Performance Analysis of a Selective Decode-and-Forward Cooperative System with SFBC and BICM-OFDM

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Abstract—In this paper, we analyze the bit error rate (BER) performance of selective decode-and-forward (DF) cooperative communications incorporating space-frequency block coding (SFBC), bit-interleaved coded modulation (BICM) and orthogonal frequency division multiplexing (OFDM). For the classical 3-node cooperation model, a tight upper bound of BER for the coded system under Rayleigh fading channels is derived. By uniting the analysis for multi-antenna cooperative schemes and the analysis for BICM under a common framework, the upper bound is proved valid for moderate SNR regime. With reasonable simplifications in the analysis, an asymptotic PEP expression that takes into account the effect of space-frequency coding, channel coding and the channel parameters is also derived, and gives insights on the diversity gain of the system. The results are validated by simulations under various settings.

I. INTRODUCTION

Cooperative diversity techniques offer additional diversity gains than conventional wireless communication systems, by exploiting one or more relay nodes in the transmission process. It has drawn extensive interests in academia, and has become a promising option in future cellular and ad-hoc networks. Two basic yet effective protocols, amplify-and-forward (AF) and decode-and-forward (DF), are proposed in [1]. And contributions from information theoretic views have been abundant, e.g. [2] and references therein. Moreover, Multiple-Input-Multiple-Output (MIMO) techniques have been extensively incorporated in the research of cooperative systems, to enhance performance by exploiting extra spatial dimensions. Space-Time Block Coding (STBC) and Space-Frequency Block Coding (SFBC) [3] are among the common choices for diversity gains due to their simple structure.

In the existing literatures on cooperative diversity, most assume un-coded transmission and/or frequency flat fading channels for ease of analysis, but the results can hardly be applied in realistic wideband channels with severe frequency selectivity. Others consider the realistic scenarios, but concentrated on the analysis for asymptotic performance. In [4] and [5], the authors give an upper bound of diversity gains for the AF and DF cooperative BICM-OFDM systems, respectively, and similar analysis is extended to multi-antenna settings in [6]. Based on the asymptotic results, optimizations on power allocation and code puncturing are considered respectively in [7] and [8]. However, these results are insufficient in bounding the practical performance. To the best of our knowledge, the analysis for non-asymptotic performance of coded MIMO cooperative communications is still absent, and cannot be done with the methods in the mentioned works.

Our work aims at the non-asymptotic analysis of BICM-OFDM cooperative systems under frequency selective channels. Specifically, in this paper we consider a selective DF scheme with SFBC and BICM-OFDM, such that results for single antenna settings can be easily acquired. In this scheme, the relay node forwards to the destination if and only if it successfully decodes the message from the source. The destination performs simple Maximum Ratio Combining (MRC) when it detects the participation of relay node (this might induce minor overhead in signaling but negligible). Such a strategy eliminates the necessity of passing the source-relay CSI on to the destination as in [4]. We develop the mathematical framework in [9] for our cooperative system in practical SNR regimes, and derive a BER upper bound under given channel conditions, which works for realistic deployment of such systems. We also perform asymptotic analysis on the worst-case pairwise error probability (PEP) for high signal-to-noise ratio (SNR), which enables insightful observations on how the space-frequency block coding, channel coding and channel parameters altogether affect the diversity gain of the system. The observations proves that the scheme can extract the spatial diversity offered by relay as well as multiple antennas, and the frequency diversity of the wireless channels simultaneously.

The remainder of this paper is organized as follows. In Section II, the system model is established. Based on the model, non-asymptotic and asymptotic performance analysis is given in Section III. In Section IV, simulation results are presented, followed by a brief conclusion in Section V.

Notations: In this paper, ⊗ denotes the Kronecker product.
of two matrices. \((\cdot)^T\), \((\cdot)^H\), \(\|\cdot\|_p\), \(\text{tr} (\cdot)\), and \(\text{rank}(\cdot)\) denote transposition, Hermitian transposition, Frobenius norm, trace, and rank of a matrix, respectively. \((\cdot)_{ij}\) denotes the \((i,j)\)-th element of a matrix. \(E\{\cdot\}\) denotes the statistical expectation. \(Q(\cdot)\) denotes the tail probability of the standard normal distribution.

