all of its reflecting elements. Inspired by the decomposition structure in (2), we propose to design both the common phase shift φ and precoding vector $\bar{\boldsymbol{\theta}}$ of IRS to simultaneously achieve transmit diversity for high-mobility users without CSI and passive precoding for low-mobility users with CSI, respectively.

III. Proposed Scheme and Problem Formulation

In this paper, we conduct linear transmit precoding at the BS, where each data stream is assigned with one dedicated precoding vector. Accordingly, the complex baseband transmitted signal at the BS can be expressed as

TABLE I
THE SPACE-TIME CODE DESIGN FOR IRS-AIDED TRANSMIT DIVERSITY SCHEME VERSUS TRANSMIT BEAMFORMED ALAMOUTI'S SCHEME.

	IRS-aided transmit diversity scheme (for the IRS-integrated BS)		Transmit beamformed Alamouti's scheme [7] (for the multi-antenna BS without IRS)	
Symbol	BS's transmit beam	IRS's common phase	BS's transmit beam 1	BS's transmit beam 2
1	$\boldsymbol{w}_{L+1}\tilde{s}_1$	$\varphi_1 = \angle\tilde{s}_2 - \angle\tilde{s}_1$	$\boldsymbol{w}_{L+1}\tilde{s}_1$	$\boldsymbol{w}'_{L+1}\tilde{s}_2$
2	$-\boldsymbol{w}_{L+1}\tilde{s}_2^*$	$\varphi_2 = \angle\tilde{s}_2 - \angle\tilde{s}_1 + \pi$	$-\boldsymbol{w}_{L+1}\tilde{s}_2^*$	$\boldsymbol{w}'_{L+1}\tilde{s}_1^*$

$$\boldsymbol{x} = \sum_{l=1}^{L} \boldsymbol{w}_l s_l + \boldsymbol{w}_{L+1} \tilde{s} \tag{3}$$

where s_l denotes the independent data symbol for low-mobility user l with $l \in \mathcal{L}$, $\tilde{s}$ denotes the common data symbol for high-mobility users in $\mathcal{K}$, and $\boldsymbol{w}_l \in \mathbb{C}^{M \times 1}$ and $\boldsymbol{w}_{L+1} \in \mathbb{C}^{M \times 1}$ are the corresponding transmit precoding vectors for symbols s_l and $\tilde{s}$, respectively. It is assumed that $\{s_l\}_{l=1}^{L}$ and $\tilde{s}$ are independent random variables with zero mean and unit variance (i.e., normalized power).

A. Transmit Diversity for High-mobility Users

In this subsection, we focus on the transmit diversity design at the IRS-integrated BS without requiring any CSI of high-mobility users. Specifically, we propose an IRS-aided transmit diversity scheme at the BS to deliver common information to high-mobility users in $\mathcal{K}$. In our proposed IRS-aided transmit diversity scheme, the BS and IRS collaboratively encode consecutive pairs of modulated data symbols under a new space-time code design with transmit precoding $\boldsymbol{w}_{L+1}$. We let $\tilde{s}_1$ and $\tilde{s}_2$ denote any pair of two consecutive modulated symbols, which are assumed to be independently drawn from an M-ary (differential) phase shift keying (PSK) constellation. In Table I, we present the space-time code design of the proposed transmit diversity scheme, as compared to the transmit beamformed Alamouti's scheme [7] at the multi-antenna BS without IRS, where $\boldsymbol{w}_{L+1}$ and $\boldsymbol{w}'_{L+1}$ denote two orthogonal transmit precoding vectors. Specifically, during the first symbol period, the BS sends $\boldsymbol{w}_{L+1}\tilde{s}_1$ (multiplexed with other data symbols $\left\{\boldsymbol{w}_l s_l^{(1)}\right\}_{l=1}^{L}$ for the low-mobility users) and the IRS sets its reflection vector as $\boldsymbol{\theta}_1 = e^{j\varphi_1}\bar{\boldsymbol{\theta}}$ with its common phase shift being the phase difference between the two modulated symbols, i.e., $\varphi_1 = \angle\tilde{s}_2 - \angle\tilde{s}_1$. During the second symbol period, the BS sends $-\boldsymbol{w}_{L+1}\tilde{s}_2^*$ (multiplexed with other data symbols $\left\{\boldsymbol{w}_l s_l^{(2)}\right\}_{l=1}^{L}$ for the low-mobility users) and the IRS sets its reflection vector as $\boldsymbol{\theta}_2 = e^{j\varphi_2}\bar{\boldsymbol{\theta}}$ with its common phase shift being $\varphi_2 = \angle\tilde{s}_2 - \angle\tilde{s}_1 + \pi$.

According to the above, the received signal at each high-mobility user during the first symbol period is expressed as

$$\begin{aligned} y_k^{(1)} &= \left(\boldsymbol{h}_k^H + \boldsymbol{g}_k^H \operatorname{diag}(\boldsymbol{\theta}_1) \boldsymbol{R}\right) \left(\sum_{l=1}^{L} \boldsymbol{w}_l s_l^{(1)} + \boldsymbol{w}_{L+1}\tilde{s}_1\right) + n_k^{(1)} \\ &= \left(\boldsymbol{h}_k^H + \boldsymbol{g}_k^H \operatorname{diag}(\boldsymbol{\theta}_1) \boldsymbol{R}\right) \boldsymbol{w}_{L+1}\tilde{s}_1 \\ &\quad + \underbrace{\left(\boldsymbol{h}_k^H + \boldsymbol{g}_k^H \operatorname{diag}(\boldsymbol{\theta}_1) \boldsymbol{R}\right) \sum_{l=1}^{L} \boldsymbol{w}_l s_l^{(1)} + n_k^{(1)}}_{\text{interference-plus-noise: } \tilde{n}_k^{(1)}} \end{aligned} \tag{4}$$

