FULL-DUPLEX TWO-WAY MIMO RELAYING SYSTEM: TRANSCEIVER DESIGN AND REVIEW OF PHYSICAL LAYER SECURITY

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FULL-DUPLEX TWO-WAY MIMO RELAYING SYSTEM: TRANSCEIVER DESIGN AND REVIEW OF PHYSICAL LAYER SECURITY

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ABSTRACT

In this paper, we consider the design of optimal transceiver and relay processing algorithms for a full-duplex (FD) two-way amplify-and-forward (AF) multiple-input multiple-output (MIMO) relaying system. We assume the channel state information of loopback self-interference (SI) channels to be imperfect. The nodes employ precoders and receive filters for suppressing the residual SI. The optimal precoders at transceivers are designed by equalising the signal-to-interference-plus-noise ratio at the relay and transceiver. Using the transceiver precoders thus designed, we then design the optimal relay precoder and transceiver receive filters by minimizing the sum of mean square error at the transceivers. The optimal precoders and receive filters are continuously updated to account for the cumulative interference effect caused by AF operation of the FD relay. The secrecy performance of this system in the presence of a passive eavesdropper is analyzed by considering the relay signal to be artificial noise for the eavesdropper. The effectiveness of the proposed design is demonstrated in our simulation results.

Index Terms— In-band full-duplex, imperfect SI cancellation, two-way relaying, amplify-and-forward, physical layer security.

1. INTRODUCTION

Due to the recent advancements in self-interference (SI) cancellation techniques [1],[2] for multiple-input multiple-output (MIMO) full-duplex systems, in-band full-duplex (FD) wireless communication is regarded as one of the emerging technologies for the next generation wireless networks. Relays have been traditionally used in cellular networks to provide improved link capacity and enhanced coverage for cell edge users. Recent research in FD relays involves use cases in physical layer security [3], wherein the relay transmits jamming signals towards the eavesdropper to improve secrecy.

Two-way in-band FD relaying [4] combines the benefits of both these technologies and has the potential to be a driving technology for 5G. Owing to their broadcast nature, wireless networks are always prone to eavesdropping. The aim of physical layer security is to prevent the eavesdropper from decoding the data of the legitimate users. In [5], a two-way FD amplify-and-forward (AF) relaying system is considered with MIMO relay, legitimate transceivers and a passive eavesdropper with single antenna. The relay does two separate tasks, generating beamforming matrix for legitimate channel and generating artificial noise (AN) for the eavesdropper. Here, secrecy is achieved at the cost of reduction in data rate.

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To the best of authors' knowledge, there is no previous work which considers the problem of residual SI mitigation simultaneously at the transceivers and at the relay. The work in [6] proposed a feedback term to mitigate the residual SI in a FD two-way MIMO relaying system. However, their proposed computation of this continuously evolving feedback term was computationally complex and does not scale well with time. A subpar alternative to compute the feedback term was also proposed in [6]. The FD operation of a two-way AF relaying system makes it inherently secure due to the presence of multiple signals combined on the same frequency. This makes it difficult for the eavesdropper to extract a signal from a particular user. However, this inherent secrecy has not been analyzed yet in literature. The main contributions of this paper are threefold:
(*i*) precoding at end-user transceivers for mitigating residual SI,
(*ii*) computing the feedback term, required for residual SI mitigation at the AF relay, in a recursive manner with minimal memory requirements,
(*iii*) analysis of physical layer security of the proposed FD system in terms of the achievable sum-secrecy rate.

$Notations$: A scalar, vector, and matrix are denoted by italic lowercase, boldface lowercase, and boldface uppercase character, respectively. For a matrix $\mathbf{X}$, its transpose, conjugate, conjugate transpose, inverse, determinant, vectorization operation, trace, and Frobenius norm are denoted by $\mathbf{X}^T, \mathbf{X}^c, \mathbf{X}^H, \mathbf{X}^{-1}, |\mathbf{X}|, \mathrm{vec}(\mathbf{X}), \mathrm{tr}(\mathbf{X})$, $\|\mathbf{X}\|_{\mathbb{F}}$, respectively. $\mathbf{X} \otimes \mathbf{Y}$ represents the kronecker product of matrices $\mathbf{X}$ and $\mathbf{Y}$. $\mathbb{E}\{\cdot\}$ is the expectation operator, $\|\mathbf{x}\|$ represents 2-norm of $\mathbf{x}$, $\mathrm{mat}(\cdot)$ performs the inverse operation of $\mathrm{vec}(\cdot)$. If a variable $i = 1$, then $\underline{i} = 2$ and vice versa. Time-slot dependent variables $a, \mathbf{a}, \mathbf{A}$ denote scalar, vector, and matrix, respectively, in the current time slot t, while $a^{(k)}, \mathbf{a}^{(k)}, \mathbf{A}^{(k)}$ denote a scalar, vector, matrix,respectively, in any time slot k.

