Linear Transceiver Design in Nonregenerative Relays With Channel State Information

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***Abstract*—This paper deals with the design of nonregenerative relaying transceivers in cooperative systems where channel state information (CSI) is available at the relay station. The conventional nonregenerative approach is the amplify and forward (A&F) approach, where the signal received at the relay is simply amplified and retransmitted. In this paper, we propose an alternative linear transceiver design for nonregenerative relaying (including pure relaying and the cooperative transmission cases), making proper use of CSI at the relay station. Specifically, we design the optimum linear filtering performed on the data to be forwarded at the relay. As optimization criteria, we have considered the maximization of mutual information (that provides an information rate for which reliable communication is possible) for a given available transmission power at the relay station. Three different levels of CSI can be considered at the relay station: only first hop channel information (between the source and relay); first hop channel and second hop channel (between relay and destination) information, or a third situation where the relay may have complete cooperative channel information including all the links: first and second hop channels and also the direct channel between source and destination. Despite the latter being a more unrealistic situation, since it requires the destination to inform the relay station about the direct channel, it is useful as an upper benchmark. In this paper, we consider the last two cases relating to CSI. We compare the performance so obtained with the performance for the conventional A&F approach, and also with the performance of regenerative relays and direct noncooperative transmission for two particular cases: narrowband multiple-input multiple-output transceivers and wideband single input single output orthogonal frequency division multiplex transmissions.**

***Index Terms*—Amplify and forward (A&F), channel state information (CSI), cooperative transmission schemes, decode and forward (D&F), multiple-input multiple-output (MIMO), orthogonal frequency division multiplex (OFDM).**

I. Introduction

COOPERATION among users at the physical layer level has shown to be a promising approach for capacity and/or range increase [3], [8], [9], [10], [12], [13], [17]. In these schemes, the signal received from the source and the signal received from the relay station are combined at the destination. Therefore, cooperative schemes can be seen as a generalization of the typical multihop approach where a relaying terminal retransmits the symbols received from the base station or central controller (thus providing range extension). The main advantage of the cooperative schemes, with respect to classical relaying strategies, is that cooperation creates a "virtual" multiple-input multiple-output (MIMO) system that may offer significant capacity gains in shadowing propagation conditions.

There are two different approaches for cooperative transmission, according to the role played by the relaying terminal: the amplify and forward (A&F) scheme and the decode and forward scheme (D&F) [9], [10], [17]. The simplest approach is the A&F approach which is a nonregenerative approach where the relay amplifies and retransmits the signal received from the source. The most complex approach is the D&F scheme where the relay station decodes the received signal and retransmits the decoded and regenerated symbols. The D&F is also known as regenerative approach.

This paper investigates the performance of linear transceivers in cooperative systems where channel state information (CSI) is available at the relay station. By making good use of the CSI, the relay becomes able to carry out some further signal shaping. We assume, therefore, some additional intelligence at the relay station. Consequently, the relay units are no longer simple amplify and forward units, but they are still not required to either decode or re-encode the symbols transmitted by the base station, as in regenerative relaying schemes. Three different degrees of channel state information can be considered at the relay station. The CSI available at the relay station can be only information about the first hop channel (between the source and relay) or information about the first hop channel and the second hop channel (between relay and destination). This second situation is feasible for time division duplex (TDD) systems if both user and relay share a previous dialogue and so channel state information can be obtained by reciprocity. Third, we can consider that the relay has knowledge about all the links involved in the communication, including the direct channel (channel between the source and the destination). Despite this being a more unrealistic situation, since it requires the destination to inform the relay station about the direct channel between source and destination, it is useful as an upper benchmark.

The matrix channel considered in this paper provides a unified way to deal with physical channels of different natures in relaying communications such as flat fading multiple antenna systems or single input single output (SISO) frequency selective systems. Within this general framework, an info-theoretical analysis of MIMO relay channels with linear processing at the relay is presented. Following the guidelines in [11], a convex

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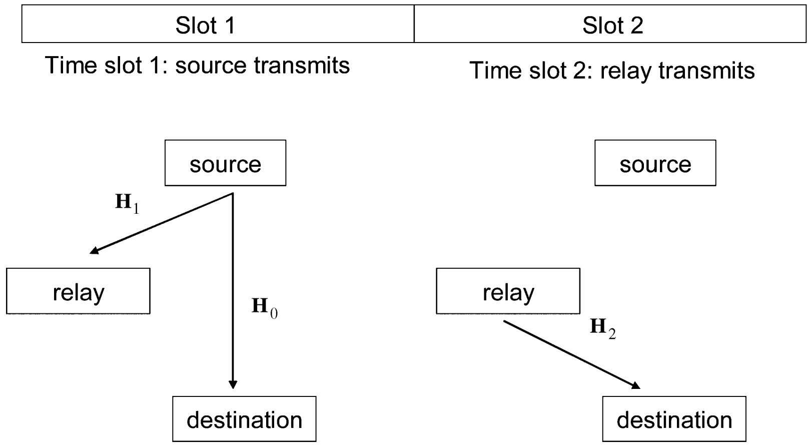

Fig. 1. TDMA scheme for cooperative transmission using one relay.

optimization solution is given in this paper. This solution can be iteratively computed for any value of signal-to-noise ratio (SNR). Recently, in [15] an independent work approached the problem in a different way, reaching a specific solution in multiple antenna channels for certain SNR values, when the number of antennas is the same in all the stations involved.

In Section II, the signal model is presented along with a lower and an upper bound of the mutual information in cooperative relaying schemes. In Section III, the optimal linear processing is found analytically based on the lower bound of the total (cooperative) mutual information with an unknown direct (source-destination) channel. This lower bound turns out to be the mutual information in the relaying path (including first and second hop channels). For the benchmark, we also look at the case of a known direct channel in Section IV. In all cases, a limited available transmission power at the relay station is considered. Results are given in Section V. Finally, in Section VI, the conclusions are given.

II. Mutual Information Bounds for Nonregenerative Cooperative Schemes

A. Signal Model for Cooperative Schemes

Assuming half duplex relaying, the scenario under analysis consists of a source and a relay terminal transmitting through two orthogonal channels, for instance two separate time slots as in time division multiple access (TDMA) systems (see Fig. 1). During the first slot, the source transmits $\mathbf{x}$ and the signals received by the end user and the relay are $\mathbf{y}_0$ and $\mathbf{y}_1$, respectively

$$\begin{aligned}\mathbf{y}_0 &= \mathbf{H}_0\mathbf{x} + \mathbf{w}_0\\ \mathbf{y}_1 &= \mathbf{H}_1\mathbf{x} + \mathbf{w}_1\end{aligned} \tag{1}$$

where $\mathbf{x} \in \mathbb{C}^{M\times 1}$ is the transmitted vector and $\mathbf{H}_0 \in \mathbb{C}^{N\times M}$ is the channel matrix, with M transmit and N receive dimensions, between the source and the destination (direct channel). $\mathbf{H}_1 \in \mathbb{C}^{R\times M}$ denotes the channel matrix between the source and the relay station (first hop channel). Finally, $\mathbf{w}_0 \in \mathbb{C}^{N\times 1}$ and $\mathbf{w}_1 \in \mathbb{C}^{R\times 1}$ are zero-mean circularly symmetric Gaussian noise vectors received at the destination and the relay, respectively, during the first slot.

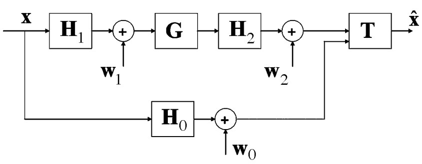

Fig. 2. Cooperative A&F signal model.

During the second slot, the relay transmits the signal using a $R \times R$ linear precoding matrix $\mathbf{G}$ that has to be designed. The signal received at the destination during this second slot is

$$\mathbf{y}_2 = \mathbf{H}_2\mathbf{G}[\mathbf{H}_1\mathbf{x} + \mathbf{w}_1] + \mathbf{w}_2. \tag{2}$$

In the previous expression, $\mathbf{H}_2 \in \mathbb{C}^{N\times R}$ denotes the channel matrix between the relay and the destination (second hop channel) and $\mathbf{w}_2$ is the noise vector at the destination during the second slot.

We will now rewrite in a more compact way the signal model for a nonregenerative relaying system

$$\mathbf{y} = \begin{bmatrix}\mathbf{y}_0\\ \mathbf{y}_2\end{bmatrix} = \begin{bmatrix}\mathbf{H}_0\\ \mathbf{H}_2\mathbf{G}\mathbf{H}_1\end{bmatrix}\mathbf{x} + \begin{bmatrix}\mathbf{I}_N & \mathbf{0} & \mathbf{0}\\ \mathbf{0} & \mathbf{H}_2\mathbf{G} & \mathbf{I}_N\end{bmatrix}\begin{bmatrix}\mathbf{w}_0\\ \mathbf{w}_1\\ \mathbf{w}_2\end{bmatrix}. \tag{3}$$

The signal model is illustrated in Fig. 2 and the optimum minimum mean-square error (MMSE) receiving scheme for such a model is given in Annex III.

By using (3), the mutual information (that provides an information rate for which reliable communication is possible) for a single cooperative connection using nonregenerative relays can be written as follows [6]:

$$I(\mathbf{y};\mathbf{x}) = \frac{1}{2}\log_2\left|\mathbf{I}_{2N} + \mathbf{H}\mathbf{R}_x\mathbf{H}^H\mathbf{R}_w^{-1}\right|. \tag{4}$$

The factor 1/2 in (4) comes from the fact that the signal vector is actually transmitted in two time instances, so the efficiency drops by one half. The mutual information is measured in bits per second per Hertz (bps/Hz).