where $n_k^{(1)} \sim \mathcal{N}_c(0, \sigma^2)$ is the zero-mean additive white Gaussian noise (AWGN) at high-mobility user k and $\tilde{n}_k^{(1)} \sim \mathcal{N}_c(0, \tilde{\sigma}_k^2)$ is the equivalent interference-plus-noise with its variance given by

$$\tilde{\sigma}_k^2 = \mathbb{E}\left\{\sum_{l=1}^{L}\left|\left(\boldsymbol{h}_k^H + \boldsymbol{g}_k^H \mathrm{diag}\left(\boldsymbol{\theta}_1\right)\boldsymbol{R}\right)\boldsymbol{w}_l\right|^2\right\} + \sigma^2$$
$$= \sum_{l=1}^{L}\boldsymbol{w}_l^H\left(\frac{\beta}{d_k^\alpha}\boldsymbol{I}_M + \frac{2\beta}{d_k^\alpha}\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{w}_l + \sigma^2. \quad (5)$$

By substituting $\boldsymbol{\theta}_1 = e^{j\varphi_1}\bar{\boldsymbol{\theta}}$ into (4), we have

$$y_k^{(1)} = \underbrace{\boldsymbol{h}_k^H\boldsymbol{w}_{L+1}}_{\tilde{h}_k}\tilde{s}_1 + \underbrace{\boldsymbol{g}_k^H\mathrm{diag}\left(\bar{\boldsymbol{\theta}}\right)\boldsymbol{R}\boldsymbol{w}_{L+1}}_{\tilde{g}_k}e^{j\varphi_1}\tilde{s}_1 + \tilde{n}_k^{(1)}$$
$$\overset{(a)}{=} \tilde{h}_k\tilde{s}_1 + \tilde{g}_k\tilde{s}_2 + \tilde{n}_k^{(1)} \quad (6)$$

where $\tilde{g}_k$ and $\tilde{h}_k$ denote the effective gains of the IRS-reflected and direct channels, respectively, and (a) holds since $\tilde{s}_2 = \tilde{s}_1 e^{j(\angle\tilde{s}_2 - \angle\tilde{s}_1)} = e^{j\varphi_1}\tilde{s}_1$.

Similarly, during the second symbol period, the received signal at each high-mobility user is given by

$$y_k^{(2)} = \left(\boldsymbol{h}_k^H + \boldsymbol{g}_k^H\mathrm{diag}\left(\boldsymbol{\theta}_2\right)\boldsymbol{R}\right)\left(\sum_{l=1}^{L}\boldsymbol{w}_l s_l^{(2)} - \boldsymbol{w}_{L+1}\tilde{s}_2^*\right) + n_k^{(2)}$$
$$= -\left(\boldsymbol{h}_k^H + \boldsymbol{g}_k^H\mathrm{diag}\left(\boldsymbol{\theta}_2\right)\boldsymbol{R}\right)\boldsymbol{w}_{L+1}\tilde{s}_2^*$$
$$+ \underbrace{\left(\boldsymbol{h}_k^H + \boldsymbol{g}_k^H\mathrm{diag}\left(\boldsymbol{\theta}_2\right)\boldsymbol{R}\right)\sum_{l=1}^{L}\boldsymbol{w}_l s_l^{(2)} + n_k^{(2)}}_{\text{interference}-\text{plus}-\text{noise: }\tilde{n}_k^{(2)}} \quad (7)$$

where $n_k^{(2)} \sim \mathcal{N}_c(0, \sigma^2)$ is the zero-mean AWGN at high-mobility user k and $\tilde{n}_k^{(2)} \sim \mathcal{N}_c(0, \tilde{\sigma}_k^2)$ is the equivalent interference-plus-noise with its variance given in (5) by replacing $\boldsymbol{\theta}_1$ with $\boldsymbol{\theta}_2$. Likewise, by substituting $\boldsymbol{\theta}_2 = e^{j\varphi_2}\bar{\boldsymbol{\theta}}$ into (7), we have

$$y_k^{(2)} = -\underbrace{\boldsymbol{h}_k^H\boldsymbol{w}_{L+1}}_{\tilde{h}_k}\tilde{s}_2^* - \underbrace{\boldsymbol{g}_k^H\mathrm{diag}\left(\bar{\boldsymbol{\theta}}\right)\boldsymbol{R}\boldsymbol{w}_{L+1}}_{\tilde{g}_k}e^{j\varphi_2}\tilde{s}_2^* + \tilde{n}_k^{(2)}$$
$$\overset{(b)}{=} -\tilde{h}_k\tilde{s}_2^* + \tilde{g}_k\tilde{s}_1^* + \tilde{n}_k^{(2)} \quad (8)$$

where (b) holds since $\tilde{s}_1^* = -\tilde{s}_2^* e^{j(\angle\tilde{s}_2 - \angle\tilde{s}_1 + \pi)} = -e^{j\varphi_2}\tilde{s}_2^*$. For the downlink MC, each high-mobility user aims to decode $\tilde{s}_1$ and $\tilde{s}_2$ while treating other symbols for low-mobility users as interference. As such, let $\boldsymbol{y}_k \triangleq \left[y_k^{(1)}, (y_k^{(2)})^*\right]^T$ denote the received signal vector for each transmitted pair of $\tilde{s}_1$ and $\tilde{s}_2$, (6) and (8) can be rewritten in a compact form as

