Coding for the MIMO multi-hop amplify-and-forward fading channel

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***Abstract*— In this paper, we study coding techniques for the multiple-input multiple-output multi-hop amplify-and-forward (MIMO-MH-AF) block-fading channel. We first derive conditions on coding rates and space-time precoders that allow a coded modulation to achieve the available diversity of the channel. We then derive a bound on the diversity order of a coded modulation so as to attain maximum diversity through parallel partitioning of the channel. Finally, word error rate performances are compared to outage probabilities.**

I. Introduction

Transmission over wireless channels suffers from severe degradations due to effects such as path loss, shadowing, fading, and interference from other transmitters. When fading varies slowly, one way to achieve reliable communication is to provide signal diversity in either time, frequency, or space [1]. For this purpose, multiple-antenna systems that provide spatial diversity together with high capacity have been extensively studied [2]. More recently, based on the seminal works in [3] and [4] on relay channels, the authors in [5][6] provide a framework for *cooperative communications*, where multiple terminals use the resources of each other to form a virtual antenna array. The main protocols that have been proposed are the *amplify-and-forward* protocol, where relays amplify the signals received from other terminals before retransmitting, and the *decode-and-forward* protocol, where relays decode the received signals before retransmitting. In networks where sources and destinations have no direct links, wireless multi-hop techniques where data is conveyed by short consecutive transmissions through terminals have been considered. Multi-hop techniques can avoid the deployment of wired backhaul links in networks, and they can help increase the coverage of a network. When employing amplify-and-forward cooperation, they can enhance the throughput with short hops and extend battery life of terminals due to lower power consumption. Capacity issues of large relay networks have been extensively addressed [7][8][9], but there are few works that consider diversity analysis of wireless multi-hop networks. Among the works on MIMO multi-hop channels, the authors in [10] [11] propose distributed space-time coding techniques for the MIMO two-hop channel, and the authors in [12] provide a thorough diversity-multiplexing tradeoff [13] analysis of the MIMO multi-hop channel, and they also propose methods to achieve the highest possible diversity available in the channel. In this paper, we study the performance of space-time bit-interleaved coded modulations (ST-BICM) schemes [14] [15] for the MIMO multi-hop amplify-and-forward (MIMO-MH-AF) channels. Based on the work in [12], we propose conditions and bounds on the coding rate and space-time spreading factor for a ST-BICM to achieve the highest possible diversity of the channel. The paper is organized as follows: Section II gives the system model and notations, and Section III gives the conditions and bounds on the diversity attainable by a ST-BICM over the MIMO-MH-AF channel. Section IV show word error rate performances, and Section V gives the concluding remarks.

II. System Model

We consider transmission from a source with n_0 antennas to a destination with n_β antennas through $\beta-1$ relays having each n_i antennas, $i = 1 \cdots \beta - 1$. Each relay scales and retransmits the symbols it received to the following relay until the information is conveyed to the destination in β hops, thus direct source-destination transmission is not considered. We consider full-duplex transmission, where terminals can transmit and receive simultaneously, knowing that half-duplex nodes only introduce delay without affecting the performance. The channel between any two nodes is block-fading, *i.e.* a codeword spans one temporal channel realization, and the channel realizations are only known to the destination. The vector of transmitted symbols $\mathbf{x}_0$ at the source is given by:

$$\mathbf{x}_0 = \mathbf{z}\mathbf{S} \quad (1)$$

where $\mathbf{z}$ is the length-sn_0 vector of 2^m-Quadrature Amplitude Modulation (QAM) modulation symbols, and $\mathbf{S}$ is a $sn_0 \times sn_0$ space-time rotation with time spreading s. In the absence of space-time precoding, we have that $s = 1$ and $\mathbf{S}$ is the $n_0 \times n_0$ identity matrix. The received signal at node i is written as:

$$\mathbf{y}_i = \mathbf{x}_{i-1}\mathbf{H}_{i-1} + \eta_i, \quad i = 1 \cdots \beta - 1 \quad (2)$$

where $\mathbf{y}_i$ is the length-sn_i vector of complex symbols received at node i, $\mathbf{x}_{i-1}$ is the length-sn_{i-1} vector of complex symbols transmitted from node $i-1$, $\mathbf{H}_{i-1}$ is the $sn_{i-1} \times sn_i$ matrix of complex Gaussian fading coefficients with zero mean and variance $1/2$, and η_i is the length-sn_i vector of circularly symmetric complex additive white Gaussian noise (AWGN) coefficients with zero mean and variance $2N_0$. At each relay, the $sn_i \times sn_i$ complex scaling matrix $\mathbf{B}_i$ performs the following operation :

