OSTBC Transmission in MIMO AF Relaying with $M$-FSK Modulation

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Abstract—This paper investigates multiple-input multiple-output (MIMO) amplify-and-forward (AF) relaying networks with orthogonal space-time block codes (OSTBC) and $M$-ary frequency-shift-keying (FSK) modulation. A non-coherent detection scheme is proposed and analyzed in a situation where the fading channels undergo temporal correlation. Specifically, the destination employs the linear minimum mean square error (LMMSE) estimation algorithm to estimate the overall channels by properly exploiting the implicit pilot-symbol-assist property of FSK transmission. It then utilizes the maximal ratio combining (MRC) to detect the transmitted information. An upper bound on the probability of errors is derived for a network with arbitrary numbers of transceiver antennas and relays. Based on the obtained bit-error-rate, the full achievable diversity order is verified. Simulation results are presented to show the validity of the analytical results.

Index Terms—Cooperative diversity, relay communications, frequency-shift-keying, fading channel, amplify-and-forward protocol, multiple-input multiple-output, orthogonal space-time block codes.

I. INTRODUCTION

Noncoherent transmission techniques have received a lot of attention due to their potential improvement in complexity by eliminating the need of channel state information at the receiver. Consequently, employing those non-coherent techniques is preferable in wireless relay networks since many wireless fading channels are involved in the networks [1], [2], which makes the task of channel estimation very complex and expensive to implement.

In recent years, much more research works has focused on non-coherent wireless relay networks [3]–[10], i.e., the wireless relay networks in which channel state information (CSI) is assumed to be unknown at the receivers (relays and destination). Among them, non-coherent amplify-and-forward (AF) has received more attention since it further puts a less processing burden on the relays due to the AF protocol [3]–[10]. However, only suboptimal non-coherent AF receivers have been studied due to the complicated deployment in practice [5], [8]. Especially, when the channels undergo temporally-correlated Rayleigh flat fading, reference [9] is the only work to investigate a non-coherent detection technique for amplify-and-forward (AF) relay networks. It would be emphasized that all the above-mentioned works study a network in which all nodes are equipped with a single antenna.

Multiple-input multiple-output (MIMO) relaying techniques, that use multiple antennas at all nodes in the network, have been known to improve considerably performance in terms of data transmission rates as well as reliability over wireless channels. However, most existing works study a coherent network, i.e., CSIs of all the transmission links are available at the receivers to perform a detection [11]. Hence, this paper studies non-coherent MIMO relaying networks to make the MIMO relaying techniques more applicable. In fact, the work in [12] preliminarily develops a detection framework for multi-antenna AF relay networks. However, the work only considers a special network with two-antenna equipped source and single-antenna equipped relays and destination. An Alamouti space-time block code is employed for this special case.

This work studies a more generalized multi-antenna AF relay network, i.e., all the nodes in the network are equipped with multiple antennas. OSTBC with $M$-FSK modulation is employed at the source to transmit a signal to the destination. Basically, the detection technique employed in this work is similar in [9], [12]. The destination first estimates the overall channels by using LMMSE estimation algorithm. Then it employs the maximal ratio combining (MRC) to detect the transmitted information. The main contribution of this paper is to develop a general framework for a network with arbitrary numbers of nodes and arbitrary numbers of transceiver antennas equipped at each node. Moreover, a unified upper-bound BER expression is derived. It is further shown that the proposed detection scheme achieves a full diversity order.

II. SYSTEM MODEL

Consider a wireless relay network in which the source, denoted by node 0, communicates with the destination, denoted by node $K+1$, with the assistance of $K$ half-duplex relays, denoted by node $i$, $i = 1, \ldots, K$. It is assumed that the $K$ relays retransmit signals to the destination over orthogonal channels. All the nodes are MIMO devices, i.e., node $i$ is equipped with $N_i$ antennas. Assume that the transmit and receive antennas at a relay node are the same. An orthogonal space-time block code is employed at the source to transmit the signal to the destination. Fixed-gain AF protocol is employed at the relays. In the first phase, i.e., $T_c$ time slots, the source
broadcasts an OSTBC designed for $N_0$ antennas to the relays and destination. In the second phase, i.e., the next $\sum_{i=1}^{K} N_i T_c$ time slots\(^1\), the relays amplify the received signals and forward to the destination. The destination then estimate the overall channels of all the links from the source to the destination and performs a MRC with $\left( \sum_{i=1}^{K} N_i + 1 \right) N_{K+1} T_c$ received signals for the final detection decision based on the estimated overall channels.