$$\boldsymbol{y}_k = \underbrace{\begin{bmatrix}\tilde{h}_k & \tilde{g}_k \\ \tilde{g}_k^* & -\tilde{h}_k^*\end{bmatrix}}_{\boldsymbol{H}}\underbrace{\begin{bmatrix}\tilde{s}_1 \\ \tilde{s}_2\end{bmatrix}}_{\tilde{\boldsymbol{s}}} + \underbrace{\begin{bmatrix}\tilde{n}_k^{(1)} \\ \left(\tilde{n}_k^{(2)}\right)^*\end{bmatrix}}_{\tilde{\boldsymbol{n}}_k} \quad (9)$$

where $\boldsymbol{H}$ denotes the equivalent channel matrix, $\tilde{\boldsymbol{s}}$ is the modulated symbol vector, and $\tilde{\boldsymbol{n}}_k$ is the interference-plus-noise vector. Then, we left-multiply the received signal vector $\boldsymbol{y}_k$ in (9) by $\boldsymbol{H}^H$, which yields

$$\bar{\boldsymbol{y}}_k = \boldsymbol{H}^H\boldsymbol{y}_k = \boldsymbol{H}^H\boldsymbol{H}\tilde{\boldsymbol{s}} + \bar{\boldsymbol{n}}_k \quad (10)$$

with $\bar{\boldsymbol{y}} \triangleq [\bar{y}_1, \bar{y}_2]^T$ and $\bar{\boldsymbol{n}}_k \triangleq \boldsymbol{H}^H\tilde{\boldsymbol{n}}_k$, where we have $\boldsymbol{H}^H\boldsymbol{H} = \left(|\tilde{h}_k|^2 + |\tilde{g}_k|^2\right)\boldsymbol{I}_2$ and it can be verified that $\bar{\boldsymbol{n}}_k$ is the equivalent interference-plus-noise vector with $\bar{\boldsymbol{n}}_k \sim \mathcal{N}_c(\boldsymbol{0}, (|\tilde{h}_k|^2 + |\tilde{g}_k|^2)\tilde{\sigma}_k^2\boldsymbol{I}_2)$.

Accordingly, the average received SINR at each high-mobility user is given by

$$\frac{\mathbb{E}\{|\tilde{h}_k|^2 + |\tilde{g}_k|^2\}}{\tilde{\sigma}_k^2} = \frac{\mathbb{E}\left\{|\boldsymbol{h}_k^H\boldsymbol{w}_{L+1}|^2 + |\boldsymbol{g}_k^H\mathrm{diag}\left(\bar{\boldsymbol{\theta}}\right)\boldsymbol{R}\boldsymbol{w}_{L+1}|^2\right\}}{\sum_{l=1}^{L}\boldsymbol{w}_l^H\left(\frac{\beta}{d_k^\alpha}\boldsymbol{I}_M + \frac{2\beta}{d_k^\alpha}\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{w}_l + \sigma^2}$$
$$= \frac{\boldsymbol{w}_{L+1}^H\left(\boldsymbol{I}_M + 2\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{w}_{L+1}}{\sum_{l=1}^{L}\boldsymbol{w}_l^H\left(\boldsymbol{I}_M + 2\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{w}_l + d_k^\alpha\sigma^2/\beta}, \quad \forall k \in \mathcal{K}. \quad (11)$$

B. Active/Passive Precoding for Low-mobility Users

Apart from serving high-mobility users via the IRS-aided transmit diversity, the IRS-integrated BS also performs active/passive precoding simultaneously to serve low-mobility users with known CSI . Specifically, given the pre-determined common IRS phase shift which helps achieve transmit diversity for high-mobility users, we carefully design the precoding vector $\bar{\boldsymbol{\theta}}$ in (2) for low-mobility users, as shown in Fig. 1.

The signal received at low-mobility user l from both the BS→user and BS→IRS→user channels is then expressed as

$$y_l = \left(\boldsymbol{h}_l^H + \boldsymbol{g}_l^H\mathrm{diag}\left(\boldsymbol{\theta}\right)\boldsymbol{R}\right)\left(\sum_{j=1}^{L}\boldsymbol{w}_j s_j + \boldsymbol{w}_{L+1}\tilde{s}\right) + n_l$$
$$= \left(\boldsymbol{h}_l^H + e^{j\varphi}\boldsymbol{g}_l^H\mathrm{diag}\left(\bar{\boldsymbol{\theta}}\right)\boldsymbol{R}\right)\boldsymbol{w}_l s_l + \left(\boldsymbol{h}_l^H + e^{j\varphi}\boldsymbol{g}_l^H\mathrm{diag}\left(\bar{\boldsymbol{\theta}}\right)\boldsymbol{R}\right)$$
$$\left(\sum_{j\neq l}^{L}\boldsymbol{w}_j s_j + \boldsymbol{w}_{L+1}\tilde{s}\right) + n_l \quad (12)$$

where $n_l \sim \mathcal{N}_c(0, \sigma^2)$ is the AWGN at low-mobility user l with σ^2 being the equivalent noise power. According to (12), the SINR of low-mobility user l is given by

$$\mathrm{SINR}_l = \frac{\left|\left(\boldsymbol{h}_l^H + e^{j\varphi}\boldsymbol{g}_l^H\mathrm{diag}\left(\bar{\boldsymbol{\theta}}\right)\boldsymbol{R}\right)\boldsymbol{w}_l\right|^2}{\sum_{j\neq l}^{L+1}\left|\left(\boldsymbol{h}_l^H + e^{j\varphi}\boldsymbol{g}_l^H\mathrm{diag}\left(\bar{\boldsymbol{\theta}}\right)\boldsymbol{R}\right)\boldsymbol{w}_j\right|^2 + \sigma^2}. \quad (13)$$

