Full-Duplex-Enabled Joint Communications and Sensing with Reconfigurable Intelligent Surfaces

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***Abstract*—The full-duplex (FD) technology has the potential to radically evolve wireless systems, facilitating the integration of both communications and radar functionalities into a single device, thus, enabling joint communication and sensing (JCAS). In this paper, we present a novel approach for JCAS that incorporates a reconfigurable intelligent surface (RIS) in the near-field of an FD multiple-input multiple-output (MIMO) node, which is jointly optimized with the digital beamformers to enable JSAC and efficiently handle self-interference (SI). We propose a novel problem formulation for FD MIMO JCAS systems to jointly minimize the total received power at the FD node's radar receiver, while maximizing the sum rate of downlink communications subject to a Cramér-Rao bound (CRB) constraint. In contrast to the typically used CRB in the relevant literature, we derive a novel, more accurate, target estimation bound that fully takes into account the RIS deployment. The considered problem is solved using alternating optimization, which is guaranteed to converge to a local optimum. The simulation results demonstrate that the proposed scheme achieves significant performance improvement both for communications and sensing. It is showcased that, jointly designing the FD MIMO beamformers and the RIS phase configuration to be SI aware can significantly loosen the requirement for additional SI cancellation.**

***Index Terms*—Full duplex, Reconfigurable Intelligent Surface, Joint Communication and Sensing**

I. Introduction

Full-duplex (FD) systems refer to wireless systems in which the same frequency band is used for both transmitting and receiving signals simultaneously, allowing for bi-directional communications [1]. This is in contrast to traditional half-duplex systems, in which a separate frequency band is used for transmitting and receiving signals, and the system can only perform one function at a time [2]. FD systems have the potential to significantly improve the performance and capacity of wireless networks, by allowing for more efficient use of the available spectrum, thus, enabling higher data rates.

Self-interference (SI) is a major challenge for FD systems and SI cancellation (SIC) is viable to make FD a reality [3], [4]. Due to simultaneous transmission and reception, FD is currently receiving significant interest as it has the potential to enable joint communication and sensing (JCAS), implying that next-generation base stations (BSs) and radar functionalities could be integrated into a single device. In [5], a novel JSAC system transceiver design for an FD MIMO base station (BS) equipped with hybrid analog and digital beamformers is presented. Recently, reconfigurable intelligent surfaces (RISs) JCAS has gained significant interest due to its potential to increase communications and sensing performance [6]. In [7], RIS-assisted JCAS under the Cramér-Rao bound (CRB) constraint was derived. However, the derived CRB neglected the RIS contribution in enhancing the sensing performance. In [8], RIS-assisted JCAS to maximize the sum rate under a radar beam-pattern similarity constraint was investigated. In [9], a CRB minimization-based beamforming design for non-line-of-sight (LoS) RIS-assisted JCAS was studied. Note that the designs presented above do not consider the effect of SI on JCAS which may overwhelm the receiver. Moreover, the CRB derived in [7] did not consider any effect of the RIS and, in [9], only the CRB for the non-LoS sensing case was derived.

In this paper, we focus on the optimization of an FD JCAS system comprising one MIMO downlink (DL) user and 1 target to be detected by the FD node's radar receiver. Firstly, we derive the exact CRB for RIS-assisted FD JCAS, by considering both the LoS and the non-LoS contributions. Then, we propose a novel formulation to jointly minimizing the effective SI power received in uplink (UL) at the radar receiver while maximizing the sum rate for the DL user, subject to the derived exact CRB constraint. However, we remark that the optimization of RIS with exact CRB constraint is extremely challenging and would require significant additional space to be elaborated in detail. Due to space limitations, we consider imposing the CRB constraint only on the digital beamformer. The considered joint optimization problem is transformed into its equivalent minimization of the mean squared error (MMSE) problem [10], which results to be composed of two terms: the SI power at the radar and the MSE at the DL user. A novel alternating optimization method is devised to minimize the overall objective function. Our goal is to tackle SI with both the digital beamformer at the MIMO transmitter and the passive beamforming offered by the RIS, hence, enabling close to ideal FD operation for JCAS.

Simulation results show that the proposed scheme achieves significant performance gain compared to the conventional JCAS scheme with no RIS. Moreover, we also show that RIS can be beneficial to improve both the communications performance and the sensing performance, while also assisting in lowering the SI.

