Hindawi Publishing Corporation
EURASIP Journal on Advances in Signal Processing
Volume 2008, Article ID 128592, 14 pages
doi:10.1155/2008/128592

Research Article

Duplex Schemes in Multiple Antenna Two-Hop Relaying

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Received 31 July 2007; Accepted 20 January 2008

Recommended by Thomas Kaiser

A novel scheme for two-hop relaying defined as space division duplex (SDD) relaying is proposed. In SDD relaying, multiple antenna beamforming techniques are applied at the intermediate relay station (RS) in order to separate downlink and uplink signals of a bi-directional two-hop communication between two nodes, namely, S1 and S2. For conventional amplify-and-forward two-hop relaying, there appears a loss in spectral efficiency due to the fact that the RS cannot receive and transmit simultaneously on the same channel resource. In SDD relaying, this loss in spectral efficiency is circumvented by giving up the strict separation of downlink and uplink signals by either time division duplex or frequency division duplex. Two novel concepts for the derivation of the linear beamforming filters at the RS are proposed; they can be designed either by a three-step or a one-step concept. In SDD relaying, receive signals at S1 are interfered by transmit signals of S1, and receive signals at S2 are interfered by transmit signals of S2. An efficient method in order to combat this kind of interference is proposed in this paper. Furthermore, it is shown how the overall spectral efficiency of SDD relaying can be improved if the channels from S1 and S2 to the RS have different qualities.

1. INTRODUCTION

There exists much ongoing work in the promising research field of two-hop relaying [1, 2]. This paper focuses on bidirectional two-hop communication between two nodes, namely, S1 and S2 via an intermediate relay station (RS). It is assumed that the downlink traffic load from S1 to S2 via the RS is the same as the uplink traffic load from S2 to S1 via the RS. Due to the high dynamic range between the signal powers of downlink and uplink signals, typical transceivers at S1, S2, and the RS cannot receive and transmit simultaneously on the same channel resource. In single-hop communication, where S1 and S2 can communicate directly with each other, this problem is typically solved by time division duplex (TDD) or frequency division duplex (FDD) [3]. In TDD, there exist two orthogonal time-slots, one for the downlink and another for the uplink. In FDD, there exist two orthogonal frequency bands, one for the downlink and another for the uplink.

Most two-hop relaying schemes also assume a strict separation of downlink and uplink signals by either TDD or FDD. These schemes are defined as one-way relaying schemes since downlink and uplink can be regarded independently. In amplify-and-forward (AF) two-hop relaying [4], the RS receives a signal from either S1 or S2 on a first hop, amplifies this signal, and retransmits it to either S2 or S1 on a second hop. Due to the fact that a half-duplex RS cannot receive and transmit simultaneously on the same channel resource, two orthogonal channel resources are required, one for the first hop and another for the second hop. If downlink and uplink signals are separated by either TDD or FDD, the number of required channel resources is doubled compared to a single-hop communication. Regarding the spectral efficiency of two-hop relaying, this leads to a trade-off between the improved receive signal quality due to the reduced overall pathloss between S1 and S2 and the increase in required channel resources due to the two-hop approach. In literature, there exist many one-way relaying schemes which try to overcome this conceptual drawback of two-hop relaying. However, there also exist schemes which relax the strict separation of downlink and uplink by TDD or FDD which are also promising and in the focus of this work.

1.1. Related work

One approach is to design two-hop relaying schemes which improve the spectral efficiency by allowing a smart reuse of channel resources among multiple one-way relaying connections [5–7].

In [5], multiple RSs are divided into two groups that alternately receive and transmit signals, that is, while one group is receiving signals from the source node, the other group is transmitting signals to the destination node. Since the source transmits all the time in this scheme, the number of required channel resources is the same as in the single-hop case. However, the performance can be significantly degraded by co-channel interference between the two groups of RSs. In [6], one source node communicates with K different destination nodes via K different RSs. Firstly, the source node transmits consecutively to the K RSs using K time slots. Secondly, all RSs transmit simultaneously to their assigned destination nodes in the relay time slot $K + 1$. Obviously, this protocol does not require double the resources compared to the single-hop network, but only $(K + 1)/K$. However, the performance may be significantly degraded by co-channel interference from the RSs at the destination nodes. The problem of co-channel interference is also addressed in [7], where the co-channel interference is kept low by a smart selection of simultaneously transmitting RSs in the relay time slot.

Two other schemes which consider only one source and one destination node are proposed in [8]. For the first scheme, the communication between source and destination node is assisted by two RSs. In the first time slot, one RS receives from the source node and the other RS transmits to the destination node. In the second time slot, the RSs switch their roles. Since the source may transmit in every time slot, the number of required channel resources is the same as in the single-hop case. However, since the two RSs use the same channel resources, there still exists co-channel interference. Further work on this first scheme considering the direct link between S1 and S2 is presented in [9].

The second scheme from [8] which is termed two-way relaying is of particular interest for this work. It has been first introduced in [10] and it has attracted many similar works. In contrast to all previous schemes, two-way relaying is especially developed for bidirectional communication. For the first time, it relaxes the constraint that downlink and uplink signals are transmitted on orthogonal time slots and/or orthogonal frequency bands. Hence, it uses neither TDD nor FDD. In two-way relaying, S1 and S2 transmit simultaneously on a first channel resource to an RS which receives a superposition of both signals. In general, there are two different approaches of how to process the receive signal at the RS. For the decode-and-forward (DF) approach, the receive signal at the RS is decoded and the two separated signals from S1 and S2 are jointly re-encoded before retransmission. For the AF approach, the receive signal is only amplified at the RS before retransmission. For both approaches, a second channel resource is used for the retransmission, and S1 and S2 may utilize their knowledge about the interference term which is coming from their own transmitted signal in order to detect the desired signal. In [11], the rate regions of DF two-way relaying are investigated. This work gives the optimal relative sizes of the first and second channel resources in order to maximize the achievable rate of DF two-way relaying. Two-way relaying is closely connected to network coding [12]. Actually, in network coding data packets from different sources in a multinode computer network are jointly encoded at intermediate network nodes, thus saving network resources, that is, DF two-way relaying can be interpreted as network coding in the original sense with the extension of allowing wireless links. In [13], the interconnection between AF two-way relaying and network coding is also established. Like in [14], it is assumed that for DF two-way relaying three orthogonal channel resources are required. The first two resources are required for the transmission from S1 and S2 to the RS, respectively. This scheme guarantees that both signals can be decoded separately at the RS. The third resource is required for the retransmission of the jointly re-encoded signal from the RS. It is shown in [13] that AF two-way relaying provides a higher throughput for low noise levels at the RS than the considered DF two-way relaying which requires three instead of two orthogonal channel resources.

Another technique which promises to improve the spectral efficiency of two-hop relaying is the application of multiple antennas leading to multiple-input multiple-output (MIMO) relaying [15]. In [16–18], it is shown that the performance of a single AF two-hop relaying connection can be significantly improved if channel state information (CSI) is exploited at a multiple antenna RS allowing single-user beamforming at the RS. In [19], multiuser beamforming is applied at multiple RSs in order to supply multiple destinations with their desired signals, that is, multiple AF two-hop relaying connections are separated spatially. However, [16–19] only assume one-way relaying schemes for the multiple antenna RSs. In [20], multiple antennas and CSI at the RS are applied in the context of DF two-way relaying. It is assumed that the signals from S1 and S2 are decoded at the RS, and two different schemes for the spatial precoding at the RS before the retransmission are proposed. For the first scheme, both decoded signals are re-encoded separately and linearly combined by applying a spatial precoding matrix coming from the singular value decomposition of the channel. For the second scheme, both decoded signals are combined by a bitwise exclusive or (XOR) operation, and the spatial precoding is applied to the new single bit stream. It is shown that the second approach outperforms the first approach in terms of achievable rate. Although the schemes in [20] apply multiple antennas in two-way relaying for the first time, decoding and re-encoding are still required at the RS.

1.2. Own contribution

In this paper, an AF two-way relaying scheme with multiple antennas and linear signal processing at the RS without decoding and re-encoding is proposed leading to a new duplex scheme, defined as space division duplex (SDD). In SDD relaying, downlink and uplink are transmitted on the same channel resources in time and frequency but separated in space. This scheme circumvents the increase in required channel resources for two-hop relaying. Since the RS in the two-way relay channel is a receiver as well as a transmitter, linear receive and transmit beamforming can both be applied at the RS if CSI is available at the RS. The resulting spatial filter matrix at the RS is termed transceive filter matrix. Two novel concepts for the design of this transceive filter are proposed. It can be designed either in three independent steps or in one step. For both concepts, the linear transceive filters fulfilling the zero forcing (ZF) and the minimum mean square error (MMSE) criteria are derived and compared regarding their bit error rate (BER) performance. In SDD relaying, receive signals at S1 are interfered by transmit signals of S1, and receive signals at S2 are interfered by transmit signals of S2. This interference is defined as duplex interference. Since duplex interference can be perfectly determined at the receivers S1 and S2, it can be subtracted leading to a very simple and efficient method, namely, subtraction of duplex interference (SDI) which is proposed in this paper. Furthermore, it is shown how the spectral efficiency of SDD relaying may be improved for the case of different channel qualities on the two channels from the RS to S1 and S2, respectively.

Regarding its spectral efficiency, SDD relaying is compared to other relaying schemes which require the same effort in terms of number of antennas, achieving CSI, and applied signal processing. Assuming multiple antennas, CSI availability at the RS, and linear signal processing, one could also exploit spatial diversity [21] at the RS instead of applying beamforming in SDD relaying. For that purpose, a one-way relaying scheme applying receive and transmit maximum ratio combining (MRC) [22, 23] at the RS is proposed which is defined as MRC relaying. For MRC relaying, double the resources are required as for SDD relaying since downlink and uplink have to be transmitted separately by either TDD or FDD. However, it provides diversity gain which can compensate the increase in required channel resources by allowing higher transmission rates. Furthermore, a relaying scheme applying a combination of receive MRC and transmit beamforming (BF) at the RS is proposed which is defined as MRC-BF relaying. In MRC-BF relaying, spatial diversity is exploited for the reception from S1 and S2 at the RS and the number of required channel resources for the transmission from the RS to S1 and S2 is reduced.

The channel resource requirements and the applied signal processing at the RS for SDD relaying, MRC relaying, and MRC-BF relaying are summarized in Figure 1. In this paper, the spectral efficiencies of all proposed relaying schemes are investigated and compared to each other.