For convenience, we adopt the convention that epoch $k$ is a period of time to complete a transmission from the source to the destination. With the above-mentioned transmission protocol, epoch $k$ starts at $t = k(\sum_{i=1}^{K} N_i + 1)T_c T$ and ends at $(k+1)(\sum_{i=1}^{K} N_i + 1)T_c T$ where $T$ is the symbol duration (or time slot duration). The channel fading coefficient between the $m$th transmit antenna of node $i$ and the $n$th receive antenna of node $j$ at epoch $k$ is denoted by $h_{i,j,m,n}[k]$. Those channel coefficients are modeled as circularly symmetric complex Gaussian random variables and assumed to be constant over $\sum_{i=1}^{K} N_i + 1$ time slots but varies dependently every period of $\sum_{i=1}^{K} N_i + 1$ time slots. The temporally-correlated fading environment is modeled with the following Jake’s autocorrelation

$$\phi_{i,j,m,n}[p] = (\sigma_{i,j,m,n}^2)^2 J_0 \left( 2\pi f_{i,j,m,n}^p \right) \quad (1)$$

where $f_{i,j,m,n}$ and $(\sigma_{i,j,m,n}^2)^2$ are the maximum Doppler frequency and the average signal strength of the channel corresponding the connection between the $m$th transmit antenna of node $i$ and the $n$th receive antenna of node $j$, respectively.

The received signal during the $l$th ($l = 1, \ldots, T_c$) time slot of the first phase at the $n$th antenna of node $j$, $n = 1, \ldots, N_i$, $j = 1, \ldots, K + 1$, at epoch $k$ is written as

$$y_{0,j}^{n,l}(t) = \frac{E_0}{N_0} \sum_{m=1}^{N_0} h_{0,j,m,n}^{m,n}[k] x_j^m(t) + w_{0,j}^{n,l}(t), \quad t \in T_{k,l} \quad (2)$$

where $T_{k,l} = [k(\sum_{i=1}^{K} N_i + 1)T_c T + (l-1)T, k(\sum_{i=1}^{K} N_i + 1)T_c T + lT]$ denotes the interval of time slot $l$ of epoch $k$, $w_{0,j}^{n,l}(t)$ is zero-mean additive white Gaussian noise (AWGN) at the $n$th antenna of node $j$ with two-sided power spectral density (PSD) of $M_0/2$ during the $l$th time slot. In the above expression, $E_0$ represents the average symbol energy available at the source and $x_j^m(t)$ is the transmitted waveform sent from the $j$th antenna of node 0 during time slot $l$. This waveform is chosen from an $M$-ary FSK constellation and therefore is written in complex baseband as

$$x_j^m(t) = \frac{1}{\sqrt{T}} \exp \left( j2\pi f^m t \right), \quad m = 1, \ldots, M$$

The following amplifying factor is chosen at the $n$th antenna of relay node $j$ before retransmitting:

$$\beta_j^n = \sqrt{\frac{E_j/N_j}{E_0/N_0 \sum_{m=1}^{N_0} (\sigma_{0,j,m,n}^2)^2 + M_0}}, \quad (3)$$

where $E_j$ is the average transmitted symbol energy allocated to node $j$. The received signal at the $g$th antenna of the destination via the $n$th antenna of relay node $j$ at epoch $k$, i.e., during the time interval $t \in T_{k,g}$, $l = T_c + 1, T_c + 2, \ldots, (\sum_{i=1}^{K} N_i + 1)T_c$, can be written as

$$y_{g,j}^{n,l}(t) = \frac{N_0}{\beta_j^n} \sqrt{E_0 h_{0,j,m,n}^{m,n}[k] x_j^m(t) - (\sum_{i=1}^{K} N_i + 1)T_c (t - (l - (\sum_{i=1}^{K} N_i + 1)T_c T) + w_{0,j}^{n,g,l}(t) \quad (4)$$

where $h_{0,j,m,n}^{m,n}[k] = h_{0,j,m,n}[k]$ is the overall channels from the $m$th antenna of the source to the $g$th antenna of the destination via the $n$th antenna of node $j$ at epoch $k$. The waveform $w_{0,j}^{n,g,l}(t)$ is the total additive noise corrupting the received signal. The noise $w_{0,j}^{n,g,l}(t)$ is also a zero-mean AWGN with two-side PSD of $M_0/2$.