In (4), the dependence on matrix $\mathbf{G}$ is in

$$\mathbf{H} = \begin{bmatrix}\mathbf{H}_0\\ \mathbf{H}_2\mathbf{G}\mathbf{H}_1\end{bmatrix} \tag{5}$$

and in

$$\mathbf{R}_w = \begin{bmatrix} \mathbf{R}_{w_0} & \mathbf{0} \\ \mathbf{0} & \mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2} \end{bmatrix}. \quad (6)$$

In (4), matrix $\mathbf{R}_x$ is the covariance matrix of the signal vector transmitted from the source

$$\mathbf{R}_x = E\{\mathbf{x}\mathbf{x}^H\}. \quad (7)$$

In (6), $\mathbf{R}_{w_0}$ and $\mathbf{R}_{w_1}$ are the covariance matrices of the noise vectors received at the destination and the relay, respectively, during the first slot. $\mathbf{R}_{w_2}$ is the covariance matrix of the noise vector at the destination during the second slot.

For the conventional A&F approach, the relay terminal retransmits the signal as it arrives, giving an equal share of the power to every channel eigenmode. Therefore, matrix $\mathbf{G}$ is given by $\mathbf{G} = \alpha\mathbf{I}_R$, where α is a proportionality factor to fulfill the relay power restriction. The problem addressed in the following is the design of a different gain matrix at the relay, when several receive/transmit dimensions are available, in order to maximize the instantaneous mutual information under different levels of channel state information. First, we derive an upper and a lower bound for the mutual information.

B. Upper and Lower Bounds for Mutual Information

From the definitions of $\mathbf{H}$ and $\mathbf{R}_w$, the determinant in (4) can be rewritten as follows [more details are given in Appendix I; see (8), shown at the bottom of the page]. From Hadamard's inequality [7]

$$\begin{vmatrix} \mathbf{X}_{11} & \mathbf{X}_{12} \\ \mathbf{X}_{21} & \mathbf{X}_{22} \end{vmatrix} \leq |\mathbf{X}_{11}||\mathbf{X}_{22}| \quad (9)$$

(with equality if and only if $\mathbf{X}_{12} = \mathbf{0}$ and $\mathbf{X}_{21} = \mathbf{0}$, provided that both $\mathbf{X}_{11}$ and $\mathbf{X}_{22}$ are both positive definite matrices), it follows, from (4) and (8), that the mutual information of the nonregenerative relaying scheme is upper bounded by

$$\begin{aligned} I(\mathbf{y};\mathbf{x}) \leq \frac{1}{2}\log_2\Big(&\left|\mathbf{I}_N + \mathbf{H}_0\mathbf{R}_x\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1}\right| \\ &\times \left|\mathbf{I}_N + \mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H \right. \\ &\left. \times \left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}\right|\Big) \end{aligned} \quad (10)$$

with equality if and only if

$$\begin{aligned} \mathbf{H}_0\mathbf{R}_x\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1} &= \mathbf{0} \\ \mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1} &= \mathbf{0}. \end{aligned} \quad (11)$$

The upper bound, therefore, is not achievable in practice, since it requires the cascaded relaying channel (composed by $\mathbf{H}_1$, $\mathbf{G}$, and $\mathbf{H}_2$) and the direct channel ($\mathbf{H}_0$) to be orthogonal.

In order to find a lower bound, we first expand the determinant of the block matrix as follows:

$$\begin{vmatrix} \mathbf{X}_{11} & \mathbf{X}_{12} \\ \mathbf{X}_{21} & \mathbf{X}_{22} \end{vmatrix} = |\mathbf{X}_{22}|\left|\mathbf{X}_{11} - \mathbf{X}_{12}\mathbf{X}_{22}^{-1}\mathbf{X}_{21}\right|. \quad (12)$$

After some matrix manipulation of the argument of the second determinant in (12), the mutual information can be written as follows [see Appendix I for more details; see (13), shown at the bottom of the page]. In Appendix II, it is shown that matrix $\mathbf{R}_{w_0}^{-1/2}\mathbf{H}_0(\mathbf{R}_x^{-1} + \mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2})^{-1}\mathbf{H}_2\mathbf{G}\mathbf{H}_1)^{-1}\mathbf{H}_0^H\mathbf{R}_{w_0}^{-H/2}$ can be decomposed as $\mathbf{Q}\mathbf{\Lambda}\mathbf{Q}^H$ where $\mathbf{Q}$ is a unitary matrix and $\mathbf{\Lambda}$ is diagonal with real nonnegative values. Therefore, the second determinant in (13) can be written as

$$\begin{aligned} |\mathbf{I}_N + \mathbf{Q}\mathbf{\Lambda}\mathbf{Q}^H| &= |\mathbf{Q}\mathbf{Q}^H + \mathbf{Q}\mathbf{\Lambda}\mathbf{Q}^H| \\ &= \left|\mathbf{Q}(\mathbf{I}_N + \mathbf{\Lambda})\mathbf{Q}^H\right| = |\mathbf{I}_N + \mathbf{\Lambda}| \geq 1 \end{aligned} \quad (14)$$

and the logarithm of (14) is nonnegative. This physically makes sense, since the second determinant in (13) represents the gain due to cooperation, instead of using pure relaying. As a consequence, the mutual information in (13) is lower bounded by

$$\begin{aligned} I_R = \frac{1}{2}\log_2\Big|&\mathbf{I}_N + \mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H \\ &\times \left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}\Big|. \end{aligned} \quad (15)$$

The lower bound is, therefore, the mutual information in the noncooperative relaying channel, that is, when the term $\mathbf{y}_0$ is not present in (3).

From now on, we will assume omnidirectional transmission from the source. The reason is that for decentralized schemes the information should arrive at the destination and also at potential relay stations about which the source has no *a priori* information. We will assume, also, that the source terminal is constrained in its average transmission power P_{ST}. Therefore, the covariance matrix of the transmitted signal is $\mathbf{R}_x = E\{\mathbf{x}\mathbf{x}^H\} = (P_{\mathrm{ST}}/M)\mathbf{I}_M$.

$$\begin{vmatrix} \mathbf{I}_N + \mathbf{H}_0\mathbf{R}_x\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1} & \mathbf{H}_0\mathbf{R}_x\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1} \\ \mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1} & \mathbf{I}_N + \mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1} \end{vmatrix} \quad (8)$$

$$\begin{aligned} I(\mathbf{y};\mathbf{x}) = \frac{1}{2}\log_2\Big\{&\left|\mathbf{I}_N + \mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}\right| \\ &\times\left|\mathbf{I}_N + \mathbf{R}_{w_0}^{-1/2}\mathbf{H}_0\left(\mathbf{R}_x^{-1} + \mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}\mathbf{H}_2\mathbf{G}\mathbf{H}_1\right)^{-1}\mathbf{H}_0^H\mathbf{R}_{w_0}^{-H/2}\right|\Big\} \end{aligned} \quad (13)$$

III. Optimum Transceiver for the Purely Relaying Channel Case

In this section, we focus on the design of matrix $\mathbf{G}$ that maximizes I_R in (15), when the relay is constrained in its average transmission power $P_{\rm RT}$. That is, we maximize the instantaneous mutual information of the noncooperative relaying channel. Only the channels $\mathbf{H}_1$ and $\mathbf{H}_2$ are known at the relay terminal. As mentioned in Section II, this maximizes the lower bound on the mutual information for the cooperative relaying channel, when the direct channel is unknown. The relay power constraint is expressed as a trace constraint on the covariance matrix of the signal transmitted by the relay, that is

$$\mathbf{R}_{Gy_1} = E\left[\mathbf{G}\mathbf{y}_1\mathbf{y}_1^H\mathbf{G}^H\right] = \mathbf{G}\left[\mathbf{H}_1\mathbf{R}_x\mathbf{H}_1^H + \mathbf{R}_{w_1}\right]\mathbf{G}^H. \tag{16}$$

At this point, in order to leave the formulation and solution as general as possible, we do not impose any structure on $\mathbf{H}_0$, $\mathbf{H}_1$, and $\mathbf{H}_2$. In the ensuing sections, we will apply the solutions to particular transmit/receive structures and propagation characteristics.

Considering $\mathbf{R}_x = E\{\mathbf{x}\mathbf{x}^H\} = (P_{\rm ST}/M)\mathbf{I}_M$ and $\mathbf{R}_{w0} = \mathbf{R}_{w2} = \sigma^2\mathbf{I}_N; \mathbf{R}_{w1} = \sigma^2\mathbf{I}_R$, (15) can be written as

$$I_R = \frac{1}{2}\log_2\left|\mathbf{I}_N + \frac{P_{\rm ST}}{M\sigma^2}\mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H \times\left(\mathbf{I}_N + \mathbf{H}_2\mathbf{G}\mathbf{G}^H\mathbf{H}_2^H\right)^{-1}\right|. \tag{17}$$

The function we want to maximize can be written in terms of the MMSE matrix for the pure relaying case as follows:

$$I_R = -\frac{1}{2}\log_2|\mathbf{M}| \tag{18}$$

since for this case, the MMSE is (see Appendix III, where MMSE receivers are derived)

$$\mathbf{M} = \left(\mathbf{I}_N + \mathbf{R}_x^{1/2}\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H \times\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}\mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x^{H/2}\right)^{-1}. \tag{19}$$

Different criteria may be adopted to derive the optimum expression for the transceiving matrix $\mathbf{G}$ at the relay, each leading to a different optimizing strategy. Many of these criteria depend on the entries of the matrix $\mathbf{M}$ as a function of $\mathbf{G}$: $\mathbf{M}(\mathbf{G})$. In particular, when the objective function $f(.)$ is Schur-concave, the following bound is used to minimize the function

$$f\left(\boldsymbol{\lambda}\left(\mathbf{M}(\mathbf{G})\right)\right) \le f\left(\mathbf{d}\left(\mathbf{M}(\mathbf{G})\right)\right) \tag{20}$$

where $\boldsymbol{\lambda}(\mathbf{M}(\mathbf{G}))$ and $\mathbf{d}(\mathbf{M}(\mathbf{G}))$ denote the vectors containing the eigenvalues and diagonal elements of $\mathbf{M}(\mathbf{G})$, respectively, in decreasing order. Therefore, from (20), it turns out that the minimum is reached when the elements in the diagonal of $\mathbf{M}(\mathbf{G})$ are the eigenvalues of $\mathbf{M}(\mathbf{G})$ [4, Ch. 6]. This means that the optimum transceiving matrix is such that $\mathbf{M}(\mathbf{G})$ is diagonal.

