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MIMO cooperative communication network design with relay selection and CSI feedback

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ABSTRACT

In this paper, we investigate an amplify-and-forward (AF) multiple-input multiple-output – spatial division multiplexing (MIMO-SDM) cooperative wireless networks, where each network node is equipped with multiple antennas. In order to deal with the problems of signal combining at the destination and cooperative relay selection, we propose an improved minimum mean square error (MMSE) signal combining scheme for signal recovery at the destination. Additionally, we propose two distributed relay selection algorithms based on the minimum mean squared error (MSE) of the signal estimation for the cases where channel state information (CSI) from the source to the destination is available and unavailable at the candidate nodes. Simulation results demonstrate that the proposed combiner together with the proposed relay selection algorithms achieve higher diversity gain than previous approaches in both flat and frequency-selective fading channels.

1. Introduction

Next-generation wireless networks will rely on the cooperation among network nodes in order to reduce outage probability of communication links within the networks, to improve spectral and power efficiency, as well as to enhance network coverage. Similarly to communication links using multiple-antenna transceivers, cooperative communication is able to provide diversity by creating multiple replicas of the desired signal via the use of the spatially distributed relay nodes, which act as a virtual antenna array. Many previous publications focused on the research of cooperative communication networks, where network nodes utilize a single antenna for signal transmission and reception [1–6]. However, it is well known that the data rate of the cooperative system employing single-antenna system is still limited due to lack of multiplexing gain.

Meanwhile, multiple-input multiple-output (MIMO) wireless communication systems have been shown to be capable of increasing the channel capacity over rich scattering fading environment [7]. Among MIMO transmission schemes, spatial division multiplexing (SDM) is a typical approach that attains high spectral efficiency thanks to achievable high multiplexing gain [8]. Consequently, combining MIMO-SDM transmission with cooperative communications is a natural research trend in order to exploit the strong points of these schemes [9–16].

A high-performance amplify-and-forward (AF) cooperative network requires effective approaches for relay selection, signal amplification at the relay and signal combination at the destination. In [12] and [13], the problem of signal amplification was solved almost completely by providing the closed-form amplification factors to compensate for the energy loss of the received signal. Since a relay plays a critical role in the performance of cooperative networks, the problem of selecting a relay among a number of candidate nodes is of particular interest to the researchers and was widely considered [3–6], [16]. In [3], two distributed node selection schemes based on the maximum channel gains and the maximum harmonic mean of the channel gains were proposed for cooperative nodes equipped with single antenna. In [4] a relay selection scheme based on the partial channel state information (CSI) was developed so that a compromise between the bandwidth efficiency and diversity order could be attained. Relay selection schemes for a decode-and-forward (DF) cooperative network, which take the energy consumption into consideration, were investigated in [5]. In another work [6], Jing and Jafarkhani derived the diversity of various single-relay selection schemes in the literature and subsequently generalized the problem of single relay selection to the

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scenario of multiple cooperative relay selection. Several results related to the relay selection in MIMO cooperative networks were reported in [11,13,15], and [16]. In [11], dual-hop non-coherent AF wireless MIMO relay networks were considered and a distributed orthogonal relay selection based on the maximum harmonic mean of the channel gains was constructed. The problem of combining relay and antenna selection was studied and a greedy antenna selection based on the minimum mean square error (MMSE) criterion was proposed in the work of [13]. In [15], the authors proposed a set of joint transmit diversity selection and multi-relay selection algorithms for the uplink of cooperative MIMO systems. Both DF and AF multirelay schemes with linear MMSE, successive interference cancellation, and adaptive reception were investigated. Relays were selected such that the MSE at the receiver was minimized or the mutual information of the relay channels was maximized. Regarding the problem of combining and detecting the received signals within two time slots at the destination, effective approaches should be developed so that diversity gain from the two communication links could be exploited. For single-antenna cooperative networks, maximal ratio combining (MRC) was proposed to use as the optimal combiner for maximizing the diversity gain [3]. Nevertheless, MRC cannot be utilized in MIMO-SDM cooperative networks due to the inherent co-channel interference (CCI) among transmit streams.

In the recent work of Tran et al. [16], the problems of distributed relay selection and signal combining for MIMO-SDM cooperative communication networks were considered. Specifically, the authors proposed three distributed relay selection algorithms based on the maximum channel gains, the maximum harmonic mean of the channel gains, and the mean square error (MSE) of the signal estimation. In addition, an MMSE signal combining scheme was also proposed to combine the signal from the direct and relaying path. In the MMSE combining scheme, the general combining weight matrix is decoupled into separate combining matrices associated with the direct and relaying paths, whereby reducing computational complexity of the signal combiner at the destination. However, the decomposition of the general combining weight matrix is probably the main reason that leads to the following limitations: (1) the system is unable to achieve significant diversity gain; and (2) the bit error rate (BER) performance of the system could not be improved when the number of candidate nodes increases.

Inspired by the above problems, in this paper we revisit the problem of MIMO-SDM cooperative communication network design and try to overcome the limitations of the previous approaches in [16]. We first derive a general MMSE signal combiner at the destination, taking into account the cross-correlation between the received signal from the source to the destination and that from the relay to the destination. With the additional cross-correlation term, the proposed MMSE combiner can achieve more signal energy to obtain improved BER performance as compared to its former counterpart in [16]. The trade-off for this improvement is, however, the cost of increased computational load since the proposed combiner can no longer be decoupled. Next, we evaluate the estimated MSE using the proposed combiner at the destination. The result is then utilized to develop two relay selection schemes under the following assumptions: (1) a candidate node knows not only the channels from the source to it and from it to the destination but also the one from the source to the destination; and (2) a candidate node only knows the channels from the source to it and from it to the destination. Simulation results show that the proposed relay selection schemes in combination with the proposed MMSE combiner outperform those in [16] remarkably under both flat fading and selective fading channels. The remainder of the paper is organized as follows. The system model is presented in Section 2. The proposed MMSE combiner is derived in Section 3. Section 4 and Section 5 respectively presents the derivation of MSE and the proposed relay selection schemes. Simulation results are provided and discussed in Section 6, and finally conclusions are given in Section 7.