2. SYSTEM MODEL

As shown in Fig. 1, the system comprises two FD transceivers S_1 and S_2, each equipped with N_s antennas, an eavesdropper E with N_e antennas, and a FD AF relay R with N_r antennas, of which N_s are used as receivers while all N_r antennas are used as transmitters, such that $N_\mathrm{r} \geq N_\mathrm{s}$. Node E is a passive eavesdropper. We assume that there is no direct link between the nodes S_1 and S_2. The matrices $\mathbf{H}_m \in \mathbb{C}^{N_\mathrm{s}\times N_\mathrm{s}}, \mathbf{G}_m \in \mathbb{C}^{N_\mathrm{s}\times N_\mathrm{r}}, \mathbf{K}_m \in \mathbb{C}^{N_\mathrm{e}\times N_\mathrm{s}}, \mathbf{K}_r \in \mathbb{C}^{N_\mathrm{e}\times N_\mathrm{r}}$, $m \in \{1,2\}$, represent the MIMO channel gains as shown in Fig. 1. These matrices are assumed to be perfectly known. The matrices $\mathbf{H}_{\mathrm{rr}} \in \mathbb{C}^{N_\mathrm{s}\times N_\mathrm{r}}, \mathbf{H}_{mm} \in \mathbb{C}^{N_\mathrm{s}\times N_\mathrm{s}}, m \in \{1,2\}$, represent the loopback SI MIMO channel gains. We assume that the available channel state information (CSI) of the loopback channels is imperfect such that

$$\mathbf{H}_{\mathrm{jj}} = \widehat{\mathbf{H}}_{\mathrm{jj}} + \mathbf{\Phi}_{\mathrm{j}}, \quad j = 1, 2, \mathrm{r}, \tag{1}$$

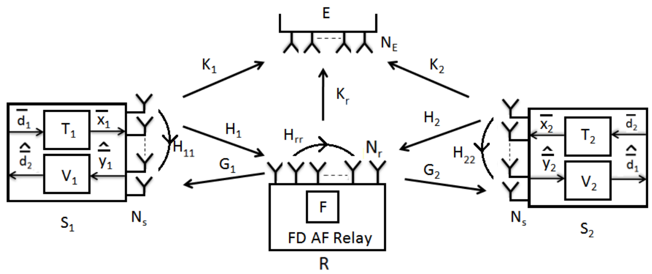


Fig. 1: System Diagram

where $\widehat{\mathbf{H}}_{\mathrm{jj}}$ is the available estimate and $\mathbf{\Phi}_{\mathrm{j}}$ is the CSI error with zero mean and covariance $\mathbb{E}\{\mathbf{\Phi}_j \mathbf{\Phi}_j^H\} = N_{\mathrm{s}}\sigma_{\mathrm{ej}}^2 \mathbf{I}_{N_{\mathrm{s}}}$. All the above channel links are modeled as independent and frequency-flat Rayleigh fading channels assumed to be static for the duration of a time slot. The vector $\mathbf{n}_k, k \in \{1, 2, \mathrm{r}, \mathrm{e}\}$ represents a circularly symmetric complex Gaussian random noise vector with zero mean and covariance $\mathbb{E}\{\mathbf{n}_m \mathbf{n}_m^H\} = \sigma_{nm}^2 \mathbf{I}_{N_{\mathrm{s}}}, k \in \{1, 2, \mathrm{r}\}$, $\mathbb{E}\{\mathbf{n}_{\mathrm{e}} \mathbf{n}_{\mathrm{e}}^H\} = \sigma_{n\mathrm{e}}^2 \mathbf{I}_{N_{\mathrm{e}}}$. The symbol α_i represents path loss for $\mathrm{S_i} - \mathrm{R}$ link. $i \in \{1, 2\}$, while $\alpha_3, \alpha_4, \alpha_5$ represent path loss for $\mathrm{S_1 - E}, \mathrm{R - E}, \mathrm{S_2 - E}$ links, respectively. We assume unity pathloss for loopback channels.

The following events occur during each time slot t :

(*i*) Each transceiver precodes data vector $\mathbf{d}_m \in \mathbb{C}^{N_s \times 1}$ of covariance $\mathbb{E}\{\mathbf{d}_m \mathbf{d}_m^H\} = \mathbf{I}_{N_{\mathrm{s}}}$ with precoding matrix $\mathbf{T}_m \in \mathbb{C}^{N_{\mathrm{s}} \times N_{\mathrm{s}}}$ to generate $\mathbf{x}_m, m \in \{1, 2\}$, and transmits it to the relay node. The signal at the relay after imperfect SI cancellation is given by

$$\widetilde{\mathbf{y}}_r = \sqrt{\alpha_1}\mathbf{H}_1\mathbf{T}_1\mathbf{d}_1 + \sqrt{\alpha_2}\mathbf{H}_2\mathbf{T}_2\mathbf{d}_2 + \mathbf{\Phi}_{\mathrm{r}}\mathbf{x}_{\mathrm{r}} + \mathbf{n}_{\mathrm{r}}. \quad (2)$$

Since the available loopback CSI is erroneous as modeled in (1), perfect SI cancellation is not possible and the consequent residual SI is given by the third term in (2).