The rest of the paper is organized as follows. Section II presents the system model and the exact CRB for FD JCAS. Section III presents a novel joint beamforming design. Finally, Sections IV and V present simulations results and conclusions,

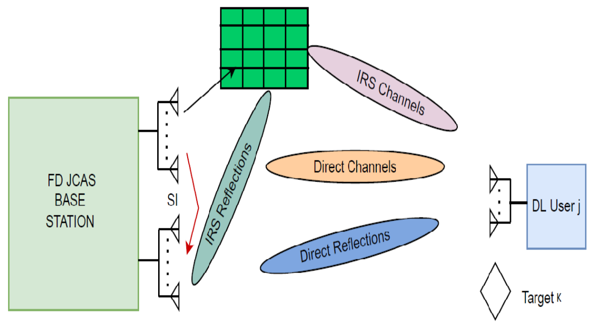

Fig. 1: The proposed FD MIMO JCAS system with RIS.

respectively.[1]

II. System Model

We consider an FD JCAS system consisting of one FD MIMO BS b equipped with M_b and N_b transmit and receive antennas, respectively, which communicates with one MIMO DL user j having N_j antenna elements, as shown in Fig. 1. The DL signal is also used to detect targets/scatterers (via their induced reflections) randomly distributed within the communication/sensing environment at the radar receiver of the FD MIMO node. We assume that the BS is assisted by a RIS placed within its near-field region, whose role is to jointly improve communication and sensing performance while offering strong SIC. Let $\mathbf{V}_j \in \mathbb{C}^{N_j \times M_j}$ denote the digital beamformer for the unit-variance data stream $\mathbf{s}_j \in \mathbb{C}^{d_j \times 1}$ intended for the DL user j. The RIS is assumed to have R rows and C columns of reflection-tunable elements, whose phase response is denoted by the diagonal matrix $\mathbf{\Phi}_i \in \mathbb{C}^{RC \times RC}$, containing $\phi_i \in \mathbb{C}^{RC \times 1}$ on its main diagonal.

Let $\mathbf{H}_{j,b} \in \mathbb{C}^{N_j \times M_b}$ and $\mathbf{H}_{j,i} \in \mathbb{C}^{N_j \times RC}$ denote the channels from the BS and the RIS to the DL user j, respectively. The channel from the RIS to the BS and from the BS to RIS are denoted with $\mathbf{H}_{b,i} \in \mathbb{C}^{N_b \times RC}$ and $\mathbf{H}_{i,b} \in \mathbb{C}^{RC \times M_b}$, respectively. Let $\mathbf{H}_{b,b} \in \mathbb{C}^{N_b \times M_b}$ denote the SI channel for the MIMO FD JCAS node b which can be decomposed as $\mathbf{H}_{b,b} = \mathbf{H}^l_{b,b} + \mathbf{H}^r_{b,b}$, where $\mathbf{H}^l_{b,b}$ and $\mathbf{H}^r_{b,b}$ denote the LoS and non-LoS contributions for the SI channel, respectively. Since the transmit and receive antennas of the FD JCAS are in the near-field, we consider a spherical wavefront and model each element of $\mathbf{H}^l_{b,b}$ as

$$\mathbf{H}^l_{b,b}(m,n) = \frac{\rho}{r_{m,n}} e^{-i2\pi \frac{d_{m,n}}{\lambda}}, \quad \forall m,n, \tag{1}$$

where the scalars λ and ρ denote the wavelength and the power normalization constant to assure $\mathbb{E}(||\mathbf{H}^l_{b,b}||^2_F) = M_b N_b$, respectively, and the scalar $d_{m,n}$ denotes the distance between m-th receive and n-th transmit antenna. Depending on the size of RIS, the channels $\mathbf{H}_{b,i}$ and $\mathbf{H}_{i,b}$ can also be in the near-field. Therefore, we consider modelling them similarly as (1).

[1] *Notations:* Boldface lower and upper case characters denote vectors and matrices, respectively. $\mathbb{E}\{\cdot\}$, $\mathrm{Tr}\{\cdot\}$, and $\mathbf{I}$ denote expectation, trace, and identity matrix, respectively. The superscripts $(\cdot)^T$ and $(\cdot)^H$ denote transpose and conjugate-transpose (Hermitian) operators, respectively.

We assume perfect CSI, which can be obtained by exploiting channel reciprocity via time division duplexing (TDD).

A. Communication Model

Let $\mathbf{y}_j$ denote the received signal at the DL user j, which can be written as

$$\mathbf{y}_j = (\mathbf{H}_{j,b} + \mathbf{H}_{j,i}\mathbf{\Phi}\mathbf{H}_{i,b})\mathbf{V}_j\mathbf{s}_j + \mathbf{n}_j, \tag{2}$$

where $\mathbf{n}_j \sim \mathcal{CN}(\mathbf{0}, \sigma_j^2\mathbf{I})$ denote the noise with variance σ_j^2 at the DL user j. Let $\mathbf{R}_j = \mathbb{E}[\mathbf{y}_j\mathbf{y}_j^H]$ denote the signal-plus noise covariance matrix at the DL user j. Let $\mathbf{R}_{\bar{j}} = \mathbb{E}[\mathbf{n}_j\mathbf{n}_j^H]$ denote its noise covariance matrix. The rate at the DL user j is given by

$$\mathcal{R}_j = \log\big[\det(\mathbf{R}_{\bar{j}}^{-1}\mathbf{R}_j)\big]. \tag{3}$$