$$\mathbf{x}_i = \mathbf{y}_i \mathbf{B}_i \tag{3}$$

subject to the power constraint $\mathbb{E}\left(\|\mathbf{y}_i \mathbf{B}_i\|^2\right) \leq sn_i$. The channel from the source (node 0) to the destination (node β) is then given as:

$$\mathbf{y}_\beta = \mathbf{x}_0 \mathcal{H} + \eta_c \tag{4}$$

with:

$$\mathcal{H} = \left(\prod_{i=1}^{\beta} \mathbf{H}_i \mathbf{B}_i \right) \tag{5}$$

is of dimensions $n_0 \times n_\beta$ and

$$\eta_c = \eta_j \sum_{j=1}^{\beta} \left(\prod_{i=j}^{\beta} \mathbf{H}_{i+1} \mathbf{B}_i \right) \tag{6}$$

Now let $\mathbf{\Gamma}$ be the covariance matrix of the colored noise, given by:

$$\mathbf{\Gamma} = \mathbb{E}\left[\eta_c^\dagger \eta_c\right] = 2N_0 \mathbf{\Theta} \tag{7}$$

where the † operator denotes transpose conjugate. By performing a Cholesky decomposition on $\mathbf{\Theta}$, we get:

$$\mathbf{\Theta} = \mathbf{\Psi}^\dagger \mathbf{\Psi} \tag{8}$$

Thus the equivalent channel model becomes:

$$\mathbf{y}_\beta \mathbf{\Psi}^{-1} = \mathbf{x}_0 \mathcal{H} \mathbf{\Psi}^{-1} + \eta \tag{9}$$

We consider a bit-interleaved coded modulation (BICM) scheme, that operates as shown in Fig. 1: a convolutional encoder of rate $R_c = M/N \leq 1$ encodes M information bits into N coded bits c_i that are first interleaved, then Gray mapped into QAM symbols, rotated using a space-time precoder and transmitted over the channel as in (4).

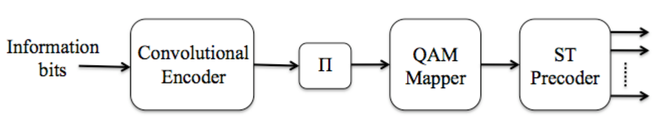

Fig. 1. ST-BICM encoder.

At the destination, the soft-input soft-output (SISO) *a posteriori* probability (APP) detector computes extrinsic information $\xi(c_i)$ on coded bits based on the channel matrix $\mathcal{H}$, the received vector $\mathbf{y}_\beta$, and *a priori* probabilities fed back from the SISO decoder, as shown in Fig. 2. The SISO decoder uses the forward-backward [16] algorithm to compute maximum *a posteriori* probabilities on information bits. The set of vectors Ω of size $|\Omega| = 2^{sn_0 m}$ generated by the QAM modulator determines the complexity at the detector. The transmitted information rate is given by $R = R_c n_0 m$.

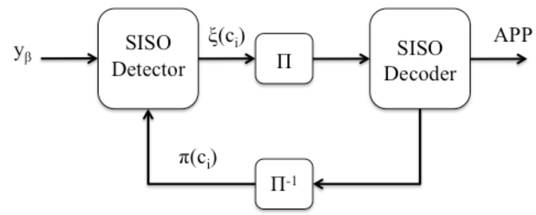

Fig. 2. ST-BICM iterative detector and decoder.

III. Diversity of coded modulations over the MIMO-MH-AF channel

We will start by recalling some results on the diversity of the MIMO-MH-AF scheme from [12]. Let the set $\hat{\mathbf{n}} = (\hat{n}_0, \hat{n}_1, \cdots, \hat{n}_\beta)$ be the set of antennas $\mathbf{n} = (n_0, n_1, \cdots, n_\beta)$ ordered in an increasing way. The achievable diversity d_{ach} of the MIMO-MH-AF channel is bounded as:

$$\frac{\hat{n}_0 (\hat{n}_1 + 1)}{2} \leq d_{ach} \leq d_{max} = \hat{n}_0 \hat{n}_1 \tag{10}$$

where:

$$d_{ach} = \sum_{i=1}^{\hat{n}_0} c_i \tag{11}$$

$$c_i = 1 - i + \min_{j=1,\cdots,\beta} \lfloor \frac{\sum_{\ell=0}^{j} \hat{n}_\ell - i}{j} \rfloor, \quad i = 1, \cdots, \hat{n}_0 \tag{12}$$