### III. DATA DETECTION

#### A. Channel Estimation

The destination correlates the received signals in (2) and (4) with the following sum waveform $r(t)$ to estimate the overall channels [9], [13]:

$$r(t) = \sum_{l=1}^{M} \frac{2}{\sqrt{T}} \cos \left( (2l - 1) \pi t \right) \quad (5)$$

The output of the correlators can be stacked and re-organized as

$$\chi_{0,K+1}(t) = \frac{E_0}{N_0} X_{0,K+1}^{(E)} h_{0,K+1} + w_{0,K+1}^{(E)}$$

where the channel vectors $h_{0,K+1} \in \mathbb{C}^{N_0 N_{K+1} \times 1}$ and $w_{0,K+1}^{(E)} \in \mathbb{C}^{N_0 N_{K+1} \times 1}$ are

$$h_{0,K+1} = \left( h_{0,0,1}^{1,1}, \ldots, h_{0,0,1}^{N_0,1}, \ldots, h_{0,N_{K+1},1}^{1,1} \right)^T,$$

$$w_{0,K+1}^{(E)} = \left( w_{0,0,1}^{1,1}, \ldots, w_{0,0,1}^{N_0,1}, \ldots, w_{0,N_{K+1},1}^{1,1} \right)^T.$$
Meanwhile, the signal vectors $y_{0,K+1}^{(E)} \in \mathbb{C}^{N_{K+1}T_{c}\times 1}$ and $y_{j,K+1}^{(E)} \in \mathbb{C}^{N_{j}N_{K+1}T_{c}\times 1}$ are

$$y_{0,K+1}^{(E)} = \begin{pmatrix} y_{0,K+1}^{1} & \ldots & y_{0,K+1}^{N_{K+1}} \end{pmatrix}^{\top},$$

$$y_{j,K+1}^{(E)} = \begin{pmatrix} y_{j,K+1}^{1} & \ldots & y_{j,K+1}^{N_{K+1}} \end{pmatrix}^{\top}.$$

On the other hand, $X_{0,K+1}^{(E)}$ and $X_{j,K+1}^{(E)}$ are defined as two $N_{K+1}$ and $N_{j}N_{K+1}$ block diagonal matrices, respectively, i.e.,

$$X_{0,K+1}^{(E)} = \text{diag}\{X_{0},X_{0},\ldots\},$$

$$X_{j,K+1}^{(E)} = \text{diag}\{\beta_{j}X_{0},\beta_{j}X_{0},\ldots\} \in \mathbb{C}^{N_{K+1}\times 1} \text{ elements},$$

where $X_{0}$ is the $T_{c} \times N_{0}$ matrix code with the elements of 1 or -1. For example, if an Alamouti code is employed at the source, then $X_{0} = \begin{pmatrix} 1 & -1 \end{pmatrix}$.

Using LMMSE estimators, the LMMSE estimations of $y_{0,K+1}^{(E)}$ and $y_{j,K+1}^{(E)}$ can be obtained as follows [14]:

$$\hat{h}_{0,K+1} = \Psi_{0,K+1}^{(E)} \left( \Phi_{0,K+1}^{(E)} \right)^{-1} \hat{y}_{0,K+1},$$

$$\hat{h}_{0,j,K+1} = \Psi_{0,j,K+1}^{(E)} \left( \Phi_{0,j,K+1}^{(E)} \right)^{-1} \hat{y}_{0,j,K+1}.$$