B. Radar Model

We consider the radar to estimate one angle of arrival (AoA), denoted as θ_k. Let ω_0 denote the angle between the FD BS and the RIS, which is perfectly known at the BS. Let $\mathbf{y}_b$ denote its total received signal, and let $\mathbf{A}$ denote a matrix, containing contributions from all the paths, defined as

$$\begin{aligned}\mathbf{A} =& \psi_k \mathbf{a}_r(\theta_k)\mathbf{a}_t(\theta_k)^T + \xi_{1,k}\mathbf{a}_r(\omega_0)\mathbf{a}_t(\omega_0)^T\mathbf{\Phi}\mathbf{a}_i(\theta_k)\mathbf{a}_i(\theta_k)^T \\ & \mathbf{\Phi}\mathbf{a}_i(\omega_0)\mathbf{a}_t(\omega_0)^T + \xi_{2,k}\ \mathbf{a}_r(\theta_k)\ \mathbf{a}_i(\theta_k)^T\mathbf{\Phi}\mathbf{a}_i(\omega_0)\mathbf{a}_t(\omega_0)^T \\ & + \psi_{b,i}\mathbf{a}_r(\omega_0)\mathbf{a}_t(\omega_0)^T\mathbf{\Phi}\xi_{3,k}\ \mathbf{a}_i(\theta_k)\ \mathbf{a}_t(\theta_k)^T,\end{aligned} \tag{4}$$

where ψ_k denotes the reflection coefficient for the LoS path between the BS and target k, $\mathbf{n}_b \sim \mathcal{CN}(\mathbf{0}, \sigma_r^2\mathbf{I})$ denotes the noise at the radar with variance σ_r^2, and $\mathbf{a}_r$ and $\mathbf{a}_t$ denote the transmit and receive antenna steering vectors for JCAS node, respectively. The scalars $\xi_{1,k}, \xi_{2,k}$ and $\xi_{3,k}$ denote the reflection coefficients for the signal from RIS to target k and back to RIS, from RIS to the radar via target k and from the transmitter to the RIS via target k, respectively. Given $\mathbf{A}$, we can write the received signal $\mathbf{y}_b$ as

$$\mathbf{y}_b = \mathbf{A}\mathbf{V}_j\mathbf{s}_j + \mathbf{H}^l_{b,b}\mathbf{V}_j\mathbf{s}_j + \mathbf{H}_{b,i}\mathbf{\Phi}\mathbf{H}_{i,b}\mathbf{V}_j\mathbf{s}_j + \mathbf{n}_b, \tag{5}$$

which contains the effective SI. Assuming uniform linear arrays at the FD node, the antenna steering vector $\mathbf{a}_r$ at the receiver, appearing in $\mathbf{A}$, can be modelled as

$$\mathbf{a}_r(\theta_k) = \frac{1}{\sqrt{N_b}}[1, e^{j\frac{2\pi}{\lambda}d sin(\theta_k)}, ..., e^{j\frac{2\pi}{\lambda}d(N_b-1)sin(\theta_k)}]^T, \tag{6}$$

and similar modelling also holds for $\mathbf{a}_t$. The scalars d and λ denote the distance and wavelength, respectively. Let ϕ_k and φ_k denote the elevation and the azimuth angles between the RIS and the target k. The RIS response can be modelled as a uniform planer array (UPA) as

$$\mathbf{a}_i(\phi_k, \varphi_k) = \frac{1}{\sqrt{RC}}[1, e^{j\frac{2\pi}{\lambda}\varpi_1}, ..., e^{j\frac{2\pi}{\lambda}\varpi_{RC-1}}]^T, \tag{7}$$

where $\varpi_i = d_{i_x} sin(\phi_k)cos(\varphi_k) + d_{i_z} sin(\varphi_k)$ [11], $0 < i < RC$, and d_{i_x} and d_{i_z} denote the distance of RIS's element i from its first element on the x and z axis, respectively. As the position and the orientation of RIS are known at the FD

BS, the azimuth and elevation angles can be expressed as a function of θ_k to be estimated as

$$\phi_k = arcos(\frac{l_k cos(\theta_k) - r_1 cos(\omega_0)}{r_2}), \quad (8a)$$

$$\varphi_k = arcos(\frac{l_k cos(\theta_k) - r_1 cos(w_0)}{r_r cos(\theta_k)}), \quad (8b)$$

where l_k is the distance of the target k, ω_0 is the angle between the FD BS and RIS, and the scalars r_1, l_k, r_2, r_r, assuming the RIS to be placed on the (x,z) plane, denote the distance between the FD BS and RIS with the relative angle ω_0, the distance between RIS and target k, the distance between RIS and target k on the (x,y) plane and the distance on the (x,z) plane, respectively.