C. Problem Formulation

In this paper, we aim to minimize the total transmit power at the BS by jointly optimizing the transmit precoding at the BS and reflect precoding at the IRS, adhering to individual SINR constraints for all users. For simplicity of notation, we let $\boldsymbol{W} = [\boldsymbol{w}_1, \cdots, \boldsymbol{w}_L, \boldsymbol{w}_{L+1}]$ denote the transmit precoding matrix at the BS. Accordingly, the problem is formulated as

$$\text{(P1)}: \min_{\boldsymbol{W}, \bar{\boldsymbol{\theta}}} \quad \sum_{j=1}^{L+1}\|\boldsymbol{w}_j\|^2 \quad (14)$$
$$\text{s.t.} \quad \frac{\left|\left(\boldsymbol{h}_l^H + e^{j\varphi}\boldsymbol{g}_l^H\mathrm{diag}\left(\bar{\boldsymbol{\theta}}\right)\boldsymbol{R}\right)\boldsymbol{w}_l\right|^2}{\sum_{j\neq l}^{L+1}\left|\left(\boldsymbol{h}_l^H + e^{j\varphi}\boldsymbol{g}_l^H\mathrm{diag}(\bar{\boldsymbol{\theta}})\boldsymbol{R}\right)\boldsymbol{w}_j\right|^2 + \sigma^2} \geq \gamma_l, \forall l \in \mathcal{L} \quad (15)$$
$$\frac{\boldsymbol{w}_{L+1}^H\left(\boldsymbol{I}_M + 2\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{w}_{L+1}}{\sum_{l=1}^{L}\boldsymbol{w}_l^H\left(\boldsymbol{I}_M + 2\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{w}_l + \bar{\sigma}_{L+1}^2} \geq \gamma_{L+1} \quad (16)$$
$$|\bar{\theta}_n| = 1, \forall n = 1, \ldots, N. \quad (17)$$

where $\gamma_l > 0, l \in \mathcal{L}$ is the minimum individual SINR requirement of each low-mobility user and $\gamma_{L+1} > 0$ is the minimum SINR requirement of all high-mobility users in $\mathcal{K}$ with $\bar{\sigma}_{L+1}^2 \triangleq \max_{k\in\mathcal{K}} \frac{d_k^{\alpha}\sigma^2}{\beta}$. Although the objective function in (14) is convex, it is tricky to solve (P2) due to the non-convex constraints. Specifically, the unit-modulus constraint in (17) is non-convex, and the transmit precoding $\boldsymbol{W}$ and reflect precoding $\bar{\boldsymbol{\theta}}$ are coupled, which thus need to be optimized jointly. In addition, (15) and (16) indicate that all the low-mobility and high-mobility users are coupled by their mutual interference, which poses even more challenges to this problem. Although there is no standard method for optimally solving such a non-convex optimization problem, we employ the AO technique to solve (P2) efficiently, as elaborated in the next section.

IV. Proposed Solution to Problem (P1)

In this section, we propose an AO algorithm for solving (P1). Specifically, we alternately optimize the transmit precoding $\boldsymbol{W}$ at the BS and the reflect precoding $\bar{\boldsymbol{\theta}}$ at the IRS in an iterative manner, until the convergence is achieved. For any given reflect precoding $\bar{\boldsymbol{\theta}}$, the combined channel from the BS to each low-mobility user is given by

$$\bar{\boldsymbol{h}}_l^H = \boldsymbol{h}_l^H + e^{j\varphi}\boldsymbol{g}_l^H \mathrm{diag}\left(\bar{\boldsymbol{\theta}}\right)\boldsymbol{R} \tag{18}$$

and thus the problem (P1) can be reduced to

$$\text{(P2):}\min_{\boldsymbol{W}} \quad \sum_{j=1}^{L+1} \|\boldsymbol{w}_j\|^2 \tag{19}$$

$$\text{s.t.} \quad \frac{\left|\bar{\boldsymbol{h}}_l^H \boldsymbol{w}_l\right|^2}{\sum_{j\neq l}^{L+1}\left|\bar{\boldsymbol{h}}_l^H\boldsymbol{w}_j\right|^2+\sigma^2} \geq \gamma_l, \forall l \in \mathcal{L} \tag{20}$$

$$\frac{\boldsymbol{w}_{L+1}^H\left(\boldsymbol{I}_M + 2\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{w}_{L+1}}{\sum_{l=1}^{L}\boldsymbol{w}_l^H\left(\boldsymbol{I}_M+2\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{w}_l + \bar{\sigma}_{L+1}^2} \geq \gamma_{L+1} \tag{21}$$

Note that the problem (P2) has an additional average SINR constraint for high-mobility users as compared to the conventional power minimization problem in the multiuser multiple-input single-output (MISO) downlink broadcast system. It is found that after reformulating problem (P2) as a convex problem, we can apply the second-order cone program (SOCP) [8] or semidefinite program (SDP) [9] to solve it efficiently. Nevertheless, this numerical solution has few insights into the structure of its optimal/suboptimal solution and has a relatively high complexity. To address this issue, we present the zero-forcing (ZF) transmit precoding in the following.