In the above expressions, $\Psi_{0,K+1}^{(E)} \in \mathbb{C}^{(2P+1)N_{K+1}T_{c}\times 1}$ and $\Psi_{0,j,K+1}^{(E)} \in \mathbb{C}^{(2P+1)N_{j}N_{K+1}T_{c}\times 1}$, $j = 1,\ldots,K$, are formed by stacking $2P + 1$ consecutive vectors $y_{0,K+1}^{(E)}[k+l]$ and $y_{0,j,K+1}^{(E)}[k+l]$, $l = 1,\ldots,P$, respectively. $\Psi_{0,K+1}^{(E)}$ and $\Psi_{0,j,K+1}^{(E)}$ are the auto-correlation and cross-correlation matrices of $y^{(E)}$. As mentioned in [9], there is a trade-off between complexity and performance, i.e., increasing $P$ may improve the performance but also increase the complexity.

The matrices $\Psi_{0,K+1}^{(E)}$ and $\Psi_{0,j,K+1}^{(E)}$ are computed, respectively, as follows

$$\Psi_{0,K+1}^{(E)} = \sqrt{E_{0}} N_{0} \Theta_{0,K+1}[k-P] \Theta_{0,K+1}[k-P+1] \ldots \Theta_{0,K+1}[k+P],$$

$$\Psi_{0,j,K+1}^{(E)} = \sqrt{E_{0}} N_{0} \Theta_{0,j,K+1}[k-P] \Theta_{0,j,K+1}[k-P+1] \ldots \Theta_{0,j,K+1}[k+P],$$

where

$$\Theta_{0,K+1}[l] = \text{diag}\{\phi_{0,K+1}^{1}[l],\phi_{0,K+1}^{2}[l],\ldots\} \left( X_{0,K+1}^{(E)} \right)^{\top},$$

$$\Theta_{0,j,K+1}[l] = \text{diag}\{\phi_{0,j,K+1}^{1}[l],\phi_{0,j,K+1}^{2}[l],\ldots\} \left( X_{0,j,K+1}^{(E)} \right)^{\top}. \quad (8)$$

The estimation errors $e_{0,K+1}[k] = h_{0,K+1}[k] - \hat{h}_{0,K+1}[k]$ and $e_{0,j,K+1}[k] = h_{0,j,K+1}[k] - \hat{h}_{0,j,K+1}[k]$ are zero-mean with covariance matrices given as [14]

$$C_{e_{0,K+1}} = \mathbf{C}_{0,K+1},$$

$$C_{e_{0,j,K+1}} = \mathbf{C}_{0,j,K+1}.$$

The auto-correlation function of the overall channel $h_{0,j,K+1}^{(E)}$, is given as

$$(\hat{h}_{0,j,K+1}^{(E)})^{\top} \hat{y}_{0,j,K+1} \approx \mathbf{C}_{h_{0,j,K+1}^{(E)}}.$$
\[
\begin{align*}
\left( A^g_{0,K+1} \right)_{p,q} &= X \text{diag} \left\{ \frac{E_0}{N_0} \phi_{0,K+1}[p-q], \ldots, \frac{E_0}{N_0} \phi_{0,K+1}[p-q] \right\} X^t + MN_0 \delta[p-q] I_{N_0 \times N_0}, \\
\left( A^g_{0,j,K+1} \right)_{p,q} &= \text{diag} \left\{ \left( A^g_{0,j,K+1} \right)_{p,q}, \left( A^g_{0,j,K+1} \right)_{p,q}, \ldots, \left( A^g_{0,j,K+1} \right)_{p,q} \right\}, \\
\left( A^{n,g}_{0,j,K+1} \right)_{p,q} &= X \text{diag} \left\{ \left( \beta^n_j \right)^2 \frac{E_0}{N_0} \phi_{0,n,g}[p-q], \ldots, \left( \beta^n_j \right)^2 \frac{E_0}{N_0} \phi_{0,n,g}[p-q] \right\} X^t \\
&\quad + \left( \left( \beta^n_j \sigma^{n,g}_{0,j,K+1} \right)^2 MN_0 \delta[n-m] + MN_0 \delta[p-q] \right) I_{N_0 \times N_0}, \quad g = 1, \ldots, N, \quad j = 1, \ldots, J, \quad n = 1, \ldots, N_j. 
\end{align*}
\]