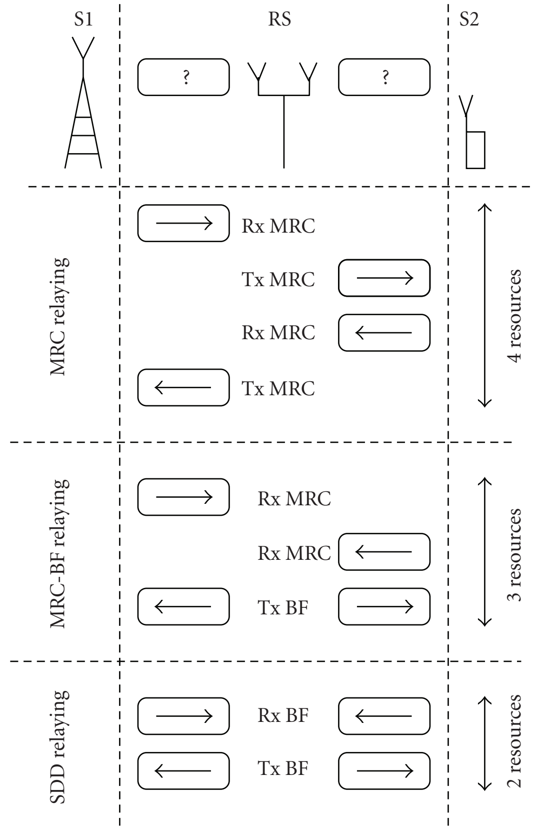

Figure 1: Channel resource requirements of different relaying schemes with applied signal processing at the RS: receive MRC (Rx MRC), transmit MRC (Tx MRC), receive beamforming (Rx BF), and transmit beamforming (Tx BF).

1.3. Notation

Throughout the paper, complex baseband transmission is assumed. Let $[\cdot]^T$, $[\cdot]^*$, $[\cdot]^H$, $\|\cdot\|_2^2$, $(\cdot)^{-1}$, $\det[\cdot]$, $\mathrm{diag}[\cdot]$, and $\mathrm{tr}\{\cdot\}$ denote the transpose, the conjugate, the conjugate transpose, the Euclidean norm, the inverse, the determinant of the matrix argument, a diagonal matrix consisting of the main diagonal elements of the matrix argument, and the sum of the main diagonal elements of the matrix argument, respectively. An identity matrix of size M and a null matrix of size $M \times M$ are denoted by $\mathbf{I}_M$ and $\varnothing_M$, respectively. $E\{\cdot\}$, $\mathrm{Re}\{\cdot\}$, and $\log_2(\cdot)$ denote the expectation, the real part, and the logarithm to the basis two, respectively.

1.4. Outline

The system model of SDD relaying is given in Section 2. Section 3 introduces the ZF and MMSE transceive filters which are firstly given for a three-step design concept, and secondly they are derived by a one-step design concept. In Section 4, the duplex interference in SDD relaying is considered. Section 5 shortly introduces MRC and MRC-BF relaying. The required amount of CSI for the different relaying schemes and extensions of these schemes are discussed in Section 6. In Section 7, the sum rate for SDD relaying is given. Simulation results regarding the BER performance and the spectral efficiency of SDD relaying are presented in Section 8. Section 9 concludes this work.

2. SYSTEM MODEL SDD RELAYING

In the following, the communication between two nodes, namely, S1 and S2, which exchange information via an intermediate RS, is considered. The nodes cannot exchange information directly, for example, due to shadowing conditions. Due to the half-duplex constraint, all stations cannot transmit and receive simultaneously on the same channel resource. S1 and S2 are equipped with $M^{(1)}$ and $M^{(2)}$ antennas, respectively. For SDD relaying, it is required that S1 and S2 are equipped with the same number of antennas, that is,

$$M^{(1)} = M^{(2)} = M, \tag{1}$$

and the RS has to be equipped with

$$M^{(\mathrm{RS})} \geq M^{(1)} + M^{(2)} = 2M, \tag{2}$$

antennas in order to be able to separate down and uplink signals by spatial beamforming.

The data vector $\mathbf{x}^{(1)} = [x_1^{(1)}, \ldots, x_M^{(1)}]^T$ of data symbols $x_n^{(1)}$, $n = 1, \ldots, M$ will be transmitted from S1 to S2, and the data vector $\mathbf{x}^{(2)} = [x_1^{(2)}, \ldots, x_M^{(2)}]^T$ of data symbols $x_n^{(2)}$, $n = 1, \ldots, M$ will be transmitted from S2 to S1. The corresponding transmit covariance matrices are given by $\mathbf{R}_{\mathbf{x}^{(k)}} = E\{\mathbf{x}^{(k)}\mathbf{x}^{(k)^H}\}$, $k = 1, 2$. The overall data vector is defined as $\mathbf{x} = [\mathbf{x}^{(1)^T}, \mathbf{x}^{(2)^T}]^T$ with covariance matrix $\mathbf{R}_{\mathbf{x}} = E\{\mathbf{x}\mathbf{x}^H\}$. For simplicity, the wireless channel is assumed to be flat fading so that all following considerations are applicable, for example, to multicarrier systems. Hence, the channel between Sk, $k = 1, 2$, and the RS may be described by the channel matrix,

$$\mathbf{H}_R^{(k)} = \begin{bmatrix} h_{1,1}^{(k)} & \cdots & h_{1,M}^{(k)} \\ \vdots & \ddots & \vdots \\ h_{M^{(\mathrm{RS})},1}^{(k)} & \cdots & h_{M^{(\mathrm{RS})},M}^{(k)} \end{bmatrix}, \tag{3}$$

where $h_{m,n}^{(k)}$, $m = 1, \ldots, M^{(\mathrm{RS})}$, and $n = 1, \ldots, M$ are complex fading coefficients. The overall channel matrix for the transmission from S1 and S2 to the RS is defined as

$$\mathbf{H}_R = \left[\mathbf{H}_R^{(1)}\mathbf{H}_R^{(2)}\right]. \tag{4}$$

The channel between the RS and Sk, $k = 1, 2$ is described by the channel matrix,

$$\mathbf{H}_T^{(k)} = \begin{bmatrix} \widetilde{h}_{1,1}^{(k)} & \cdots & \widetilde{h}_{1,M^{(\mathrm{RS})}}^{(k)} \\ \vdots & \ddots & \vdots \\ \widetilde{h}_{M,1}^{(k)} & \cdots & \widetilde{h}_{M,M^{(\mathrm{RS})}}^{(k)} \end{bmatrix}, \tag{5}$$

where $\widetilde{h}_{n,m}^{(k)}$, $n = 1, \ldots, M$, and $m = 1, \ldots, M^{(\mathrm{RS})}$ are complex fading coefficients. Assuming channel reciprocity, channel matrix $\mathbf{H}_T^{(k)}$ is the transpose of $\mathbf{H}_R^{(k)}$, that is, $\mathbf{H}_T^{(k)} = \mathbf{H}_R^{(k)^T}$ if the channel is constant during one transmission cycle which includes the transmission from S1 to S2 and the transmission from S2 to S1. For the following considerations, the more general case of $\mathbf{H}_T^{(k)} \neq \mathbf{H}_R^{(k)^T}$ is regarded. The overall channel matrix for the transmission from the RS to S2 and S1 is defined as

$$\mathbf{H}_T = \begin{bmatrix} \mathbf{H}_T^{(2)} \\ \mathbf{H}_T^{(1)} \end{bmatrix}. \tag{6}$$

In SDD relaying, the data vectors $\mathbf{x}^{(1)}$ and $\mathbf{x}^{(2)}$ are exchanged between S1 and S2 during two orthogonal time slots. During the first time slot, S1 and S2 transmit simultaneously to the RS. Since spatial filtering will only be applied at the RS, only scalar transmit filters $\mathbf{Q}^{(1)} = q^{(1)}\mathbf{I}_M$ and $\mathbf{Q}^{(2)} = q^{(2)}\mathbf{I}_M$ are applied at S1 and S2. These transmit filters are required in order to fulfill the transmit energy constraints at S1 and S2. Assuming that $E^{(1)}$ and $E^{(2)}$ are the transmit energies of nodes S1 and S2, the transmit energy constraints are given by

$$E\left\{\|q^{(k)}\mathbf{x}^{(k)}\|_2^2\right\} = E^{(\mathrm{k})}, \quad k = 1, 2. \tag{7}$$

Assuming positive and real scalar transmit filters, the transmit energy constraints from (7) lead to

$$q^{(k)} = \sqrt{\frac{E^{(k)}}{\mathrm{tr}\{\mathbf{R}_{\mathbf{x}^{(k)}}\}}} \quad k = 1, 2, \tag{8}$$

that is, the transmit energy of each node is equally shared among all transmit antennas of the node. The overall transmit filter is given by the block diagonal matrix,

$$\mathbf{Q} = \begin{bmatrix} \mathbf{Q}^{(1)} & \mathbf{0}_M \\ \mathbf{0}_M & \mathbf{Q}^{(2)} \end{bmatrix}. \tag{9}$$

The receive vector $\mathbf{y}_{\mathrm{RS}}$ at the RS is given by

$$\mathbf{y}_{\mathrm{RS}} = \mathbf{H}_R\mathbf{Q}\mathbf{x} + \mathbf{n}_{\mathrm{RS}}, \tag{10}$$

where $\mathbf{n}_{\mathrm{RS}}$ is an additive white Gaussian noise vector with covariance matrix $\mathbf{R}_{\mathbf{n}_{\mathrm{RS}}} = E\{\mathbf{n}_{\mathrm{RS}}\mathbf{n}_{\mathrm{RS}}^H\}$. The covariance matrix of the RS receive vector $\mathbf{y}_{\mathrm{RS}}$ results in

$$\mathbf{R}_{\mathbf{y}_{\mathrm{RS}}} = E\{\mathbf{y}_{\mathrm{RS}}\mathbf{y}_{\mathrm{RS}}^H\} = \mathbf{H}_R\mathbf{Q}\mathbf{R}_{\mathbf{x}}\mathbf{Q}^H\mathbf{H}_R^H + \mathbf{R}_{\mathbf{n}_{\mathrm{RS}}}. \tag{11}$$

At the RS, a linear transceive filter $\mathbf{G}$ is designed in order to ensure that S1 receives an estimate of data vector $\mathbf{x}^{(2)}$ and S2 receives an estimate of data vector $\mathbf{x}^{(1)}$. There are several possibilities of how $\mathbf{G}$ can be designed which will be discussed in Section 3. After applying transceive filter $\mathbf{G}$, the RS transmit vector is given by

$$\mathbf{x}_{\mathrm{RS}} = \mathbf{G}\mathbf{y}_{\mathrm{RS}} = \mathbf{G}(\mathbf{H}_R\mathbf{Q}\mathbf{x} + \mathbf{n}_{\mathrm{RS}}). \tag{12}$$