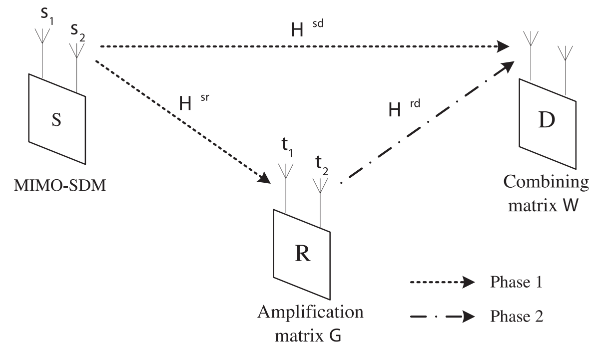


Fig. 1. MIMO cooperative network model, $N = 2$ for simple presentation.

2. System model

We consider an ad hoc wireless communications network which consists of a source (S), a destination (D), and K capable neighbouring nodes which are randomly distributed between the source and the destination as illustrated in Fig. 1. It is assumed that the distance d between the source and the destination is close enough such that the direct communications between the nodes is possible. All nodes including the source, the destination and neighbouring nodes have $N > 1$ antennas for both transmission and reception. The cooperative communications in the network is done in two consecutive phases, namely, *relay selection* and *cooperative transmission*. During the first phase, a distributed relay selection algorithm such as presented in [3,16] is used to select the best neighbouring node to act as the relay to support the transmission from the source to the destination. In order to facilitate the distributed selection algorithm, the neighbouring nodes are assumed to perfectly know the CSI of the links from them to the destination as well as from them to the source. Depending on the considered situations, the neighbouring nodes are further assumed to know or not to know the CSI from the source to the destination. These two cases are referred to as the one with CSI feedback (CSIF) and non-CSI feedback (NCSI). In the second phase, with the help of the relay, communications from the source to the destination can be done in the cooperative mode, i.e., via both the direct and relaying path. The cooperative transmission from the source to the destination is conducted in two time slots. In the first time slot, the source broadcasts its data to both the destination and the relay. During the next time slot, the relay forwards its received data to the destination. In this paper, we only focus on the popular AF strategy. For simplicity, half-duplex relaying is assumed in our system. We further assume that there is no transmission delay in the second hop, i.e., second time slot. Although the cooperative communications happens in two phases and two time slots, the signal model is the same. The only difference is that during the relay selection phase, all the neighbouring nodes k, ($k = 1, 2, \ldots, K$) are involved in the signal exchange while during the cooperative transmission only the relay r is active. Therefore, in the next sections we simplify the cooperative model to only two time slots: from the source to the destination and from the relay to the destination. In our general model, k is used to denote the neighbouring node index, which in fact relates to the first phase. This model can certainly be used in the cooperative transmission phase by replacing k by the relay index r. The channels between the nodes are assumed to be affected by flat Rayleigh fading.

We first define the signal vector transmitted from the source as $\boldsymbol{s} = [s_1, s_2, \ldots, s_N]^T$, where s_i, $(i = 1, 2, \ldots N)$ is the modulated signal symbol transmitted from the ith antenna of the source, and i is the antenna index. The covariance matrix of the transmit vector is equal to $\boldsymbol{R}_{ss} = \mathrm{E}\{\boldsymbol{ss}^H\} = E_s \boldsymbol{I}_N$, where E_s is the symbol energy, $\boldsymbol{I}_N$ is the $N \times N$ identity matrix, and $\mathrm{E}\{\cdot\}$ denotes the expectation operation. The total transmit power of the source is therefore equal to $P_T = \mathrm{tr}(\boldsymbol{R}_{ss}) = NE_s$, where $\mathrm{tr}(\cdot)$ denotes the trace of matrix.

During the first time slot, the transmit vector $\boldsymbol{s}$ is broadcast to all network nodes including the destination and K neighbouring nodes. The signal vectors received at the destination and at node k are respectively given by

$$\boldsymbol{y}_1 = \sqrt{\frac{\gamma_{sd}}{NE_s}} \boldsymbol{H}^{sd} \boldsymbol{s} + \boldsymbol{n}_1 = \rho_{sd} \boldsymbol{H}^{sd} \boldsymbol{s} + \boldsymbol{n}_1, \tag{1}$$

$$\boldsymbol{x}_k = \sqrt{\frac{\gamma_{sk}}{NE_s}} \boldsymbol{H}^{sk} \boldsymbol{s} + \boldsymbol{n}_k = \rho_{sk} \boldsymbol{H}^{sk} \boldsymbol{s} + \boldsymbol{n}_k. \tag{2}$$

Here γ_{sd} and γ_{sk} are the signal-to-noise power ratio (SNR) at the receiver of the link from the source to the destination and from the source to node k, respectively. These SNRs take into account the path loss of the respective links and are assumed to be the same at each receive branch of the receiver. $\boldsymbol{H}^{sd}$ and $\boldsymbol{H}^{sk}$ denote the $N \times N$ MIMO channel matrices from the source to the destination and from the source to neighbouring node k, respectively. Entries of the channel matrices can be modelled using independent and identically distributed (iid) complex Gaussian random variables with zero mean and unit variance. This channel model reflects the effect of uncorrelated flat Rayleigh fading. $\boldsymbol{n}_1 \in \mathbb{C}^{N\times 1}$ and $\boldsymbol{n}_k \in \mathbb{C}^{N\times 1}$ represent the additive white Gaussian noise (AWGN) vectors affecting the receiver of the destination and the neighbouring node k during the first time slot, respectively. The entries $n_{1,i}$ and $n_{k,i}$ of the noise vectors can also be modelled using iid complex Gaussian random variables with zero mean and unit variance, i.e., $n_{1,i} \in \mathcal{N}_c(0, 1)$ and $n_{k,i} \in \mathcal{N}_c(0, 1)$.