One example of Schur-concave optimization is the maximization of (17). According to this criterion, $\mathbf{G}$ must be such that

$$\mathbf{G} = \arg\min_{\mathbf{G}}\left\{\log_2|\mathbf{M}(\mathbf{G})|\right\}. \tag{21}$$

Following (20), the matrix $\mathbf{G}$ that turns $\mathbf{M}(\mathbf{G})$ into a diagonal matrix is given by

$$\mathbf{G} = \mathbf{V}_{d,2}\mathbf{P}_2\tilde{\mathbf{G}}\mathbf{P}_1^H\mathbf{U}_{d,1}^H = \mathbf{V}_2\tilde{\mathbf{G}}\mathbf{U}_1^H \tag{22}$$

where $\tilde{\mathbf{G}}$ is a diagonal matrix, $\mathbf{P}_i$ are arbitrary permutation matrices and $\mathbf{U}_d$, $\mathbf{V}_d$ are left and right eigenvectors of the matrix channels, when the channel eigenvalues are in decreasing order

$$\begin{aligned}\mathbf{H}_1 &= \mathbf{U}_{d,1}\boldsymbol{\Lambda}_{d,1}^{1/2}\mathbf{V}_{d,1}^H = \mathbf{U}_{d,1}\mathbf{P}_1\mathbf{P}_1^H\boldsymbol{\Lambda}_{d,1}^{1/2}\mathbf{P}_1\mathbf{P}_1^H\mathbf{V}_{d,1}^H\\ &= \mathbf{U}_1\boldsymbol{\Lambda}_1^{1/2}\mathbf{V}_1^H\\ \mathbf{H}_2 &= \mathbf{U}_{d,2}\boldsymbol{\Lambda}_{d,2}^{1/2}\mathbf{V}_{d,2}^H = \mathbf{U}_{d,2}\mathbf{P}_2\mathbf{P}_2^H\boldsymbol{\Lambda}_{d,2}^{1/2}\mathbf{P}_2\mathbf{P}_2^H\mathbf{V}_{d,2}^H\\ &= \mathbf{U}_2\boldsymbol{\Lambda}_2^{1/2}\mathbf{V}_2^H.\end{aligned} \tag{23}$$

Without loss of generality and for the purpose of optimizing the mutual information, the permutation matrix $\mathbf{P}_1$ can be taken as the identity, as can be verified in (17). Given this, the optimization turns into both the determination of the diagonal elements of $\tilde{\mathbf{G}}$, under a power constraint at the relay station, and the choice of the permutation matrix $\mathbf{P}_2$ maximizing (17).

Let us now obtain the optimum elements of $\tilde{\mathbf{G}}$. Using the singular value decomposition (SVD) of the channels in (23), (17) can be rewritten as follows:

$$I_R = \log_2\left|\mathbf{I}_N + \frac{P_{\rm ST}}{M\sigma^2}\mathbf{U}_2\boldsymbol{\Lambda}_2^{1/2}\mathbf{V}_2^H\mathbf{G}\mathbf{U}_1\boldsymbol{\Lambda}_1\mathbf{U}_1^H\mathbf{G}^H\mathbf{V}_2 \times\boldsymbol{\Lambda}_2^{1/2}\mathbf{U}_2^H\left(\mathbf{I}_N + \mathbf{U}_2\boldsymbol{\Lambda}_2^{1/2}\mathbf{V}_2^H\mathbf{G}\mathbf{G}^H\mathbf{V}_2\boldsymbol{\Lambda}_2^{1/2}\mathbf{U}_2^H\right)^{-1}\right|. \tag{24}$$

Note that the solution given above for $\mathbf{G}$ in (22) is consistent with the Hadamard determinant theorem [6] that establishes that the matrix within the determinant should be diagonal to maximize (24). By including (22) in (24) and applying the inverse lemma to the noise matrix and the commutative property of the determinant, we arrive at

$$I_R = \log_2\left|\mathbf{I}_R + \frac{P_{\rm ST}}{M\sigma^2}\boldsymbol{\Lambda}_2^{1/2}\tilde{\mathbf{G}}\boldsymbol{\Lambda}_1\tilde{\mathbf{G}}^H\boldsymbol{\Lambda}_2^{1/2} \times\left(\mathbf{I}_R + \boldsymbol{\Lambda}_2^{1/2}\tilde{\mathbf{G}}\tilde{\mathbf{G}}^H\boldsymbol{\Lambda}_2^{1/2}\right)^{-1}\right|. \tag{25}$$

The restriction under the relay power can also be rewritten as a function of $\tilde{\mathbf{G}}$ as follows:

$$\mathrm{trace}\left(\frac{P_{\rm ST}}{M}\tilde{\mathbf{G}}\Lambda_1\tilde{\mathbf{G}}^H + \sigma^2\tilde{\mathbf{G}}\tilde{\mathbf{G}}^H\right) \le P_{\rm RT}. \tag{26}$$

Then, to compute the elements of diagonal matrix $\tilde{\mathbf{G}}$ we need to solve the following scalar problem, where the unknowns are $|g_r|^2$ for $r = 1,\ldots,R$

$$\begin{aligned}&\text{Maximize } \sum_{r=1}^{R}\log_2\left(1 + \frac{P_{\rm ST}}{M\sigma^2}\frac{\lambda_{1,r}\lambda_{2,r}|g_r|^2}{1+\lambda_{2,r}|g_r|^2}\right)\\ &\text{subject to } \quad |g_r|^2 \ge 0 \quad r = 1,\ldots,R\\ &\qquad\qquad \sum_{r=1}^{R}\left(\frac{P_{\rm ST}}{M}\lambda_{1,r} + \sigma^2\right)|g_r|^2 = P_{\rm RT}\end{aligned} \tag{27}$$

with $\lambda_{1,r}$ and $\lambda_{2,r}$ are the rth eigenvalues of the first and second hop channels, respectively, and g_r is the rth diagonal component of matrix $\tilde{\mathbf{G}}$. Note that, if the power of the relay could be increased without bound, the mutual information would be limited by the eigenvalues of the first hop channel. The permutation matrix $\mathbf{P}_2$ is implicitly included in the ordering of $\lambda_{2,r}$, and it also needs to be optimized.

The problem in (27) is a standard concave optimization problem (the objective function and the inequality constraint functions are concave, while the equality constraint function is affine with respect to $|g_r|^2$), which can be solved by means of the Karush-Kuhn-Tucker conditions [5] to obtain the optimum value for $|g_r|^2$ $r = 1, \ldots, R$. After some tedious calculations, the optimum solution for the rth component in the diagonal matrix $\tilde{\mathbf{G}}$ is shown to be the square root of

$$|g_r|^2 = \frac{p_r}{(P_{\mathrm{ST}}/M)\lambda_{1,r} + \sigma^2} \tag{28}$$

with p_r the power assigned by the relay terminal to each channel eigenmode

$$p_r = \left[\sqrt{\mu\frac{P_{\mathrm{ST}}\lambda_{1,r}}{M\lambda_{2,r}} + \left(\frac{P_{\mathrm{ST}}\lambda_{1,r}}{2M\lambda_{2,r}}\right)^2} - \frac{P_{\mathrm{ST}}\lambda_{1,r}}{2M\lambda_{2,r}} - \frac{\sigma^2}{\lambda_{2,r}}\right]^+ \tag{29}$$

where μ is a common constant to fulfill the relay power restriction.

This optimal solution for conventional relaying schemes is a suboptimal solution for cooperative schemes, as it may be obtained through the maximization of the lower bound in (15). The solution finally obtained [linear precoding matrix given by (22) with the elements of the diagonal matrix diagonal matrix $\tilde{\mathbf{G}}$ computed as the square root of (28)] is actually quite intuitive. To maximize mutual information, the relay station should first perform a filter operation on the received signal, matched to the first hop channel eigenmodes. Then, the first hop channel eigenmodes are retransmitted having been matched to the second hop channel eigenmodes. The optimal gain for each mode retransmission in (28) and (29) indicates that the "bad" modes of the second hop channel must be penalized by means of the last term proportional to $(-\sigma^2/\lambda_{2,r})$. This term also appears in the conventional waterfilling solution (see, for instance, the optimal power assignment in a conventional MIMO system [16], when CSI is available at the transmitter). Note, that the sum of the other three terms is always greater than or equal to zero and enhances those modes of the first hop channel whose ratio, with respect to the corresponding mode in the second hop channel, is greater.