The RS transmit vector $\mathbf{x}_{\mathrm{RS}}$ has to fulfill the transmit energy constraint at the RS, that is,

$$E\left\{\|\mathbf{x}_{\mathrm{RS}}\|_2^2\right\} \leq E^{(\mathrm{RS})}, \tag{13}$$

where $E^{(\mathrm{RS})}$ is the maximum transmit energy at the RS. In the following, the estimate for data vector $\mathbf{x}^{(1)}$ at S2 is termed $\hat{\mathbf{x}}^{(1)}$, and the estimate for data vector $\mathbf{x}^{(2)}$ at S1 is termed $\hat{\mathbf{x}}^{(2)}$. For each receiving node, the scalar receive filters $\mathbf{P}^{(1)} = p^{(1)}\mathbf{I}_M$ at S2 and $\mathbf{P}^{(2)} = p^{(2)}\mathbf{I}_M$ at S1 with filter coefficients $p^{(1)}$ and $p^{(2)}$ are assumed. The overall receive filter matrix results in

$$\mathbf{P} = \begin{bmatrix} \mathbf{P}^{(1)} & \mathbf{0}_M \\ \mathbf{0}_M & \mathbf{P}^{(2)} \end{bmatrix}. \tag{14}$$

The overall estimated data vector $\hat{\mathbf{x}} = [\hat{\mathbf{x}}^{(1)^T}, \hat{\mathbf{x}}^{(2)^T}]^T$ is given by

$$\hat{\mathbf{x}} = \mathbf{P}(\mathbf{H}_T\mathbf{G}\mathbf{H}_R\mathbf{Q}\mathbf{x} + \mathbf{H}_T\mathbf{G}\mathbf{n}_{\text{RS}} + \mathbf{n}_{\text{R}}), \tag{15}$$

where $\mathbf{n}_R = [\mathbf{n}_R^{(2)^T}\ \mathbf{n}_R^{(1)^T}]^T$ is the combined additive white Gaussian noise vector of S2 and S1 with $\mathbf{n}_R^{(2)}$ and $\mathbf{n}_R^{(1)}$ being the noise vector at S2 and S1, respectively. The covariance matrix of $\mathbf{n}_R$ is defined by $\mathbf{R}_{\mathbf{n}_R} = E\{\mathbf{n}_R\mathbf{n}_R^H\}$.

3. LINEAR TRANSCEIVE FILTERS FOR SDD RELAYING

In the following, it is assumed that instantaneous CSI about $\mathbf{H}_R$ and $\mathbf{H}_T$ is available at the RS. In this case, there are two concepts of how the transceive filter $\mathbf{G}$ at the RS can be designed. For the first concept, $\mathbf{G}$ is assumed as a combination of a linear receive filter $\mathbf{G}_R$, a weight matrix $\mathbf{G}_\Pi$, and a linear transmit filter $\mathbf{G}_T$ where all filters can be determined independently, that is, the transceive filter is designed in three steps. For the second concept, $\mathbf{G}$ is designed in one step without separating it into a receive, a weight, and a transmit filter part.

3.1. Three-step design for the linear transceive filter

In the first step, the RS receive vector $\mathbf{y}_{\text{RS}}$ is multiplied with the linear receive filter matrix $\mathbf{G}_R$ resulting in the RS estimation vector,

$$\hat{\mathbf{x}}_{\text{RS}} = \left[\hat{\mathbf{x}}_{\text{RS}}^{(1)^T}, \hat{\mathbf{x}}_{\text{RS}}^{(2)^T}\right]^T = \mathbf{G}_R\mathbf{y}_{\text{RS}} \tag{16}$$

with the estimate $\hat{\mathbf{x}}_{\text{RS}}^{(1)}$ for $\mathbf{x}^{(1)}$ and the estimate $\hat{\mathbf{x}}_{\text{RS}}^{(2)}$ for $\mathbf{x}^{(2)}$, respectively.

In the second step, $\hat{\mathbf{x}}_{\text{RS}}$ is multiplied with the RS weight matrix

$$\mathbf{G}_\Pi = \begin{bmatrix} \sqrt{\dfrac{\beta}{M}}\mathbf{I}_M & \mathbf{0}_M \\ \mathbf{0}_M & \sqrt{\dfrac{(1-\beta)}{M}}\mathbf{I}_M \end{bmatrix}, \tag{17}$$

where the parameter β with $0 \le \beta \le 1$ is a weight factor which is applied to the RS estimation vectors before retransmission. For $\beta = 0.5$, the RS estimation vectors are equally weighted while for $\beta = 1$ only $\hat{\mathbf{x}}_{\text{RS}}^{(1)}$ is transmitted and for $\beta = 0$ only $\hat{\mathbf{x}}_{\text{RS}}^{(2)}$ is transmitted.

In the third step, the weighted RS estimation vector is multiplied with the transmit filter matrix $\mathbf{G}_T$ leading to the RS transmit vector,

$$\mathbf{x}_{\text{RS}} = \mathbf{G}_T\mathbf{G}_\Pi\hat{\mathbf{x}}_{\text{RS}}, \tag{18}$$

from (12). The transmit filter $\mathbf{G}_T$ separates the vectors designated to S1 and S2 before retransmission and substitutes receive processing at S1 and S2. The overall transceive filter matrix is given by

$$\mathbf{G} = \mathbf{G}_T\mathbf{G}_\Pi\mathbf{G}_R. \tag{19}$$

In the following, two different linear transceive filters $\mathbf{G}$ based on the ZF and MMSE criteria are considered, respectively. The derivation of the filters is exactly like in a single-hop MIMO system and can be verified in [24]. Hence, only the resulting filters are summarized here:

(1) ZF transceive filter

(a) ZF receive filter:

$$\mathbf{G}_{R,\text{ZF}} = (\mathbf{Q}_R^H\mathbf{H}_R^H\mathbf{R}_{\mathbf{n}_{\text{RS}}}^{-1}\mathbf{H}_R\mathbf{Q})^{-1}\mathbf{Q}^H\mathbf{H}_R^H\mathbf{R}_{\mathbf{n}_{\text{RS}}}^{-1}; \tag{20}$$

(b) ZF transmit filter:

$$\mathbf{G}_{T,\text{ZF}} = \frac{1}{p_{\text{ZF}}}\mathbf{H}_T^H(\mathbf{H}_T\mathbf{H}_T^H)^{-1}, \tag{21}$$

with the scalar receive filters,

$$p_{\text{ZF}}^{(1)} = p_{\text{ZF}}^{(2)} = p_{\text{ZF}} = \sqrt{\frac{\text{tr}\{(\mathbf{H}_T\mathbf{H}_T^H)^{-1}\mathbf{G}_\Pi\mathbf{G}_R\mathbf{R}_{\mathbf{y}_{\text{RS}}}\mathbf{G}_R^H\mathbf{G}_\Pi^H\}}{E^{(\text{RS})}}}. \tag{22}$$

(2) MMSE transceive filter

(a) MMSE receive filter:

$$\mathbf{G}_{R,\text{MMSE}} = \mathbf{R}_\mathbf{x}\mathbf{Q}^H\mathbf{H}_R^H(\mathbf{H}_R\mathbf{Q}\mathbf{R}_\mathbf{x}\mathbf{Q}^H\mathbf{H}_R^H + \mathbf{R}_{\mathbf{n}_{\text{RS}}})^{-1}; \tag{23}$$

(b) MMSE transmit filter:

$$\mathbf{G}_{T,\text{MMSE}} = \frac{1}{p_{\text{MMSE}}}\left(\mathbf{H}_T^H\mathbf{H}_T + \frac{\text{tr}\{\mathbf{R}_{\mathbf{n}_R}\}}{E^{(\text{RS})}}\mathbf{I}\right)^{-1}\mathbf{H}_T^H, \tag{24}$$

with the scalar receive filters,

$$\begin{aligned} p_{\text{MMSE}}^{(1)} &= p_{\text{MMSE}}^{(2)} = p_{\text{MMSE}} \\ &= \sqrt{\frac{\text{tr}\{\boldsymbol{\Upsilon}^{-2}\mathbf{H}_T^H\mathbf{G}_\Pi\mathbf{G}_R\mathbf{R}_{\mathbf{y}_{\text{RS}}}\mathbf{G}_R^H\mathbf{G}_\Pi^H\mathbf{H}_T\}}{E^{(\text{RS})}}}, \end{aligned} \tag{25}$$

where $\boldsymbol{\Upsilon} = \mathbf{H}_T^H\mathbf{H}_T + (\text{tr}\{\mathbf{R}_{\mathbf{n}_R}\}/E^{(\text{RS})})\mathbf{I}$. Since the derived receive and transmit filters $\mathbf{G}_R$ and $\mathbf{G}_T$ require the same channel coefficients in case of channel reciprocity, processing effort at the RS could be saved. For example, the calculation of the inverse of $\mathbf{H}_T\mathbf{H}_T^H$ in (21) may be reused for the calculation of the inverse of $\mathbf{H}_R^H\mathbf{H}_R$ in (20) if $\mathbf{R}_{\mathbf{n}_{\text{RS}}}$ and $\mathbf{Q}$ are diagonal matrices with equal entries on their main diagonal.

3.2. One-step design for the linear transceive filter

In the following, the ZF and MMSE criteria are applied directly to the estimate of (15), that is, the transceive filter design is not separated into an independent receive and transmit filter design as introduced in the previous section. For the one-step concept, there exist no RS estimation vectors. Hence, it is not possible to give different weights to each direction of communication before the retransmission as introduced in (17). Since the one-step concept is not based on former results for receive and transmit beamforming, the optimization problems are formulated and solved in the following.

(1) ZF transceive filter

For the ZF criterion, the transceive filter $\mathbf{G}$ at the RS has to be designed such that the mean-squared error of the estimate vector $\hat{\mathbf{x}}$ for data vector $\mathbf{x}$ is minimized. With the ZF constraint and the RS transmit power constraint of (13), the ZF optimization may be formulated as

$$\left\{\mathbf{G}_{\mathrm{ZF}}, p_{\mathrm{ZF}}^{(1)}, p_{\mathrm{ZF}}^{(2)}\right\} = \underset{\{\mathbf{G}, p^{(1)}, p^{(2)}\}}{\arg\min} E\{\|\hat{\mathbf{x}} - \mathbf{x}\|_2^2\}, \tag{26a}$$

$$\text{subject to}: \hat{\mathbf{x}} = \mathbf{x} \quad \text{for } \mathbf{n}_{\mathrm{RS}} = \mathbf{0}_{M^{(\mathrm{RS})}\times 1}, \ \mathbf{n}_R = \mathbf{0}_{M\times 1}, \tag{26b}$$

$$E\{\|\mathbf{x}_{\mathrm{RS}}\|_2^2\} \le E^{(\mathrm{RS})} \quad k = 1, 2. \tag{26c}$$