II. SYSTEM MODEL

The system consists of one source node \((S)\), one relay node \((R)\) and one destination node \((D)\), and performs one-way transmission. All the three links, namely source-destination (SD) link, source-relay (SR) link, and relaydestination (RD) link are considered. One or more of the nodes are equipped with multiple antennas and SFBC is performed wherever possible. The protocol is a selective DF protocol, consisting of two phases. In the first phase, the source node broadcasts and the other two nodes listen. In the second phase, the relay node forwards if and only if it correctly decodes the message from the source node (via CRC or other means), and the destination node combines the received signal in both phases for detection. Otherwise the destination tries to decode the data only with the signal received in the first phase. Perfect synchronization is assumed in the transmission.

Denote the number of antennas at the source, the relay and the destination by \(N_S\), \(N_R\) and \(N_D\), respectively. The \(N_R\) antennas at the relay node are used for both the reception in the first phase and the transmission in the second phase. The number of channel impulse response (CIR) taps on the SD, SR and RD links are denoted by \(L_{SD}\), \(L_{SR}\) and \(L_{RD}\). Our discussion is limited to complex orthogonal SFBC [10]. For \(N_S\) antennas at the source, an \(R_S/J_S\)-rate SFBC is defined as the complex orthogonal block code that transmits \(R_S\) symbols across \(J_S\) subcarriers. The code generation matrix is denoted by an \(N_S \times J_S\) matrix \(G_S\) and satisfies

\[
G_S G_S^H = \kappa_S \sum_{i=1}^{R_S} |x_i|^2 \cdot I_{N_S}
\]

where \(\kappa_S\) is a positive constant, \(\{x_i\}_{i=1}^{R_S}\) is the set of symbols transmitted, and \(I_{N_S}\) is the \(N_S \times N_S\) identity matrix. A simple case is the Alamouti code,

\[
G_S = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix}.
\]

In practice, the subcarriers involved in a space-frequency code block is adequately small, such that the channel frequency response (CFR) on these adjacent subcarriers are approximately the same. At the relay node, the SFBC matrix \(G_R\) is also a complex orthogonal design but not necessarily the same as \(G_S\).

Assume a convolutional code with free distance \(d_{\text{free}}\) is used for channel coding. At the source node, the bit interleaver is denoted by a mapping \(c_k \rightarrow (K_S, n_S, i_k)\), indicating that the \(k\)-th input bit \(c_k\) is mapped to the \(i_k\)-th bit of the \(n_S\)-th symbol, in a space-frequency code block designated by its first subcarrier \(K_S\), where \(n_S = 1, 2, ..., R_S\), \(i = 0, 1, ..., \log_2 M - 1\), and \(M\) is the cardinality of the constellation set. At the relay node the interleaver \(c_k \rightarrow (K_R, n_R, i_k)\), is defined similarly. For a convolutional code, the \(d_{\text{free}}\) distinct bits between two codewords may span \(d\) consecutive bits. The interleaver shall guarantee that any such \(d\) consecutive bits are mapped onto different SFBC code blocks with sufficient spacing in frequency. For convenience of analysis, a common channel code and \(M\)-ary quadrature amplitude modulation (M-QAM) scheme is assumed at the source and relay node, while the space-frequency coding might be different to fully exploit the antennas at each node.

The wireless channel is modeled as a quasi-static multipath fading channel. For simplicity we assume that the CIR between different antenna pairs are uncorrelated. Each multipath tap is statistically independent and modeled as a zero-mean complex Gaussian random variable with variance \(1/L\), where \(L\) can be \(L_{SD}, L_{SR}\) or \(L_{RD}\), representing the number of multipath taps on SD, SR, RD links respectively.

A cyclic prefix (CP) of proper length is appended to each OFDM symbol before transmission to avoid inter-symbol interference. In the first phase, after CP removal and fast Fourier transform (FFT) at the destination, the received signal at the \(K_S\)-th space-frequency code block is given by

\[
Y_{SD} = \sqrt{\gamma_{SD}} \cdot H_{SD} \cdot C_S + N_{SD}
\]

where \(Y_{SD}\) is an \(N_D \times J_S\) matrix, \(\gamma_{SD}\) is the power gain considering transmit power and large-scale pathloss, \(C_S = G_S(x_1, x_2, ..., x_{R_S})\), is attained by applying the symbols \(x_1, x_2, ..., x_{R_S}\) to the code generation matrix \(G_S\), and \(N_{SD}\) is a \(N_D \times J_S\) complex AWGN matrix with mutually independent elements of zero mean and variance \(N_0\). Therefore the average signal-to-noise ratio (SNR) at a receiving antenna satisfies \(\text{SNR} = \gamma_{SD}/N_0\). The \(N_D \times N_S\) channel matrix \(H_{SD}\) is given as in [9]