For those null channel eigenvalues (with independence if they are from the first or second hop channel), the assigned power will be zero. This means that for $N = 1$ (with M and $R > 1$), as could be intuitively expected, the relay should retransmit only the best first hop eigenmode. Then, all the available relay power is assigned to this eigenmode, which is retransmitted using a matched filter to the single second hop channel eigenmode.

When there are several nonzero eigenvalues in both hops, we need to compute the best assignment between the modes of $\mathbf{H}_1$ and $\mathbf{H}_2$ through the permutation matrix $\mathbf{P}_2$, since the ordering can impact in the final overall mutual information performance. From lemma 1 in Appendix IV, it follows that for optimum performance, $\lambda_{1,r}$ and $\lambda_{2,r}|g_r|^2$ for $r = 1, \ldots, R$, must be ordered in the same way in order to maximize mutual information. Nevertheless, while the term $\lambda_{2,r}|g_r|^2$ increases monotonically with respect to $\lambda_{2,r}$, this is not the case for $\lambda_{1,r}$ [see (30)]

$$\lambda_{2,r}|g_r|^2 = \frac{1}{(P_{\mathrm{ST}}/M)\lambda_{1,r} + \sigma^2} \times \left[\sqrt{\mu\frac{P_{\mathrm{ST}}\lambda_{1,r}\lambda_{2,r}}{M} + \left(\frac{P_{\mathrm{ST}}\lambda_{1,r}}{2M}\right)^2} - \frac{P_{\mathrm{ST}}\lambda_{1,r}}{2M} - \sigma^2\right]^+. \tag{30}$$

In particular, for high SNR_1 values, (30) will decrease with $\lambda_{1,r}$. For those cases, where $\lambda_{2,r}|g_r|^2$ decreases with $\lambda_{1,r}$, the optimization problem being considered turns into both the determination of the diagonal elements of $\tilde{\mathbf{G}}$, under a power constraint at the relay station, and also the best choice of the permutation matrix for the second hop eigenvalues that maximize (25).

IV. Optimum Solution When Direct Channel and Relaying Channel Known at the Relay Station

In Section III, a procedure to maximize mutual information making use of the knowledge about $\mathbf{H}_1$ and $\mathbf{H}_2$ was proposed. Despite the solution being the optimal one for conventional relaying channels, it does not maximize the mutual information for cooperative schemes but a lower bound of the mutual information in such systems. When knowledge about the direct channel is available, maximization of mutual information itself can be accomplished. Despite the results obtained being interesting and useful as a benchmark, in practice, having knowledge about $\mathbf{H}_0$ at the relay station may pose some practical problems.

By expanding the determinant of the block matrix in (4) using

$$\begin{vmatrix} \mathbf{X}_{11} & \mathbf{X}_{12} \\ \mathbf{X}_{21} & \mathbf{X}_{22} \end{vmatrix} = |\mathbf{X}_{11}|\left|\mathbf{X}_{22} - \mathbf{X}_{21}\mathbf{X}_{11}^{-1}\mathbf{X}_{12}\right| \tag{31}$$

an exact expression for mutual information can be obtained

$$I_F(\mathbf{y};\mathbf{x}) = \frac{1}{2}\log_2\left|\mathbf{I}_N + \frac{P_{\mathrm{ST}}}{M\sigma^2}\mathbf{H}_0\mathbf{H}_0^H\right| + I \tag{32}$$

where the subscript F stands for full channels knowledge, and I is given by

$$I = \frac{1}{2}\log_2\left|\mathbf{I}_N + \frac{P_{\mathrm{ST}}}{M\sigma^2}\mathbf{H}_2\mathbf{G}\mathbf{H}_1\left(\mathbf{I}_N + \frac{P_{\mathrm{ST}}}{M\sigma^2}\mathbf{H}_0\mathbf{H}_0^H\right)^{-1} \times\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\left(\mathbf{I}_N + \mathbf{H}_2\mathbf{G}\mathbf{G}^H\mathbf{H}_2^H\right)^{-1}\right|. \tag{33}$$

In the previous equations, it has been assumed that $\mathbf{R}_x = (P_{\mathrm{ST}}/M)\mathbf{I}_M$, $\mathbf{R}_{w_0} = \mathbf{R}_{w_2} = \sigma^2\mathbf{I}_N$ and $\mathbf{R}_{w_1} = \sigma^2\mathbf{I}_R$.

Note that maximizing mutual information with respect to $\mathbf{G}$ is equivalent to maximizing I. Conceptually, the solution to the maximization of (33) is similar to the one obtained in the previous section. Nevertheless, if information about the direct hop channel $\mathbf{H}_0$ is available, the optimal solution uses, instead of the eigenvalues of the first hop channel $\mathbf{H}_1$, the eigenvalues of the equivalent prefiltered channel $\mathbf{H}_1(\mathbf{I}_N+(P_{\rm ST}/M/\sigma^2)\mathbf{H}_0\mathbf{H}_0^H)^{-1/2}$. Note that the eigenvalues λ_1 in (28) will now be the eigenvalues of

$$\mathbf{H}_1\left[\mathbf{I}_M-\mathbf{H}_0^H\left(\mathbf{I}_N+\frac{P_{\rm ST}}{M\sigma^2}\mathbf{H}_0\mathbf{H}_0^H\right)^{-1}\mathbf{H}_0\frac{P_{\rm ST}}{M\sigma^2}\right]\mathbf{H}_1^H \tag{34}$$

obtained by applying the matrix inversion lemma to $\mathbf{H}_1(\mathbf{I}_N+(P_{\rm ST}/M/\sigma^2)\mathbf{H}_0\mathbf{H}_0^H)^{-1}\mathbf{H}_1^H$.

When the SNR in the direct hop channel (SNR_0) is high, the matrix in brackets can be approximated by the matrix that projects onto the subspace orthogonal to the direct hop channel $\mathbf{P}_\perp=\mathbf{I}_M-\mathbf{H}_0^H(\mathbf{H}_0\mathbf{H}_0^H)^{-1}\mathbf{H}_0$.

On the contrary, for low SNR_0, the optimal linear gain will be the solution obtained in Section III; that is the solution for the relay station in pure relaying schemes.

V. Applications

The solution given in the previous section has been obtained using a general setting. The solution will now be particularized to specific operative conditions. In this section, we consider two different physical environments in order to provide results for the proposed solutions. First, a flat fading multiple antenna system is considered. Second, a SISO frequency selective system is assumed.

The factor 1/2 in (4) can be dropped if a high spatial reuse of the second slot is used [1], [14]. Spatial reuse means the following. Let us assume a TDMA strategy for the downlink (DL), with the base station serving K users. First, the K DL transmissions are allocated. At the end of the frame, K' simultaneous retransmissions from the corresponding relays are allocated in a single slot, with $K'\le K$. Therefore, under a multiuser TDMA scenario the effective mutual information of a single cooperative connection has to be multiplied by a factor $K/(K+1)$, instead of a factor 1/2, corresponding to the relay slot nonreuse case. This amounts to saying that, in a wireless communication scenario, if the radio-resource management strategy allows K to be high enough, the term $K/(K+1)$ approaches 1. For those reasons, the factor 1/2 in (4) has been removed and the results have to be carefully interpreted. With the omission of the factor 1/2, we show what the multiuser benefits of such a system would be if a perfect spatial reuse of the relay slot were possible. To that end, a suitable radio resource management strategy needs to be setup in order to achieve reduced levels of interference (see [1] and [2] for practical approaches).

A. Results for Flat Fading Multiple Antenna Systems

The proposed approach has been compared in terms of ergodic and outage mutual information with other strategies such as noncooperative and cooperative approaches in flat fading multiple antenna systems. For cooperative approaches, both conventional nonregenerative (A&F) and regenerative relaying schemes (D&F-Unconstrained Coding [9]) have been considered. The simulations have been carried out assuming flat fading Rayleigh channels in all links. Different situations

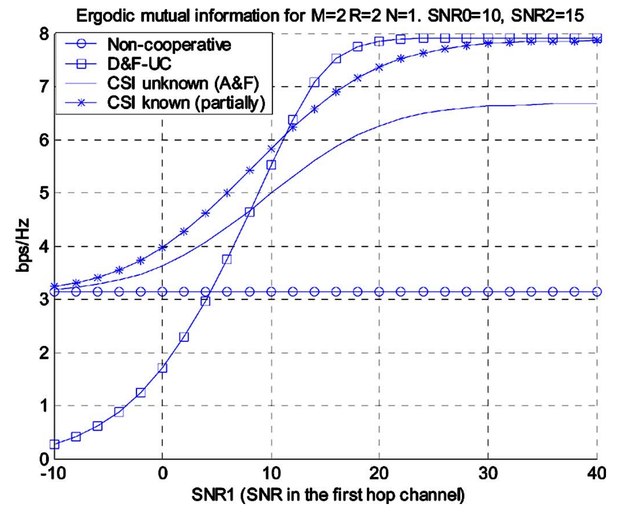

Fig. 3. Ergodic mutual information [bps/Hz] as a function of the mean SNR_1 (SNR in the first hop channel), for $M=2$, $R=2$, $N=1$. $\mathrm{SNR}_0=10$ dB, $\mathrm{SNR}_2=15$ dB.