(*ii*) The relay precodes the signal received in the previous time slot, $\mathbf{y}_{\mathrm{r}}^{(t-1)}$, using matrix $\mathbf{F} \in \mathbb{C}^{N_{\mathrm{r}} \times N_{\mathrm{s}}}$ and transmits signal

$$\mathbf{x}_{\mathrm{r}} = \mathbf{F}\widetilde{\mathbf{y}}_{\mathrm{r}}^{(t-1)}, \quad t \geq 2. \quad (3)$$

(*iii*) Each transceiver receives relay's transmit signal. After canceling out its own signal and also imperfect SI cancellation, we get

$$\begin{aligned}\widetilde{\mathbf{y}}_m =& \sqrt{\alpha_m}\mathbf{G}_m\mathbf{F}[\sqrt{\alpha_{\underline{m}}}\mathbf{H}_{\underline{m}}^{(t-1)}\mathbf{T}_{\underline{m}}^{(t-1)}\mathbf{d}_{\underline{m}}^{(t-1)} + \mathbf{\Phi}_{\mathrm{r}}^{(t-1)}\mathbf{x}_{\mathrm{r}}^{(t-1)} \\ &+ \mathbf{n}_{\mathrm{r}}^{(t-1)}] + \mathbf{\Phi}_m\mathbf{x}_m + \mathbf{n}_m. \end{aligned} \quad (4)$$

The residual SI is given by the terms containing $\mathbf{\Phi}_r$ and $\mathbf{\Phi}_m$. Each transceiver then applies a receive filter $\mathbf{V}_m \in \mathbb{C}^{N_{\mathrm{s}} \times N_{\mathrm{s}}}$ to $\widetilde{\mathbf{y}}_m$ to obtain an estimate of the data, $\widehat{\mathbf{d}}_{\underline{m}}$, transmitted by the other transceiver.

The overall end-to-end communication requires two time slots.

3. DESIGN OF OPTIMAL PRECODERS AND RECEIVE FILTERS

We address the design of transceiver precoders $\mathbf{T}_1, \mathbf{T}_2$ at current time slot t, followed by the design of the relay precoder $\mathbf{F}$ and the transceiver receive filters $\mathbf{V}_1$ and $\mathbf{V}_2$ at the next time slot.

3.1. Transceiver Precoder Design

The precoder matrices $\mathbf{T}_1, \mathbf{T}_2$ are designed to essentially nullify the effect of transceiver-relay channel apart from balancing the signal-to-interference-plus-noise ratio (SINR) at the relay and transceiver nodes. Moreover, the transceivers operate under a transmit power constraint $P_m \leq P_m^{max}$. We begin with the QR-decomposition of hermitian of the channel matrices given by $\mathbf{H}_m^H = \mathbf{Q}_m\mathbf{R}_m$, $m = 1, 2$. Thus, $\mathbf{T}_m$ is given by

$$\mathbf{T}_m = \beta_m\mathbf{Q}_m\mathbf{W}_m, \quad m = 1, 2, \quad (5)$$

and the resultant transmit power is given by

$$P_m = \mathbb{E}\{\|\mathbf{T}_m\mathbf{d}_m\|^2\} = \beta_m^2 tr(\mathbf{W}_m\mathbf{W}_m^H), \; m = 1, 2, \quad (6)$$

where the scaling factor β_m will be derived in the sequel and $\mathbf{W}_m$ is derived from $\mathbf{R}_m$ such that

$$\mathbf{H}_m\mathbf{T}_m = \beta_m\mathbf{I}_{N_{\mathrm{s}}}. \quad (7)$$

Apart from its regular benefits [7], channel inversion also simplies the design of the precoder in the following way. From (5), it's clear that for a given $\mathbf{H_m}$, Since $SINR_{\mathrm{R}}$ is an increasing function of β_m^2 whereas $SINR_m$ is a decreasing function of β_m^2, the optimal value of β_m is obtained by solving:

$$SINR_{\mathrm{R}} = SINR_m.$$

Using (7), (2) and (4), we equate them as:

$$\frac{\alpha_m\beta_m^2 + \alpha_{\underline{m}}\beta_{\underline{m}}^2}{\sigma_{nr}^2 + P_{\mathrm{r}}\sigma_{er}^2} = \frac{\alpha_m\alpha_{\underline{m}}\beta_{\underline{m}}^{(t-1)^2} tr[\mathbf{V}_m^H\mathbf{G}_m\mathbf{F}\mathbf{F}^H\mathbf{G}_m^H\mathbf{V}_m]}{X_m}. \quad (8)$$

where $P_{\mathrm{r}} = \mathbb{E}\{\|\mathbf{x}_r\|^2\}$,

$$\begin{aligned} X_m =& tr[\mathbf{V}_m^H\{\alpha_m\mathbf{G}_m\mathbf{F}(\mathbf{\Gamma}_c + \sigma_{nr}^2\mathbf{I}_{N_{\mathrm{s}}})\mathbf{F}^H\mathbf{G}_m^H\}\mathbf{V}_m] + \\ & \sigma_{nm}^2 tr(\mathbf{V}_m\mathbf{V}_m^H) + \beta_m^2\sigma_{em}^2 tr[\mathbf{V}_m\mathbf{V}_m^H] tr[\mathbf{W}_m\mathbf{W}_m^H]. \\ \text{Let } u =& \beta_m^2, \quad z_1 = \alpha_m^{-1}\alpha_{\underline{m}}\beta_{\underline{m}}^2, \quad z_2 = \alpha_m^{-1}(\sigma_{nr}^2 + P_{\mathrm{r}}\sigma_{er}^2), \\ z_3 =& \alpha_m\alpha_{\underline{m}}\beta_{\underline{m}}^{(t-1)^2} tr[\mathbf{V}_m^H\mathbf{G}_m\mathbf{F}\mathbf{F}^H\mathbf{G}_m^H\mathbf{V}_m], \\ z_4 =& tr[\mathbf{V}_m^H\{\sigma_{nm}^2 + \alpha_m\mathbf{G}_m\mathbf{F}(\mathbf{\Gamma}_c + \sigma_{nr}^2\mathbf{I}_{N_{\mathrm{s}}})\mathbf{F}^H\mathbf{G}_m^H\}\mathbf{V}_m], \\ z_5 =& \sigma_{em}^2 tr[\mathbf{V}_m\mathbf{V}_m^H] tr[\mathbf{W}_m\mathbf{W}_m^H]. \end{aligned}$$