During the second time slot, if it is selected as the relay the neighbouring node k will forward the amplified version of the received signal vector $\boldsymbol{x}_k$ to the destination while the source stops transmitting. The amplification factor is chosen such that it can compensate for the power loss occurred in the link from the source to the relay. The amplification function of the relay can be represented by a diagonal matrix $\boldsymbol{G}_k \in \mathbb{C}^{N\times N}$, whose the diagonal scalar entry $g_{k,i}$ is the amplification factor of the ith relay branch. The amplified signal vector can be expressed as

$$\boldsymbol{t} = \boldsymbol{G}_k \boldsymbol{x}_k = \rho_{sk} \boldsymbol{G}_k \boldsymbol{H}^{sk} \boldsymbol{s} + \boldsymbol{G}_k \boldsymbol{n}_k. \tag{3}$$

Under the assumption that each relay has the maximum total transmit power of P_k, we have the following condition: $\mathrm{E}\{\mathrm{tr}(\boldsymbol{tt}^H)\} = \mathrm{tr}\left(E_s \rho_{sk}^2 \boldsymbol{G}_k^2 \boldsymbol{H}^{sk} (\boldsymbol{H}^{sk})^H + \boldsymbol{G}_k^2 \boldsymbol{I}_N\right) \le P_k$. Assume that full power P_k is available and that the power is divided equally among N transmit antennas, we have $E_s \rho_{sk}^2 g_{k,i}^2 \|\boldsymbol{h}_i^{sk}\|_F^2 + g_{k,i}^2 = P_k/N$, where $\boldsymbol{h}_i^{sk}$ is the ith row of the channel matrix $\boldsymbol{H}^{sk}$, and $\|\cdot\|_F^2$ denotes the Frobenius norm. Equivalently, we can express the ith diagonal element g_i of $\boldsymbol{G}_k$ as:

$$g_{k,i} = \sqrt{\frac{P_k/N}{E_s \rho_{sk}^2 \|\boldsymbol{h}_i^{sk}\|_F^2 + 1}} = \sqrt{\frac{P_k/N}{\gamma_{sk}/N \|\boldsymbol{h}_i^{sk}\|_F^2 + 1}}. \tag{4}$$

Provided that $P_k = NE_s$, we obtain the amplification factor

$$g_{k,i} = \sqrt{\frac{NE_s}{\gamma_{sk} \|\boldsymbol{h}_i^{sk}\|_F^2 + N}}. \tag{5}$$

As the source stops transmitting, the received signal at the destination contains only the signal transmitted from the relay and is given by

$$\begin{aligned} \boldsymbol{y}_2 &= \sqrt{\frac{\gamma_{kd}}{P_k}} \boldsymbol{H}^{kd} \boldsymbol{G}_k \boldsymbol{x}_k + \boldsymbol{n}_2 \sqrt{\frac{\gamma_{kd}}{P_k}} \boldsymbol{H}^{kd} \boldsymbol{G}_k (\rho_{sk} \boldsymbol{H}^{sk} \boldsymbol{s} + \boldsymbol{n}_k) + \boldsymbol{n}_2, \\ \boldsymbol{y}_2 &= \rho_{kd} \rho_{sk} \boldsymbol{H}^{kd} \boldsymbol{G}_k \boldsymbol{H}^{sk} \boldsymbol{s} + \rho_{kd} \boldsymbol{H}^{kd} \boldsymbol{G}_k \boldsymbol{n}_k + \boldsymbol{n}_2 = \boldsymbol{H}^{skd} \boldsymbol{s} + \tilde{\boldsymbol{n}}_2, \end{aligned} \tag{6}$$

where $\boldsymbol{n}_2$ is the AWGN vector affecting the destination receiver at the second time slot, and

$$\boldsymbol{H}^{skd} = \rho_{kd} \rho_{sk} \boldsymbol{H}^{kd} \boldsymbol{G}_k \boldsymbol{H}^{sk}, \tag{7}$$

$$\mathrm{n}_2 = \rho_{kd} \boldsymbol{H}^{kd} \boldsymbol{G}_k \boldsymbol{n}_k + \boldsymbol{n}_2. \tag{8}$$

At the end of the second time slot, the destination will combine the two received signal vectors from the first time slot given by (1) and that from the second time slot given by (6) to have the total received signal vector in the following form:

$$\begin{bmatrix} \boldsymbol{y}_1 \\ \boldsymbol{y}_2 \end{bmatrix} = \begin{bmatrix} \rho_{sd} \boldsymbol{H}^{sd} \\ \boldsymbol{H}^{skd} \end{bmatrix} \boldsymbol{s} + \begin{bmatrix} \boldsymbol{n}_1 \\ \tilde{\boldsymbol{n}}_2 \end{bmatrix}. \tag{9}$$

Using the following notations

$$\boldsymbol{y} \triangleq \begin{bmatrix} \boldsymbol{y}_1 \\ \boldsymbol{y}_2 \end{bmatrix}, \boldsymbol{H} \triangleq \begin{bmatrix} \rho_{sd} \boldsymbol{H}^{sd} \\ \boldsymbol{H}^{skd} \end{bmatrix}, \boldsymbol{n} \triangleq \begin{bmatrix} \boldsymbol{n}_1 \\ \tilde{\boldsymbol{n}}_2 \end{bmatrix}, \tag{10}$$

we have the system equation in the short form as

$$\boldsymbol{y} = \boldsymbol{Hs} + \boldsymbol{n}. \tag{11}$$

3. Linear combining weight matrix at the receiver

The principle of the linear combining is to use a weight matrix $\boldsymbol{W}$ to combine the received signal vector $\boldsymbol{y}$ to estimate the transmit signal vector $\boldsymbol{s}$, i.e.,

$$\hat{\boldsymbol{s}} = \boldsymbol{Wy}. \tag{12}$$

Using the MMSE method, $\boldsymbol{W}$ should be designed based on the following cost function

$$\boldsymbol{W} = \underset{\boldsymbol{W}}{\mathrm{argmin}} \mathrm{E}\{\|\boldsymbol{s} - \boldsymbol{Wy}\|^2\}. \tag{13}$$

Solving (13) for $\boldsymbol{W}$ amounts to the following orthogonal principle $\mathrm{E}\{(\boldsymbol{Wy} - \boldsymbol{s})\boldsymbol{y}^H\} = \boldsymbol{0}$, or equivalently $\boldsymbol{WR}_{yy} = \boldsymbol{R}_{sy}$, where $\boldsymbol{R}_{yy} = \mathrm{E}\{\boldsymbol{yy}^H\}$ is the covariance matrix of the received signal vector $\boldsymbol{y}$ and $\boldsymbol{R}_{sy} = \mathrm{E}\{\boldsymbol{sy}^H\}$ the cross-correlation matrix between the transmit vector $\boldsymbol{s}$ and the received vector $\boldsymbol{y}$. The MMSE weight matrix is now given by