\[
y^{n,g,l}_{j,K+1}[k] = \sum_{m=1}^{N_0} \beta^n_j \sqrt{\frac{E_0}{N_0}} \left( \hat{I}^{n,g,l}_{0,j,K+1}[k] + \epsilon^{n,g,l}_{0,j,K+1}[k] \right) x^n_{m}[k] + w^{n,g,l}_{0,j,K+1}[k], 
\]

where \( x^n_{m}[k] \) is the \( M \times 1 \) vector that represents the transmit symbol from the \( m \)th antenna of node 0 at epoch \( k \). Note that \( x^n_{m}[k] \) has a 1 at an element and 0 at others. The elements of noise vectors \( w^{n,g,l}_{0,j,K+1}[k] \) and \( w^{n,g,l}_{0,j,K+1}[k] \) with size \( M \times 1 \) are i.i.d. zero-mean random variables with variance \( M_0 \) and \( \left( \beta^n_j \sigma^{n,g}_{0,j,K+1} \right)^2 + 1 \) \( M_0 \), respectively.

By using the property of complex orthogonal designs, one can stack and rewrite the input/output relations as

\[
\begin{align*}
\left( \begin{array}{c}
y^1_{0,K+1} \\
y^2_{0,K+1} \\
\vdots \\
y^{N_0}_{0,K+1}
\end{array} \right) &= \sqrt{\frac{E_0}{N_0}} \left( \begin{array}{c}
\hat{H}^1_{0,K+1} \\
\hat{H}^2_{0,K+1} \\
\vdots \\
\hat{H}^{N_0}_{0,K+1}
\end{array} \right) \mathbf{x} \\
&\quad + \sqrt{\frac{E_0}{N_0}} \left( \begin{array}{c}
w^1_{0,K+1} \\
w^2_{0,K+1} \\
\vdots \\
w^{N_0}_{0,K+1}
\end{array} \right), \\
\left( \begin{array}{c}
y^{n,1}_{0,j,K+1} \\
y^{n,2}_{0,j,K+1} \\
\vdots \\
y^{n,N_0}_{0,j,K+1}
\end{array} \right) &= \beta^n_j \sqrt{\frac{E_0}{N_0}} \left( \begin{array}{c}
\hat{H}^{n,1}_{0,j,K+1} \\
\hat{H}^{n,2}_{0,j,K+1} \\
\vdots \\
\hat{H}^{n,N_0}_{0,j,K+1}
\end{array} \right) \mathbf{x} \\
&\quad + \beta^n_j \sqrt{\frac{E_0}{N_0}} \left( \begin{array}{c}
w^{n,1}_{0,j,K+1} \\
w^{n,2}_{0,j,K+1} \\
\vdots \\
w^{n,N_0}_{0,j,K+1}
\end{array} \right),
\end{align*}
\]

where \( y^g_{0,K+1} = \left( y^g_{0,K+1} \ldots y^g_{0,K+1} \right)^t \in \mathbb{C}^{M_0 \times 1} \) and \( y^{n,g}_{0,j,K+1} = \left( y^{n,g}_{0,j,K+1} \ldots y^{n,g}_{0,j,K+1} \right)^t \in \mathbb{C}^{M_0 \times 1} \), represent the output of the correlators at the \( g \)th antenna of the destination. Similarly, \( w^g_{0,K+1} = \left( w^g_{0,K+1} \ldots w^g_{0,K+1} \right)^t \in \mathbb{C}^{M_0 \times 1} \) and \( w^{n,g}_{0,j,K+1} = \left( w^{n,g}_{0,j,K+1} \ldots w^{n,g}_{0,j,K+1} \right)^t \in \mathbb{C}^{M_0 \times 1} \) are the noise vectors. Note that \( \hat{y}^{g,l}_{0,K+1} = y^{g,l}_{0,K+1} \) or \( \hat{y}^{g,l}_{0,K+1} = \left( \hat{y}^{g,l}_{0,K+1} \right)^* \) depends on the structure of OSTBCs. It is similar to \( \tilde{y}^{n,g,l}_{0,K+1}, z^{n,g,l}_{0,K+1}, \hat{H}^g_{0,K+1} \) (or \( E^g_{0,K+1} \)) and \( \tilde{H}^{n,g}_{0,j,K+1} \) (or \( E^{n,g}_{0,j,K+1} \)) denote the \( M_T \times M_c \) matrices containing estimated channel gains (or channel estimation errors). Note that the matrices are uniquely obtained from any OSTBC.