From the derivation in Appendix A, it can be seen that the ZF transceive filter is given by

$$\mathbf{G}_{\mathrm{ZF}} = \frac{1}{p_{\mathrm{ZF}}}\left(\mathbf{H}_T^H \mathbf{H}_T\right)^{-1} \mathbf{H}_T^H \mathbf{Q}^H \mathbf{H}_R^H \left(\mathbf{H}_R \mathbf{Q}\mathbf{Q}^H \mathbf{H}_R^H\right)^{-1} \tag{27}$$

with the scalar receive filters,

$$p_{\mathrm{ZF}}^{(1)} = p_{\mathrm{ZF}}^{(2)} = p_{\mathrm{ZF}} = \sqrt{\frac{\mathrm{tr}\{\boldsymbol{\Gamma}^{-2}\mathbf{H}_T^H \mathbf{Q}^H \mathbf{H}_R^H \boldsymbol{\Phi}^{-1}\mathbf{R}_{\mathbf{y}_{\mathrm{RS}}}\boldsymbol{\Phi}^{-1}\mathbf{H}_R \mathbf{Q}\mathbf{H}_T\}}{E^{(\mathrm{RS})}}}, \tag{28}$$

where $\boldsymbol{\Gamma} = \mathbf{H}_T^H \mathbf{H}_T$ and $\boldsymbol{\Phi} = \mathbf{H}_R \mathbf{Q}\mathbf{Q}^H \mathbf{H}_R^H$. Comparing (27) with the single filters in (20) and (21) shows that both solutions are very similar since both concepts simply reverse the two channels $\mathbf{H}_R$ and $\mathbf{H}_T$;

(2) MMSE transceive filter

The MMSE transceive filter $\mathbf{G}_{\mathrm{MMSE}}$ at the RS has to be designed such that the mean-squared error of the estimate vector $\hat{\mathbf{x}}$ for transmit vector $\mathbf{x}$ is minimized. With the RS transmit power constraint of (13), the MMSE optimization may be formulated as

$$\{\mathbf{G}_{\mathrm{MMSE}}, p_{\mathrm{MMSE}}^{(1)}, p_{\mathrm{MMSE}}^{(2)}\} = \underset{\{\mathbf{G}, p^{(1)}, p^{(2)}\}}{\arg\min} E\{\|\hat{\mathbf{x}} - \mathbf{x}\|_2^2\}, \tag{29a}$$

$$\text{subject to: } E\{\|\mathbf{x}_{\mathrm{RS}}\|_2^2\} \le E^{(\mathrm{RS})}. \tag{29b}$$

From the derivation in Appendix B it can be seen that the MMSE transceive filter is given by

$$\mathbf{G}_{\mathrm{MMSE}} = \frac{1}{p_{\mathrm{MMSE}}}\left(\mathbf{H}_T^H \mathbf{H}_T\right)^{-1} \mathbf{H}_T^H \mathbf{R}_{\mathbf{x}}^H \mathbf{Q}^H \mathbf{H}_R^H \left(\mathbf{H}_R \mathbf{Q}\mathbf{R}_{\mathbf{x}}\mathbf{Q}^H \mathbf{H}_R^H + \mathbf{R}_{\mathbf{n}_{\mathrm{RS}}}\right)^{-1} \tag{30}$$

with the scalar receive filters,

$$p_{\mathrm{MMSE}}^{(1)} = p_{\mathrm{MMSE}}^{(2)} = p_{\mathrm{MMSE}} = \sqrt{\frac{\mathrm{tr}\{\boldsymbol{\Gamma}^{-2}\mathbf{H}_T^H \mathbf{R}_{\mathbf{x}}^H \mathbf{Q}^H \mathbf{H}_R^H \left(\mathbf{R}_{\mathbf{y}_{\mathrm{RS}}}^H\right)^{-1}\mathbf{H}_R \mathbf{Q}\mathbf{R}_{\mathbf{x}}\mathbf{H}_T\}}{E^{(\mathrm{RS})}}}, \tag{31}$$

where $\boldsymbol{\Gamma} = \mathbf{H}_T^H \mathbf{H}_T$. The solution in (30) is somehow different from the solutions in (23) and (24). This comes from the fact that the RS transmit energy constraint has to be relaxed in order to get an analytical solution for the MMSE one-step concept. For a detailed description on this circumstance, please see Appendix B. Due to this difference in both solutions, different BER performances of the one-step and the three-step designs are expected. The three-step concept should outperform the one-step concept since it does not require a relaxation of its constraints.

4. SUBTRACTION OF DUPLEX-INTERFERENCE IN SDD RELAYING

In the following, knowledge about the own transmitted vectors $\mathbf{x}^{(1)}$ and $\mathbf{x}^{(2)}$ will be exploited at S1 and S2, respectively, in order to improve the performance of SDD relaying. For that purpose, $\hat{\mathbf{x}}$ from (15) is decomposed into an overall useful receive signal vector $\mathbf{x}_{\mathrm{uf}} = [\mathbf{x}_{\mathrm{uf}}^{(1)^T}, \mathbf{x}_{\mathrm{uf}}^{(2)^T}]^T$, an overall intersymbol-interference vector $\mathbf{x}_{\mathrm{is}} = [\mathbf{x}_{\mathrm{is}}^{(1)^T}, \mathbf{x}_{\mathrm{is}}^{(2)^T}]^T$, and an overall duplex interference vector $\mathbf{x}_{\mathrm{di}} = [\mathbf{x}_{\mathrm{di}}^{(1)^T}, \mathbf{x}_{\mathrm{di}}^{(2)^T}]^T$ each consisting of the corresponding vectors at S1 and S2. Furthermore, a matrix $\mathbf{A} = \mathbf{P}\mathbf{H}_T \mathbf{G}\mathbf{H}_R \mathbf{Q}$ is defined as

$$\mathbf{A} = \begin{bmatrix} \mathbf{A}^{(1)} & \mathbf{A}_{\mathrm{di}}^{(2)} \\ \mathbf{A}_{\mathrm{di}}^{(1)} & \mathbf{A}^{(2)} \end{bmatrix} \tag{32}$$

with matrices $\mathbf{A}^{(1)}$, $\mathbf{A}_{\mathrm{di}}^{(1)}$, $\mathbf{A}^{(2)}$, and $\mathbf{A}_{\mathrm{di}}^{(2)}$ each of size $M \times M$. Matrices $\mathbf{A}_{\mathrm{uf}}^{(1)} = \mathrm{diag}[\mathbf{A}^{(1)}]$ and $\mathbf{A}_{\mathrm{uf}}^{(2)} = \mathrm{diag}[\mathbf{A}^{(2)}]$ correspond to the useful receive signal vectors containing $\mathbf{x}^{(1)}$ at S2 and containing $\mathbf{x}^{(2)}$ at S1, respectively. Matrices $\mathbf{A}_{\mathrm{is}}^{(1)} = \mathbf{A}^{(1)} - \mathbf{A}_{\mathrm{uf}}^{(1)}$ and $\mathbf{A}_{\mathrm{is}}^{(2)} = \mathbf{A}^{(2)} - \mathbf{A}_{\mathrm{uf}}^{(2)}$ correspond to the intersymbol interference between the data symbols of $\mathbf{x}^{(2)}$ at S1 and the data symbols of $\mathbf{x}^{(1)}$ at S2, respectively. Matrices $\mathbf{A}_{\mathrm{di}}^{(2)}$ and $\mathbf{A}_{\mathrm{di}}^{(1)}$ correspond to the duplex interference from $\mathbf{x}^{(2)}$ at S2 and from $\mathbf{x}^{(1)}$ at S1, respectively. Applying this notation, (15) can be rewritten as

$$\hat{\mathbf{x}} = \underbrace{\begin{bmatrix} \mathbf{A}_{\mathrm{uf}}^{(1)} & \mathbf{0}_M \\ \mathbf{0}_M & \mathbf{A}_{\mathrm{uf}}^{(2)} \end{bmatrix}\mathbf{x}}_{\mathbf{x}_{\mathrm{uf}}} + \underbrace{\begin{bmatrix} \mathbf{A}_{\mathrm{is}}^{(1)} & \mathbf{0}_M \\ \mathbf{0}_M & \mathbf{A}_{\mathrm{is}}^{(2)} \end{bmatrix}\mathbf{x}}_{\mathbf{x}_{\mathrm{is}}} + \underbrace{\begin{bmatrix} \mathbf{0}_M & \mathbf{A}_{\mathrm{di}}^{(2)} \\ \mathbf{A}_{\mathrm{di}}^{(1)} & \mathbf{0}_M \end{bmatrix}\mathbf{x}}_{\mathbf{x}_{\mathrm{di}}} + \mathbf{P}\mathbf{H}_T \mathbf{G}\mathbf{n}_{\mathrm{RS}} + \mathbf{P}\mathbf{n}_R. \tag{33}$$

Subtracting the overall duplex interference vector $\mathbf{x}_{\mathrm{di}}$ from the estimation vector $\hat{\mathbf{x}}$ at S1 and S2, the improved overall estimation vector in SDD relaying is given by

$$\hat{\mathbf{x}}_{\mathrm{imp}} = \hat{\mathbf{x}} - \mathbf{x}_{\mathrm{di}}. \tag{34}$$

Since the duplex interfence is eliminated, the overall signal-to-noise-and-interference ratio (SINR) at S1 and S2 is increased for the estimate in (34) compared to the estimate in (15). This corresponds to a signal-to-noise ratio (SNR) gain in the BER performance which is analyzed in the simulations. Note that this improvement can only be verified for linear transceive filters which introduce interference among simultaneously received and transmitted data symbols like the MMSE transceive filter, for example. A linear filter which fulfills the ZF constraint does not introduce duplex interference at S1 and S2, that is, for the linear ZF transceive filter no SNR gain can be achieved due to subtraction of duplex interference (SDI).

Furthermore, only the duplex interference coming from signal vector $\mathbf{x}^{(1)}$ can be eliminated at S1, and only the duplex interference coming from signal vector $\mathbf{x}^{(2)}$ can be eliminated at S2 by applying SDI. This means that for $M \geq 2$ antennas at S1 and S2, the intersymbol interference $\mathbf{x}_{\text{is}}$ between data symbols of the same vector $\mathbf{x}^{(k)}$ cannot be eliminated since S1 does not know $\mathbf{x}^{(2)}$ and S2 does not know $\mathbf{x}^{(1)}$.

5. RELATED RELAYING SCHEMES

In SDD relaying, the receive and transmit signals at the RS are neither decoded nor encoded. Therefore, SDD relaying can still be interpreted as an AF relaying scheme which applies linear signal processing at the RS. The downlink and uplink signals are separated by multiple antenna beamforming techniques. Due to the proposed linear transceive filters from Section 3, no further signal processing is required at S1 and S2. In this section, two other relaying schemes are proposed, namely, MRC relaying and MRC-BF relaying which are already known from Figure 1. Compared to SDD relaying, the same effort in terms of number of antennas, achieving CSI, and applied signal processing is required in MRC and MRC-BF relaying. Since both schemes apply state-of-the-art signal processing at the RS, they are only shortly summarized here.