$$\boldsymbol{W} = \boldsymbol{R}_{sy} (\boldsymbol{R}_{yy})^{-1}. \tag{14}$$

In order to find $\boldsymbol{W}$, we note that: $\mathrm{E}\{\boldsymbol{ss}^H\} = E_s \boldsymbol{I}_N$, $\mathrm{E}\{\boldsymbol{n}_1 \boldsymbol{n}_1^H\} = \mathrm{E}\{\boldsymbol{n}_2 \boldsymbol{n}_2^H\} = \mathrm{E}\{\boldsymbol{n}_k \boldsymbol{n}_k^H\} = \boldsymbol{I}_N$. Under assumption that the noises affecting the relay and the destination at the two time slots are uncorrelated, we have $\mathrm{E}\{\boldsymbol{n}_1 \boldsymbol{n}_2^H\} = \mathrm{E}\{\boldsymbol{n}_2 \boldsymbol{n}_1^H\} = \mathrm{E}\{\boldsymbol{n}_1 \boldsymbol{n}_k^H\} = \mathrm{E}\{\boldsymbol{n}_2 \boldsymbol{n}_k^H\} = \mathrm{E}\{\boldsymbol{n}_k \boldsymbol{n}_1^H\} = \mathrm{E}\{\boldsymbol{n}_k \boldsymbol{n}_2^H\} = \boldsymbol{0}_N$. In addition, using (8) we can write $\mathrm{E}\{\mathrm{n}_2 \mathrm{n}_2^H\} = \gamma_{kd}/P_k \boldsymbol{H}^{kd} \boldsymbol{G}_k^2 (\boldsymbol{H}^{kd})^H + \boldsymbol{I}_N$. We now can easily obtain the covariance matrix $\boldsymbol{R}_{nn}$ of the noise vector $\boldsymbol{n}$ at the destination as

$$\boldsymbol{R}_{nn} = \mathrm{E}\{\boldsymbol{nn}^H\} = \begin{bmatrix} \boldsymbol{I}_N & \boldsymbol{0}_N \\ \boldsymbol{0}_N & \frac{\gamma_{kd}}{P_k} \boldsymbol{H}^{kd} \boldsymbol{G}_k^2 (\boldsymbol{H}^{kd})^H + \boldsymbol{I}_N \end{bmatrix}. \tag{15}$$

The covariance matrix $\boldsymbol{R}_{yy}$ is now given by $\boldsymbol{R}_{yy} = \mathrm{E}\{\boldsymbol{yy}^H\} = E_s \boldsymbol{HH}^H + \boldsymbol{R}_{nn}$. The cross-correlation matrix between $\boldsymbol{s}$ and $\boldsymbol{y}$ can be similarly given by $\boldsymbol{R}_{sy} = \mathrm{E}\{\boldsymbol{s}\boldsymbol{y}^H\} = E_s\boldsymbol{H}^H$. As a result, the combining matrix is obtained as:

$$\boldsymbol{W} = E_s\boldsymbol{H}^H(E_s\boldsymbol{H}\boldsymbol{H}^H + \boldsymbol{R}_{nn})^{-1} = \boldsymbol{H}^H\left(\boldsymbol{H}\boldsymbol{H}^H + \frac{1}{E_s}\boldsymbol{R}_{nn}\right)^{-1}. \quad (16)$$

We can see that the weight matrix is expressed in a more compact form than that in [16]. More importantly, the covariance matrix $\boldsymbol{R}_{yy}$ is no longer block diagonal as the cross-correlation term $\mathrm{E}\{\boldsymbol{y}_1\boldsymbol{y}_2^H\}$ is taken into consideration. Consequently, it is not possible to decompose the weight matrix $\boldsymbol{W}$ into two separate weight matrices associated with the direct and relaying path as in [16]. Besides, the linear combination of the signals from two paths using a single weight matrix $\boldsymbol{W}$ of size $N\times 2N$, as shown above, requires higher computational complexity as compared to the combination method proposed in [16]. Considering the effect of signal combining, however, the proposed combination method gives us more advantages. First, the weight matrix is jointly optimized for the two received signal vectors, i.e., $\boldsymbol{y}_1$ and $\boldsymbol{y}_2$, and used to combine $\boldsymbol{y}_1$ and $\boldsymbol{y}_2$ for the estimation of the transmit vector $\boldsymbol{s}$. This results in a combiner with higher energy gain, and hence better detection performance. Second, since the $N\times 2N$ instead of the $N\times N$ weight matrix is used, the joint combination is expected to provide double diversity gain over the independent combination as proposed in [16]. These statements will be justified by the simulation results in the next section.

4. Evaluation of mean square error

In this section, we will derive the estimated MSE used for relay selection in our proposed system. Note that the MSE based relay selection was initially proposed in [16]. Our derivation, however, results in a different solution of MSE associated with each node k. This new solution of MSE will facilitate our proposed scheme of relay selection with CSIF.

Using the MMSE detection, the covariance matrix of the estimated signal vector $\hat{\boldsymbol{s}}$ in (12) is given by $\boldsymbol{R}_{\hat{s}\hat{s}} = \mathrm{E}\{\hat{\boldsymbol{s}}\hat{\boldsymbol{s}}^H\} = \boldsymbol{W}\boldsymbol{R}_{yy}\boldsymbol{W}^H$. The MSE matrix is defined as the covariance matrix of the error vector given by $\boldsymbol{e} = (\hat{\boldsymbol{s}} - \boldsymbol{s})$, i.e., $\boldsymbol{E} \triangleq \mathrm{E}\{(\hat{\boldsymbol{s}} - \boldsymbol{s})(\hat{\boldsymbol{s}} - \boldsymbol{s})^H\}$. After straightforward mathematical manipulations, we can easily have

$$\boldsymbol{E} = E_s[\boldsymbol{I}_N - \boldsymbol{H}^H\left(\boldsymbol{H}\boldsymbol{H}^H + \frac{1}{E_s}\boldsymbol{R}_{nn}\right)^{-1}\boldsymbol{H}]. \quad (17)$$

Using the matrix inversion lemma $(\boldsymbol{A}+\boldsymbol{B}\boldsymbol{C}\boldsymbol{D})^{-1} = \boldsymbol{A}^{-1} - \boldsymbol{A}^{-1}\boldsymbol{B}(\boldsymbol{D}\boldsymbol{A}^{-1}\boldsymbol{B} + \boldsymbol{C}^{-1})^{-1}\boldsymbol{D}\boldsymbol{A}^{-1}$ with $\boldsymbol{A} = \boldsymbol{I}_N$, $\boldsymbol{B} = \boldsymbol{H}^H$, $\boldsymbol{C} = \frac{1}{E_s}\boldsymbol{R}_{nn}$, $\boldsymbol{D} = \boldsymbol{H}$, and note that $\boldsymbol{I}_N = \boldsymbol{I}_N^{-1}$, $\boldsymbol{I}_N^{-1}\boldsymbol{H}^H = \boldsymbol{H}^H$ and $\boldsymbol{H}\boldsymbol{I}_N^{-1} = \boldsymbol{H}$ we can rewrite (17) as follows:

$$\boldsymbol{E} = E_s(\boldsymbol{I}_N + E_s\boldsymbol{H}^H\boldsymbol{R}_{nn}^{-1}\boldsymbol{H})^{-1}. \quad (18)$$