So (8) becomes: $\dfrac{\alpha_m u + z_1}{z_2} = \dfrac{z_3}{z_4 + u \times z_5}$,

$$\Rightarrow u = \frac{-(z_4 + z_1 z_5) \pm \sqrt{(z_4 + z_1 z_5)^2 + 4z_5(z_1 z_4 - z_2 z_3)}}{2z_5}$$

$\Rightarrow \beta_m = |\sqrt{u}|$, where β_m is the smallest positive solution. The optimal value of β_m is chosen as $\beta_m^* = \min(\beta_m, \beta_m^{max})$, where β_m^{max} is obtained by putting $P_m = P_m^{max}$ in (6).

3.2. Relay Precoder and Transceiver Receive Filter Design

The optimal design is obtained by solving the following sum of mean square error (SMSE) optimization problem:

$$\begin{aligned} \min_{\rho, \bar{\mathbf{F}}, \mathbf{V}_1, \mathbf{V}_2} \quad & f(\rho, \bar{\mathbf{F}}, \mathbf{V}_1, \mathbf{V}_2) \\ \text{subject to} \quad & \mathbb{E}\{\|\mathbf{x}_{\mathrm{r}}\|^2\} \leq P_{\mathrm{r}}^{max}. \end{aligned} \quad (9)$$

The relay precoder matrix $\mathbf{F}$ is decomposed as $\mathbf{F} = \rho\bar{\mathbf{F}}$, where ρ is a positive scaling factor and $\|\bar{\mathbf{F}}\|_{\mathbb{F}} = 1$. This decomposition simplifies the derivation of $\mathbf{F}$ in closed form. For the same reason, the term ρ^{-1} is introduced in the SMSE equation. Due to space constraints, we'll denote $(\sigma_{nr}^2 + \alpha_1\beta_1^{(j)^2} + \alpha_2\beta_2^{(j)^2})\mathbf{I}_{N_{\mathrm{S}}}$ by $\mathbf{\Delta}^{(j)}$ henceforth.

The SMSE of the two transceivers is given by

$$\begin{aligned} f(\rho, \bar{\mathbf{F}}, \mathbf{V}_1, \mathbf{V}_2) =& \sum_{i=1}^{2} \mathbb{E}\{\|\mathbf{d}_i^{(t-1)} - \rho^{-1}\hat{\mathbf{d}}_i\|^2\} \\ =& \sum_{i=1}^{2} \mathbb{E}\{\|\mathbf{d}_i^{(t-1)}\|^2\} - \rho^{-1} tr(\mathbb{E}\{\mathbf{V}_{\underline{i}}^H\widetilde{\mathbf{y}}_{\underline{i}}\mathbf{d}_i^{(t-1)^H}\}) + \\ & \rho^{-1} tr(\mathbb{E}\{\mathbf{d}_i^{(t-1)}\widetilde{\mathbf{y}}_{\underline{i}}^H\mathbf{V}_{\underline{i}}\}) + \rho^{-2}\mathbb{E}\{\|\mathbf{V}_{\underline{i}}^H\widetilde{\mathbf{y}}_{\underline{i}}\|^2\}. \end{aligned} \quad (10)$$

We now express each term of (10) in terms of the relevant optimization variables. Thus, we have

$$\mathbb{E}\{\|\mathbf{d}_i\|^2\} = N_{\mathrm{s}},\ \mathbb{E}\{\mathbf{d}_{\mathbf{i}}\widetilde{\mathbf{y}}_{\underline{i}}^H\mathbf{V}_{\underline{i}}\} = \sqrt{\alpha_i\alpha_{\underline{i}}}\beta_i tr(\mathbf{F}^H\mathbf{G}_{\underline{i}}^H\mathbf{V}_{\underline{i}}). \quad (11)$$

Further, using (4) and (7), we have

$$\begin{aligned}\mathbb{E}\{\|\mathbf{V}_i^H\widetilde{\mathbf{y}}_i\|^2\} =&\alpha_i tr[\mathbf{V}_i^H\mathbf{G}_i\mathbf{F}(\mathbb{E}\{\mathbf{\Gamma}_0\} + \alpha_{\underline{i}}\beta_{\underline{i}}^{(t-1)^2}\mathbf{I}_{N_{\mathrm{s}}}+\\ &\sigma_{nr}^2\mathbf{I}_{N_{\mathrm{S}}})\mathbf{F}^H\mathbf{G}_i^H\mathbf{V}_i] + \sigma_{ni}^2 tr(\mathbf{V}_i\mathbf{V}_i^H)+\\ &\mathbb{E}\{tr(\mathbf{V}_i^H\mathbf{\Phi}_i\mathbf{T}_i\mathbf{d}_i\mathbf{d}_i^H\mathbf{T}_i^H\mathbf{\Phi}_i^H\mathbf{V}_i)\}, \end{aligned}\quad (12)$$