To achieve an accurate estimation, the CRBs for AoA θ_k should be below the threshold ζ_k, which imposes the constraint

$$\frac{1}{\zeta_k} \leq \frac{1}{\mathrm{CRB}(\theta_k)}, \quad (9)$$

where $\mathrm{CRB}(\theta_k)$ is derived in the Appendix.

C. Problem Formulation

We embark on the task of jointly optimizing the performance of both communication and radar systems. While the concept of communication rate is well-defined for the DL user, there is no such concept that exists for the UL. The sum rate maximization problem can be formulated as a function of MMSE [10]. Conversely, even though the notion of UL rate in the context of FD JCAS is not well-established, the challenge of SI cancellation can be framed as a minimization problem, specifically minimizing the squared Frobenius norm of the SI power, thereby improving the accuracy of radar detection.

For the DL user, we assume that it deploys the combiner $\mathbf{F}_j$ to estimate its data streams $\mathbf{s}_j$ as $\hat{\mathbf{s}}_j = \mathbf{F}_j \mathbf{y}_j$. Assuming that the combiner $\mathbf{F}_j$ is optimized based on the minimization of the MSE criteria, its closed-form solution is

$$\begin{aligned}\mathbf{F}_j =& \mathbf{V}_j^H (\mathbf{H}_{j,b} + \mathbf{H}_{j,i}\mathbf{\Phi}\mathbf{H}_{i,b})^H ((\mathbf{H}_{j,b} + \mathbf{H}_{j,i}\mathbf{\Phi}\mathbf{H}_{i,b})\mathbf{V}_j\mathbf{V}_j^H \\ & (\mathbf{H}_{j,b} + \mathbf{H}_{j,i}\mathbf{\Phi}\mathbf{H}_{i,b}) + \sigma_r^2 \mathbf{I})^{-1}.\end{aligned} \quad (10)$$

Given $\mathbf{F}_j$ as (10), the MSE of the DL user j becomes

$$\mathbf{E}_j = (\mathbf{I} + \mathbf{V}_j^H \mathbf{H}_j^H \mathbf{R}_{\bar{j}}^{-1} \mathbf{H}_j \mathbf{V}_j)^{-1}, \quad (11)$$

where $\mathbf{R}_j = \sigma_j^2 \mathbf{I}$ is the noise covariance matrix with variance σ_j^2. Let $\mathbf{W}_j$ denote the weight computed as [10]

$$\mathbf{W}_j = \frac{w_j}{\ln 2} \mathbf{E}_j^{-1}. \quad (12)$$

The joint minimization problem under the total sum-power, CRB and unit-modulus constraint for the RIS can be stated as

$$\min_{\mathbf{V}_j, \mathbf{\Phi}} \quad \mathrm{Tr}(\mathbf{E}_{SI}) + \mathrm{Tr}(\mathbf{W}_j \mathbf{E}_j) \quad (13a)$$

$$\text{s.t.} \quad \mathrm{Tr}\big(\mathbf{V}_j \mathbf{V}_j^H\big) \leq p_o, \quad (13b)$$

$$|\phi(i)| = 1, \forall i, \quad (13c)$$

$$(9) \quad (13d)$$

where $\mathbf{E}_{SI}$ is a matrix defined as

$$\begin{aligned}\mathbf{E}_{SI} =& \mathbf{H}_{b,b}^l \mathbf{V}_j \mathbf{V}_j^H {\mathbf{H}_{b,b}^l}^H + \mathbf{H}_{b,b}^l \mathbf{V}_j \mathbf{V}_j^H \mathbf{H}_{i,b}^H \mathbf{\Phi}^H \mathbf{H}_{b,i}^H + \mathbf{H}_{b,i}\mathbf{\Phi} \\ & \mathbf{H}_{i,b}\mathbf{V}_j\mathbf{V}_j^H {\mathbf{H}_{b,b}^l}^H + \mathbf{H}_{b,i}\mathbf{\Phi}\mathbf{H}_{i,b}\mathbf{V}_j\mathbf{V}_j^H \mathbf{H}_{i,b}^H \mathbf{\Phi}^H \mathbf{H}_{b,i}^H\end{aligned} \quad (14)$$

obtained by writing the Frobenius norm squared as a function of the trace operator, whose diagonal elements capture the effective SI power and (13b) and (13c) denote the total sum-power constraint p_0 and the unit-modulus constraint on RIS.