Now, let us express (15) in the following form:

$$\boldsymbol{R}_{nn} = \begin{bmatrix} \boldsymbol{R}_{nn,1} & \boldsymbol{0} \\ \boldsymbol{0} & \boldsymbol{R}_{nn,k} \end{bmatrix}, \quad (19)$$

where $\boldsymbol{R}_{nn,1} = \boldsymbol{I}_N$, and $\boldsymbol{R}_{nn,k} = \frac{\gamma_{kd}}{P_k}\boldsymbol{H}^{kd}\boldsymbol{G}_k^2(\boldsymbol{H}^{kd})^H + \boldsymbol{I}_N$. Since $\boldsymbol{R}_{nn}$ is a block diagonal matrix we have

$$\boldsymbol{R}_{nn}^{-1} = \begin{bmatrix} \boldsymbol{R}_{nn,1}^{-1} & \boldsymbol{0} \\ \boldsymbol{0} & \boldsymbol{R}_{nn,k}^{-1} \end{bmatrix}, \quad (20)$$

and it follows that: $\boldsymbol{H}^H\boldsymbol{R}_{nn}^{-1}\boldsymbol{H} = \rho_{sd}^2(\boldsymbol{H}^{sd})^H\boldsymbol{H}^{sd} + (\boldsymbol{H}^{skd})^H\boldsymbol{R}_{nn,k}^{-1}\boldsymbol{H}^{skd}$. Therefore, the error covariance matrix in (18) becomes

$$\boldsymbol{E} = E_s[\boldsymbol{I}_N + E_s\left(\rho_{sd}^2(\boldsymbol{H}^{sd})^H\boldsymbol{H}^{sd} + (\boldsymbol{H}^{skd})^H\boldsymbol{R}_{nn,k}^{-1}\boldsymbol{H}^{skd}\right)]^{-1}. \quad (21)$$

The MSE associated with detection of the ith transmit substream can be expressed as:

$$\begin{aligned} MSE_i &= E_s\left[\boldsymbol{I}_N + E_s\boldsymbol{H}^H\boldsymbol{R}_{nn}^{-1}\boldsymbol{H}\right]_{ii}^{-1} \\ &= E_s\left[\boldsymbol{I}_N + E_s\left(\rho_{sd}^2(\boldsymbol{H}^{sd})^H\boldsymbol{H}^{sd} + (\boldsymbol{H}^{skd})^H\boldsymbol{R}_{nn,k}^{-1}\boldsymbol{H}^{skd}\right)\right]_{ii}^{-1}. \end{aligned} \quad (22)$$

Note from (22) that the MSE_i differs from that of [16] in that it also contains the channel matrix $\boldsymbol{H}^{sd}$ of the direct link from the source to the destination.

5. Proposed relay selection algorithms

5.1. Relay selection based on MSE criterion with CSI feedback

In this approach, we assume that each neighbouring node k perfectly knows all the CSI including $\boldsymbol{H}^{sd}$, $\boldsymbol{H}^{sk}$ and $\boldsymbol{H}^{kd}$, the SNRs γ_{sd}, γ_{sk} and γ_{kd}. In practice, the CSI of the local links from a neighbouring node to the source and from a neighbouring node to the destination can be easily estimated during the exchange of RTS/CTS control frames between the source and the destination. The assumption of the CSI of the direct link from the source to the destination available at a neighbouring node is more ideal and can be realized by an additional feedback of $\boldsymbol{H}^{sd}$ and γ_{sd} from the destination to K neighbouring nodes. Our proposal of the feedback realization is to modify unused fields such as Address 3 (6 bytes) and Address 4 (6 bytes) in the ad hoc mode of the IEEE 802.11 standards to carry out this information. When the destination broadcasts a CTS frame to the source, neighbouring nodes can overhear it and extract the expected information.

Under the assumptions of CSI availability, a neighbouring node k will calculate the MSE matrix for the case if it is used as the relay based on (17) as follows:

$$\boldsymbol{E}_k \triangleq E_s(\boldsymbol{I}_N - \boldsymbol{W}_k\boldsymbol{H}_k) = E_s\left[\boldsymbol{I}_N - \boldsymbol{H}_k^H\left(\boldsymbol{H}_k\boldsymbol{H}_k^H + \frac{1}{E_s}\boldsymbol{R}_{nn}\right)^{-1}\boldsymbol{H}_k\right], \quad (23)$$

where

$$\boldsymbol{H}_k = \begin{bmatrix} \rho_{sd}\boldsymbol{H}^{sd} \\ \boldsymbol{H}^{skd} \end{bmatrix} = \begin{bmatrix} \rho_{sd}\boldsymbol{H}^{sd} \\ \rho_{kd}\rho_{sk}\boldsymbol{H}^{kd}\boldsymbol{G}_k\boldsymbol{H}^{sk} \end{bmatrix}. \quad (24)$$

Using the MSE criterion, a neighbouring node k will select the largest MSE among N diagonal entries of $\boldsymbol{E}_k$ given in (21) as

$$e_k = \max_i[\boldsymbol{E}_k]_{ii} = \max_i E_s[\boldsymbol{I}_N - \boldsymbol{W}_k\boldsymbol{H}_k]_{ii}. \quad (25)$$

The corresponding channel quality index (CQI) is calculated as follows:

$$Q_k = \frac{1}{e_k}. \quad (26)$$

The relay selection algorithm based on MSE-CSIF criterion is summarized in Algorithm 1.