where,

$$\mathbf{\Gamma}_0 = \sum_{k=0}^{t-2}\prod_{l=1}^{t-k-1}\mathbf{\Phi}_{\mathrm{r}}^{(t-l)}\mathbf{F}^{(t-l)}\mathbf{\Delta}^{(k)}\prod_{l=1}^{t-k-1}\mathbf{F}^{(k+l)^H}\mathbf{\Phi}_{\mathrm{r}}^{(k+l)^H},$$

for time slots $t \geq 3$ and $\mathbf{\Gamma}_0 = \mathbf{0}_{N_s}$ for $t < 3$. Throughout the paper, it is assumed that $\prod_{i=a}^{b}(\cdot) = 1$, if $b < a$. Using Lemma 1 from [8] and proceeding as in Theorem 1 from [6], we obtain $\mathbb{E}\{\mathbf{\Gamma}_0\}$ as:

$$\begin{aligned}\mathbf{\Gamma}_p = \mathbb{E}\{\mathbf{\Gamma}_0\} =&\sum_{k=1}^{t-1}(\sigma_{er}^2)^k tr(\mathbf{F}^{(t-k)}\mathbf{\Delta}^{(t-k-1)}\mathbf{F}^{(t-k)^H})\mathbf{I}_{N_{\mathrm{S}}}\times\\ &\prod_{l=1}^{k-1}tr(\mathbf{F}^{(t-l)}\mathbf{F}^{(t-l)^H}).\end{aligned}\quad (13)$$

The feedback term $\mathbf{\Gamma}_p$ represents the contribution of all the previous relay precoders which is required to suppress the residual SI caused by FD operation. After a few algebraic manipulations on (13), we can show that $\mathbf{\Gamma}_p$ can be recursively computed as

$$\mathbf{\Gamma}_p = \sigma_{er}^2[tr(\mathbf{F}^{(q)}\mathbf{\Delta}^{(t-2)}\mathbf{F}^{(q)^H})\mathbf{I}_{\mathbf{N_s}}+\mathbf{\Gamma}_p^{(q)}tr(\mathbf{F}^{(q)}\mathbf{F}^{(q)^H})]. \quad (14)$$

where $q = t-1$. This recursive structure of $\mathbf{\Gamma}_p$ assures that the nodes need not store all the previous relay precoders and transceiver receive filter matrices. This greatly reduces the memory requirement at the relay as well as the computation time for $\mathbf{\Gamma}_p$. Moreover, the filters designed using (14) will lead to better performance than that of the system proposed in [6] where only the n latest time slots are used for computing $\mathbf{\Gamma}_p$.

Using Lemma 1 from [8] and equation (6), we can rewrite

$$\mathbb{E}\{tr(\mathbf{V}_i^H\mathbf{\Phi}_i\mathbf{T}_i\mathbf{d}_i\mathbf{d}_i^H\mathbf{T}_i^H\mathbf{\Phi}_i^H\mathbf{V}_i)\} = \sigma_{ei}^2 P_i tr(\mathbf{V}_i\mathbf{V}_i^H). \quad (15)$$

Using (2), (3), (5), (13), we express the relay transmit power as

$$\mathbb{E}\{\|\mathbf{x}_{\mathrm{r}}\|^2\} = tr[\mathbf{F}(\mathbf{\Gamma}_p + \mathbf{\Delta}^{(t-1)})\mathbf{F}^H]. \quad (16)$$

Now, the Lagrangian for the problem in (9) is given by

$$\mathcal{L} = f(\rho, \bar{\mathbf{F}}, \mathbf{V}_1, \mathbf{V}_2) + \lambda(\mathbb{E}\{\|\mathbf{x}_{\mathrm{r}}\|^2\} - P_{\mathrm{r}}^{max} + b^2), \quad (17)$$

where λ is the lagrangian variable and b is the slack variable. On substituting the results of (10), (11), (12), (16), (13),(15) in (17) and putting $\mathbf{F} = \rho\bar{\mathbf{F}}$, we get

$$\begin{aligned}&\mathcal{L}(\rho, \bar{\mathbf{F}}, \mathbf{V}_1, \mathbf{V}_2, \lambda) = \lambda[\rho^2 tr(\bar{\mathbf{F}}\mathbf{\Gamma}_{\mathrm{r}}\bar{\mathbf{F}}^H) - P_{\mathrm{r}}^{max} + b^2]+\\ &2N_{\mathrm{s}} + \sum_{i=1}^{2}\{-\sqrt{\alpha_i\alpha_{\underline{i}}}\beta i^{(t-1)}tr(\mathbf{V}_{\underline{i}}^H\mathbf{G}_{\underline{i}}\bar{\mathbf{F}} + \bar{\mathbf{F}}^H\mathbf{G}_{\underline{i}}^H\mathbf{V}_{\underline{i}}) + \rho^{-2}\sigma_{n\underline{i}}^2\\ &\rho^{-2}\sigma_{e\underline{i}}^2 P_{\underline{i}} tr(\mathbf{V}_{\underline{i}}\mathbf{V}_{\underline{i}}^H + \alpha_{\underline{i}} tr(\mathbf{V}_{\underline{i}}^H\mathbf{G}_{\underline{i}}\bar{\mathbf{F}}\mathbf{\Gamma}_{\underline{i}}\bar{\mathbf{F}}^H\mathbf{G}_{\underline{i}}^H\mathbf{V}_{\underline{i}}\},\end{aligned}\quad (18)$$