5.1. *MRC relaying*

MRC relaying is a one-way relaying protocol, that is, the bidirectional communication between S1 and S2 requires four orthogonal channel resources. MRC is a well-known approach for combating and fading of the wireless channel [22]. Originally, signals which are received via multiple diversity branches are combined that way that the SNR at the receiver is maximized. MRC can also be applied to the transmit signal [23]. In two-hop relaying, one may apply both receive and transmit MRC since each antenna at the RS represents a diversity branch for reception as well as for transmission. In MRC relaying, on the first channel resource, S1 transmits $\mathbf{x}^{(1)}$ to the RS. Firstly, receive MRC is applied to the receive vector at the RS, that is, the MRC receive filter at the RS is matched to channel $\mathbf{H}_R^{(1)}$ from S1 to the RS. Secondly, transmit MRC is applied at the RS, that is, the MRC transmit filter at the RS is matched to channel $\mathbf{H}_T^{(2)}$ from the RS to S2. On the second channel resource, the RS retransmits the filtered vector to S2 leading to the estimate $\hat{\mathbf{x}}^{(1)}$. Using the third and fourth channel resource, the same scheme is applied for the transmission of $\mathbf{x}^{(2)}$ from S2 to S1 via the RS.

In contrast to SDD relaying, downlink and uplink signals are separated conventionally by either TDD or FDD in MRC relaying.

5.2. *MRC-BF relaying*

For MRC-BF relaying, three orthogonal channel resources are required for the bidirectional communication between S1 and S2. On the first channel resource, S1 transmits $\mathbf{x}^{(1)}$ to the RS. Receive MRC is applied to the receive vector at the RS, that is, the MRC receive filter at the RS is matched to channel $\mathbf{H}_R^{(1)}$ from S1 to the RS. The estimate $\hat{\mathbf{x}}_{\text{RS}}^{(1)}$ is stored at the RS for further signal processing. On the second channel resource, S2 transmits $\mathbf{x}^{(2)}$ to the RS. Receive MRC is applied to the receive vector at the RS, that is, the MRC receive filter at the RS is matched to channel $\mathbf{H}_R^{(2)}$ from S2 to the RS. The two estimates $\hat{\mathbf{x}}_{\text{RS}}^{(1)}$ and $\hat{\mathbf{x}}_{\text{RS}}^{(2)}$ after the MRC receive filters are spatially separated by a linear transmit beamforming filter which can be taken from the set of transmit filters in Section 3.1. On the third channel resource, the filtered estimates at the RS are simultaneously retransmitted to S1 and S2.

In MRC-BF relaying on the first two channel resources, downlink and uplink signals are separated by either TDD or FDD, but on the third channel resource, downlink and uplink signals are separated by SDD. This means that MRC-BF relaying is a mixture of different duplex schemes.

Note that the order of MRC and beamforming could also be reversed which would lead to another relaying scheme. In this scheme, firstly receive beamforming and secondly transmit MRC would be applied at the RS. Since this scheme is very similar to MRC-BF relaying and provides no new results, it is not considered in the following.

6. CSI REQUIREMENTS

Throughout the paper, it is assumed that the considered CSI is instantaneously and perfectly known. However, there exists much space for future work which investigates the impact of noninstantaneous and imperfect CSI to the proposed relaying schemes. In this section, SDD relaying is analyzed concerning the location where CSI is required, and how it can be achieved at this location.

SDD relaying without SDI requires CSI only at the RS. CSI of the channels $\mathbf{H}_R^{(1)}$ and $\mathbf{H}_R^{(2)}$ from S1 and S2 to the RS, respectively, can be obtained by inserting a pilot signal into the transmit signal of each node and estimating each channel at the RS independently. For a sufficiently long channel coherence time which allows to assume channel reciprocity, the same channel coefficients can be used for the retransmission from the RS to S1 and S2, that is, $\mathbf{H}_T^{(k)} = \mathbf{H}_R^{(k)^T}$, $k = 1, 2$. This means that no CSI feedback channels are required for SDD relaying without SDI.

The performance of SDD relaying may be improved if CSI is also available at S1 and S2. In this case, SDD relaying with SDI as introduced in Section 4 can be applied. For SDD relaying with SDI, it is assumed that the RS still estimates both channels $\mathbf{H}_R$ and $\mathbf{H}_T$ in order to design the transceive filter. Furthermore, the matrices $\mathbf{A}_{\text{di}}^{(1)}$ and $\mathbf{A}_{\text{di}}^{(2)}$ from (32) are determined at the RS and signaled to S1 and S2, respectively, via a feedback channel. Knowing these matrices and the own transmitted vectors $\mathbf{x}^{(1)}$ and $\mathbf{x}^{(2)}$ at S1 and S2, respectively, it

Table 1: CSI requirements for the proposed relaying schemes.

	CSI estimation at RS	CSI signaling: RS → S1/S2
MRC relaying	X	
MRC-BF relaying	X	
SDD relaying	X	
MRC-BF relaying with SDI	X	X
SDD relaying with SDI	X	X

is possible to subtract the duplex interference $\mathbf{x}_{\text{di}}^{(2)}$ at S1 and $\mathbf{x}_{\text{di}}^{(1)}$ at S2.

In MRC relaying, CSI about the same channels like in SDD relaying is required at the RS. Therefore, CSI can be achieved in the same way. Due to the separation of downlink and uplink signals by either TDD or FDD, there exists no duplex interference in MRC relaying, that is, CSI signaling from the RS to S1 and S2 cannot improve the performance.

In MRC-BF relaying, CSI about the same channels like in SDD relaying is required at the RS. Therefore, CSI can be achieved in the same way. Like in SDD relaying, duplex interference is generated at S1 and S2 due to the transmit beamforming filter in MRC-BF relaying. The required CSI for SDI can be achieved via a feedback channel like in SDD relaying.

Table 1 gives an overview whose schemes require CSI estimation at the RS and whose schemes additionally require CSI signaling from the RS to S1 and S2.

A final remark will be given on SDD relaying combined with cooperative relaying [1]. Since S1 and S2 always receive and transmit simultaneously in SDD relaying, it is not possible to exploit the direct channel between S1 and S2 for a cooperative relaying approach. Hence, SDD relaying is a relaying scheme which is especially developed for scenarios where the direct channel between S1 and S2 is not available, for example, due to shadowing or limited transmit power. Since S1 and S2 receive and transmit on different channel resources, cooperation is possible for MRC and MRC-BF relaying in general. However, additional effort would be required in this case, and cooperative relaying goes beyond the scope of this paper.

7. SUM RATE OF SDD RELAYING

In the following, the sum rate of a system is defined as the sum of the mutual information values for all transmissions using the same channel resources. It is a measure for the spectral efficiency of the considered relaying schemes. In [25], it is shown that for a MIMO system with

$$\tilde{\mathbf{y}} = \tilde{\mathbf{A}}\tilde{\mathbf{x}} + \tilde{\mathbf{B}}\tilde{\mathbf{n}}, \tag{35}$$

the mutual information is given by

$$C_{\text{MIMO}} = \log_2\left(\det\left[\mathbf{I} + \frac{\tilde{\mathbf{A}}\mathbf{R}_{\tilde{\mathbf{x}}}\tilde{\mathbf{A}}^H}{\tilde{\mathbf{B}}\mathbf{R}_{\tilde{\mathbf{n}}}\tilde{\mathbf{B}}^H}\right]\right), \tag{36}$$

where $\tilde{\mathbf{A}}$ and $\tilde{\mathbf{B}}$ depend on the underlying MIMO system, and $\mathbf{R}_{\tilde{\mathbf{x}}}$ and $\mathbf{R}_{\tilde{\mathbf{n}}}$ are the transmit vector and receive noise vector covariance matrices, respectively.

In the following, the intersymbol interference and the duplex interference in SDD relaying are regarded as additional noise, leading to the overall interference and noise vector:

$$\mathbf{n}^{(k)} = \left[\mathbf{x}^{(k)^T}\ \mathbf{x}^{(i)^T}\ \mathbf{n}_{\text{RS}}^T\ \mathbf{n}_R^{(k)^T}\right]^T, \quad k = \begin{cases}1 & \text{for } i = 2,\\ 2 & \text{for } i = 1,\end{cases} \tag{37}$$

at node Si, with covariance matrix $\mathbf{R}_{\mathbf{n}^{(k)}} = E\{\mathbf{n}^{(k)}\mathbf{n}^{(k)^H}\}$. Furthermore, the overall interference and noise matrix $\mathbf{B}_{\text{TW}}^{(k)}$ is given by

$$\mathbf{B}_{\text{TW}}^{(k)} = \left[\mathbf{A}_{\text{is}}^{(k)}\ \ \mathbf{A}_{\text{di}}^{(i)}\ \ \mathbf{P}^{(k)}\mathbf{H}_T^{(k)}\mathbf{G}\ \ \mathbf{P}^{(k)}\right], \quad k = \begin{cases}1 & \text{for } i = 2,\\ 2 & \text{for } i = 1,\end{cases} \tag{38}$$

at node Si. Under these assumptions, the mutual information in SDD relaying at each node is given by

$$C_{\text{TW}}^{(k)} = \frac{1}{2}\log_2\left(\det\left[\mathbf{I}_M + \frac{\mathbf{A}_{\text{uf}}^{(k)}\mathbf{R}_{\mathbf{x}^{(k)}}\mathbf{A}_{\text{uf}}^{(k)^H}}{\mathbf{B}_{\text{TW}}^{(k)}\mathbf{R}_{\mathbf{n}^{(k)}}\mathbf{B}_{\text{TW}}^{(k)^H}}\right]\right) \quad \text{for } k = 1, 2, \tag{39}$$

where $C_{\text{TW}}^{(1)}$ is the mutual information at node S2, and $C_{\text{TW}}^{(2)}$ is the mutual information at node S1. The pre-log factor 1/2 is introduced in order to indicate the increase in required channel resources for each direction of communication due to the two-hop relaying approach. Because of the simultaneous transmission of downlink and uplink signals, the sum rate of SDD relaying results in

$$C_{\text{TW}} = C_{\text{TW}}^{(1)} + C_{\text{TW}}^{(2)}. \tag{40}$$

Note that in case of SDI at S1 and S2 as introduced in Section 4, matrices $\mathbf{A}_{\text{di}}^{(i)}$, $i = 1, 2$, are set to be zero, that is, $\mathbf{A}_{\text{di}}^{(i)} = \mathbf{0}_M$, since there exists no duplex interference for this scheme.

7.1. Maximizing the sum rate

Both mutual information values $C_{\text{TW}}^{(1)}$ and $C_{\text{TW}}^{(2)}$ depend on the quality of both channels, $\mathbf{H}_{R/T}^{(1)}$ between S1 and the RS and $\mathbf{H}_{R/T}^{(2)}$ between S2 and the RS, that is, even if one channel is much better than the other channel, both the downlink and uplink signals are degraded by the worse channel.