Algorithm 1. Relay Selection Algorithm Based on MSE-CSIF

1: **for** $k = 1$ to K **do**
2: $\boldsymbol{E}_k \leftarrow E_s(\boldsymbol{I}_N - \boldsymbol{W}_k\boldsymbol{H}_k)$
3: **for** $i = 1$ to N **do**
4: $\boldsymbol{e}(i) \leftarrow [\boldsymbol{E}_k]_{ii}$
5: **end for**
6: $e_{k,\max} \leftarrow \max(\boldsymbol{e})$
7: $Q_k \leftarrow 1/e_{k,\max}$
8: **end for**
9: $k_{\mathrm{sel}} \leftarrow \arg\max_k(Q_k)$

$r \leftarrow k_{\text{sel}}$
return r as the relay

5.2. Relay selection based on MSE criterion without CSI feedback

In this approach, we assume that a neighbouring node k perfectly knows the CSI $\boldsymbol{H}^{sk}$, $\boldsymbol{H}^{kd}$ and the SNRs γ_{sk} and γ_{kd} of the local links. However, it does not have any information related to $\boldsymbol{H}^{sd}$ and γ_{sd}. The latter assumption relaxes the requirement of the CSI feedback and reserves the CTS frame structure. From (22), we can write

$$\text{MSE}_i < E_s\left[\boldsymbol{I}_N + E_s(\boldsymbol{H}^{skd})^H \boldsymbol{R}_{nn,k}^{-1}\boldsymbol{H}^{skd}\right]_{ii}^{-1}. \tag{27}$$

Using again the matrix inversion lemma again, we have

$$\begin{aligned}&\left[\boldsymbol{I}_N + E_s(\boldsymbol{H}^{skd})^H \boldsymbol{R}_{nn,k}^{-1}\boldsymbol{H}^{skd}\right]^{-1}\\ &\quad = \boldsymbol{I}_N - (\boldsymbol{H}^{skd})^H\left[\boldsymbol{H}^{skd}(\boldsymbol{H}^{skd})^H + \frac{1}{E_s}\boldsymbol{R}_{nn,k}\right]^{-1}\boldsymbol{H}^{skd} = \boldsymbol{I}_N - \hat{\boldsymbol{W}}_k\boldsymbol{H}^{skd},\end{aligned} \tag{28}$$

where $\hat{\boldsymbol{W}}_k \triangleq (\boldsymbol{H}^{skd})^H[\boldsymbol{H}^{skd}(\boldsymbol{H}^{skd})^H + 1/E_s\boldsymbol{R}_{nn,k}]^{-1}$. Eq. (27) now can be re-expressed as: $\text{MSE}_i < \text{MSE}'_i$, where $\text{MSE}'_i = E_s[\boldsymbol{I}_N - \hat{\boldsymbol{W}}_k\boldsymbol{H}^{skd}]_{ii}$. Under the above assumption, the node k will calculate the MSE matrix using the following equation: $\boldsymbol{E}_k \triangleq E_s[\boldsymbol{I}_N - \hat{\boldsymbol{W}}_k\boldsymbol{H}^{skd}]$. In the MSE-NCSI criterion, each node k selects the largest MSE, among N diagonal entries of $\boldsymbol{E}_k$ and compute the corresponding CQI as follows:

$$e_k = \max_i[\boldsymbol{E}_k]_{ii} = \max_i E_s[\boldsymbol{I}_N - \hat{\boldsymbol{W}}_k\boldsymbol{H}^{skd}]_{ii}, \tag{29}$$

$$Q_k = \frac{1}{e_k}. \tag{30}$$

The relay selection based on MSE-NCSI criterion is summarized in Algorithm 2.

Algorithm 2. Relay Selection Algorithm Based on MSE-NCSI

for $k = 1$ to K **do**
$\boldsymbol{E}_k \leftarrow E_s(\boldsymbol{I}_N - \hat{\boldsymbol{W}}_k\boldsymbol{H}^{skd})$
for $i = 1$ to N **do**
$\boldsymbol{e}(i) \leftarrow [\boldsymbol{E}_k]_{ii}$
end for
$e_{k,\max} \leftarrow \max(\boldsymbol{e})$
$Q_k \leftarrow 1/e_{k,\max}$
end for
$k_{\text{sel}} \leftarrow \arg\max_k (Q_k)$
$r \leftarrow k_{\text{sel}}$
return r as the relay

5.3. Extension to multiple-relay selection

In some situations, the selection of multiple relays could be of particular interest. In such circumstances, our proposed scheme can easily be extended to achieve the goal. Thanks to the distributed characteristic of the relay selection algorithms, the destination is able to know the channel quality indices Q_k, ($k = 1, \ldots, K$), of the neighboring nodes and able to select the best nodes for relaying the signals. As soon as the relays are selected, they can start amplifying their received signals from the source as in (3) and forwarding them to the destination via different time slots. At the destination, the received signal vectors from the source and the relays will be arranged in a proper way to get the system equation similar to the one in (11) for signal recovery.

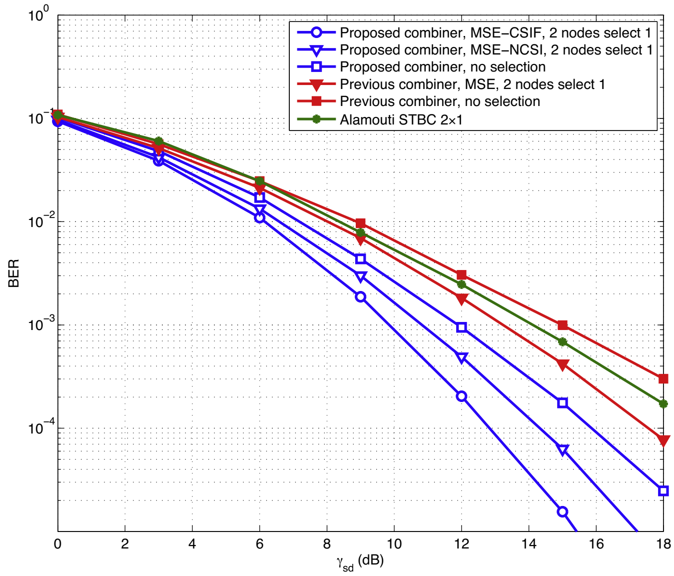


Fig. 2. BER performance comparison of the proposed MMSE combiner with the previous one in [16], $N = 2$, one relay, $\gamma_{sd} = \gamma_{sk} = \gamma_{kd}$, and BPSK modulation.