A. ZF Transmit Precoding Optimization

For the transmit precoding design at BS, we consider applying the ZF criterion to eliminate the multiuser interference among all low-mobility users, with the closed-form expression given by

$$\bar{\boldsymbol{W}}_{ZF} = \bar{\boldsymbol{H}}\left(\bar{\boldsymbol{H}}^H\bar{\boldsymbol{H}}\right)^{-1}\bar{\boldsymbol{P}} \tag{49}$$

where $\bar{\boldsymbol{H}}^H = \left[\bar{\boldsymbol{h}}_1,\ldots,\bar{\boldsymbol{h}}_L\right]^H$ denotes the equivalent channel matrix and $\bar{\boldsymbol{P}} = \mathrm{diag}\left(\sqrt{P_1},\ldots,\sqrt{P_L}\right)$ denotes the diagonal transmit power matrix for the L low-mobility users. Furthermore, we can set $P_l = \gamma_l\sigma^2$ to meet the constraints of (20) with equality to minimize the transmit power for each low-mobility user.

Moreover, to avoid extra interference to low-mobility users, the transmit precoding $\boldsymbol{w}_{L+1}$ for high-mobility users can be designed by exploiting the null space of the channel matrix $\bar{\boldsymbol{H}}^H$. Specifically, let $\boldsymbol{F} \in \mathbb{C}^{M\times T}$ with $M - L \leq T \leq M$ denote an orthonormal basis for the null space of $\bar{\boldsymbol{H}}^H$, we can design the transmit precoding for high-mobility users as $\boldsymbol{w}_{L+1} = \boldsymbol{F}\boldsymbol{v}$, with $\boldsymbol{v} \in \mathbb{C}^{T\times 1}$ being a linear combination vector. Apparently, we have $\bar{\boldsymbol{H}}^H\boldsymbol{w}_{L+1} = \bar{\boldsymbol{H}}^H\boldsymbol{F}\boldsymbol{v} = \boldsymbol{0}_{L\times 1}$, which implies that the transmit precoding $\boldsymbol{w}_{L+1}$ for high-mobility users will not cause any interference to low-mobility users, i.e., $\boldsymbol{h}_l^H\boldsymbol{w}_{L+1} = 0, \forall l \in \mathcal{L}$ in (20). Thus, given the ZF transmit precoding in (49), problem (P2) is reduced to

$$\text{(P3):}\min_{\boldsymbol{v}} \quad \|\boldsymbol{v}\|^2 \tag{50}$$

$$\text{s.t.} \quad \frac{\boldsymbol{v}^H\boldsymbol{F}^H\left(\boldsymbol{I}_M+2\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{F}\boldsymbol{v}}{\mathrm{tr}\left(\bar{\boldsymbol{W}}_{ZF}^H(\boldsymbol{I}_M+2\boldsymbol{R}^H\boldsymbol{R})\bar{\boldsymbol{W}}_{ZF}\right)+\bar{\sigma}_{L+1}^2} \geq \gamma_{L+1} \tag{51}$$

where $\|\boldsymbol{w}_{L+1}\|^2 = \|\boldsymbol{F}\boldsymbol{v}\|^2 = \boldsymbol{v}^H\boldsymbol{F}^H\boldsymbol{F}\boldsymbol{v} = \boldsymbol{v}^H\boldsymbol{v} = \|\boldsymbol{v}\|^2$ due to the orthonormality of $\boldsymbol{F}$, i.e., $\boldsymbol{F}^H\boldsymbol{F} = \boldsymbol{I}_{T\times T}$. Moreover, one can easily show that the inequality constraint in (51) is satisfied with equality at the optimum; otherwise, the optimal $\boldsymbol{v}$ could be scaled down to satisfy the constraint with equality, thereby decreasing the objective function and contradicting optimality. Similar to [10], it follows that the optimal $\boldsymbol{v}$ should be selected as the scaled principal eigenvector of $\boldsymbol{F}^H\left(\boldsymbol{I}_M + 2\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{F}$ associated with the largest eigenvalue, which is given by

$$\boldsymbol{v}^\star = \sqrt{p_{L+1}}\bar{\boldsymbol{v}} \quad \Rightarrow \quad \boldsymbol{w}_{L+1}^\star = \boldsymbol{F}\boldsymbol{v}^\star \tag{52}$$

where $\bar{\boldsymbol{v}} = \mathbb{P}\left\{\boldsymbol{F}^H\left(\boldsymbol{I}_M+2\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{F}\right\}$ is the normalized principal eigenvector and thus $\boldsymbol{F}\bar{\boldsymbol{v}}$ is the normalized transmit precoding direction for high-mobility users, and p_{L+1} denotes the precoding power, which is chosen to meet the equality of the constrain in (51), i.e.,

$$p_{L+1} = \frac{\gamma_{L+1}\left(\mathrm{tr}\left(\bar{\boldsymbol{W}}_{ZF}^H\left(\boldsymbol{I}_M+2\boldsymbol{R}^H\boldsymbol{R}\right)\bar{\boldsymbol{W}}_{ZF}\right)+\bar{\sigma}_{L+1}^2\right)}{\bar{\boldsymbol{v}}^H\boldsymbol{F}^H\left(\boldsymbol{I}_M+2\boldsymbol{R}^H\boldsymbol{R}\right)\boldsymbol{F}\bar{\boldsymbol{v}}}. \tag{53}$$