where $\mathbf{\Gamma}_{\underline{i}} = \mathbf{\Gamma}_p + (\alpha_{\underline{i}}\beta_{\underline{i}}^{(t-1)^2} + \sigma_{nr}^2)\mathbf{I}_{N_s}$, $\quad \mathbf{\Gamma}_{\mathrm{r}} = (\mathbf{\Gamma}_p + \mathbf{\Delta}^{(t-1)})$. The optimization problem in (18) can be solved using the Karush-Kuhn-Tucker (KKT) conditions [9]. Since the SMSE function is not jointly convex in the optimization variables, we use the coordinate descent method, wherein the minimization is performed with respect to one variable while keeping the other variables fixed. First, keeping $\mathbf{V}_i$ fixed, we apply the KKT conditions $\frac{\partial\mathcal{L}}{\partial\rho} = 0$, $\frac{\partial\mathcal{L}}{\partial\bar{\mathbf{F}}^c} = \mathbf{0}_{N_r\times N_s}$, $\frac{\partial\mathcal{L}}{\partial\lambda} = 0$, $\frac{\partial\mathcal{L}}{\partial b} = 0$ to respectively get:

$$\lambda\rho tr(\bar{\mathbf{F}}\mathbf{\Gamma}_{\mathrm{r}}\bar{\mathbf{F}}^H) = \rho^{-3}(w_1 + w_2), \quad (19)$$

$$\begin{aligned}&\alpha_1\mathbf{G}_1^H\mathbf{V}_1\mathbf{V}_1^H\mathbf{G}_1\bar{\mathbf{F}}\mathbf{\Gamma}_1 + \alpha_2\mathbf{G}_2^H\mathbf{V}_2\mathbf{V}_2^H\mathbf{G}_2\bar{\mathbf{F}}\mathbf{\Gamma}_2 + \lambda\rho^2\bar{\mathbf{F}}\mathbf{\Gamma}_{\mathrm{r}}\\ &= \sqrt{\alpha_1\alpha_2}(\beta_1^{(t-1)}\mathbf{G}_2^H\mathbf{V}_2 + \beta_2^{(t-1)}\mathbf{G}_1^H\mathbf{V}_1),\end{aligned}\quad (20)$$

$$tr(\bar{\mathbf{F}}\mathbf{\Gamma}_{\mathrm{r}}\bar{\mathbf{F}}^H) = \rho^{-2}(P_{\mathrm{r}}^{max} - b^2), \quad (21)$$

$$\lambda[\rho^2 tr(\bar{\mathbf{F}}\mathbf{\Gamma}_{\mathrm{r}}\bar{\mathbf{F}}^H) - P_{\mathrm{r}}^{max}] = 0, \quad (22)$$

where, $\quad w_i = (\sigma_{ni}^2 + \sigma_{ei}^2 P_i)tr(\mathbf{V}_i\mathbf{V}_i^H), \quad i = 1, 2.$

Now, from (19) and (22), we observe that at optimal point, $\lambda \neq 0$ and so the constraint on $\mathbb{E}\{\|\mathbf{x}_{\mathrm{r}}\|^2\}$ becomes equality constraint, i.e., the relay $\mathbf{R}$ transmits at full power P_{r}^{max}. So, (21) becomes:

$$tr(\bar{\mathbf{F}}\mathbf{\Gamma}_{\mathrm{r}}\bar{\mathbf{F}}^H) = \rho^{-2}(P_{\mathrm{r}}^{max}). \quad (23)$$

From (19) and (21), we get: $\lambda\rho^2 = \frac{(w_1+w_2)}{P_{\mathrm{r}}^{max}}$. Substituting this value in (20) and following Theorem 2 of [6], we get:

$$\bar{\mathbf{F}}^* = \mathrm{mat}(\mathbf{M}^{-1}\sqrt{\alpha_1\alpha_2}\mathrm{vec}[\beta_1^{(t-1)}\mathbf{G}_2^H\mathbf{V}_2 + \beta_2^{(t-1)}\mathbf{G}_1^H\mathbf{V}_1]),$$

where

$$\mathbf{M} = (\mathbf{\Gamma}_{\mathrm{r}}^T \otimes \frac{(w_1+w_2)\mathbf{I}_{N_r}}{P_{\mathrm{r}}^{max}}) + \sum_{m=1}^{2}(\mathbf{\Gamma}_m^T \otimes \alpha_m\mathbf{G}_m^H\mathbf{V}_m\mathbf{V}_m^H\mathbf{G}_m).$$

The optimal ρ^* is obtained from (23). So, the optimal transmit precoder for B is given by $\mathbf{F}^* = \rho^*\bar{\mathbf{F}}^*$. Using the optimal $\mathbf{F}^*$, we compute optimal receive filters by applying the KKT condition: $\frac{\partial\mathcal{L}}{\partial\mathbf{V}_i^c} = 0_{N_s}$, for $i \in \{1, 2\}$, which gives the optimal receive filter:

$$\mathbf{V}_i^* = [w_i\mathbf{I}_{N_s} + \alpha_i\mathbf{G}_i\mathbf{F}^*\mathbf{\Gamma}_i\mathbf{F}^{*^H}\mathbf{G}_i^H]^{-1}\rho^*\sqrt{\alpha_i\alpha_{\underline{i}}}\beta_{\underline{i}}^{(t-1)}\mathbf{G}_i\mathbf{F}^*.$$

The matrices $\mathbf{F}^*, \mathbf{V}^*$ are obtained by iterative computation until a stable value of SMSE is reached. The initial value of $\mathbf{V}$ for computing $\mathbf{F}$ can be any random $N_{\mathrm{s}} \times N_{\mathrm{s}}$ complex matrix.