III. Novel Algorithm Design for FD JCAS

In this section, we provide a novel algorithm to solve the optimization problem (13) based on alternating optimization. Let $\mathcal{L}$ denote the Lagrangian of (13) and let λ_0 and μ_k denote the Lagrange multipliers for the total sum-power constraint at the FD BS and for the CRB constraint of target k, respectively.

A. Digital Beamformer Optimization

To optimize the digital beamformer $\mathbf{V}_j$, which jointly optimizes the DL rate and handles the SI for JCAS, we take the derivative of $\mathcal{L}$ with respect to the conjugate of $\mathbf{V}_j$, which leads to the following optimal WMMSE digital beamformer

$$\begin{aligned}\mathbf{V}_j =& \Big((\mathbf{H}_{j,b} + \mathbf{H}_{j,i}\mathbf{\Phi}\mathbf{H}_{i,b})^H \mathbf{F}_j^H \mathbf{W}_j \mathbf{F}_j (\mathbf{H}_{j,b} + \mathbf{H}_{j,i}\mathbf{\Phi}\mathbf{H}_{i,b}) \\ & + (\mathbf{H}_{b,b}^l + \mathbf{H}_{b,i}\mathbf{\Phi}\mathbf{H}_{i,b})^H (\mathbf{H}_{b,b}^l + \mathbf{H}_{b,i}\mathbf{\Phi}\mathbf{H}_{i,b}) \\ & + \mu_k 2 \bar{\mathbf{A}}_{\theta_k}^H \mathbf{\Sigma}^{-1} \bar{\mathbf{A}}_{\theta_k} + \lambda_0 \mathbf{I}\Big)^{-1} (\mathbf{H}_{j,b} + \mathbf{H}_{j,i}\mathbf{\Phi}\mathbf{H}_{i,b}) \mathbf{F}_j^H \mathbf{W}_j.\end{aligned} \quad (15)$$

where $\bar{\mathbf{A}}_{\theta_k}^H$ is defined in the Appendix. To find the optimal values of λ_0 and μ_k, a linear search method can be adopted. In this study, we employ the Bisection method.

B. RIS Optimization

We consider optimizing the RIS to jointly achieve quasi-ideal SIC on the UL side (which can be seen as a virtual UL rate optimization problem) and also to maximize the rate at the DL side. For such a purpose, we consider minimizing the Frobenius Norm squared of the effective SI. Let $\mathbf{B}, \mathbf{C}, \mathbf{D}$ be

$$\mathbf{B} = \mathbf{H}_{b,i}^H \mathbf{H}_{b,i} + \mathbf{H}_{j,i}^H \mathbf{F}_j^H \mathbf{W}_j \mathbf{F}_j \mathbf{H}_{j,i}, \quad (16a)$$

$$\mathbf{C} = \mathbf{H}_{i,b}\mathbf{H}_{i,b}^H + \mathbf{H}_{i,b}\mathbf{V}_j\mathbf{V}_j^H \mathbf{H}_{i,b}^H, \quad (16b)$$

$$\mathbf{D} = \mathbf{H}_{i,b} {\mathbf{H}_{b,b}^l}^H \mathbf{H}_{b,i} + \mathbf{H}_{i,b}\mathbf{V}_j\mathbf{V}_j^H \mathbf{H}_{j,b}^H \mathbf{F}_j^H \mathbf{W}_j \mathbf{F}_j \mathbf{H}_{j,i}, \quad (16c)$$

Given the auxiliary matrices, problem (13) can be restated with respect to diagonal elements of RIS ϕ as

$$\min_{\phi} \quad \phi^H \mathbf{\Lambda} \phi + \mathbf{d}^T \phi + \phi^H \mathbf{d}^* \quad (17a)$$

$$(13c) \quad (17b)$$

where $\mathbf{\Lambda} = (\mathbf{B} \odot \mathbf{C})$, and $\mathbf{d}$ is a vector made of the diagonal elements of the matrix $\mathbf{D}$. To render a feasible solution, we adopt the majorization-maximization optimization method [12] by constructing an upper bound denoted as $g(\cdot)$ for the objective function (17), denoted as $f(\cdot)$. For a problem of type

Algorithm 1 Optimization of RIS

Initialize: iteration index $n = 1$, accuracy ϵ.
Evaluate: $f(\phi(0))$.
Repeat until convergence
Calculate $\mathbf{q}^{(n)} = (\lambda_i^{max}\mathbf{I} - \boldsymbol{\Lambda})\phi^{(n)} - \mathbf{d}^*$
Update $\phi^{(n+1)}$ as $\phi^{(n+1)} = e^{i\angle\mathbf{q}_i^{(n)}}$.
if $|f(\phi_i^{(n+1)}) - f(\phi_i^{(n)})|/f(\phi_i^{(n+1)}) \leq \epsilon$
Stop and return $\phi^{(n+1)}$.
else n=n+1 and repeat.