B. Reflect Precoding Optimization

For fixed transmit precoding $\boldsymbol{W}$ at the BS, problem (P1) reduces to a feasibility-check problem. Specifically, let $a_{l,j}^* = \boldsymbol{h}_l^H\boldsymbol{w}_j$ and $\boldsymbol{b}_{l,j}^H = e^{j\varphi}\boldsymbol{g}_l^H\mathrm{diag}\left(\boldsymbol{R}\boldsymbol{w}_j\right)$, problem (P1) can be reduced to

$$\text{(P4):Find} \quad \bar{\boldsymbol{\theta}} \tag{54}$$

$$\text{s.t.} \quad \frac{\left|\boldsymbol{b}_{l,l}^H\bar{\boldsymbol{\theta}}+a_{l,l}^*\right|^2}{\sum_{j\neq l}^{L+1}\left|\boldsymbol{b}_{l,j}^H\bar{\boldsymbol{\theta}}+a_{l,j}^*\right|^2+\sigma^2} \geq \gamma_l, \forall l \in \mathcal{L} \tag{55}$$

$$|\bar{\theta}_n| = 1, \forall n = 1,\ldots,N. \tag{56}$$

Problem (P4) remains non-convex due to the non-convex unit-modulus constraints in (56). However, by transforming the constraints (55) and (56) into quadratic forms, the SDR technique can be applied to solve problem (P4) efficiently.

Specifically, problem (P4) can be transformed into

$$\text{(P4.1):Find} \quad \bar{\boldsymbol{\theta}} \tag{57}$$

$$\text{s.t.} \quad \tilde{\boldsymbol{\theta}}^H\boldsymbol{B}_{l,l}\tilde{\boldsymbol{\theta}} + |a_{l,l}|^2 \geq$$

Algorithm 1 The AO Algorithm for Solving Problem (P1)

1: Initialization: $\bar{\boldsymbol{\theta}} := \bar{\boldsymbol{\theta}}^{(0)}$, $\boldsymbol{W} := \boldsymbol{W}^{(0)}$, and the iteration number $i = 0$.
2: **repeat**
3: Solve problem (P3) for given $\bar{\boldsymbol{\theta}}^{(i)}$ via the ZF transmit precoding design in Section IV-A, and denote the solution as $\boldsymbol{W}^{(i)}$.
4: Solve problem (P4.2) for given $\boldsymbol{W}^{(i)}$ via the SDR method in Section IV-B, and denote the solution after performing Gaussian randomization as $\bar{\boldsymbol{\theta}}^{(i+1)}$.
5: Update $i := i + 1$.
6: **until** The fractional increase of the objective value in (14) is below a threshold $\xi > 0$ or the iteration number i reaches the pre-designed number of iterations I.

$$\gamma_l\left(\sum_{j\neq l}^{L+1}\tilde{\boldsymbol{\theta}}^H\boldsymbol{B}_{l,j}\tilde{\boldsymbol{\theta}}+\sum_{j\neq l}^{L+1}|a_{l,j}|^2+\sigma^2\right),\forall l\in\mathcal{L} \tag{58}$$

$$|t| = 1, |\bar{\theta}_n| = 1, \forall n = 1, \ldots, N. \tag{59}$$

where

$$\boldsymbol{B}_{l,j} = \begin{bmatrix} \boldsymbol{b}_{l,j}\boldsymbol{b}_{l,j}^H & a_{l,j}^*\boldsymbol{b}_{l,j} \\ a_{l,j}\boldsymbol{b}_{l,j}^H & 0 \end{bmatrix}, \quad \tilde{\boldsymbol{\theta}} = \begin{bmatrix} \bar{\boldsymbol{\theta}} \\ t \end{bmatrix} \tag{60}$$

with t being an auxiliary variable. As $\tilde{\boldsymbol{\theta}}^H\boldsymbol{B}_{l,j}\tilde{\boldsymbol{\theta}} = \text{tr}\left(\boldsymbol{B}_{l,j}\tilde{\boldsymbol{\theta}}\tilde{\boldsymbol{\theta}}^H\right)$, we further define $\tilde{\boldsymbol{\Theta}} = \tilde{\boldsymbol{\theta}}\tilde{\boldsymbol{\theta}}^H$, which is required to satisfy the condition of $\tilde{\boldsymbol{\Theta}} \succeq \boldsymbol{0}$ and rank $\left(\tilde{\boldsymbol{\Theta}}\right) = 1$. However, the rank-one constraint is non-convex and solving the feasibility-check problem may not yield an efficient reflect precoding for minimizing the total transmit power at the BS. To obtain a more efficient solution, we relax the rank-one constraint and reformulate problem (P4.1) with an explicit objective, which is given by

$$\text{(P4.2)}: \max_{\tilde{\boldsymbol{\Theta}},\{c_l\}_{l=1}^L} \quad \sum_{l=1}^{L} c_l \tag{61}$$

$$\text{s.t.} \quad \text{tr}\left(\boldsymbol{B}_{l,l}\tilde{\boldsymbol{\Theta}}\right) + |a_{l,l}|^2 \geq$$

$$\gamma_l\left(\sum_{j\neq l}^{L+1}\text{tr}\left(\boldsymbol{B}_{l,j}\tilde{\boldsymbol{\Theta}}\right)+\sum_{j\neq l}^{L+1}|a_{l,j}|^2+\sigma^2\right)+c_l,\forall l\in\mathcal{L} \tag{62}$$

$$\left[\tilde{\boldsymbol{\Theta}}\right]_{n,n} = 1, \quad \forall n = 1, \ldots, N+1 \tag{63}$$

$$\tilde{\boldsymbol{\Theta}} \succeq \boldsymbol{0}, c_l \geq 0, \quad \forall l \in \mathcal{L} \tag{64}$$

where the slack variable c_l can be interpreted as the "SINR residual" of low-mobility user l in precoding optimization. It can be verified that problem (P4.2) is a convex SDP problem, which can be optimally solved with existing convex optimization solvers such as CVX [11]. Although the SDR technique may not provide a rank-one solution, we can retrieve a high-quality rank-one solution to problem (P4.1) from the obtained higher-rank solution by using e.g., Gaussian randomization.