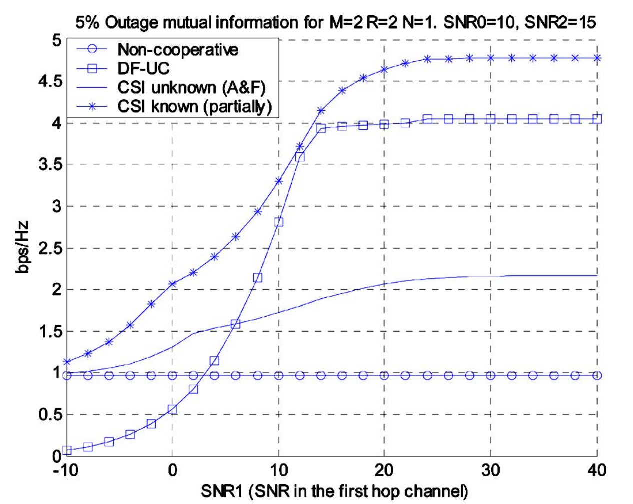

Fig. 4. 5% Outage mutual information information [bps/Hz] as a function of the mean SNR_1 (SNR in the first hop channel), for $M=2$, $R=2$, $N=1$. $\mathrm{SNR}_0=10$ dB, $\mathrm{SNR}_2=15$ dB.

relating to mean SNR at the involved links have been studied. The mean SNRs are defined as follows:

$$\mathrm{SNR}_0=\frac{P_{\rm ST}}{\sigma^2L_0};\quad \mathrm{SNR}_1=\frac{P_{\rm ST}}{\sigma^2L_1};\quad \mathrm{SNR}_2=\frac{P_{\rm RT}}{\sigma^2L_2} \tag{35}$$

with L_0, L_1, and L_2 the pathloss for the direct channel, the first hop channel and the second hop channel, respectively. Channel components (from one transmitting antenna to one receiving antenna) are generated as i.i.d zero-mean complex Gaussian variables, with a variance according to the path loss of the channel.

Figs. 3 and 4 show the ergodic mutual information and 5% outage mutual information, respectively, versus the mean SNR in the first hop channel (SNR_1) for noncooperative transmission (non-Coop), cooperative decode and forward-unconstrained code (D&F-UC), cooperative conventional A&F approach, and the proposed scheme: cooperative A&F with partial CSI. For this latter scheme, the gain matrix at the relay station has been obtained by maximizing the lower bound of the instantaneous

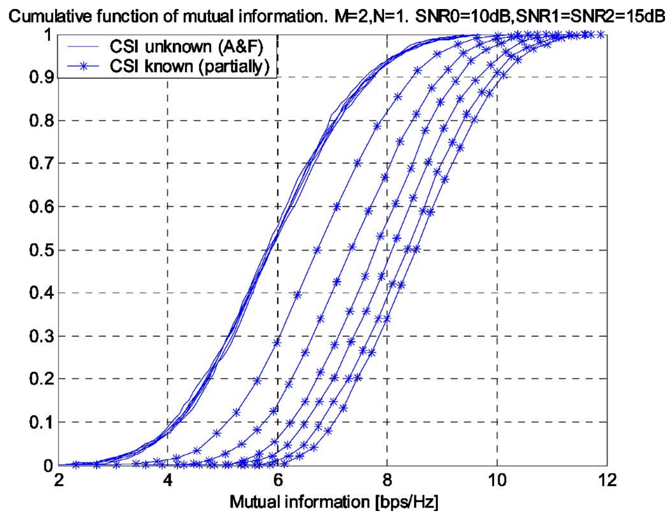

Fig. 5. Cumulative function of mutual information for conventional A&F (black lines) and suboptimal proposed approach for $M = 2$, $N = 1$, $R = 2, 3, 4, 5, 6, 7$. $\mathrm{SNR}_0 = 10$ dB, $\mathrm{SNR}_1 = \mathrm{SNR}_2 = 15$ dB.

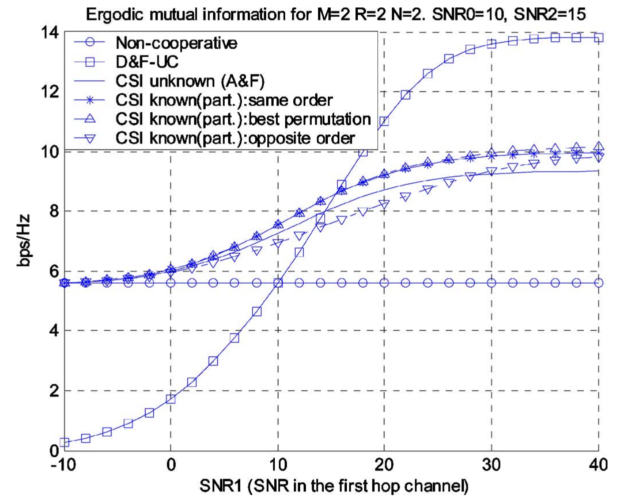

Fig. 6. Ergodic mutual information versus the mean SNR_1, for $M = 2$, $R = 2$, $N = 2$. $\mathrm{SNR}_0 = 10$ dB, $\mathrm{SNR}_2 = 15$ dB. When CSI is partially known, channel eigenvalues of the first and second hop can be matched in either the same decreasing order, opposite decreasing order or according to the optimum permutation matrix.

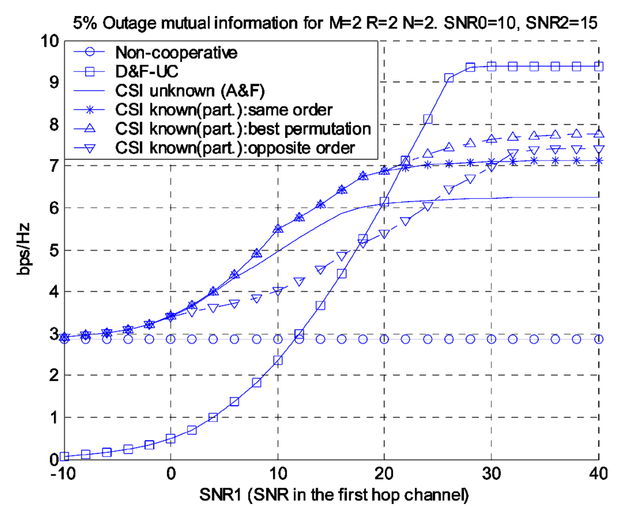

Fig. 7. 5% Outage mutual information versus the mean SNR_1, for $M = 2$, $R = 2$, $N = 2$. $\mathrm{SNR}_0 = 10$ dB, $\mathrm{SNR}_2 = 15$ dB. When CSI is partially known, channel eigenvalues of the first and second hop are matched in either the same decreasing order, opposite decreasing order or according to the optimum permutation matrix.

mutual information, but the results shown correspond to the actual values of the mutual information so obtained. The signal to noise ratio in the direct channel (SNR_0) is fixed to 10 dB and the SNR in the second hop channel is $\mathrm{SNR}_2 = 15$ dB. Regarding the number of antennas, $M = 2$, $R = 2$, and $N = 1$ antennas are considered at the source, the relay and the destination, respectively. The D&F approach needs to decode the symbols correctly at the relay terminal. Therefore, for low SNR_1 (first hop channel), the cooperative D&F-UC may offer an even worse performance than the noncooperative approach. On the other hand, nonregenerative cooperative relaying (conventional A&F and the proposed approach A&F with partial CSI) always offers a better performance than noncooperative transmission, even for low SNR_1. Nevertheless, while conventional cooperative A&F saturates to a mutual information value smaller than the cooperative D&F, the new approach converges to the cooperative D&F performance in terms of ergodic mutual information (Fig. 3). In terms of outage mutual information, the proposed approach can further improve the performance of the cooperative D&F-UC and can double the performance of the conventional cooperative A&F (Fig. 4), for SNR_1 values greater than 10 dB.

Fig. 5 shows the cumulative function of mutual information for the conventional cooperative A&F and the mutual information achieved by the proposed approach for a different number of antennas at the relay station considering only one antenna at the final destination. As was expected, for $N = 1$ and $R > 1$, mutual information can be increased with nonregenerative relays using some further processing, rather than using conventional A&F procedure. Note also that the performance gap increases when the number of relay antennas increases.

When the number of antennas at the relay station and the final destination is the same, there is a reduction in the improvement obtained from the use of CSI compared with conventional cooperative A&F (see Fig. 6 and 7), in terms of ergodic mutual information and outage mutual information. This comes from the fact that the eigenvalue dispersion reduces (there are no zero eigenvalues in the second hop channel). As a consequence, there is a reduction in the impact of giving an equal power to all these eigenvalues.

For $N = 2$ antennas at the final destination, we have two nonzero eigenvalues in the first hop channel and in the second hop channel. The SVD channel decomposition raises a question regarding the ordering of the channel eigenvalues, as explained in Section III. In Fig. 6 and 7, we have considered that the first and second hop eigenvalues are taken in decreasing order. We compare this approach with the performance obtained when the permutation matrix that maximizes (25) is introduced in the solution. The optimal permutation matrix depends on the particular values of each channel realization. Therefore, for some particular channel realizations, the optimal order is that which assigns the greatest first hop channel eigenvalues to the greatest

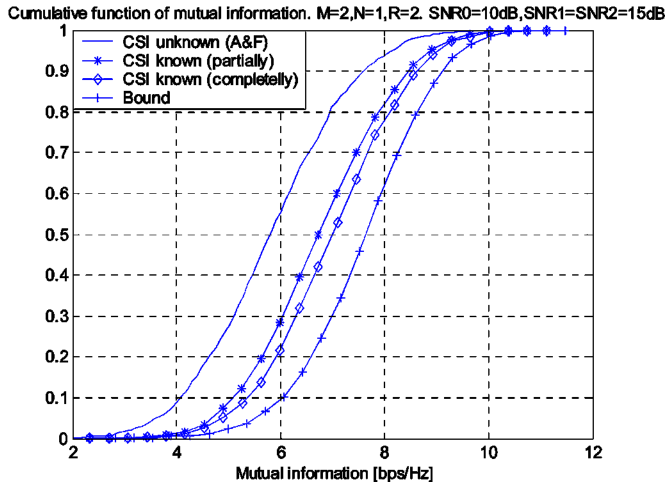

Fig. 8. Cumulative function of mutual information of A&F scheme for $\mathbf{H}_0$ unknown, $\mathbf{H}_0$ known (optimal approach) and upper bound ($\mathbf{H}_0$ and relaying channel are orthogonal). $M = 2, N = 1, R = 2$. $\text{SNR}_0 = 10$ dB, $\text{SNR}_1 = \text{SNR}_2 = 15$ dB.