Algorithm 2 JCAS Optimization with SIC

Initialize iteration index n, accuracy ϵ, digital beamformer and combiner.
Repeat until convergence
Update $\mathbf{F}_j$ with (10).
Update $\mathbf{W}_j$ with (12).
Update $\mathbf{V}_j$ with (15).
Search for the Lagrange multipliers.
Update $\boldsymbol{\Phi}_i$ with Algorithm 1.
if convergence condition is satisfied
Stop and return the optimized variables.
else repeat.

(17), it has been shown in [12] that at iteration n the following upper bound can be considered

$$g(\phi|\phi^{(n)}) = 2\mathrm{Re}\{\mathbf{s}^H\mathbf{q}^{(n)}\} + c, \tag{18}$$

where λ^{max} is the maximum eigenvalue $\mathbf{q}^{(n)} = (\lambda^{max}\mathbf{I} - \boldsymbol{\Lambda})\phi^{(n)} - \mathbf{d}^*$. Given the upper bound and $\mathbf{q}^{(n)}$, our problem can be restated as a minimization of the upper bound as,

$$\min_{\phi_r} \quad 2\mathrm{Re}\{\mathbf{d}^H\mathbf{q}^{(n)}\}, \tag{19a}$$

$$\text{s.t.} \quad |\phi(i)| = 1, \quad \forall i, \tag{19b}$$

By solving problem (19), we get the following

$$\phi^{(n+1)} = e^{i\angle\mathbf{q}^{(n)}}. \tag{20}$$

When the digital beamformer is computed under the CRB constraint at each iteration, the RIS optimization should be carried out by solving the aforementioned problem iteratively until convergence. The procedure for optimizing the phase response of RIS is given in Algorithm 1. The overall procedure to optimize and solve the joint optimization problem under the CRB is formally given in Algorithm 2.

The convergence of the proposed scheme is straightforward by combining the reasoning of the well-established WMMSE [10] and the majorization-maximization technique [12]. However, due to space limitations, we omit the extended proof.

IV. Simulation Results

In this section, we present simulation results to validate the advantage of the proposed SI-aware FD JCAS transceiver design.

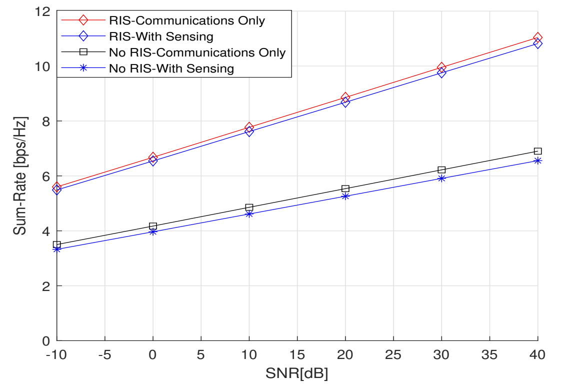

Fig. 2: Communications performance for FD JCAS with SIC.

We consider the FD BS and the DL user to be equipped with uniform linear arrays (ULA) at 80 m from the FD BS, and the FD BS to be placed in the center of the three-dimensional coordinate system with ULA aligned with the z-axis. The RIS is assumed to be placed on the (x,y) plane, with a relative angle of 30° with respect to the FD BS, with its first element being 5 m far from the first transmit antenna of the FD BS. We assume that the FD BS is assumed to be equipped $M_b = 15$ transmit antennas and $N_b = 10$ receive antennas, and the DL user is assumed to be equipped with $N_j = 5$ receiving antennas. The RIS is assumed to be of size 10×10. The channels between the FD BS and the DL user j, denoted with $\mathbf{H}_{j,b}$ and $\mathbf{H}_{b,i}$, are modelled with the line of sight (LoS) channel model. The number of data streams to be transmitted to the DL user is set to be $d_j = 2$. The digital beamformer $\mathbf{V}_j$ is initialized as the dominant eigenvectors of the effective channel covariance matrices, and the response of the NF-IRSs is initialized with random phases. We define the signal-to-noise-ratio (SNR) of our system as $SNR = p_o/\sigma_j^2$, where p_o is the total transmit power and σ_j^2 is the noise variance at the DL user. For the CRB constraint, we set $\zeta_k = 0.01$ and the AoA to be estimated to be randomly distributed on a circle of $50m$ with the angle range limited to $[-\pi/2, -\pi/2]$.

For comparison, we define the following benchmark schemes: 1) *RIS-Communications Only* - A scheme in which the beamformers and RIS are designed to maximize the performance of the communications and there is no sensing and SI (half-duplex (HD) mode), 2) *No RIS-Communications Only* - A scheme similar to scheme 1) but without RIS, and 3) *No RIS- With Sensing*- A scheme in which FD BS performs JCAS but without the aid of RIS. We label our scheme as *RIS-With Sensing*.