It is worth noting that there exist a number of works related to the multiple-relay selection problem in the literature [6,13,15]. However, few of them can be adopted in our proposed scheme. Although a multiple-relay selection scheme was proposed in [6]. This scheme is actually for the selection of the relays equipped with single antenna. Therefore, it could not directly be applied the proposed scheme. Similarly, the greedy antenna selection based on the MMSE criterion in [13] could not be utilized except for the case where the nodes in the network are equipped with $N = 2$ antennas. This is due to the fact that at each relay node a pair of antennas are selected out of N antennas. Among multiple-relay selection algorithms, the ones in [15] were devised for cooperative MIMO systems. Hence, they could be utilized in our proposed scheme. Nevertheless, the system equation at the destination must be modified appropriately for signal detection.

Detailed descriptions of how the proposed algorithms can be extended to the case of multiple-relay selection or of how to adopt existing algorithms in the proposed system as well as the comparison among different multiple-relay selection algorithms are out of this paper's scope and will be left for future works.

6. Simulation results

In this section, Monte-Carlo simulations are used to evaluate performance of the proposed MIMO-SDM cooperative systems. For simplicity but without loss of generality, we assume that the number of antennas of each node is $N = 2$. The channels between each pair of nodes are quasi-static and affected by flat Rayleigh fading.

6.1. Comparison with previously proposed MMSE combiner

In order to compare the detection performance of our proposed MMSE combiner with the previous one in [16], we have plotted the signal constellations of the 8-PSK signal obtained after the two combiners (not included here due to limited space) at the SNR $\gamma_{sd} = \gamma_{sk} = \gamma_{kd} = 18$ dB. It is clear that the constellation achieved by the proposed combiner is less scattered than that by the previous one. Thus the proposed combiner will promise better error performance. To support this statement we compare the BER performance of the two combiners in Fig. 2. The simulated model includes two neighbouring nodes with equal link SNRs, i.e., $\gamma_{sd} = \gamma_{sk} = \gamma_{kd}$, for

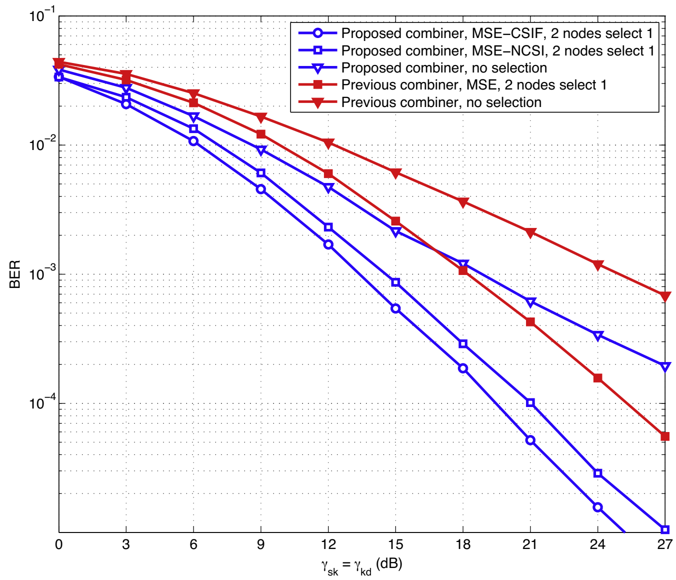
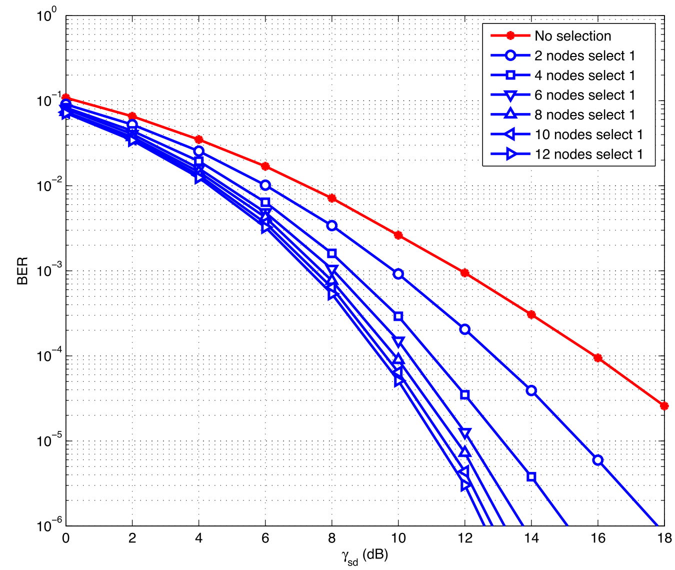


Fig. 3. BER performance when SNRs of the relaying path varies, γ_{sd} = 6dB, N = 2, one relay, BPSK modulation, and perfect CSI.

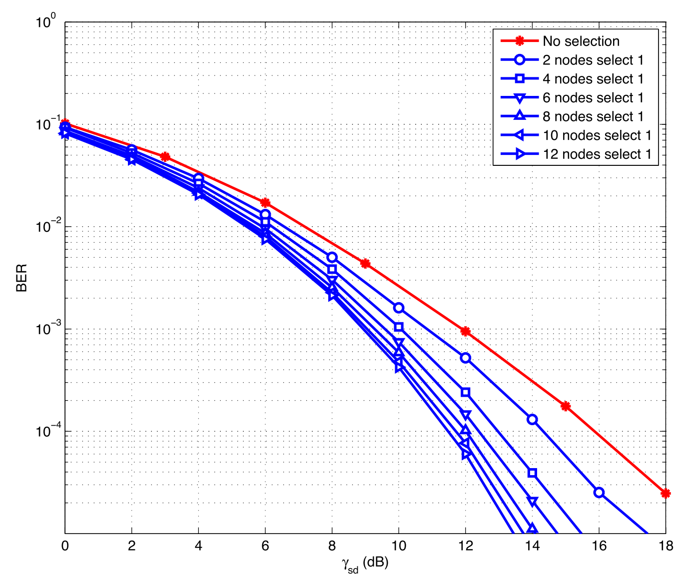
k = 1, 2. BPSK was used in the simulation. The relay is selected from the two neighbouring nodes based on the MSE-CSIF and MSE-NCSI algorithms. In case of no selection, the model includes a neighbouring node as the relay. It can be clearly seen in the figure that the proposed detector significantly outperforms the previous combiner under both the case of using and without using relay selection. The proposed combiner under the case without using relay selection is shown to even outperform the previous combiner with MSE 2-select-1 relay selection. It is also realized from the figure that the proposed MSE-CSIF relay selection provides the best performance, followed by the MSE-NCSI. Both the proposed selection algorithms outperform the previous MSE selection significantly. Comparing with the reference system of 2×1 Alamouti STBC leads us to the conclusion that the proposed system can achieve diversity order greater than 2. This conclusion is clear as the combiner can combine different versions of the transmitted signal vector $\boldsymbol{s}$ in both space and time domain via the two paths and two time slots. However, as the system is affected by CCI among N transmit streams, the precise diversity order needs more complicated analysis in the future work.