\[
H_{SD} = W_{K_S}^H \cdot \frac{K_S}{K_S}
\]

\[
W_{K_S} = I_{N_D} \otimes \hat{w}_{K_S}
\]

\[
\hat{w}_{K_S} = \begin{bmatrix} 1 & e^{i \frac{2 \pi}{K_S} K_S} & \cdots & e^{i \frac{2 \pi}{K_S} K_S(L_{SD}-1)} \\ \vdots & \vdots & \ddots & \vdots \\ \hat{h}_{N_{K_S}} & \cdots & \hat{h}_{N_{K_S}} \end{bmatrix}^T
\]

where \(\hat{h}_{SD}\) is a \(N_D L_{SD} \times N_S\) matrix, \(\hat{h}_{ij} = \begin{bmatrix} h_{ij,1} & h_{ij,2} & \cdots & h_{ij,L_{SD}} \end{bmatrix}^T\) is the \(L_{SD} \times 1\) vector representing the CIR between the transmitting antenna \(j\) at the source and the receiving antenna \(i\) at the destination, \(i = 1, 2, ..., N_D\) and \(j = 1, 2, ..., N_S\), \(N\) is the number of OFDM subcarriers.

At the relay node, the received signal at the \(K_S\)-th code block is given by

\[
Y_{SR} = \sqrt{\gamma_{SR}} H_{SR} \cdot C_S + N_{SR}
\]

and the definitions of the terms therein can be easily acquired by replacing the “SD’s” and “D’s” in (3)/(4) with “SR’s” and “R’s”, respectively. The relay node computes the bit metric for
the \(i_{S}\)-th bit of the \(n_{S}\)-th symbol in the space-frequency code block \(K_{S}\)

\[
m_{K_{S},n_{S}}^{i_{S}}(c_{k}) = \min_{c_{S,n} \in X_{i_{S}}^{n_{S}}} \left( \|Y_{SD} - \sqrt{\gamma_{SD}} H_{SD} C_{S}\|_{F}^{2} \right)
\]

(6)

where \(X_{i_{S}}^{n_{S}}\) denotes the subset of constellation points whose \(i_{S}\)-th bit is \(c_{k} \in \{0,1\}\), and \(C_{S,n}\) is the \(n\)-th symbol composing \(C_{S}\). The bit metrics are subsequently de-interleaved and decoded by the Viterbi algorithm. Two possible cases should be considered here. If the decoder outputs fail CRC, the relay node keeps silent, and the destination performs detection.

For the terminated convolutional code we consider, an improved upper bound for FER is given in [11]

\[
\text{FER}_{SR} \leq \frac{1}{L_{1}} \sum_{j=d_{\text{free}}}^{\infty} \sum_{i} a_{i,j} \cdot P_{j}
\]

(12)

where \(a_{i,j}\) is the number of error events composed of a single error event with output Hamming distance \(i\) and input Hamming distance \(j\), and \(P_{j}\) here equals the SR link PEP \(P_{SR}(c,c')\) when \(c\) and \(c'\) differ in \(j\) positions.

After acquiring \(P(c,\hat{c})\), the end-to-end BER is

\[
\text{BER} \leq \frac{L_{1} \sum_{j=d_{\text{free}}}^{\infty} \sum_{i} a_{i,j} \cdot P_{j}}{L_{1} \sum_{j=d_{\text{free}}}^{\infty} \sum_{i} a_{i,j} \cdot P_{j}}
\]

(13)

where \(L_{1}\) is the length of information bits, and \(P_{j}\) equals the end-to-end PEP \(P_{SD}(c,\hat{c})\) when \(c\) and \(\hat{c}\) differ in \(j\) positions.