The highest achievable diversity is thus determined by the product of the two smallest antenna configurations. However, it is always strictly higher than half of the same product, whatever the number of hops is. In the case of the amplify-and-forward protocol, the upper bound on the diversity is rarely reached due to the mismatch between adjacent sub-channels. In fact, it can achieve the upper bound provided that the two relays with the lower antenna configurations are in the vicinity of each others, and that:

$$\hat{n}_2 \geq \hat{n}_0 + \hat{n}_1 - 1$$

In other words, the MIMO-MH-AF channel can be seen as a point-to-point MIMO $n_0 \times n_\beta$ channel whose diversity cannot exceed $\hat{n}_0 \hat{n}_1$.

A. Achieving the achievable diversity

1) Non-precoded MIMO-MH-AF channel: Following [17], we define the *BO-channel* as the binary-oriented channel with input c_i and output $\xi(c_i)$ seen by the code. In addition, according to *Definition 1* in [17], under perfect *a priori* information in the BO-channel, the number of non-ergodic fading sub-channels is denoted by D_{st} and called the *state diversity*. Following this definition, the state diversity $d_{st}(c)$ achieved by a codeword c is the number of non-zero partial Hamming weights ω_i, where $\sum_{i=1}^{D_{st}} \omega_i = \omega_H(c)$ is the Hamming weight of the code. Without space-time precoding of the MIMO multi-hop channel, the number of the fading sub-channels of the BO-channel corresponds to the number of transmit antennas at the source, *i.e.* $D_{st} = n_0$. In addition, an interleaver is said to be optimal if it is capable of placing the bits of an error event as much uniformly as possible on the different channel states and on different time slots [15] [14]. Hence, under optimal interleaving, the condition on the coding rate R_c in order to achieve the state diversity D_{st} of a coded modulation transmitted over a non-precoded MIMO-MH-AF channel is given by:

$$d_{st} = D_{st} \quad \Longleftrightarrow \quad R_c \leq \frac{1}{D_{st}} = \frac{1}{n_0} \tag{13}$$

In fact, as the interleaver is optimal, for any pair of codewords (c, c'), the ω non-zero bits of $c - c'$ are transmitted over the D_{st} states and over different time periods. The interleaving, modulation and transmission through the channel convert the codewords c and c' onto points $\mathcal{C}$ and $\mathcal{C}'$ in a Euclidean space. For a fixed channel, the performance is directly linked to the Euclidean squared distance $|\mathcal{C} - \mathcal{C}'|^2$, that can be rewritten as a sum of ω squared Euclidean distances associated to the ω non-zero bits of $c - c'$. For each of the ω squared Euclidean distances, we can build an equivalent channel model which corresponds to the transmission of a BPSK modulation over one row of the channel matrix $\mathcal{H}$. Thus, several squared Euclidean distances appear to be transmitted on the same equivalent channel and the squared distance $|\mathcal{C} - \mathcal{C}'|^2$ can be factorized as follows: $|\mathcal{C} - \mathcal{C}'|^2 = \sum_{i=1}^{n_\beta} d_i^2$ where d_i^2 is linearly dependent on the norm of the i-th row of $\mathcal{H}$. For the decoder to recover the achievable diversity $d_{ach} = \hat{n}_0 \hat{n}_1$, the coded bits should be sent over the $D_{st} = n_0$ channel states. Hence, the performance of coded modulations over the MIMO-MH-AF channel depends on the number of antennas available at the source, while with uncoded systems transmission rate is limited by the smallest antenna configuration $\hat{n}_0$.

2) Precoded MIMO-MH-AF channel: Now let us suppose that a space-time precoder takes sn_0 modulated symbols, rotates them, and sends their combinations over the n_0 transmit antennas over s time periods. In this case, we have that $D_{st} = n_0/s$. The condition on the coding rate R_c in order to achieve full state diversity is as follows:

$$d_{st} = D_{st} \quad \Longleftrightarrow \quad R_c \leq \frac{1}{D_{st}} = \frac{s}{n_0} \tag{14}$$

As with the point-to-point MIMO channel [14], space-time rotations with full spreading (*i.e.* $s = n_0$) lead to $D_{st} = 1$ and full transmission diversity whatever the coding rate is. However, for some ranges of coding rates, tuning of the time spreading can lead to full diversity with lower detection complexity. By reducing the number of states seen by the code, coding rates that achieve full diversity are increased. As an example, we consider transmission over the four-hop $(4, 2, 2, 2, 3)$ channel. The available diversity of the channel is $d_{ach} = 3$ achievable without space-time precoding with $R_c = 1/4$, as $D_{st} = 4$. The rate can be increased to $R_c = 1/2$ with a space-time rotation with $s = 2$ that reduces state diversity to $D_{st} = 2$ and allows to achieve the highest possible diversity of the channel.