In order to gain insight into the performance improvement of the proposed combiner, we illustrate in Fig. 3 the BER for the case the SNR of the direct path is fixed to γ_{sd} = 6 dB while the SNRs of the relaying path are set $\gamma_{sk} = \gamma_{kd}$ and varies from 0 to 27 dB. For the case no relay selection is used, the proposed combiner has 6 dB SNR gain at BER = 10^{-3}. When relay selection is used with two neighbouring nodes, the combiner also achieves significant gain over the previous ones. For example, at BER = 10^{-4} the SNR gain is 4.5 dB for NCSI and 6 dB for CSIF selection. Also noted from the figure that the MSE-CSIF selection provides 1.5 dB SNR gain higher than the MSE-NCSI selection. This gain can compensate for the required modification in the control frame structure.

From Fig. 3, we can also see that there is a cross between the BER curve of the proposed combiner without relay selection with that of the previous combiner utilizing the MSE relay selection algorithm. This could be qualitatively explained as follows. In the low SNR region, the proposed combiner has higher capability of cancelling CCIs as compared to the previous combiner. Thus, it provides the destination with lower BER. On the other hand, as SNR increases, BER performance of the destination is mainly determined by the achievable diversity order. Because the previous combiner adopts the relay selection algorithm, it allows the destination to achieve higher diversity order, and hence higher BER performance in the high SNR region.



(a) MSE-SCIF selection.



(b) MSE-NCSI selection.

Fig. 4. BER performance comparison as the number of neighbouring nodes increase, N = 2, BPSK modulation, $\gamma_{sd} = \gamma_{sk} = \gamma_{kd}$, and perfect CSI.

6.2. *Performance under increased number of neighbouring nodes*

We now explore the effect of the two relay selection algorithms on the system performance as the number of candidate relays increases. Simulated BER performance of the MSE-CSIF and the MSE-NCSI using BPSK is shown in Fig. 4(a) and (b), respectively. In our simulations, the number of candidate relays is increasingly selected from the set $\{1, 2, 4, 6, 8, 10, 12\}$. It can be clearly seen from the two figures that the BER performance is remarkably improved when there are more candidate relays distributed between the source and the destination. A common trend can be observed for the two algorithms is that within the SNR range of interest the

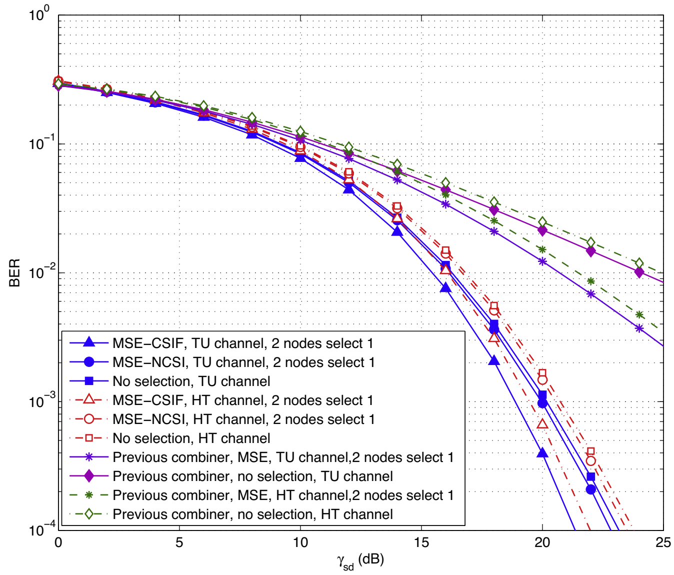


Fig. 5. BER performance under frequency-selective fading channels using OFDM, $N=4$, one relay node, $\gamma_{sd}=\gamma_{sk}=\gamma_{kd}$, 16-QAM modulation, and COST-207 channel model.

improvement in the system performance becomes less significant when the number of candidate relays reaches to 10 nodes.

6.3. *Performance evaluation over frequency-selective fading channel*

In this section we evaluate the performance of the proposed system under frequency-selective fading channel. Two typical channels, namely, Typical Urban (TU) and Hilly Terrain (HT) according to COST-207 channel model [17], where each path is modelled as an independent and uncorrelated Rayleigh fading channel with appropriate amplitudes, are evaluated. In the simulation, $N=4$ and 16-QAM mapping are used. The system uses the orthogonal frequency division multiplexing (OFDM) for multicarrier modulation with assumption that the cyclic prefix (CP) length is larger than the maximum delay of the channels, i.e., $L_p=20$. Under this assumption the cooperative MIMO-SDM-OFDM system can be decomposed into parallel sub-systems affected by flat fading sub-channels. BER performance was simulated for one typical sub-channel. Comparing the curves in Fig. 5 we can see that by using OFDM the proposed system outperforms the previous one significantly under both TU and HT channels. It can also be seen that the BER performance over the TU channel is better than that over the HT channel. The observation can be justified using the fact that the total energy of the TU channel is slightly large than that of the HT channel. The proposed system is thus particularly suitable for implementation in practical systems.

7. Conclusions

In this paper, we have proposed an MMSE combiner and two relay selection algorithms, called MSE-CSIF and MSE-NCSI, for MIMO-SDM cooperative communication networks. In the MSE-CSIF selection approach, a neighbouring node needs to perfectly know the CSIs of the communication links not only from the source to it and from it to the destination but also from the source to the destination. In the MSE-NCSI selection one, nevertheless, the requirement of knowing the CSI from the source to the destination is relaxed. It was shown that thanks to knowing the CSI between the source and destination, the MSE-CSIF algorithm was able to offer higher BER performance than the MSE-NCSI one for the same number of neighbouring nodes. Simulation results also showed that without relay selection, the proposed MMSE combiner provides noticeable performance improvement compared to its former counterpart in both flat fading and frequency-selective fading channels, yet at the cost of higher computational complexity. Moreover, the proposed selection algorithms together with the proposed MMSE combiner were demonstrated to significantly outperform the schemes previously proposed in [16] since they were able to attain higher diversity order.

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