A. Non-asymptotic Analysis

To acquire the BER bound, we will calculate \(P_{D}(c,\hat{c})\) first. Though similar to the asymptotic analysis in [9], we have to consider the joint detection in the cooperative system as in (9), and 2) deal carefully with the inequalities so that the result is effective under moderate SNRs. Suppose that \(c\) and \(\hat{c}\) differ in \(j\) positions, \(j \geq d_{\text{free}}\). Utilizing (3), (8) and (9), we have

\[
P_{D}(c,\hat{c}|H_{SD},H_{RD})
\]

\[
= P\left( \sum_{j} m_{K_{S},K_{S}}^{i_{S}}(c_{k}) \geq \sum_{j} m_{K_{S},K_{S}}^{i_{S}}(\hat{c}_{k'}) \right)
\]

\[
= P\left( \sum_{j} \|N_{SD}\|_{F}^{2} + \sum_{j} \|N_{RD}\|_{F}^{2} \right)
\]

\[
\geq \sum_{j} \|N_{SD} + \sqrt{\gamma_{SD}} H_{SD} (C_{S} - \hat{C}_{S})\|_{F}^{2}
\]

\[
+ \sum_{j} \|N_{RD} + \sqrt{\gamma_{RD}} H_{RD} (C_{R} - \hat{C}_{R})\|_{F}^{2}
\]

\[
= P\left( \sum_{j} \gamma_{SD} \|N_{SD} + \sqrt{\gamma_{SD}} H_{SD} (C_{S} - \hat{C}_{S})\|_{F}^{2}
\]

\[
+ \sum_{j} \gamma_{RD} \|N_{RD} + \sqrt{\gamma_{RD}} H_{RD} (C_{R} - \hat{C}_{R})\|_{F}^{2}
\]

\[
\leq \beta
\]

(14)

where \(c_{k'} \in \{0,1\}\), \(c_{k'} = 1 - c_{k}\), and

\[
\beta = - \sum_{j} \sqrt{\gamma_{SD}} \left( \text{tr}\{N_{SD}(C_{S} - \hat{C}_{S})H_{SD}\} + \text{tr}\{H_{SD}(C_{S} - \hat{C}_{S})N_{SD}^{H}\} \right)
\]

\[
- \sum_{j} \sqrt{\gamma_{RD}} \left( \text{tr}\{N_{RD}(C_{R} - \hat{C}_{R})H_{RD}\} + \text{tr}\{H_{RD}(C_{R} - \hat{C}_{R})N_{RD}^{H}\} \right)
\]

is a zero-mean gaussian term with variance

\[
\frac{2\gamma_{SD}}{\text{tr}\{N_{SD}H_{SD}\}} + \frac{2\gamma_{RD}}{\text{tr}\{N_{RD}H_{RD}\}}
\]

\[
\left( \text{tr}\{N_{SD}H_{SD}^{H}\} + \text{tr}\{N_{RD}H_{RD}^{H}\} \right)
\]

\[
\cdot \|H_{SD}(C_{S} - \hat{C}_{S})\|_{F}^{2}
\]

\[
+ \sum_{j} \gamma_{RD} \|H_{RD}(C_{R} - \hat{C}_{R})\|_{F}^{2}
\]

\[
\left( \text{tr}\{N_{RD}H_{RD}^{H}\} + \text{tr}\{N_{SD}H_{SD}^{H}\} \right)
\]
The summations correspond to \( j \) SFBC blocks either at the source or at the relay. Notice the fact that \( C_S(C_R) \) and \( \tilde{C}_S(\tilde{C}_R) \) differ in at least one symbol, we have

\[
P_D(c, \tilde{c}|H_{SD}, H_{RD}) = \left| \frac{\sum_{j} \gamma_{SD} H_{SD}(C_S - \tilde{C}_S)_{i_F}^2}{2N_0} + \frac{\sum_{j} \gamma_{RD} H_{RD}(C_R - \tilde{C}_R)_{i_F}^2}{2N_0} \right|^{1/2}
\]

\[
= Q\left( \frac{\sum_{j} \gamma_{SD} H_{SD}(C_S - \tilde{C}_S)_{i_F}^2}{2N_0} + \frac{\sum_{j} \gamma_{RD} H_{RD}(C_R - \tilde{C}_R)_{i_F}^2}{2N_0} \right)^{1/2}
\]

where \( d_{\min} \) is the minimum Euclidean distance between two constellation symbols, and \( \alpha \) is a constant determined by the modulation and bit mapping scheme, which is the average of squared Euclidean distance between two constellation symbols given one bit error, in units of \( d_{\min}^2 \). Since the CIRs of SD link and RD link are independent, we further have

\[
P_D(c, \tilde{c}) = E_{H_{SD}, H_{RD}} \left\{ P_D(c, \tilde{c}|H_{SD}, H_{RD}) \right\}
\]