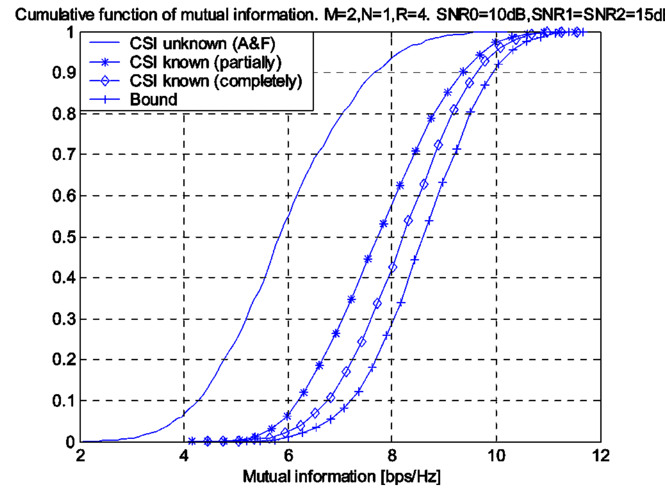

Fig. 9. Cumulative function of mutual information of A&F scheme, suboptimal proposed approach ($\mathbf{H}_0$ unknown), optimal approach ($\mathbf{H}_0$ known) and upper bound ($\mathbf{H}_0$ and relaying channel are orthogonal). $M = 2, N = 1, R = 4$. $\text{SNR}_0 = 10$ dB, $\text{SNR}_1 = \text{SNR}_2 = 15$ dB.

second hop channel eigenvalues, while for other channel realizations this is not the optimal approach. For completeness, we have also considered the performance obtained when the channel eigenvalues of both hops are in the opposite order. This solution is always worse than the other two, in terms of ergodic mutual information. In terms of outage mutual information, this solution is far from the optimal one, with the exception of extremely high SNR_1. Ordering the channel eigenvalues of the first and second hop channels in the same order coincides with the optimal ordering, until SNR_1 is greater than 20 dB. However, even for SNR_1 values greater than 20 dB the performance is close to the optimum one (optimum over the set of permutation matrices). Therefore, given that the number of permutation matrices increases as $N!$, for large values of N, a practical solution is to match channel eigenvalues of the first and second hop channels in the same decreasing order.

Let us now consider the results obtained when the cooperative channel is completely known. That is, the relay station has knowledge of the three channels involved in the cooperative communication: direct and first and second hop channels. The cumulative function of the instantaneous mutual information has been depicted for $R = 2$ and $R = 4$ in Figs. 8 and 9, respectively, and for $M = 2$ and $N = 1$ antennas at the source and the end user, respectively. If we compare the results obtained with this solution and the ones obtained when no information about $\mathbf{H}_0$ is available, it can be observed that there is a slight loss due to the lack of CSI about $\mathbf{H}_0$ (see Fig. 8). This loss grows as the number of relay antennas increases (see Fig. 9).

The cumulative function of the instantaneous mutual information upper bound has also been depicted in Figs. 8 and 9. The upper bound has been computed using $\mathbf{G}$ that maximizes (33), as was explained in Section IV when $\mathbf{H}_0$ is available at the relay station. Regarding this bound, it is interesting to note the following two observations. First, the bound is not achievable, since it requires the cascaded relaying channel (including first hop channel, linear gain matrix at the relay, and second hop channel) and the direct channel to be orthogonal. Second, if there are, however, enough degrees of freedom at the relay station (increasing the receiving/transmitting dimensions R), the achievable performance approaches the bound. This can be interpreted as if, with enough degrees of freedom, we are able to design a gain matrix $\mathbf{G}$ that orthogonalizes the cascaded relaying channel and the direct channel, and still achieves a high mutual information in the subspace orthogonal to the direct hop channel $\mathbf{H}_0$.

B. Results for SISO Frequency Selective Channels

We have also considered frequency-selective time-invariant channels, modeled as finite impulse response (FIR) filters of maximum order $L-1$. Using a block transmission with a cyclic prefix (CP), the channel can be modeled as an $N \times N$ circulant Toeplitz matrix which is diagonalized by the discrete Fourier transform (DFT) matrix. Therefore, the optimal coding strategy for linear time invariant (LTI) SISO channels is OFDM, with proper power/bit allocation across the subcarriers. The optimization problem again turns into both the computation of the diagonal elements of $\tilde{\mathbf{G}}$, under a power constraint at the relay station, and also the choice of the permutation matrix for the second hop eigenvalues that maximize (25). The power allocation will be given by (29), with the channel eigenvalues given by the discrete Fourier transform (DFT) of the FIR first hop and second hop channels. Taking, for instance, a DFT of $N = 64$ points, the choice of the permutation matrix turns into an extremely high computationally complex problem since there are $N!$ possible permutation matrices. Following the conclusions in the previous subsection, we have considered instead, as a practical solution, the channel eigenvalues of the first and second hop channels in the same decreasing order. Note that this solution implies that each bit can be sent in the second hop channel through a carrier which is different from the one used in the first hop channel. The results are reported in Figs. 10 and 11. The mutual information is measured in bps/Hz; therefore, we need to divide the expressions for the mutual information given in the paper by the number of carriers.

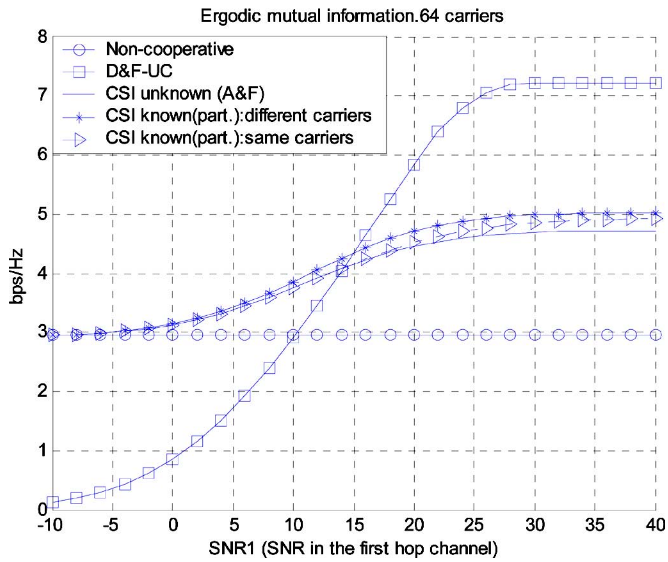

Fig. 10. Ergodic mutual information [bps/Hz] as a function of the mean SNR_1 (SNR in the first hop channel), for FIR channels with $L = 5$ taps and 64 carriers. $\mathrm{SNR}_0 = 10$ dB, $\mathrm{SNR}_2 = 15$ dB.

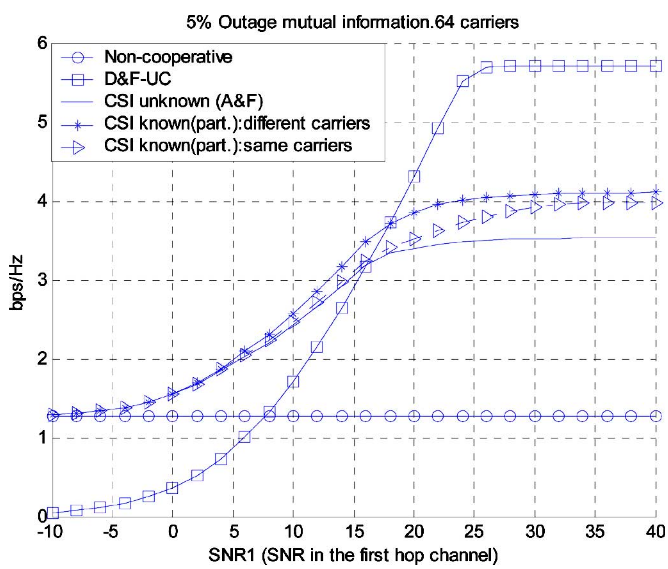

Fig. 11. Five percent outage mutual information [bps/Hz] as a function of the mean SNR_1 (SNR in the first hop channel), for FIR channels with $L = 5$ taps and 64 carriers. $\mathrm{SNR}_0 = 10$ dB, $\mathrm{SNR}_2 = 15$ dB.