The vector \( \mathbf{x}[k] \) is defined as \( \mathbf{x}[k] = (x_1[k] \ldots x_C[k]) \), where \( x_c[k], \ c = 1, \ldots, C \) is the \( M \times 1 \) vector that represents the \( c \)th data symbol that enter the OSTBC encoder at epoch \( k \). Note that \( x_c[k] \) is an unit vector.

Giving the estimated (overall) channels, the maximum SNR detector at the destination is of the following form

\[
\mathbf{r}[k] = \sum_{g=1}^{N_0} \left( \hat{H}^g_{0,K+1} \right)^H[k]\mathbf{y}^g_{0,K+1}[k]+ \sum_{j=1}^{K} \sum_{g=1}^{N_l} \sum_{n=1}^{N_i} \left( \hat{H}^{n,g}_{0,j,K+1} \right)^H[k]\mathbf{y}^{n,g}_{j,K+1}[k],
\]

where the combining weights are

\[
\varepsilon^g_{0,K+1} = \left( \frac{E_0}{N_0} \sum_{m=1}^{N_0} \left( \sigma^{m,n,g}_{0,j,K+1} \right)^2 + M_0 M_0 \right)^{-1}
\]

\[
\varepsilon^{n,g}_{j,K+1} = \left( \left( \beta^n_j \right)^2 \frac{E_0}{N_0} \sum_{m=1}^{N_0} \left( \sigma^{m,n,g}_{0,j,K+1} \right)^2 + M_0 M_0 \right)^{-1}
\]

Finally, due to the orthogonal property of OSTBCs, the transmitted symbols are decided by

\[
\tilde{m}, \tilde{x}_c[k] = \arg \max_{m=1, \ldots, M} \text{Re}(r[m+cM[k]], c = 1, \ldots, C)
\]

where \( r_i[k] \) is the \( i \)th element of the \( M \times 1 \) vector \( \mathbf{r}[k] \). \( \tilde{x}_c[k] \) is a \( M \times 1 \) unit vector with 1 at the \( m \)th element, i.e., the \( c \)th transmit waveform that enter the OSTBC encoder is decoded as using the \( m \)th \( M \)-FSK tone’s frequency.

**IV. UPPER BOUND ON BER PERFORMANCE AND DIVERSITY ORDER**

Since the decision rule in (37) is equivalent to the symbol-wise decision rule, i.e., each transmitted symbol can be decoded independently, the instantaneous SNR at the combiner’s
output can be written as
\[\gamma = \sum_{g=1}^{N_{K+1}} \gamma_{0,K+1}^g + \sum_{j=1}^{K} \sum_{n=1}^{N_j} \gamma_{j,K+1}^n\] (38)

where
\[\gamma_{0,K+1}^g = \frac{E_0}{N_0} \varepsilon_{0,K+1}^g \left( \sum_{m=1}^{N_0} \left| h_{0,m,K+1}^g \right|^2 \right)\] (39)
\[\gamma_{j,K+1}^n = \sum_{g=1}^{N_{K+1}} \frac{E_0}{N_0} \left( \beta_0^g \right)^2 \varepsilon_{j,K+1}^g \left( \sum_{m=1}^{N_0} \left| h_{j,m,n,K+1}^g \right|^2 \right)\] (40)

To simplify our analysis, we assume that BFSK is employed at the source, i.e., \(M = 2\) and the average signal strength between any two antennas of two particular nodes is identical, i.e., \(\left(\sigma_{i,j}^{m,n}\right)^2 = (\sigma_{i,j})^2, i = 1, \ldots, K\). It means that the average signal strength between any two antennas of the source-destination link via a relay is also identical, i.e., \(\left(\sigma_{i,j,K+1}^{m,n}\right)^2 = (\sigma_{i,j,K+1})^2\). The MGF of \(\gamma_{0,K+1}^g\) is
\[M_{\gamma_{0,K+1}^g}(s) = \left(1 + \frac{E_0}{N_0} (\sigma_{0,K+1}^g - \bar{\sigma}_{0,K+1}^g) \varepsilon_{0,K+1}^g s\right)^{-N_0}\] (41)