4. ANALYSIS OF THE SECRECY PERFORMANCE OF THE PROPOSED DESIGN

In this section, we review the secrecy performance of our proposed system. We consider the worst case scenario where E is able to decode the signals from transceivers using blind channel estimation techniques. Even in this case, E will not have CSI of relay-transceiver channels $\mathbf{G}_m$ and hence, it won't be able to decode the signal received from R, which is precoded with $\mathbf{F}$. Relay precoder $\mathbf{F}$ is a function of relay-transceiver channels $\mathbf{G}_m$. So, even in the worst case scenario, the relay signal acts as AN for E. Also, as seen in the previous section, the relay always transmits at full power P_r^{max}. This makes the relay signal a strong interferer for E. The signal received by E at time t is given by

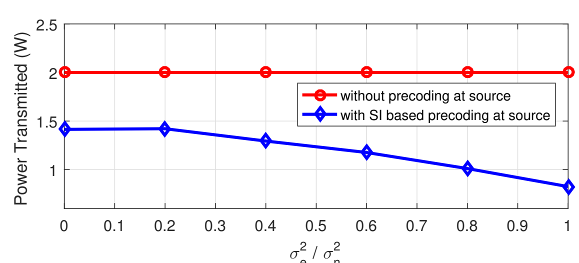
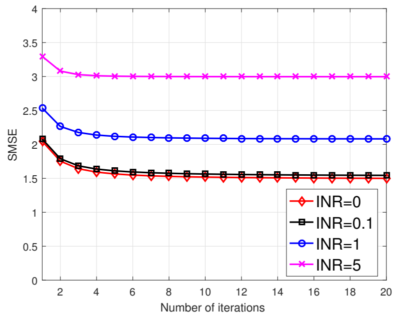
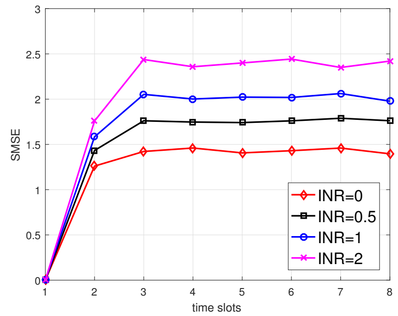


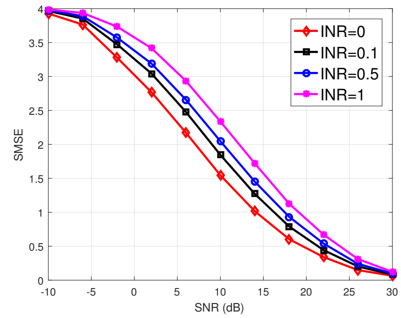
Fig. 2: Transmit power of transceiver S_1, in the 8^{th} time slot, versus SNR, with $P_1^{max} = 2$.



(a) Convergence of SMSE at each time slot.



(b) Evolution of SMSE over time at SNR=10dB.



(c) SMSE performance at different SNR, INR.

Fig. 3

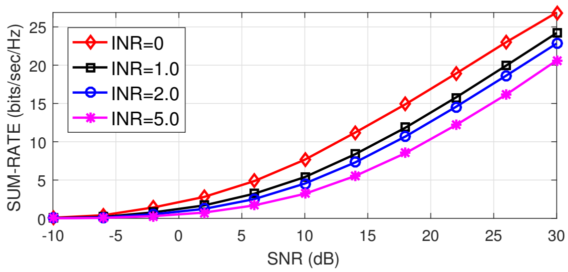


Fig. 4: Sum secrecy rate, in the 8^{th} time slot, versus SNR, for different values of INR

$$\mathbf{y}_e = \sqrt{\alpha_3}\mathbf{K}_1\mathbf{x}_1 + \sqrt{\alpha_5}\mathbf{K}_2\mathbf{x}_2 + \sqrt{\alpha_4}\mathbf{K}_{\mathrm{r}}\mathbf{x}_{\mathrm{r}} + \mathbf{n}_{\mathrm{e}}, \qquad (24)$$

We review the performance of this system using the sum-secrecy-rate [5] given by $[R_{\mathrm{S}_1} + R_{\mathrm{S}_2} - R_{\mathrm{E}}]^+$, where R_m is the rate of the m^{th} transceiver node, R_{E} is the rate of eavesdropper, in $bits/sec/Hz, m \in \{1,2\}$ and $[z]^+ = \max(z,0)$. We can express the sum-secrecy-rate as

$$R_s = \Big[\Big(\sum_{m=1}^{2}\mathbb{E}\{log_2|\mathbf{I}_{N_{\mathrm{s}}}+\mathrm{SINR}_m|\}\Big)-\mathbb{E}\{log_2|\mathbf{I}_{N_{\mathrm{e}}}+\mathrm{SINR}_{\mathrm{E}}|\}\Big]^+,$$