\[
= \left\{ \frac{\sum_{j} \gamma_{SD} H_{SD}(C_S - \tilde{C}_S)_{i_F}^2}{2N_0} + \frac{\sum_{j} \gamma_{RD} H_{RD}(C_R - \tilde{C}_R)_{i_F}^2}{2N_0} \right\}^{1/2}
\]

\[
= Q\left( \frac{\sum_{j} \gamma_{SD} H_{SD}(C_S - \tilde{C}_S)_{i_F}^2}{2N_0} + \frac{\sum_{j} \gamma_{RD} H_{RD}(C_R - \tilde{C}_R)_{i_F}^2}{2N_0} \right)^{1/2}
\]

with the last step following similar mathematical manipulations in [9], where \( \lambda_i(A_{SD}) \) is the \( i \)-th singular value of \( A_{SD} = \sum_j w_j w_j^H \), and \( r_{SD} = \min(j, L_{SD}) \). \( A_{RD} \) and \( r_{RD} \) are defined similarly. Now we can compute \( P_D(c, \tilde{c}) \) by numerical integral, with \( \lambda_i(A_{SD}) \) and \( \lambda_i(A_{RD}) \) attained by simple Monte Carlo simulations.

Due to page limit, the analysis and result for \( P_{SD}(c, \tilde{c}) \) and \( P_{SR}(c, \tilde{c}) \) is omitted. Next, we substitute the results of \( P_D(c, \tilde{c}), P_{SD}(c, \tilde{c}) \) and \( P_{SR}(c, \tilde{c}) \) into (10)-(13), and acquire the end-to-end BER upper bound. Simulation results are presented in Section IV.

### B. Asymptotic Analysis

The non-asymptotic results in the preceding subsection does not offer straightforward observations on the high SNR regime. We apply Chernoff’s inequality \( Q(x) \leq \frac{1}{2} \exp(-\frac{\lambda^2}{2}) \) in (16) instead of the integral, and ignore the difference in power gains for the three links by setting \( \frac{N_{SD}}{N_0} = \frac{N_{RD}}{N_0} = \frac{N_{SR}}{N_0} = SNR \), we have

\[
P_{SD}(c, \tilde{c}) \leq \frac{1}{2} \prod_{i=0}^{r_{SD}-1} \lambda_i(A_{SD}) \cdot \left( \frac{\kappa_S \alpha d_{\min}^2 \cdot SNR}{4L_{SD}} \right)^{N_{SD}N_{SD}r_{SD}}
\]

and

\[
P_{SR}(c, \tilde{c}) \leq \frac{1}{2} \prod_{i=0}^{r_{SR}-1} \lambda_i(A_{SR}) \cdot \left( \frac{\kappa_R \alpha d_{\min}^2 \cdot SNR}{4L_{SR}} \right)^{N_{SR}N_{SR}r_{SR}}
\]

where \( \lambda_i(A_{SR}) \) and \( r_{SR} \) are defined similarly as \( \lambda_i(A_{SD}) \) and \( r_{SD} \).

In sufficiently high SNR regime, we consider only the worst-case pairwise error event corresponding to the first term in the summation of (12). Substitute the asymptotic FER into (10) and (11), the upper bound for worst-case PEP that considers only codeword pairs with Hamming distance of \( d_{\text{free}} \), can be simplified as

\[
P(c, \tilde{c}) \leq \varphi_1 SNR^{-N_{SD}N_{SD}r_{SD} - N_{SR}N_{SR}r_{SR}} + \varphi_2 SNR^{-N_{SD}N_{SD}r_{SD} - N_{SR}N_{SR}r_{SR}}
\]

where \( \varphi_1, \varphi_2 \) are positive constants irrelevant to SNR, and \( r_{SD} = \min(d_{\text{free}}, L_{SD}), r_{SR} = \min(d_{\text{free}}, L_{SR}), r_{RD} = \min(d_{\text{free}}, L_{RD}) \). When SNR \( \to \infty \), the diversity gain can be defined as the negative slope of the PEP in (20) as a function of SNR on a double-logarithmic scale [5]. Thus the diversity gain is given by

\[
G_{\text{div}} = N_{SD}N_{SD}r_{SD} + \min(N_{SR}N_{SR}r_{SR}, N_{SR}N_{SR}r_{SR})
\]