To sum up, we solve problem (P1) by alternately solving subproblems (P2) and (P4.2) in an iterative manner. The details of the proposed AO algorithm are summarized in Algorithm 1.

V. Simulation Results

In this section, we provide simulation results to validate the performance of the IRS-aided multiuser communication system with our proposed co-design of transmit diversity and active/passive precoding. In a three-dimensional (3D) Cartesian coordinate system, the IRS locates on the $x-z$ plane with its center point at $(0, 0, 4)$ m; while the multi-antenna BS locates on the $y-z$ plane with its center point at $(0, 0.5, 4)$ m. Without loss of generality, we assume that all the low-mobility and high-mobility users are randomly located in the front half-space reflection area of IRS with the same distance of $d = d_j = 50$ m, $j \in \mathcal{L} \cup \mathcal{K}$ on the $x - y$ plane. The numbers of high-mobility and low-mobility users are both set to 3. In addition, we assume that all low-mobility users have the same SINR target, $\gamma = \gamma_l, \forall l \in \mathcal{L}$ and all high-mobility users also share the SINR target of γ_{L+1} for their common information.

We adopt the square uniform planar array (UPA) model for the IRS, configuring an equal number of reflecting elements per dimension along the x- and z-axes, i.e., $N_x = N_z = \sqrt{N}$. The reference path gain at the distance of 1 m is set as $\beta = -30$ dB for all individual links. The path loss exponent is set as $\alpha = 2$ for the link between the BS and its integrated IRS (which is modeled by the near-field line-of-sight (LoS) channel due to the short distance) and set as $\alpha = 2.5$ for the other links (due to the relatively large distance). In addition, the channels between the BS/IRS and the low-mobility users, i.e., $\left\{\boldsymbol{g}_l^H, \boldsymbol{h}_l^H\right\}, \forall l \in \mathcal{L}$, are modeled by the Rician fading with the Rician factor of 5 dB. Unless otherwise stated, the number of BS antennas is $M = 8$, the wavelength is $\lambda = 0.05$ m, the element spacing is set as $\Delta = \lambda/2 = 0.025$ m, and the noise power at each user is set equal as $\sigma^2 = -85$ dBm.

For the joint active and passive precoding design in problem (P2), we consider two benchmark schemes as follows.

- **Random phase with ZF transmit precoding**: In this scheme, the reflect coefficients in $\bar{\boldsymbol{\theta}}$ are randomly generated following a uniform distribution within $[0, 2\pi)$ at the IRS and the ZF transmit precoding in Section IV-A is applied at the BS.
- **DFT-based codebook search with ZF transmit precoding**: In this scheme, the passive precoding $\bar{\boldsymbol{\theta}}$ is searched over a discrete Fourier transform (DFT)-based codebook (denoted by $\mathcal{D}$) to maximize the minimum channel gain among all low-mobility users, i.e., $\bar{\boldsymbol{\theta}}^\star = \arg\max_{\bar{\boldsymbol{\theta}}\in\mathcal{D}} \min_{l\in\mathcal{L}}\|\bar{\boldsymbol{h}}_l^H\|^2$. Then, given $\bar{\boldsymbol{\theta}}^\star$, the ZF transmit precoding in Section IV-A is applied at the BS.

In Fig. 2, we show the transmit power at the BS versus BS/IRS-user distance, with $N = 400$, $\gamma = 10$, and $\gamma_{L+1} = 1$. It is observed that regardless of the BS/IRS-user distance, the proposed AO algorithm attains the lowest transmit power. In particular, the proposed AO algorithm requires about 1.8 dB and 2.7 dB lower transmit power than the DFT-based codebook search and the random phase schemes, respectively. Mover, by exploiting the pronounced passive precoding gain, the IRS-aided system using the proposed AO algorithm requires much lower transmit power (up to 8.2 dB) than the baseline system without IRS.

Next, we show the symbol error rate (SER) performance of our proposed IRS-aided transmit diversity scheme, where the 8-ary PSK modulation and the following benchmark schemes are adopted.

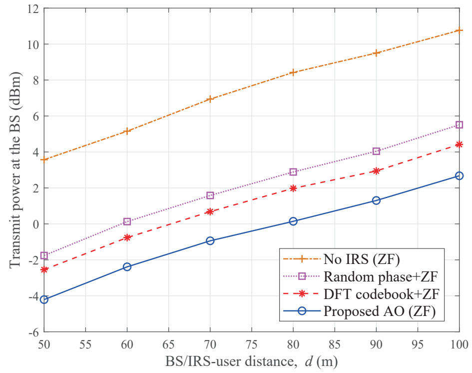

Fig. 2. BS transmit power versus BS/IRS-user distance, with $N = 400$, $\gamma = 10$, and $\gamma_{L+1} = 1$.