For the three-step concept of the transceive filter design from Section 3.1, it is possible to give different weights β to the two RS estimation vectors $\hat{\mathbf{x}}_{\text{RS}}^{(1)}$ and $\hat{\mathbf{x}}_{\text{RS}}^{(2)}$ after the receive filter $\mathbf{G}_R$. Assigning equal weights to both RS estimation vectors at the RS before retransmission may lead to a suboptimum sum rate if one RS estimation vector is received over a good channel while the other RS estimation vector is received over a bad channel. The sum rate of (40) can be maximized by optimizing β from (17). The underlying optimization problem is formulated as

$$\beta_{\text{opt}} = \arg\max_{\beta}\left\{C_{\text{TW}}^{(1)} + C_{\text{TW}}^{(2)}\right\}, \quad \text{subject to: } 0 \leq \beta \leq 1. \tag{41}$$

There exists no closed form solution to this optimization problem. However, it can be solved by numeric computer optimization.

For the one-step design from Section 3.2, this optimization is not possible since there exist no estimation vectors at the RS which could be weighted. The filter design for the one-step concept is adapted to the overall channel which is a combination of $\mathbf{H}_R$ and $\mathbf{H}_T$, but it cannot be adapted to each channel separately which is the case for the three-step design.

7.2. Approximation for maximizing the sum rate

In the following, the optimization problem in (41) is simplified leading to a closed form approximation for β_{opt} in the three-step transceive filter design. Let us assume a fading channel with an average SNR on the channel from S1 to the RS given by $\rho^{(1)}$ and an average SNR on the channel from S2 to the RS given by $\rho^{(2)}$. In this case, the overall average SNR for AF relaying at receiving node S2 results in [4],

$$\rho_{\text{ov}}^{(1)} = \frac{\beta\rho^{(1)}\rho^{(2)}}{\rho^{(1)} + \beta\rho^{(2)} + 1}, \tag{42}$$

and the overall SNR at receiving node S1 results in

$$\rho_{\text{ov}}^{(2)} = \frac{(1-\beta)\rho^{(1)}\rho^{(2)}}{(1-\beta)\rho^{(1)} + \rho^{(2)} + 1}. \tag{43}$$

Approximating the mutual information values of (39) by the single-input single-output (SISO) mutual information:

$$\widetilde{C}_{\text{TW}}^{(k)} = \frac{1}{2}\log_2\left(1 + \rho_{\text{ov}}^{(k)}\right) \quad \text{for } k = 1, 2, \tag{44}$$

the sum rate may be approximated in the high SNR region by

$$\widetilde{C}_{\text{TW}} = \frac{1}{2}\log_2\left(\rho_{\text{ov}}^{(1)}\right) + \frac{1}{2}\log_2\left(\rho_{\text{ov}}^{(2)}\right). \tag{45}$$

Substituting (42) and (43) into (45) and setting the deviation of (45) equal to zero the approximation leads to

$$\beta_{\text{app}} = \begin{cases} 0.5 & \text{for } \rho^{(1)} = \rho^{(2)}, \\ \dfrac{\rho^{(1)} + 1 - \sqrt{\left(\rho^{(1)}+1\right)\left(\rho^{(2)}+1\right)}}{\rho^{(1)} - \rho^{(2)}} & \text{for } \rho^{(1)} \neq \rho^{(2)}. \end{cases} \tag{46}$$

Note that the sum rate which is calculated by (45) is different from the exact sum rate in (40). However, in order to determine the optimum parameter β this approximation provides reasonable results with low effort, which is also confirmed by the following simulation results.

8. SIMULATION RESULTS

In this section, the BER performance and the average sum rate of SDD relaying are analyzed by means of simulations. The overall BER performance which is defined as the average over both BER values at S1 and S2, respectively, is used to compare the different design concepts for the transceive filters in SDD relaying. It is also a measure in order to indicate the gain due to SDI in SDD relaying. The BER performance strongly depends on the applied modulation and coding schemes which have to be individually adapted to the current channel conditions and the quality of service (QOS) requirements of the transmission. Due to this dependency and due to the discrete number of available modulation and coding schemes, the BER performance is no feasible measure to analyze the spectral efficiency of SDD relaying. Furthermore, SDD relaying, MRC-BF relaying, and MRC relaying provide different transmission rates for the same modulation and coding schemes due to the different number of required channel resources so that a comparison of their BER performances would not be fair. Thus, the sum rate defined in Section 7 is used to compare the spectral efficiency of the proposed relaying schemes.

For the BER performance analyses, the data symbols of S1 and S2 are QPSK modulated. For the sum rate analyses, Gaussian data signals are assumed. The channel coefficients are spatially white and Rayleigh distributed with zero mean and variance one. The noise vectors are complex zero mean Gaussian with variance σ_{RS}^2 at the RS, variance σ_1^2 at S1, and variance σ_2^2 at S2, respectively. The presented results are achieved from Monte Carlo simulations with statistically independent channel fading realizations where $\rho^{(1)} = E^{(\text{RS})}/\sigma_1^2 = E^{(1)}/\sigma_{\text{RS}}^2$ denotes the average SNR between S1 and the RS, and $\rho^{(2)} = E^{(\text{RS})}/\sigma_2^2 = E^{(2)}/\sigma_{\text{RS}}^2$ denotes the average SNR between S2 and the RS.

8.1. Comparison of one-step and three-step designs

For the following investigations, the average SNR $\rho^{(1)}$ of the first channel from S1 to the RS is fixed at a certain value, and the overall BER is depicted depending on the average SNR $\rho^{(2)}$ of the second channel from S2 to the RS. It is assumed that nodes S1 and S2 are each equipped with $M = 1$ antenna and the RS is equipped with $M^{(\text{RS})} = 2$ antennas. Figure 2 gives the overall BER performance for the linear ZF and MMSE transceive filters from Section 3 which are either designed in one step or in three steps. For the one-step design, $\beta = 0.5$ is chosen since the optimization of the sum rate is not of interest for the following investigations. For all transceive filters, the BER performance has an error floor which increases for decreasing $\rho^{(1)}$, that is, all curves show a saturation region where an increase of $\rho^{(2)}$ does no longer improve the BER performance due to the fixed value of $\rho^{(1)}$. From receive and transmit oriented spatial filters, it is known that the linear MMSE receive and transmit filters outperform the linear ZF receive and transmit filters [24]. This result is also found for the transceive filters in SDD relaying for the one-step design which applies the same receive and transmit filters like in [24]. The one-step and the three-step designs

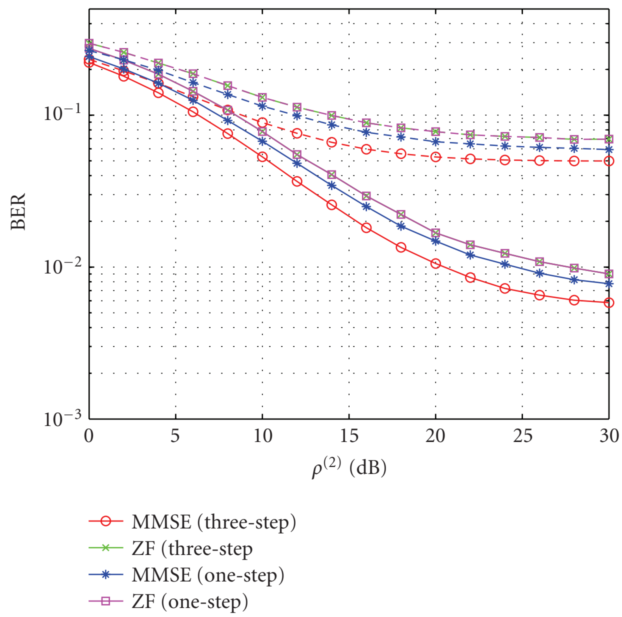

Figure 2: Comparison of overall BER performance for the ZF and MMSE transceive filters with one-step and three-step designs, $M^{(1)} = M^{(2)} = 1$, $M^{(RS)} = 2$ (dashed lines: $\rho^{(1)} = 10$ dB, solid lines: $\rho^{(1)} = 20$ dB).

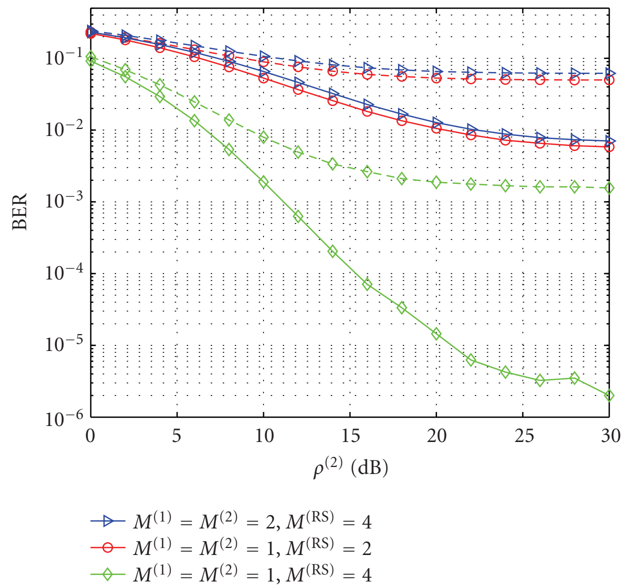

Figure 3: Comparison of overall BER performance for the MMSE transceive filter for different antenna configurations (dashed lines: $\rho^{(1)} = 10$ dB, solid lines: $\rho^{(1)} = 20$ dB).

for the linear ZF transceive filter lead exactly to the same BER performance. This has already been expected from the derivation of the filters since both solutions simply reverse the overall channels $\mathbf{H}_R$ and $\mathbf{H}_T$. For the MMSE transceive filter, the three-step design performs better than the one-step design. This could also be expected from the design of the filters since the one-step design does not consider the RS energy constraint in its optimization which leads to a suboptimum solution. Comparing (30) with (23) and (21), it can be seen that the one-step MMSE transceive filter is a combination of a MMSE receive filter and a ZF transmit filter. Thus, the BER performance of the one-step MMSE transceive filter is better than a three-step transceive filter consisting of a ZF receive and a ZF transmit filter but worse than a three-step transceive filter consisting of a MMSE receive and a MMSE transmit filter.

In the following, the BER performance of the MMSE transceive filter from the three-step design is considered since it provides the best results and its relative behavior is similar to all other introduced transceive filters. Figure 3 gives the overall BER performance depending on the number of antennas at S1, S2, and RS. The result for $M^{(RS)} = 2$ antennas at the RS and $M^{(1)} = M^{(2)} = 1$ antenna at S1 and S2 is already known from the Figure 2. Increasing the number of antennas at RS leads to a significantly improved overall BER performance which can be seen for the case $M^{(RS)} = 4$ and $M^{(1)} = M^{(2)} = 1$. For this antenna configuration, the antenna beams at the RS get tighter, that is, due to the higher degree of freedom at the RS the spatial separation of S1 and S2 by the linear MMSE transceive filter can be improved. However, increasing the number of antennas at S1 and S2 even degrades the BER performance compared to the one-antenna case, that is, $M^{(RS)} = 4$ and $M^{(1)} = M^{(2)} = 2$ provide a worse BER performance than $M^{(RS)} = 2$ and $M^{(1)} = M^{(2)} = 1$. This can be explained by the analyses from Section 4. Here, it is shown that for $M^{(1)} = M^{(2)} \geq 2$ additional intersymbol interference among symbols of the same source appears. This intersymbol interference does not exist for $M^{(1)} = M^{(2)} = 1$ and leads to a degradation of the BER performance for $M^{(1)} = M^{(2)} \geq 2$. This means that for SDD relaying, an increase of the number of antennas at the RS improves the BER performance, but a simultaneous increase of antennas at S1 and S2 will even degrade the BER performance in case of linear filtering at the RS. Of course, if the additional antennas at S1 and S2 are used for spatial diversity by space-time coding, for example, the performance can be also improved. But these considerations are beyond the scope of this paper.