Fig. 2 shows the performance of the communications in terms of the user rate as a function of the proposed novel FD JCAS transceiver design, in comparison to the benchmark schemes. We can see that when the BS act as an HD BS and there is no SI and sensing, the beamformer $\mathbf{V}_j$ and the RIS is designed to only maximize the DL performance, leading to a higher rate. However, in the FD JCAS case, the beamformers and RIS are designed to jointly handle SI, enhance the sensing performance and improve the communications.

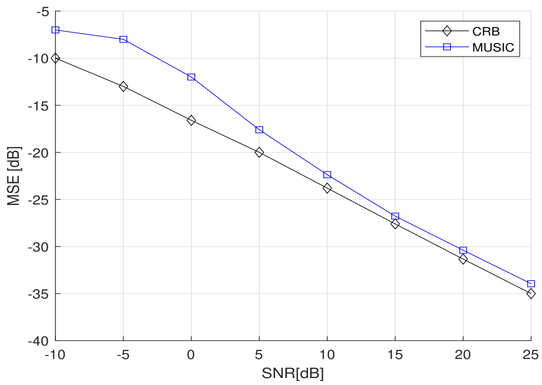

Fig. 3: MSE for the AoA estimation as a function of SNR.

In Fig. (3), we compare the performance of the estimation of AoA θ_k when using a MUSIC-based estimator. Our approach demonstrates effective management of SI, and as the SNR increases, the estimation performance approaches the CRB. While there remains a slight difference between the estimation performance and the CRB, this gap can be further reduced by incorporating the CRB constraint into the optimization of the Reconfigurable Intelligent Surface (RIS), which is a direction for future research.

V. Conclusions

This work introduces a new method to enable FD JCAS by considering the impact of SI. The authors derive the exact CRB for RIS-assisted JCAS and propose a joint optimization framework based on alternating optimization that satisfies the CRB constraint. Simulation results demonstrate that the proposed beamforming method, which accounts for SI, leads to a substantial performance improvement and effectively manages SI. Furthermore, the reduction in data transmission rate compared to a communications-only approach is negligible, which paves the path toward FD JCAS with accurate and energy-efficient SI management with RIS.

Appendix

In this section, we derive the CRB for AoA θ_k, which can be written as [13]

$$\mathrm{CRB}(\theta_k) = \frac{1}{2}\Big(\mathrm{Tr}\big(\mathbf{V}_j^H \bar{\mathbf{A}}_{\theta_k}^H \mathbf{\Sigma}^{-1} \bar{\mathbf{A}}_{\theta_k} \mathbf{V}_j\big)\Big)^{-1}. \tag{21}$$

where $\bar{\mathbf{A}}_{\theta_k} = \partial \mathbf{A}/\partial \theta_k$. We first define the derivatives of the antenna responses as

$$\partial \mathbf{a}_r = \frac{1}{\sqrt{N_b}}[0, ..., j\frac{2\pi}{\lambda} d(N_b-1) cos(\theta_k) e^{j\frac{2\pi}{\lambda} d(N_b-1) sin(\theta_k)}]^T, \tag{22a}$$

$$\partial \mathbf{a}_t = \frac{1}{\sqrt{M_b}}[0, ..., j\frac{2\pi}{\lambda} d(M_b-1) cos(\theta_k) e^{j\frac{2\pi}{\lambda} d(M_b-1) sin(\theta_k)}]^T, \tag{22b}$$

Let $\partial \varpi_i$ denote the derivative of ϖ_i, obtained by expressing ϕ_k and φ_k as a function of θ_k. Then the derivative of $\mathbf{a}_r^{RIS}$ with respect to θ_k can be written as

$$\partial \mathbf{a}_t^{RIS} = \frac{1}{\sqrt{RC}}[0, ..., j\frac{2\pi}{\lambda} \partial \varpi_i e^{j\frac{2\pi}{\lambda} \varpi_{RC-1}}] \tag{23}$$

By considering the complete deployment effect of RIS, including both the LoS and non-LoS links, $\bar{\mathbf{A}}_{\theta_k}$ can be written as