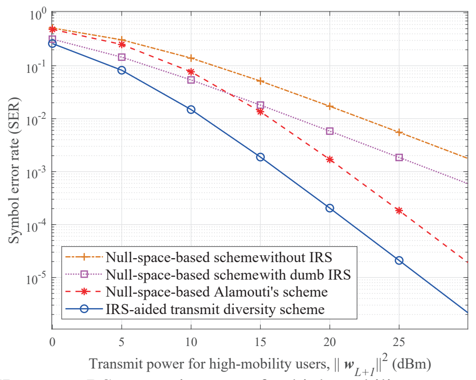

Fig. 3. SER versus BS transmit power for high-mobility users, with $N = 400$ and $\gamma = 10$.

- **Null-space-based scheme without IRS**: In this scheme, $\boldsymbol{w}_{L+1}$ is designed to project into the null space of $\bar{\boldsymbol{H}}^H$ and then the BS multicasts the modulated symbols $\boldsymbol{w}_{L+1}\tilde{s}$ to high-mobility users.
- **Null-space-based scheme with dumb IRS**: In this scheme, the IRS fixes its common phase and the BS multicasts the modulated symbols $\boldsymbol{w}_{L+1}\tilde{s}$ to high-mobility users with $\boldsymbol{w}_{L+1}$ projected into the null space of $\bar{\boldsymbol{H}}^H$.
- **Null-space-based Alamouti's scheme**: In this scheme, two orthogonal transmit precoding vectors $\boldsymbol{w}_{L+1}$ and $\boldsymbol{w}'_{L+1}$ are generated to create an orthogonal channel condition for transmitting the space-time code as in Table I to high-mobility users.

In Fig. 3, we compare the average SER versus the transmit power at the BS for high-mobility users, with $N = 400$ and $\gamma = 10$. It can be observed that our proposed IRS-aided transmit diversity scheme achieves a transmit diversity gain of order two as compared to the baseline systems without space-time code design. In addition, owing to the additional signal reflection power induced by the IRS for achieving a higher average channel gain, the IRS-aided transmit diversity scheme (null-space-based scheme with dumb IRS) achieves up to 5 dB gain over the null-space-based Alamouti's scheme (null-space-based scheme without IRS).

VI. Conclusions

In this paper, we explored a novel IRS-aided multiuser communication system with a new architecture of IRS-integrated BS, and proposed a co-design of transmit diversity and active/passive precoding to simultaneously serve multiple high-mobility and low-mobility users without and with CSI, respectively. Specifically, we designed the IRS's common phase shift to help achieve transmit diversity at the BS without any CSI to serve high-mobility users; meanwhile, the active/passive precoding gain can be simultaneously achieved for low-mobility users with their CSI known at the IRS-integrated BS. We then formulated and solved a joint reflect and transmit precoding optimization problem to minimize the total transmit power at the BS. Simulation results demonstrated the substantial performance gain achieved by the new IRS-aided multiuser communication system with the proposed co-design of transmit diversity and active/passive precoding.

References

[1] 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, May 2021.

[2] Q. Wu and R. Zhang, "Towards smart and reconfigurable environment: Intelligent reflecting surface aided wireless network," *IEEE Commun. Mag.*, vol. 58, no. 1, pp. 106–112, Jan. 2020.

[3] M. Di Renzo *et al.*, "Smart radio environments empowered by reconfigurable AI meta-surfaces: An idea whose time has come," *EURASIP J. Wireless Commun. Netw.*, vol. 2019:129, May 2019.

[4] B. Zheng, C. You, W. Mei, and R. Zhang, "A survey on channel estimation and practical passive beamforming design for intelligent reflecting surface aided wireless communications," *IEEE Commun. Surveys Tuts.*, vol. 24, no. 2, pp. 1035–1071, Second Quarter 2022.

[5] Y. Chen, Y. Wang, and L. Jiao, "Robust transmission for reconfigurable intelligent surface aided millimeter wave vehicular communications with statistical CSI," *IEEE Trans. Wireless Commun.*, vol. 21, no. 2, pp. 928–944, Feb. 2022.

[6] S. Sun and H. Yan, "Channel estimation for reconfigurable intelligent surface-assisted wireless communications considering doppler effect," *IEEE Wireless Commun. Lett.*, vol. 10, no. 4, pp. 790–794, Apr. 2021.

[7] S. X. Wu, W.-K. Ma, and A. M.-C. So, "Physical-layer multicasting by stochastic transmit beamforming and Alamouti space-time coding," *IEEE Trans. Signal Process.*, vol. 61, no. 17, pp. 4230–4245, Sept. 2013.

[8] A. Wiesel, Y. Eldar, and S. Shamai, "Linear precoding via conic optimization for fixed MIMO receivers," *IEEE Trans. Signal Process.*, vol. 54, no. 1, pp. 161–176, Dec. 2006.

[9] M. Bengtsson and B. Ottersten, "Optimal and suboptimal transmit beamforming," in *Handbook of Antennas in Wireless Communications*. L. C. Godara, Ed. CRC Press, 2001.

[10] V. Havary-Nassab, S. Shahbazpanahi, A. Grami, and Z.-Q. Luo, "Distributed beamforming for relay networks based on second-order statistics of the channel state information," *IEEE Trans. Signal Process.*, vol. 56, no. 9, pp. 4306–4316, Sept. 2008.

[11] M. Grant and S. Boyd, "CVX: Matlab software for disciplined convex programming," 2016. [Online]. Available: http://cvxr.com/cvx