8.2. *Subtraction of duplex-interference*

Figure 4 gives the overall BER performance for the three-step MMSE transceive filter with and without SDI as introduced in Section 4 for $M^{(1)} = M^{(2)} = 1$, and $M^{(RS)} = 2$, and $M^{(RS)} = 4$, respectively. Like the previous results, the BER performance has an error floor which increases for decreasing $\rho^{(1)}$. There exists a significant improvement for the BER performance for the linear MMSE transceive filter if SDI is applied. For a target BER of 10^{-2}, the SNR gain due to SDI is approximately 4 dB for $\rho^{(1)} = 20$ dB and $M^{(RS)} = 2$. For $M^{(RS)} = 4$, there also exists an improvement of the BER performance if SDI is applied. However, the SNR gain is much lower than in case of $M^{(RS)} = 2$. The higher number of antennas at the RS provides a better spatial separation of S1 and S2 which directly reduces the duplex interference. This means that for more than $M^{(RS)} = 2$ antennas at the RS, SDI does not provide a significant improvement and can be

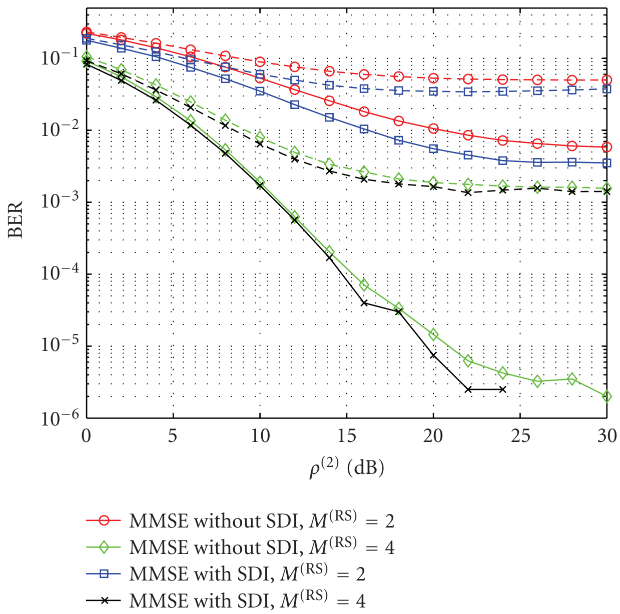

Figure 4: Comparison of the overall BER performance for the MMSE transceive filter with and without SDI for different numbers of antennas at the RS, $M^{(1)} = M^{(2)} = 1$ (dashed lines: $\rho^{(1)} = 10$ dB, solid lines: $\rho^{(1)} = 20$ dB).

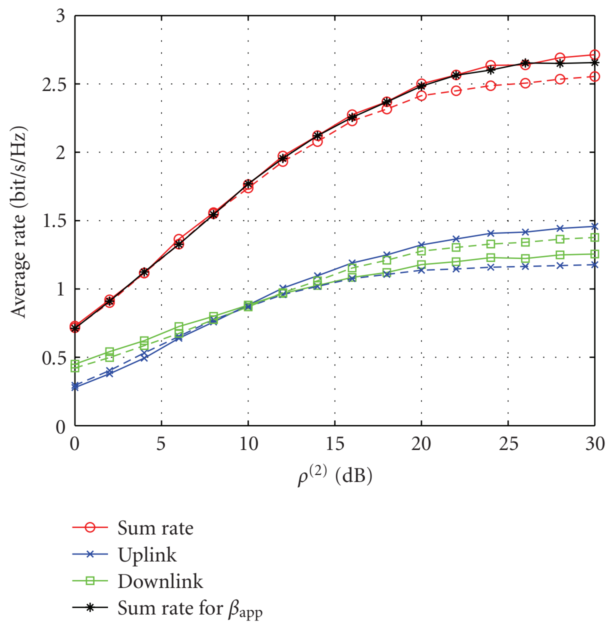

Figure 6: Average sum rate and average rate of downlink and uplink of MMSE transceive filter depending on $\rho^{(2)}$ for fixed $\rho^{(1)} = 10$ dB, $M^{(1)} = M^{(2)} = 1$, $M^{(RS)} = 2$ (solid lines: β from the optimization in (41); dashed lines: $\beta = 0.5$).

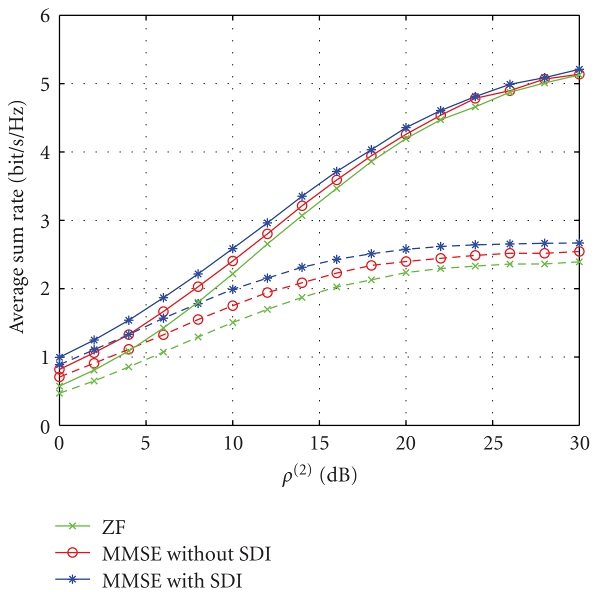

Figure 5: Comparison of average sum rate for the ZF and MMSE transceive filters with and without SDI, $M^{(1)} = M^{(2)} = 1$, $M^{(RS)} = 2$ (dashed lines: $\rho^{(1)} = 10$ dB, solid lines: $\rho^{(1)} = 20$ dB).

neglected. In Figure 5, the average sum rates of the linear ZF transceive filter, the three-step MMSE transceive filter without SDI, and the three-step MMSE transceive filter with SDI are given. Note that SDI cannot improve the performance of the linear ZF transceive filter since the two channels are perfectly orthogonalized by this filter which implicitly suppresses duplex interference at S1 and S2 at the cost of noise enhancement at the receivers. For increasing $\rho^{(2)}$, the average sum rate converges to a constant maximum which depends on $\rho^{(1)}$. For small $\rho^{(1)}$, the sum rate converges faster with increasing $\rho^{(2)}$ and the maximum sum rate is lower than for high $\rho^{(1)}$. For low $\rho^{(2)}$, the linear ZF transceive filter has a worse performance than the linear MMSE transceive filter. If $\rho^{(1)}$ and $\rho^{(2)}$ are sufficiently high, the overall noise at S1 and S2 which consists of the noise at the RS and the noise at S1 and S2 itself can be neglected. In this case, the sum rate of the linear ZF transceive filter converges to the sum rate of the MMSE transceive filter. This effect can already be seen for $\rho^{(1)} = 20$ dB and high values of $\rho^{(2)}$. It can be seen from Figure 5 that SDI increases the sum rate for the MMSE transceive filter if the SNR on both channels is low. For high values of $\rho^{(1)}$ and $\rho^{(2)}$, there exists almost no duplex interference and the sum rates of the MMSE transceive filter with and without SDI converge.

8.3. *Maximizing sum rate*

In Section 7.1, maximizing the sum rate by giving different weights to the RS estimation vectors in case of different channel qualities on the two channels is discussed. In Figure 6 for fixed $\rho^{(1)} = 10$ dB, the average sum rate depending on $\rho^{(2)}$ of a three-step MMSE transceive filter achieved for the numeric optimization of β_{opt} from (41) is compared to the value achieved for fixed $\beta = 0.5$ and the value achieved by the approximation β_{app} from (46). Additionally, the rates of downlink and uplink for β_{opt} and $\beta = 0.5$ are depicted separately. For equal channel qualities on both channels ($\rho^{(1)} = \rho^{(2)}$), all approaches provide the same average sum rate. However, for increasing difference of the

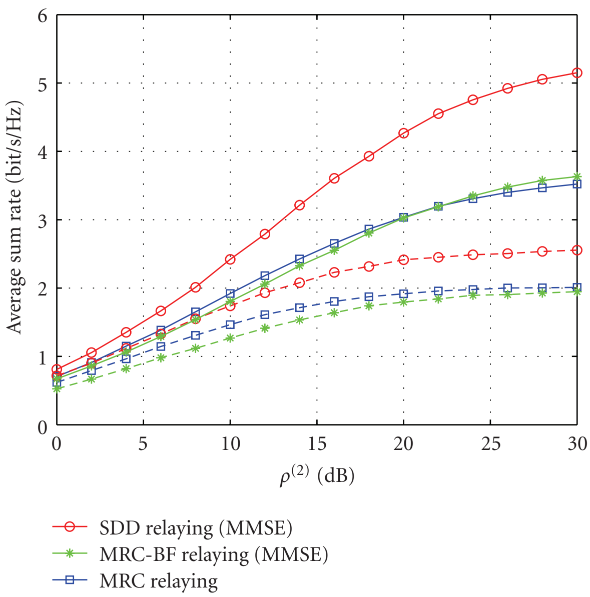

FIGURE 7: Comparison of average sum rate for SDD relaying, MRC relaying, and MRC-BF relaying, $M^{(1)} = M^{(2)} = 1$, $M^{(RS)} = 2$ (dashed lines: $\rho^{(1)} = 10$ dB, solid lines: $\rho^{(1)} = 20$ dB).

channel qualities on both channels, the sum rates diverge. The optimization of β provides a higher sum rate than the fixed approach with $\beta = 0.5$. Let us consider the uplink for $\rho^{(2)} > 10$ dB, which means that the first hop of the uplink has a higher SNR than the first hop of the downlink. Assigning equal weights ($\beta = 0.5$) to both RS estimation vectors results in a lower rate in the uplink than in the downlink since the second hop of the uplink has a lower SNR; but, if the RS estimation vector of the uplink gets a higher weight due to the optimization, then the situation changes and the uplink achieves a higher rate than the downlink. In general, this means that the sum rate may be increased by introducing a higher weight to the RS estimation vector which is received over the better channel on the first hop. This can be explained as follows. The noise at the RS is filtered by the MMSE receive filter which leads to different effective SNR values for the two receive vectors from S1 and S2 after the filter. The receive vector which comes over the better channel has a higher SNR, provides a higher mutual information, and consequently it should get a higher weight before retransmission. The approximation for β_{app} comes very close to the optimum sum rate for β_{opt}, although the approximation requires a significantly lower effort than the optimum solution.