$$\begin{aligned}\bar{\mathbf{A}}_{\theta_k} =& \psi_k \partial \mathbf{a}_r(\theta_k) \mathbf{a}_t(\theta_k)^T + \psi_k \mathbf{a}_r(\theta_k) \partial \mathbf{a}_t(\theta_k)^T \\ &+ \xi_{1,k} \mathbf{a}_r(\omega_0) \mathbf{a}_t(\omega_0)^T \mathbf{\Phi} \partial \mathbf{a}_r^{RIS}(\theta_k) \mathbf{a}_i(\theta_k)^T \mathbf{\Phi} \mathbf{a}_r^{RIS}(\omega_0) \\ &\mathbf{a}_t(\omega_0)^T + \xi_{1,k} \mathbf{a}_r(\omega_0) \mathbf{a}_t(\omega_0)^T \mathbf{\Phi} \mathbf{a}_i(\theta_k) \partial \mathbf{a}_i(\theta_k)^T \mathbf{\Phi} \mathbf{a}_i(\omega_0) \\ &\mathbf{a}_t(\omega_0)^T + \xi_{2,k}\ \partial \mathbf{a}_r(\theta_k) \mathbf{a}_t^{RIS}(\theta_k)^T \mathbf{\Phi} \mathbf{a}_i(\omega_0) \mathbf{a}_t(\omega_0)^T \\ &+ \xi_{2,k}\ \mathbf{a}_r(\theta_k) \partial \mathbf{a}_i(\theta_k)^T \mathbf{\Phi} \mathbf{a}_i(\omega_0) \mathbf{a}_t(\omega_0)^T \\ &+ \xi_{3,k} \mathbf{a}_r(\omega_0) \mathbf{a}_t(\omega_0)^T \mathbf{\Phi} \partial \mathbf{a}_i(\theta_k)\ \mathbf{a}_t(\theta_k)^T \\ &+ \xi_{3,k} \mathbf{a}_r(\omega_0) \mathbf{a}_t(\omega_0)^T \mathbf{\Phi} \mathbf{a}_i(\theta_k) \partial \mathbf{a}_t(\theta_k)^T,\end{aligned} \tag{24}$$

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References

[1] C. K. Sheemar *et al.*, "Practical hybrid beamforming for millimeter wave massive MIMO full duplex with limited dynamic range," *IEEE Open J. Commun. Society*, vol. 3, pp. 127–143, 2022.

[2] G. Liu *et al.*, "In-band full-duplex relaying: A survey, research issues and challenges," *IEEE Commun. Surveys Tuts.*, vol. 17, no. 2, pp. 500–524, 2015.

[3] G. C. Alexandropoulos *et al.*, "Full duplex hybrid A/D beamforming with reduced complexity multi-tap analog cancellation," in *Proc. IEEE SPAWC*, Atlanta, USA, 2020, pp. 1–6.

[4] C. K. Sheemar and D. Slock, "Hybrid beamforming and combining for millimeter wave full duplex massive mimo interference channel," in *2021 IEEE Global Communications Conference (GLOBECOM)*. IEEE, 2021, pp. 1–6.

[5] M. A. Islam *et al.*, "Integrated sensing and communication with millimeter wave full duplex hybrid beamforming," in *Proc. IEEE ICC*, Seoul, South Korea, 2022, pp. 1–6.

[6] C. B. Barneto, E. Rastorgueva-Foi, M. F. Keskin, T. Riihonen, M. Turunen, J. Talvitie, H. Wymeersch, and M. Valkama, "Millimeter-wave mobile sensing and environment mapping: Models, algorithms and validation," *IEEE Transactions on Vehicular Technology*, vol. 71, no. 4, pp. 3900–3916, 2022.

[7] X. Wang *et al.*, "Joint waveform and discrete phase shift design for RIS-assisted integrated sensing and communication system under cramér-rao bound constraint," *IEEE Trans. Veh. Technol.*, vol. 71, no. 1, pp. 1004–1009, 2021.

[8] H. Luo, R. Liu, M. Li, Y. Liu, and Q. Liu, "Joint beamforming design for RIS-assisted integrated sensing and communication systems," *IEEE Trans. Veh. Technol.*, vol. 71, no. 12, pp. 13 393–13 397, 2022.

[9] X. Song *et al.*, "Intelligent reflecting surface enabled sensing: Cramér-rao lower bound optimization," in *Proc. IEEE GLOBECOM*, Rio de Janerio, Brazil, 2022, pp. 413–418.

[10] S. S. Christensen *et al.*, "Weighted sum-rate maximization using weighted MMSE for MIMO-BC beamforming design," *IEEE Transactions on Wireless Communications*, vol. 7, no. 12, pp. 4792–4799, 2008.

[11] Y. Chen *et al.*, "Hybrid spherical-and planar-wave channel modeling and spatial multiplexing analysis for terahertz integrated um-mimo and irs systems," in *Proc. IEEE ICC*, Seoul, South Korea, 2022, pp. 3490–3495.

[12] C. Pan *et al.*, "Multicell mimo communications relying on intelligent reflecting surfaces," *IEEE Trans. Wireless Commun.*, vol. 19, no. 8, pp. 5218–5233, 2020.

[13] Z. Cheng *et al.*, "Transmit sequence design for dual-function radar-communication system with one-bit DACs," *IEEE Trans. Wireless Commun.*, vol. 20, no. 9, pp. 5846–5860, 2021.