Figs. 10 and 11 also depict the results obtained when each bit is carried in the second hop channel through the same carrier which is used in the first hop channel. As in previous figures, the results corresponding to direct transmission, cooperative D&F-UC, and conventional cooperative A&F are also included. Different situations relating to mean SNR at the involved links have been studied. The mean SNRs here are the mean SNRs per carrier. They are defined as follows:

$$\mathrm{SNR}_0 = \frac{P_{\mathrm{ST}}/N}{\sigma^2 L_0}; \quad \mathrm{SNR}_1 = \frac{P_{\mathrm{ST}}/N}{\sigma^2 L_1}; \quad \mathrm{SNR}_2 = \frac{P_{\mathrm{RT}}/N}{\sigma^2 L_2} \tag{36}$$

with L_0, L_1, and L_2 the pathloss for the direct channel, the first hop channel and the second hop channel, respectively. As can be observed from Figs. 10 and 11, by matching eigenvalues of the first and the second hop channels in the same decreasing order, previous to optimizing gain values, we can achieve a better performance (despite it is slight) than transmitting each bit through the same carrier in the first and second hop channels. The performance gain with respect to the conventional cooperative A&F approach is around 16% in terms of outage mutual information and around 6% in terms of ergodic mutual information for high SNR_1 values, $\mathrm{SNR}_0 = 10$ dB and $\mathrm{SNR}_2 = 15$ dB.

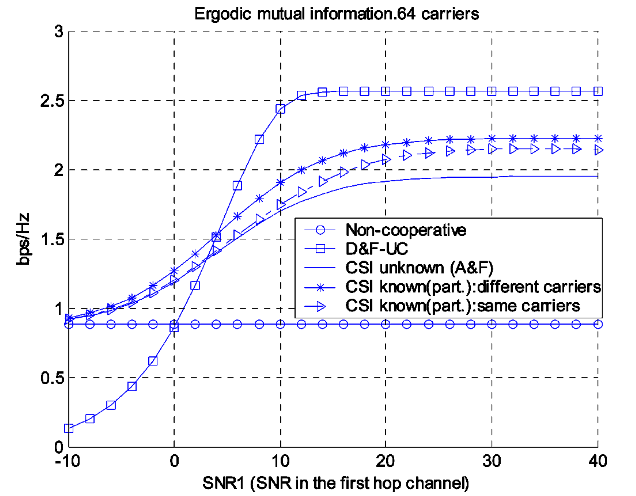

Fig. 12. Ergodic mutual information [bps/Hz] as a function of the mean SNR_1 (SNR in the first hop channel), for FIR channels with $L = 5$ taps and 64 carriers. $\mathrm{SNR}_0 = 0$ dB, $\mathrm{SNR}_2 = 5$ dB.

When $\mathrm{SNR}_2 = 5$ dB, the performance gain with respect to conventional cooperative A&F approach increases, as can be observed in Fig. 12. This behavior makes sense, since it is well known that waterfilling solutions offer mutual information improvements for low SNR. In Fig. 12, SNR_0 has also been lowered 10 dB, in order to better appreciate the performance due to cooperation. In this case, the performance gain with respect to the conventional cooperative A&F approach is around 13% in terms of ergodic mutual information for high SNR_1 values, $\mathrm{SNR}_0 = 0$ dB and $\mathrm{SNR}_2 = 5$ dB.

The channels in this section are simulated as FIR filters of six i.i.d zero-mean complex Gaussian taps, of equal variance. The block length is $N = 64$ and the channel transfer functions are computed by taking the 64-point DFT.

VI. Conclusion

In this paper, the use of CSI at the relay station has been considered for the optimum design of the gain matrix in nonregenerative cooperative schemes for a fixed power constraint. The optimal linear processing at the relay station has been found analytically, based on a lower bound of the total (cooperative) mutual information with an unknown direct (source-destination) channel. This lower bound turns out to be the mutual information for the pure relaying scheme (noncooperative). This solution can be implemented when the first and second hop channels are known at the relay station. Despite this solution not being the mutual information-maximizing solution for cooperative schemes, it lets us obtain a substantial increase in ergodic mutual information when some of the eigenvalues of the second hop channel are zero.

For SISO frequency selective systems, the proposed scheme using CSI at the relay station can also offer an improvement regarding conventional A&F scheme, when the bits are not necessarily transmitted through the same carriers in the first and second hop channels, thus making it possible to align the first and second hop channel modes in the same increasing/decreasing order.

Additionally, we have obtained the mutual information maximizing solution, which is only achievable when the direct hop channel is also known. Comparing the results obtained with this solution and the ones obtained when no information about $\mathbf{H}_0$ is available, it can be concluded that the loss due to the lack of CSI about $\mathbf{H}_0$ is not significant.

An upper bound for the mutual information of nonregenerative cooperative schemes has also been obtained. This bound is only achieved when both the direct and relaying channel (including first and second hop channels) are orthogonal. The solution obtained for the optimal linear processing when the direct channel is also known has shown that matrix $\mathbf{G}$ should perform a role to "orthogonalize" the direct and relaying channel. Such a solution approaches the upper bound as the number of relay receiving/transmitting dimensions increases.

Appendix I

The general expression for the mutual information is given by

$$I(\mathbf{y};\mathbf{x}) = \frac{1}{2}\log_2\left|\mathbf{I}_{2N} + \mathbf{H}\mathbf{R}_x\mathbf{H}^H\mathbf{R}_w^{-1}\right|. \quad (37)$$

From the definitions of $\mathbf{H}$ and $\mathbf{R}_w$, the determinant can be rewritten as shown in (38) and (39) at the bottom of the page. We can expand the determinant of the block matrix as follows:

$$\begin{vmatrix}\mathbf{X}_{11} & \mathbf{X}_{12}\\ \mathbf{X}_{21} & \mathbf{X}_{22}\end{vmatrix} = |\mathbf{X}_{22}|\left|\mathbf{X}_{11} - \mathbf{X}_{12}\mathbf{X}_{22}^{-1}\mathbf{X}_{21}\right|. \quad (40)$$

To simplify, we denote as $\mathbf{A}$ and $\mathbf{B}$ the following matrices:

$$\mathbf{A} = \mathbf{H}_2\mathbf{G}\mathbf{H}_1; \quad \mathbf{B} = \left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}. \quad (41)$$

From the definitions of $\mathbf{A}$ and $\mathbf{B}$, the second determinant is

$$\begin{aligned}
&\big|\mathbf{I}_N + \mathbf{H}_0\mathbf{R}_x\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1} \\
&\quad -\mathbf{H}_0\mathbf{R}_x\mathbf{A}^H\mathbf{B}(\mathbf{I}_N + \mathbf{A}\mathbf{R}_x\mathbf{A}^H\mathbf{B})^{-1}\mathbf{A}\mathbf{R}_x\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1}\big| \\
&\overset{(a)}{=} \big|\mathbf{I}_N + \mathbf{H}_0\mathbf{R}_x\left(\mathbf{I}_M - \mathbf{A}^H\mathbf{B}(\mathbf{I}_N + \mathbf{A}\mathbf{R}_x\mathbf{A}^H\mathbf{B})^{-1}\right. \\
&\qquad \left.\times \mathbf{A}\mathbf{R}_x\right)\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1}\big| \\
&\overset{(b)}{=} \left|\mathbf{I}_N + \mathbf{H}_0\mathbf{R}_x(\mathbf{I}_M + \mathbf{A}^H\mathbf{B}\mathbf{A}\mathbf{R}_x)^{-1}\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1}\right| \\
&\overset{(c)}{=} \big|\mathbf{I}_N + \mathbf{H}_0\mathbf{R}_x\left(\mathbf{I}_M + \mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\right. \\
&\qquad \times\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1} \\
&\qquad \left.\times\ \mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x\right)^{-1}\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1}\big| \\
&\overset{(d)}{=} \big|\mathbf{I}_N + \mathbf{R}_{w_0}^{-1/2}\mathbf{H}_0\mathbf{R}_x \\
&\qquad \times\big(\mathbf{I}_M + \mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H \\
&\qquad \times\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}\mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x\big)^{-1} \\
&\qquad \times\mathbf{H}_0^H\mathbf{R}_{w_0}^{-H/2}\big| \\
&\overset{(e)}{=} \big|\mathbf{I}_N + \mathbf{R}_{w_0}^{-1/2}\mathbf{H}_0 \\
&\qquad \times\big(\mathbf{R}_x^{-1} + \mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H \\
&\qquad \times\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}\mathbf{H}_2\mathbf{G}\mathbf{H}_1\big)^{-1} \\
&\qquad \times\mathbf{H}_0^H\mathbf{R}_{w_0}^{-H/2}\big| \qquad (42)
\end{aligned}$$

where the equality (a) comes from matrix factorization, (b) comes from the matrix inversion lemma, (c) comes from definitions for matrices $\mathbf{A}$ and $\mathbf{B}$, (d) comes from the commutative property of the determinant. In (e), we are considering that matrix $\mathbf{R}_x$ is full rank, and, therefore, it is invertible. This assumption can be taken without loss of generality, since we can always write expression (38) using a full rank matrix $\mathbf{R}_x$ with another equivalent channels $\mathbf{H}_0$ and $\mathbf{H}_1$ with reduced dimensions.