3) Vertical reduction of the MIMO-MH-AF channel: Another way for a coded modulation scheme to attain the highest possible diversity d_{ach} with higher coding rates is by vertical reduction [12] at the source, defined as follows: Let $\mathbf{n} = (n_0, n_1, \cdots, n_\beta)$ represent the set of antennas of a MIMO multi-hop block-fading channel. The channel represented as $\mathbf{n}' = \left(n_0', n_1', \cdots, n_\beta'\right)$ is a vertical reduction of $\mathbf{n}$ if $d_{ach}(\mathbf{n}) = d_{ach}(\mathbf{n}')$ and $n_i' \leq n_i$. In order to attain full diversity with higher coding rates, it is thus beneficial to perform vertical reduction to decrease the number of channel states at the source, *i.e.* finding $n_0' < n_0$ with $d_{ach}(\mathbf{n}) = d_{ach}(\mathbf{n}')$. In this case, we have that $D_{st}' = n_0'/s$ and:

$$d_{st} = D_{st}' \quad \Longleftrightarrow \quad R_c \leq \frac{1}{D_{st}'} = \frac{s}{n_0'} \tag{15}$$

Again, considering the same four-hop $(4, 2, 2, 2, 3)$ channel, only 2 out of the 4 transmit antennas can be used, allowing a coding rate $R_c = 1/2$ to achieve $d_{ach} = 3$ without precoding over the $(2, 2, 2, 2, 3)$ channel.

B. Achieving the maximal diversity: parallel partitioning of the MIMO-MH-AF channel

In order to achieve the upper bound on the achievable diversity in (10), the authors in [12] proposed to partition the channel in different AF paths. The motivation is to introduce temporal processing in addition to the space processing inherent to MIMO systems, which is accomplished by partitioning the relays in each layer. The relays coordinate to convey the information following the predefined partition in a periodic way. To obtain full-dviersity independent parallel partitions, relays are chosen as to activate a subset of their antennas per transmission time as to ensure that any two different AF paths do not share edges. The number of partitions one can carve from a MIMO-MH-AF channel and their sizes varies depending on the channel configuration. The channel model with partitioning is thus given by:

$$\mathbf{y}_\beta = \mathbf{x}_0 \mathcal{H}_{pp} + \eta_c \tag{16}$$

with:

$$\mathcal{H}_{pp} = diag\left(\underbrace{\mathcal{H}_1, \ldots, \mathcal{H}_1}_{\alpha}, \ldots, \mathcal{H}_K, \ldots, \mathcal{H}_K\right) \tag{17}$$

being the channel matrix. Here $\alpha = \frac{s}{K}$ if $s > 1$, and $\alpha = 1$ otherwise. The matrices $\mathcal{H}_k$ have dimensions $n_0 \times n_\beta$.

The MIMO-MH-AF channel with K parallel partitions having each $n_{k,0}$ transmit antennas is equivalent to an uncorrelated n_c-block fading channel with n_t transmit antennas [14]. We can thus obtain a modified Singleton bound [18] [15] on the achievable diversity order of a coded modulation under optimal interleaving as:

$$d \leq \min\left(\frac{sd_{max}}{Kn_{0,k}}\lfloor\frac{Kn_{0,k}}{s}(1-R_c)+1\rfloor, d_{ach}=d_{max}\right) \quad (18)$$

Note that optimal interleaving in this case implies that, in addition to uniformly distributing the bits of an error events on the different channel states and time instance, bits should be sent in different AF paths. To illustrate the bound in (18), we provide the following examples:

- Example 1 : Consider the channel $\mathbf{n} = (2,2,2)$, $d_{ach}=3$, achievable with $R_c = 1/2$. By decomposing the channel into 2 correlated channels with $\mathbf{n_k} = (1,2,2)$ each, a code with $R_c = 1/2$ can be used but the correlation between the paths still maintains the diversity order at $d_{ach}=3$. However, if we decompose it into 2 independent channels with $\mathbf{n_k} = (2,1,2)$ each, the diversity would be $d_{ach} = 2+2=4=d_{max}$, achievable with $R_c = 1/4$ and the same optimized interleaver for the 2×2 point-to-point MIMO channel with $n_c = 2$ blocks. Alternatively, a half-rate code with an $s=2$ space-time rotation achieves the maximal diversity as well.
- Example 2 : $\mathbf{n} = (3,3,3)$, $d_{ach} = 7$, achievable with $R_c = 1/3$. If we decompose it into 3 channels with $\mathbf{n_k} = (3,1,3)$ each, the diversity would be $d_{ach} = d_{max} = 9$, achievable with $R_c = 1/9$. However, using a rotation with $s=3$ can allow a code with $R_c = 1/3$ to achieve the maximal diversity.
- Example 3 : Consider the channel $\mathbf{n} = (2,2,2,3)$, $d_{ach} = 3$, achievable with $R_c = 1/2$. If we decompose it into 2 non-independent channels with $\mathbf{n_k} = (2,1,2,3)$ each, the diversity would be $d_{ach} = 2+2 = 4 = d_{max}$, achievable with $R_c = 1/4$. This is possible because the channels starting from the dispatching level, namely $\mathbf{n_1''} = (2,2,3)$ and $\mathbf{n_2''} = (2,2)$ have diversity order $d_{max} = 4$.

In fact, achieving maximum diversity does not require having independent parallel partitions. In some cases, full diversity can be achieved with a non-independent parallel partition, conditioned on the fact that if the partition occurs at relay i, the diversity of the k^{th} AF path is:

$$d_{max,k} = \min\{d_{(n_{k,0},\cdots,n_{k,i})}, d_{(n_{k,i},\cdots,n_{k,\beta})}\} \quad (19)$$

Although achieving full diversity, parallel partitions suffer from high outage probability, as by choosing subsets of antennas to operate at relays the other antennas are wasted resources. An alternative is the so-called “flip-and-foward” (FF) protocol where rotation matrices are applied at relays to create parallel partitions while using all the antennas. To illustrate this protocol, we consider the two-hop channel with $\mathbf{n} = (2,2,2)$. We can decompose this channel into two parallel channels $\mathcal{H}_1$ and $\mathcal{H}_2$ given by:

$$\mathcal{H}_1 = \mathbf{H}_1\mathbf{P}_1\mathbf{H}_2 \quad (20)$$

$$\mathcal{H}_2 = \mathbf{H}_1\mathbf{P}_2\mathbf{H}_2 \quad (21)$$

To obtain independent parallel partitions, we set:

$$\mathbf{P}_1 = \begin{bmatrix} 1 & 0 \\ 0 & 0 \end{bmatrix} \quad (22)$$

and

$$\mathbf{P}_2 = \begin{bmatrix} 0 & 0 \\ 0 & 1 \end{bmatrix} \quad (23)$$

With the FF protocol, we have that :

$$\mathbf{P}_1 = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \quad (24)$$

and

$$\mathbf{P}_2 = \begin{bmatrix} 1 & 0 \\ 0 & -1 \end{bmatrix} \quad (25)$$

By using all the antenna resources in all AF paths, the FF scheme is shown to be superior in terms of outage probability.

IV. Simulation results

In this section, word error rate performance of coded modulations over the MIMO-MH-AF channel is compared to outage probability [19] for $1000 < N < 1500$. We consider non-recursive non-systematic convolutional (NRNSC) codes and interleavers from [14] that are optimal for point-to-point MIMO block-fading channels. The space-time rotations are the Dispersive Nucleo Algebraic (DNA) precoders from [14] that exist for all antenna configurations and spreading factors and that are optimal with iterative detection and decoding. In Fig. 3, results for the two-hop MIMO AF channel with four-state NRNSC codes and BPSK modulation are shown. The initial channel is the $(4,2,2)$ channel with $d_{ach} = d_{max} = 4$. With half-rate coding and without precoding, maximal diversity cannot be guaranteed. In order to attain full diversity, a code with $R_c = 1/4$ should be used if no precoding is available. However, a space-time rotation with $s = 2$ is needed to achieve full diversity with a half-rate code. Another way to achieve full diversity is by vertical reduction, *i.e.* by sending symbols over three out of the four available antennas at the source. As the $(3,2,2)$ and the $(4,2,2)$ MIMO AF channels have the same diversity order, a code with $R_c = 1/3$ is sufficient to achieve full diversity. In Fig. 4, error rate performance for the two-hop $(2,2,2)$ MIMO AF channel with BPSK modulation is shown. The half-rate code is the $(133,171)_8$ 64-state NRNSC code, and the $R_c = 1/4$ code is the $(13,17,15,11)_8$ four-state code. The diversity of the $(2,2,2)$ channel is $d_{ach} = 3 < d_{max} = 4$, that is achieved with or without precoding using half-rate coding. In order to achieve full diversity, partitioning is required using either the parallel partition or the flip-and-forward protocol, that is similar to the parallel partitioning in terms of achievable diversity orders but has better coding gains. Maximum diversity is thus attained with two AF paths using either a $R_c = 1/4$ code or a half-rate code with a $s = 2$ space-time rotation according to (18). Fig. 5 shows the performance over the three-hop $(2,2,2,3)$ MIMO AF channel with four-state NRNSC codes and BPSK modulation. Without partitioning, the diversity $d_{ach} = 3$ of the channel is achievable with half-rate coding. In order to attain $d_{max} = 4$, a partition with non-independent AF paths is sufficient, as explained in Example 3 of Section III-B. Coded modulations with half-rate coding and $s = 2$ rotation or alternatively with $R_c = 1/4$ achieve this upper bound.