Meanwhile, the MGF of \(\gamma_{j,K+1}^n\) can be computed as Appendix A. Since \(\gamma_{0,K+1}^g\) and \(\gamma_{j,K+1}^n\) are statistically independent, the average BER for the OSTBC with BFSK in MIMO-AF relaying can be upper-bounded as
\[P_e \leq \frac{1}{\pi} \int_0^\pi M_{\gamma} \left( \frac{q}{\sin^2 \theta} \right) d\theta = \frac{1}{\pi} \int_0^\pi \prod_{g=1}^{N_{K+1}} M_{\gamma_{g}}(s) \left( \frac{q}{\sin^2 \theta} \right) d\theta\] (42)

where \(g = \frac{1}{2}\) for BFSK. One can obtain an upper-bound BER expression of the network by substituting (41) and the result from Appendix A into (42). Under the high SNR assumption, \(\bar{\sigma}_{0,K+1}^2 (i = 0, \ldots, K)\), and \(\bar{\sigma}_{0,j,K+1}^2 (i = 1, \ldots, K)\) approach 0. It then can be verified that a maximum diversity order of \(N_0N_{K+1}+\max\{N_0, N_{K+1}\}K\sum_{j=1}^{N_0} N_j\) is achieved.

V. SIMULATION RESULTS

In conducting the simulations, it is assumed that the source and relays have an equal transmit power, i.e., \(E_i = E, i = 0, \ldots, K\). The noise components at the receivers, i.e., relays and destination are modeled as i.i.d. \(CN(0,1)\) random variables. The path-loss follows the exponential-decay model, i.e., \(\left(\sigma_{i,j}^{(k)}\right)^2 = d_{i,j}^{-\nu}\) where \(d_{i,j}\) is the distance between node \(i\) and node \(j\). All the simulations are reported with the path loss exponent \(\nu = 4\). In addition, all the relays are assumed to have the same distances to the source and to the destination, i.e., \(d_{0,1} = d_{0,2} = \cdots = d_{0,K} = d_1, d_{1,1,K+1} = d_{2,K+1} = \cdots = d_{K,K+1} = d_2, \) and \(d_{0,K+1} = d_0\). The Doppler frequencies are set as \(10f_{0,i}T = f_{i,K+1}T = f_{0,K+1}T = 0.01, i = 1, \ldots, K\). BFSK modulation is employed at the source.

Fig. 1 shows the average BER of the proposed scheme by simulation for a single relay network with two-antenna equipped source and single-antenna equipped relay and destination. An Alamouti space-time block code is used at the source. One can observe the tightness of the derived upper-bound BER in (42) of the proposed scheme. Also, the diversity order of 3 is confirmed.³

³To our best knowledge, there are no state-of-the-art non-coherent detectors for MIMO AF relaying to compare with our scheme. The coherent detector is the only work that is close to our work. Therefore, to verify our proposed scheme, comparison with the coherent detector is considered.
space-time code for three transmit antennas is employed at the source. In this simulation, the relays are placed close to the source. The figure again confirms that one can get the estimations of the (overall) channels in MIMO AF relaying networks employed M-FSK modulation without the explicit pilot symbols to perform a detection. Note that the performance gap to the coherent scheme of the proposed scheme becomes smaller when P increases.

VI. CONCLUSION

A detection scheme for multiple-input multiple-output (MIMO) amplify-and-forward (AF) relaying networks has been proposed. The investigated networks compose of one source, K relays, and one destination. Orthogonal space-time block code (OSTBC) is employed at the source together with M-ary frequency-shift-keying (FSK) modulation to transmit the signals to the destination. By using the orthogonal property of FSK signalling, we have discussed an overall channel estimation method without the explicit pilot symbols. With the estimated overall channels, a maximal ratio combiner is employed to detect the transmitted information. An upper-bound expression on the probability of errors is obtained for the estimated overall channels, a maximal ratio combiner is employed to detect the transmitted information. An upper-bound expression on the probability of errors is obtained for the estimated overall channels, a maximal ratio combiner is employed to detect the transmitted information.