where

$$\begin{aligned}\mathbf{SINR}_m =&[\beta_{\underline{m}}^{(t-1)^2}(\mathbf{V}_m^H\mathbf{G}_m\mathbf{F}\mathbf{F}^H\mathbf{G}_m^H\mathbf{V}_m)][\mathbf{V}_m^H(\sigma_{nm}^2+\\ &\sigma_{nb}^2\mathbf{G}_m\mathbf{F}\mathbf{F}^H\mathbf{G}_m^H + \beta_m^2\mathbf{\Phi}_m\mathbf{Q}_m\mathbf{W}_m\mathbf{W}_m^H\mathbf{Q}_m^H\mathbf{\Phi}_m^H\\ &+\mathbf{G}_m\mathbf{F}\mathbf{\Gamma}_r\mathbf{F}^H\mathbf{G}_m^H)\mathbf{V}_m]^{-1}, m=1,2,\\ \mathbf{SINR}_{\mathrm{E}} =&(\alpha_3\mathbf{K}_1\mathbf{x}_1\mathbf{x}_1^H\mathbf{K}_1^H + \alpha_5\mathbf{K}_2\mathbf{x}_2\mathbf{x}_2^H\mathbf{K}_2^H)\times\\ &(\alpha_4\mathbf{K}_{\mathrm{r}}\mathbf{x}_{\mathrm{r}}\mathbf{x}_{\mathrm{r}}^H\mathbf{K}_{\mathrm{r}}^H + \sigma_{ne}^2\mathbf{I}_{N_{\mathrm{e}}})^{-1},\\ \text{with } \mathbf{\Gamma}_r =&\mathbf{\Phi}_{\mathrm{r}}^{(t-1)}\mathbf{F}^{(t-1)}(\mathbf{\Delta}^{(t-2)} + \mathbf{\Gamma}_{\mathrm{r}}^{(t-1)})\mathbf{F}^{(t-1)^H}\mathbf{\Phi}_{\mathrm{r}}^{(t-1)^H}.\end{aligned}$$

Similar to $\mathbf{\Gamma}_p$, $\mathbf{\Gamma}_r$ is also a feedback term and is $\mathbf{0}_{\mathrm{N_s}}$ for $t < 3$

5. SIMULATION RESULTS

For the simulations, we consider: $N_{\mathrm{s}} = 2$, $N_{\mathrm{r}} = 4$, $P_1^{\max} = P_2^{\max} = N_{\mathrm{s}} \times 1\mathrm{W}, P_{\mathrm{r}}^{\max} = N_{\mathrm{r}} \times 1\mathrm{W}$, We consider the path loss as defined by the 3GPP LTE for outdoor macro cells [10], viz., $\alpha = 15.3 + 37.6\log_{10}(d)$ dB, where d is the distance between nodes, in metres, at an operating frequency of 2 GHz. We consider $\alpha_1 = \alpha_2 = -92\mathrm{dBm}, \alpha_3 = \alpha_4 = \alpha_5 = -101\mathrm{dBm}$. All the channels matrices and estimation error matrices are assumed to follow Rayleigh fading and their elements are independent and identically distributed complex Gaussian random variables with zero mean and unit variance. Each result was obtained by averaging about 1000 Monte Carlo simulations.

Fig. 2 shows the transmit power of S_1 versus $\sigma_{em}^2/\sigma_{nm}^2, m \in \{1,2\}$. We compare our design with that of [6], where no precoding is done at the sources. The system in [6] is designed to mitigate the residual SI at R and doesn't consider the residual SI at $\mathrm{S}_1, \mathrm{S}_2$. As seen from Fig. 2, the transceiver in [6] always transmit at full power irrespective of the variations in residual SI. This will lead to poor system performance at medium and high values of σ_{em}^2. The precoder design proposed in this paper accounts for this residual SI, hence the transmit power is optimized according to the amount of residual SI. Also, in the case of no residual SI ($\sigma_{em}^2 = 0$), the transmit power with the proposed design is much lesser than that in [6].

The algorithm for iterative computation of $\mathbf{F}, \mathbf{V}_m$ converges in about 5 iterations as seen in Fig. 3(a), thus making it practically feasible for real-world systems. Fig. 3(b) shows the variation of SMSE over time at SNR = 10dB, for different values of INR. It can be observed from the figure that the SMSE starts to stabilize from the 3^{rd} time slot. This is due to the effect of feedback term $\mathbf{\Gamma}_p$ which begins from $t = 3$. So, we can consider any time slot after $t = 3$ for evaluation of further results. We considered the 8^{th} time slot. Fig. 3(c) shows the variation of SMSE versus SNR for different values of INR. The proposed design is found to show good performance at low residual INR and it degrades with increasing INR. This may be accredited to the contribution of the feedback term Γ_p in supressing the residual SI. This again highlights the need for precoding and receive filtering with multiple antenna nodes to suppress the SI.

The results sum secrecy-rate analysis of the system are as shown in Fig. 4. It shows the variation of sum-secrecy rate versus SNR for varying INR. It can be observed that, with sufficient suppression of residual SI, this system achieves good sum-secrecy rate. The high rate is also due the strong interference from the relay signal to $\mathbf{E}$.

6. CONCLUSION

We presented the design of optimal precoders and receive filters for a FD two-way MIMO relaying system operating under transmit power constraints. The end-user precoders optimize the transmit power to mitigate the residual SI. We presented an iterative technique for updating the optimal relay precoder and receive filters in each time slot, with low memory requirement and high computational efficiency. By analyzing the sum-secrecy rate performance, we conclude that with proper SI cancellation, this system provides good physical layer security without requiring any additional expenditure of power, computation, or loss of data rate for separate AN generation.

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