8.4. Average sum rate of different relaying schemes

In the following, the average sum rate of MRC relaying and MRC-BF relaying with a linear MMSE transmit filter is compared to the sum rate of SDD relaying with a linear MMSE transceive filter at the RS. In Figure 7, the average sum rate of the three schemes is depicted depending on $\rho^{(2)}$ for $\rho^{(1)} = 10$ dB and $\rho^{(1)} = 20$ dB. Obviously, the linear MMSE transceive filter in the SDD relaying approach outperforms the MRC and MRC-BF relaying approaches in terms of average sum rate. This comes from the fact that in SDD relaying, downlink and uplink signals are transmitted simultaneously. In MRC relaying, downlink and uplink signals require separated channel resources, and in MRC-BF relaying indeed the number of required channel resources is decreased compared to MRC relaying but still higher than in SDD relaying. There exists a trade-off between the number of required channel resources and the effective receive SNR at S1 and S2. SDD relaying provides the worst effective receive SNR since both antennas at the RS are used for spatial separation of downlink and uplink signals. However, the reduced amount of required channel resources more than compensates the worst effective receive SNR compared to the other schemes. MRC provides the best effective receive SNR since both antennas at the RS are used to improve the effective receive SNR, but the improved effective receive SNR cannot compensate the increased amount of required resources compared to SDD relaying. The effective receive SNR and the number of required channel resources of MRC-BF relaying lie in-between the values of MRC relaying and SDD relaying. However, this trade-off in MRC-BF relaying seems to provide the worst performance since it suffers from duplex interference in contrast to MRC relaying.

9. CONCLUSION

In this paper, SDD relaying is proposed as a novel relaying scheme saving channel resources for equal traffic load in downlink and uplink. In SDD relaying, downlink and uplink signals are separated by applying transceive filters at the RS which can be designed by a one-step or a three-step concept. It turns out that the three-step transceive filters provide better BER performances than the one-step filters and that the three-step transceive filters are more flexible, for example, it is only possible for the three-step transceive filters to give different weights to downlink and uplink signals which lead to an increased sum rate. For a linear MMSE transceive filter, duplex interference can be subtracted at S1 and S2 which results in an improved BER performance. The linear MMSE transceive filters always outperform the linear ZF transceive filters in terms of overall BER performance. The performance of the transceive filters can be significantly improved by increasing the number of antennas at the RS which allows a better spatial separation of S1 and S2. For the sake of the comparison, MRC and MRC-BF relaying are proposed as other relaying schemes which apply multiple antennas and linear signal processing at the RS. It is shown that SDD relaying provides a higher spectral efficiency than MRC and MRC-BF relaying due to the simultaneous transmission of downlink and uplink signals.

APPENDICES

A. DERIVATION OF ZF TRANSCEIVE FILTER

As shown in [26], the ZF optimization problem in (26a) is not convex. However, the Karush-Kuhn-Tucker (KKT) conditions [27] can be used to solve (26a) under the constraints

(26b) and (26c). For simplicity but without loss of generality, it is assumed that the scalar receive filters at S1 and S2 are the same, that is, $p^{(1)} = p^{(2)} = p$. By applying the Lagrangian function,

$$L(\mathbf{G}, p, \mu, \mathbf{\Lambda}) = \operatorname{tr}\{|p|^2(\mathbf{H}_T\mathbf{G}\mathbf{R}_{\mathbf{n}_{RS}}\mathbf{G}^H\mathbf{H}_T^H + \mathbf{R}_{\mathbf{n}_R})\} + \mu(\operatorname{tr}\{\mathbf{G}\mathbf{R}_{\mathbf{y}_{RS}}\mathbf{G}^H\} - E^{(RS)}) - 2\operatorname{Re}\{\operatorname{tr}\{\mathbf{\Lambda}(p\mathbf{H}_T\mathbf{G}\mathbf{H}_R\mathbf{Q} - \mathbf{I})\}\}, \tag{A.1}$$

the KKT conditions are given by

$$\frac{\partial L(\mathbf{G}, p, \mu, \mathbf{\Lambda})}{\partial \mathbf{G}} = |p|^2\mathbf{H}_T^*\mathbf{G}^*\mathbf{R}_{\mathbf{n}_{RS}}^*\mathbf{H}_T^T + \mu\mathbf{G}^*\mathbf{R}_{\mathbf{y}_{RS}}^* - p\mathbf{Q}\mathbf{H}_R^T\mathbf{H}_T^T\mathbf{\Lambda}^T = \mathbf{0}, \tag{A.2a}$$

$$\frac{\partial L(\mathbf{G}, p, \mu, \mathbf{\Lambda})}{\partial p} = \operatorname{tr}\{p^*\mathbf{H}_T\mathbf{G}\mathbf{R}_{\mathbf{n}_{RS}}\mathbf{G}^H\mathbf{H}_T^H - \mathbf{\Lambda}\mathbf{H}_T\mathbf{G}\mathbf{H}_R\} = 0, \tag{A.2b}$$

$$\mu(\operatorname{tr}\{\mathbf{G}\mathbf{R}_{\mathbf{y}_{RS}}\mathbf{G}^H\} - E^{(RS)}) = 0, \tag{A.2c}$$

$$\mathbf{\Lambda}(p\mathbf{H}_T\mathbf{G}\mathbf{H}_R\mathbf{Q} - \mathbf{I}) = \mathbf{0}. \tag{A.2d}$$

From the fourth KKT (A.2d), for $\det[\mathbf{\Lambda}] \neq 0$ one gets

$$\mathbf{G} = \frac{1}{p}(\mathbf{H}_T^H\mathbf{H}_T)^{-1}\mathbf{H}_T^H\mathbf{Q}^H\mathbf{H}_R^H(\mathbf{H}_R\mathbf{Q}\mathbf{Q}^H\mathbf{H}_R^H)^{-1}. \tag{A.3}$$

Applying $\mathbf{G}$ from (A.3), p may be determined from the third KKT (A.2c) for $\mu \neq 0$ leading to the solution given in (28).

B. DERIVATION OF MMSE TRANSCEIVE FILTER

As shown in [26], the MMSE optimization expression in (29a) is not convex. In this case, the KKT conditions are only required to solve (29a) under the transmit power constraint (29b). For simplicity but without loss of generality, it is assumed that the scalar receive filters at S1 and S2 are the same, that is, $p^{(1)} = p^{(2)} = p$. By applying the Lagrangian function,

$$L(\mathbf{G}, p, \mu) = \operatorname{tr}\{\mathbf{R}_\mathbf{x} + |p|^2\mathbf{H}_T\mathbf{G}\mathbf{H}_R\mathbf{Q}\mathbf{R}_\mathbf{x}\mathbf{Q}^H\mathbf{H}_R^H G^H\mathbf{H}_T^H - p\mathbf{H}_T\mathbf{G}\mathbf{H}_R\mathbf{Q}\mathbf{R}_\mathbf{x} - p^*\mathbf{R}_\mathbf{x}\mathbf{Q}^H\mathbf{H}_R^H\mathbf{G}^H\mathbf{H}_T^H\} + \operatorname{tr}\{|p|^2(\mathbf{H}_T\mathbf{G}\mathbf{R}_{\mathbf{n}_{RS}}\mathbf{G}^H\mathbf{H}_T^H + \mathbf{R}_{\mathbf{n}_R})\} + \mu(\operatorname{tr}\{\mathbf{G}\mathbf{R}_{\mathbf{y}_{RS}}\mathbf{G}^H\} - E^{(RS)}), \tag{B.1}$$

the KKT conditions are given by

$$\frac{\partial L(\mathbf{G}, p, \mu)}{\partial \mathbf{G}} = |p|^2\mathbf{H}_T^*\mathbf{G}^*\mathbf{R}_{\mathbf{y}_{RS}}^*\mathbf{H}_T^T - p\mathbf{R}_\mathbf{x}^T\mathbf{Q}^T\mathbf{H}_R^T\mathbf{H}_T^T + \mu\mathbf{G}^*\mathbf{R}_{\mathbf{y}_{RS}}^* = \mathbf{0}, \tag{B.2a}$$

$$\frac{\partial L(\mathbf{G}, p, \mu)}{\partial p} = \operatorname{tr}\{-\mathbf{H}_T\mathbf{G}\mathbf{H}_R\mathbf{Q}\mathbf{R}_\mathbf{x} + p^*\mathbf{H}_T\mathbf{G}\mathbf{R}_{\mathbf{y}_{RS}}\mathbf{G}^H\mathbf{H}_T^H + p^*\mathbf{R}_{\mathbf{n}_R}\} = 0, \tag{B.2b}$$

$$\mu(\operatorname{tr}\{\mathbf{G}\mathbf{R}_{\mathbf{y}_{RS}}\mathbf{G}^H\} - E^{(RS)}) = 0. \tag{B.2c}$$

From the first KKT condition (B.2a) it can bee seen that there exists no analytical solution for $\mathbf{G}$ due to the existence of the power constraint (29b). Neglecting the power constraint in (29b), the first KKT can be rewritten as

$$\frac{\partial L(\widetilde{\mathbf{G}})}{\partial \widetilde{\mathbf{G}}} = \mathbf{H}_T^*\widetilde{\mathbf{G}}^*\mathbf{R}_{\mathbf{y}_{RS}}^*\mathbf{H}_T^T - \mathbf{R}_\mathbf{x}^T\mathbf{Q}^T\mathbf{H}_R^T\mathbf{H}_T^T = \mathbf{0}. \tag{B.3}$$

Solving (B.3) for $\widetilde{\mathbf{G}}$, finally leads to

$$\widehat{\mathbf{G}} = (\mathbf{H}_T^H\mathbf{H}_T)^{-1}\mathbf{H}_T^H\mathbf{R}_\mathbf{x}^H\mathbf{Q}^H\mathbf{H}_R^H\mathbf{R}_{\mathbf{y}_{RS}}^{-1}. \tag{B.4}$$

Introducing the normalization factor $1/p$ to (B.4) in order to fulfill the power constraint (29b) at the RS, subsequently one gets

$$\mathbf{G} = \frac{1}{p}\widehat{\mathbf{G}} \tag{B.5}$$

with p from (31). Note that $1/p$ does not come from the optimization process itself, but it is a somehow artificial weighting factor similar to the approach in [28].

ACKNOWLEDGEMENTS

Parts of this work have been presented at IST Mobile and Wireless Communications Summit 2007 and at IEEE 18th Symposium on Personal, Indoor, and Mobile Radio Communications s(PIMRC 2007).

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