APPENDIX A

DERIVATION OF MGF OF $\hat{z}_{j,K+1}$ IN (40)

Let $Y = \sum_{g=1}^{N_{K+1}} \sum_{m=1}^{N_0} |\hat{h}_{0,j,K+1}|^2 = \left(\sum_{m=1}^{N_0} |\hat{h}_{0,j,0}|^2\right)^2 \left(\sum_{g=1}^{N_{K+1}} |\hat{h}_{j,g,K+1}|^2\right)^2 = X_1X_2$ where

$X_1 = \sum_{m=1}^{N_0} |\hat{h}_{0,j,0}|^2$ and $X_2 = \sum_{g=1}^{N_{K+1}} |\hat{h}_{j,g,K+1}|^2$.

Due to the fact that $|\hat{h}_{0,j,0}|^2$ and $|\hat{h}_{j,g,K+1}|^2$ are the estimates of $|h_{0,j,0}|^2$ and $|h_{j,g,K+1}|^2$, one can approximate that the pdfs of $|\hat{h}_{0,j,0}|^2$ and $|\hat{h}_{j,g,K+1}|^2$ have the same form as the pdfs of $|h_{0,j,0}|^2$ and $|h_{j,g,K+1}|^2$, respectively. Since $|\hat{h}_{0,j,0}|^2$ and $|\hat{h}_{j,g,K+1}|^2$ are Rayleigh distributed, the pdfs of $|\hat{h}_{0,j,0}|^2$ and $|\hat{h}_{j,g,K+1}|^2$ can be approximated as $f_{|\hat{h}_{0,j,0}|^2}(x) = \frac{1}{\sigma_{0,j}^2} e^{-\frac{x}{\sigma_{0,j}^2}}$, respectively. It is clear that

$$f_{X_1}(x) = \frac{x^{N_0-1} e^{-\frac{x}{\sigma_{0,j}^2}}}{\Gamma(N_0)} \frac{1}{\Gamma(N_{K+1})}$$

and

$$f_{X_2}(x) = \frac{x^{N_{K+1}-1} e^{-\frac{x}{\sigma_{j,K+1}^2}}}{\Gamma(N_{K+1}) \Gamma(N_{K+1})}$$

The MGF of $Y$ can then be computed as follows:

**Case 1** ($N_0 > N_{K+1}$): $\mathcal{M}_Y(s) = \int_0^{\infty} f_{X_1}(x)M_{X_2}(sx)dx$

$\approx \int_0^{\infty} x^{N_0-1} e^{-\frac{x}{\sigma_{0,j}^2}} \frac{1}{\Gamma(N_0)} \frac{1}{\Gamma(N_{K+1})} \frac{1}{\Gamma(N_{K+1})} \frac{1}{\Gamma(N_{K+1})} dx$

$= \int_0^{\infty} \frac{1}{\Gamma(N_0)} \frac{1}{\Gamma(N_{K+1})} \frac{1}{\Gamma(N_{K+1})} \frac{1}{\Gamma(N_{K+1})} dx$

$\approx \int_0^{\infty} \frac{1}{\Gamma(N_{K+1})} \frac{1}{\Gamma(N_{K+1})} \frac{1}{\Gamma(N_{K+1})} dx$

**Case 2** ($N_0 < N_{K+1}$): Similar to Case 1, but $\sigma_{0,j}^2$ and $\sigma_{j,K+1}^2$ are interchanged.

**Case 3** ($N_0 = N_{K+1}$):

$$\mathcal{M}_Y(s) = \int_0^{\infty} x^{N_{K+1}-1} e^{-\frac{x}{\sigma_{j,K+1}^2}} \frac{1}{\Gamma(N_{K+1})} \frac{1}{\Gamma(N_{K+1})} \frac{1}{\Gamma(N_{K+1})} dx$$

$$\approx \int_0^{\infty} \frac{1}{\Gamma(N_{K+1})} \frac{1}{\Gamma(N_{K+1})} \frac{1}{\Gamma(N_{K+1})} dx$$

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