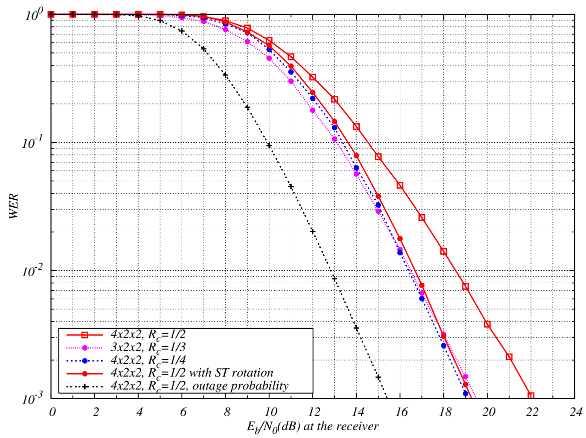

Fig. 3. Two-hop MIMO-AF channel, BPSK modulation, NRNSC codes.

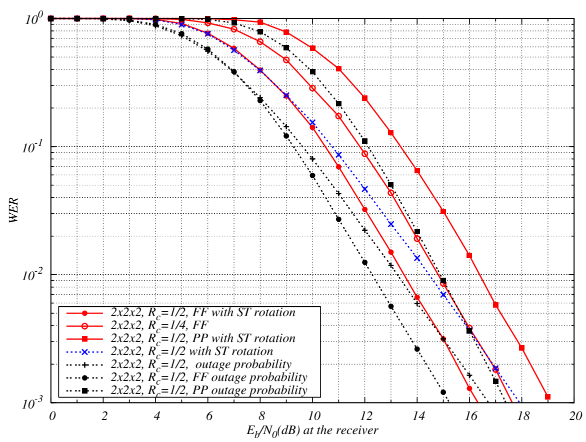

Fig. 4. Two-hop MIMO-AF channel, QPSK modulation, NRNSC codes.

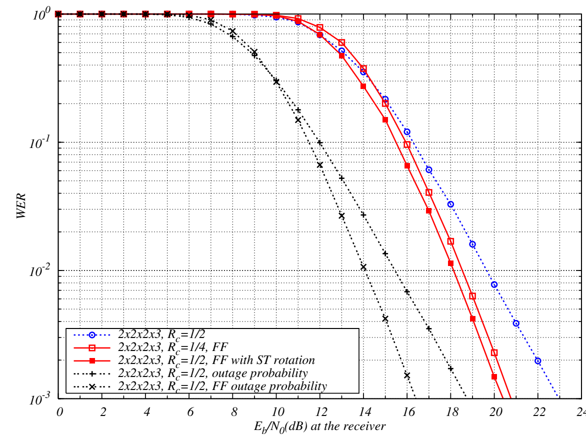

Fig. 5. Three-hop MIMO-AF channel, BPSK modulation, NRNSC codes.

V. Conclusions

We studied coding strategies for the multiple-antenna multi-hop amplify-and-forward channel. We proposed conditions on coding rates to achieve full-diversity over these channels. Moreoever, we derived a bound on diversity orders a coded modulation can achieve with low decoding complexity over the paritioned multi-hop channel. Finally, performances close to outage probabilities for different channel configurations, coding rates, and constellation sizes are shown.

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