Appendix II

Let us illustrate that the following matrix is positive semidefinite

$$\begin{aligned}
&\mathbf{R}_{w_0}^{-1/2}\mathbf{H}_0\big(\mathbf{R}_x^{-1} + \mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H \\
&\qquad \times\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}\mathbf{H}_2\mathbf{G}\mathbf{H}_1\big)^{-1} \\
&\quad \times\mathbf{H}_0^H\mathbf{R}_{w_0}^{-H/2} \\
&\overset{(a)}{=} \mathbf{R}_{w_0}^{-1/2}\mathbf{H}_0\big(\mathbf{V}_x\mathbf{D}_x^{-1}\mathbf{V}_x^H \\
&\qquad +\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\mathbf{V}_w\mathbf{D}_w^{-1}\mathbf{V}_w^H\mathbf{H}_2\mathbf{G}\mathbf{H}_1\big)^{-1} \\
&\quad \times\mathbf{H}_0^H\mathbf{R}_{w_0}^{-H/2} \\
&\overset{(b)}{=} \mathbf{R}_{w_0}^{-1/2}\mathbf{H}_0\left(\mathbf{V}_x\mathbf{D}_x^{-1}\mathbf{V}_x^H + \mathbf{Q}_w\boldsymbol{\Lambda}_w\mathbf{Q}_w^H\right)^{-1} \\
&\quad \times\mathbf{H}_0^H\mathbf{R}_{w_0}^{-H/2} \\
&\overset{(c)}{=} \mathbf{R}_{w_0}^{-1/2}\mathbf{H}_0(\mathbf{W}\boldsymbol{\Sigma}\mathbf{W}^H)^{-1}\mathbf{H}_0^H\mathbf{R}_{w_0}^{-H/2} \\
&= \mathbf{R}_{w_0}^{-1/2}\mathbf{H}_0\mathbf{W}\boldsymbol{\Sigma}^{-1}\mathbf{W}^H\mathbf{H}_0^H\mathbf{R}_{w_0}^{-H/2} \\
&\overset{(d)}{=} \mathbf{Q}\boldsymbol{\Lambda}\mathbf{Q}^H \qquad (43)
\end{aligned}$$

where (a) comes from the fact that $\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}$ and $\mathbf{R}_x$ are noise and signal correlation matrices, and as such they can be decomposed as $\mathbf{V}\mathbf{D}\mathbf{V}^H$, where $\mathbf{V}$ is an unitary matrix and $\boldsymbol{\Lambda}$ is a diagonal matrix with real positive entries,

$$\left|\mathbf{I}_{2N} + \begin{bmatrix}\mathbf{H}_0\\ \mathbf{H}_2\mathbf{G}\mathbf{H}_1\end{bmatrix}\mathbf{R}_x\begin{bmatrix}\mathbf{H}_0^H & \mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\end{bmatrix}\begin{bmatrix}\mathbf{R}_{w_0} & \mathbf{0}\\ \mathbf{0} & \mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\end{bmatrix}^{-1}\right| \quad (38)$$

$$= \begin{vmatrix}\mathbf{I}_N + \mathbf{H}_0\mathbf{R}_x\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1} & \mathbf{H}_0\mathbf{R}_x\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}\\ \mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x\mathbf{H}_0^H\mathbf{R}_{w_0}^{-1} & \mathbf{I}_N + \mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{R}_x\mathbf{H}_1^H\mathbf{G}^H\mathbf{H}_2^H\left(\mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\right)^{-1}\end{vmatrix} \quad (39)$$

provided that the noise and signal correlation matrices are full rank matrices. In the second equation above, the subindexes x and w correspond to signal and noise, respectively. Actually, from the SVD decompostion, any matrix than can be written as $\mathbf{A}\mathbf{A}^H$ can be decomposed as $\mathbf{Q}\boldsymbol{\Lambda}\mathbf{Q}^H$, where $\mathbf{Q}$ is a unitary matrix and $\boldsymbol{\Lambda}$ is a diagonal matrix with real nonnegative entries on its main diagonal. This is where (b) comes from. The third equality (c) comes from the fact that the matrix in brackets, $\mathbf{V}_x\mathbf{D}_x^{-1}\mathbf{V}_x^H + \mathbf{Q}_w\boldsymbol{\Lambda}_w\mathbf{Q}_w^H$, is hermitian. Therefore, it can be written as $\mathbf{W}\boldsymbol{\Sigma}\mathbf{W}^H$, with $\mathbf{W}$ an unitary matrix and $\boldsymbol{\Sigma}$ a diagonal matrix with real entries. Since $\mathbf{V}_x\mathbf{D}_x^{-1}\mathbf{V}_x^H + \mathbf{Q}_w\boldsymbol{\Lambda}_w\mathbf{Q}_w^H$ is the result of the addition of one positive definite matrix and one positive semidefinite matrix, it can be shown that all the elements in the diagonal matrix $\boldsymbol{\Sigma}$ are positive. To show this, let us consider that there exist an eigenvalue of the sum matrix λ_i which is nonpositive, then its associated eigenvector $\mathbf{w}_i$ is such that $\mathbf{w}_i^H(\mathbf{V}_x\mathbf{D}_x^{-1}\mathbf{V}_x^H + \mathbf{Q}_w\boldsymbol{\Lambda}_w\mathbf{Q}_w^H)\mathbf{w}_i \leq 0$. Nevertheless, this cannot be true because of the positiveness of the matrix. Finally (d) comes again from the fact that the matrix can be written as $\mathbf{A}\mathbf{A}^H$.

Appendix III

A. Linear MMSE Receiver

Let us assume a linear signal model

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{w}. \tag{44}$$

The mutual information between the transmitted signal $\mathbf{x}$ and the received signal $\mathbf{y}$ is [6]

$$I(\mathbf{y};\mathbf{x}) = \log_2 \left|\mathbf{I} + \mathbf{H}\mathbf{R}_x\mathbf{H}^H\mathbf{R}_w^{-1}\right|. \tag{45}$$

Assume the matrix $\mathbf{T}$ as the linear receiver, $\hat{\mathbf{x}} = \mathbf{T}\mathbf{y}$. The error between the transmitted and the received symbols is

$$\mathbf{M} = E\left\{(\mathbf{x} - \mathbf{T}\mathbf{y})(\mathbf{x} - \mathbf{T}\mathbf{y})^H\right\}. \tag{46}$$

Assuming that symbols and noise are statistically independent, the matrix $\mathbf{T}$ that minimizes the error between each of the transmitted and received symbols

$$\text{MMSE}_i = (\mathbf{M})_{ii} = E\left\{\left|x_i - \mathbf{t}_i^H\mathbf{y}\right|^2\right\} \tag{47}$$

is [4, Ch.6]

$$\mathbf{T} = \mathbf{R}_x\mathbf{H}^H(\mathbf{H}\mathbf{R}_x\mathbf{H}^H + \mathbf{R}_w)^{-1} \tag{48}$$

and the MMSE matrix

$$\mathbf{M} = \left(\mathbf{I} + \mathbf{R}_x^{1/2}\mathbf{H}^H\mathbf{R}_w^{-1}\mathbf{H}\mathbf{R}_x^{H/2}\right)^{-1}. \tag{49}$$

Therefore, the MMSE receiver preserves the mutual information [4, p.243]

$$I(\mathbf{y};\mathbf{x}) = I(\hat{\mathbf{x}};\mathbf{x}) = -\log_2|\mathbf{M}| \tag{50}$$

and the MMSE is sufficient statistics for capacity. Now, let us apply this expression to the case of the relay A&F link and the cooperative A&F link so as to derive optimum MMSE receivers.

Relay A&F link: For the cascade relay channel (model in Fig. 2 without $\mathbf{H}_0$ branch), it is assumed that there is no linear precoding in the transmitter, but a linear precoding matrix $\mathbf{G}$ at the relay terminal that has to be designed. In this case, the definition of the signal model and matrices $\mathbf{H}$ and $\mathbf{R}_w$ is

$$\mathbf{y}_2 = \mathbf{H}_2\mathbf{G}\mathbf{H}_1\mathbf{x} + \mathbf{H}_2\mathbf{G}\mathbf{w}_1 + \mathbf{w}_2 \tag{51}$$

$$\mathbf{H} = \mathbf{H}_2\mathbf{G}\mathbf{H}_1 \quad \mathbf{R}_w = \mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}. \tag{52}$$

Cooperative A&F link: For the cooperative A&F link shown in Fig. 2, the receiver derives an estimate of the transmitted symbol out of the signal received from the relay and the direct links. In this case, the definition of the matrices $\mathbf{H}$ and $\mathbf{R}_w$ is

$$\mathbf{H} = \begin{bmatrix}\mathbf{H}_0 \\ \mathbf{H}_2\mathbf{G}\mathbf{H}_1\end{bmatrix} \quad \mathbf{R}_w = \begin{bmatrix}\mathbf{R}_{w_0} & \mathbf{0} \\ \mathbf{0} & \mathbf{H}_2\mathbf{G}\mathbf{R}_{w_1}\mathbf{G}^H\mathbf{H}_2^H + \mathbf{R}_{w_2}\end{bmatrix}. \tag{53}$$

Appendix IV

Lemma 1: The function $\sum_i \log(1 + x_i(y_i/(1+y_i)))$ (assuming $x_i \geq x_{i+1}$) is maximized when the y_i's are in decreasing order $y_i \geq y_{i+1}$.

Proof: Assume for a moment that for $i < j$ $(x_i \geq x_j)$, y_i are such that $y_i < y_j$. It follows that the term $(1 + (x_iy_i/(1+y_i)))(1 + (x_jy_j/(1+y_j)))$ can be maximized by simply swapping indices

$$\begin{aligned}
&1 + \frac{x_iy_j}{1+y_j} + \frac{x_jy_i}{1+y_i} + \frac{x_iy_j}{1+y_j}\frac{x_jy_i}{1+y_i} \\
&\geq 1 + \frac{x_iy_i}{1+y_i} + \frac{x_jy_j}{1+y_j} + \frac{x_iy_i}{1+y_i}\frac{x_jy_j}{1+y_j} \\
&\Leftrightarrow x_iy_j(1+y_i) + x_jy_i(1+y_j) \\
&\quad \geq x_iy_i(1+y_j) + x_jy_j(1+y_i) \\
&\Leftrightarrow x_i(y_j - y_i) \geq x_j(y_j - y_i) \\
&\Leftrightarrow x_i \geq x_j
\end{aligned}.$$

Since, the logarithm is monotonic, the proof is completed.

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