Some important observations can be made with (21). It reveals that the maximum diversity gain is limited by the antenna settings of all nodes, the free distance of channel code, and the number of multipath taps on each link. The frequency diversity can be extracted by applying channel codes with proper free distance. It is worthwhile to note that the diversity gains in (21) is an upper bound, which may not be fully achieved for various channel conditions. Besides, if we omit the diversity gain \( N_{SD}N_{SR}r_{SD} \) offered by the SD link in (21), we may readily find that the maximum diversity gain is the same as that in the AF beamforming case discussed.
In this section, we present simulation results to illustrate the analytical results derived in the preceding section. We specify the settings that the number of antennas per node is up to 2, and limit the channel code choices to rate 1/2 convolutional codes, one with generator polynomials $(1,3)_8$ and free distance $d_{\text{free}} = 3$, and the other with generator polynomials $(5,7)_8$ and $d_{\text{free}} = 5$. The modulation is 16-QAM with Gray labeling, and the total number of subcarriers for OFDM is 256, of which 240 subcarriers are dedicated to data transmission. A proper rectangular interleaver is used.

In Figure 1, we examine the non-asymptotic results in a realistic downlink scenario. The power gains for SD and RD links are balanced, and the power gain for SR link is 10dB higher than the other two links. Simulated BER curves and corresponding analytical upper bound under the antenna setting $(N_S, N_R, N_D) = (2, 1, 1)$ are given. The number of multipath taps on all links is given by $(L_{\text{SD}}, L_{\text{SR}}, L_{\text{RD}}) = (3, 3, 3)$ for the $d_{\text{free}} = 3$ code, and $(L_{\text{SD}}, L_{\text{SR}}, L_{\text{RD}}) = (5, 5, 5)$ for the $d_{\text{free}} = 5$ code. It can be observed that the gap between the actual BER curve and the analytical upper bound narrows quickly as SNR increases, and when BER goes down to the scale of $10^{-6}$, the two curves almost converge, showing that the upper bound is quite tight at the SNR regime for reliable data transmissions. It shows that the upper bound can be used for prediction of realistic system performance.

In the following, the asymptotic performance is illustrated, assuming the power gains of all links are the same. In Figure 2, to observe the influence of antenna, three typical antenna settings with $(N_S, N_R, N_D) = (2, 1, 1)$, $(N_S, N_R, N_D) = (2, 2, 1)$ and $(N_S, N_R, N_D) = (2, 2, 2)$ are chosen, and a $(N_S, N_R, N_D) = (1, 1, 1)$ setting as reference. The number of multipath taps on all links is given by $(L_{\text{SD}}, L_{\text{SR}}, L_{\text{RD}}) = (9, 9, 9)$. The channel code is fixed to the code $(1,3)_8$. It can be observed that the diversity gain $G_d$, which is indicated by the negative slope of the BER curve, increases as more antennas participate in the cooperative transmission. In the simulation, the negative slope exceeds 5 for the $G_d = 6$ curve, and for the $G_d = 9$ curve it exceeds 8 at SNR $= 10^{-3}$, which confirms (21) very well. For larger $G_d$, as in $G_d = 12$ or $G_d = 24$ cases, the diversity gain can only be observed at very low BER. In fact, in the $G_d = 24$ case the negative slope remains not more than 20 at the SNR $= 10^{-8}$.

In Figure 3, we focus on the influence of multipath environment and thus use the specific antenna setting $(N_S, N_R, N_D) = (2, 1, 1)$, and the $(1,3)_8$ code. One or two of the links are with poor multipath environment (and with a flat fading case in the uppermost curve). Four cases with diversity gains 6, 7 and 8 are depicted. The diversity gains are consistent with our prediction quite well. Comparing these cases with the $G_d = 9$ case in Fig. 1, we see that given a certain channel code, a poor multipath environment may restrict the total diversity gain that can be achieved, and that all the links may have some influence on the total performance. One thing to note here is the comparison of the two cases with $G_d = 7$. When the multipath taps on the SR link increase to 9 from 2, the frequency diversity on the SR link is better exploited, but the total diversity gain is the same. This can be readily explained by (21). In (21), only the smaller value of $N_S N_R T_{\text{SR}}$ and $N_R N_D T_{\text{RD}}$ is involved in the final result, i.e., the diversity gain provided by the relay (the SR and RD links) is determined by the link with smaller diversity gain.

V. CONCLUSION

In this paper, the performance of the selective DF cooperative system equipped with SFBC and BICM-OFDM is analyzed. In the non-asymptotic analysis, an upper bound for BER is presented, which is proved tight under moderate SNR regime for reliable data transmissions by simulations, and can be used for performance prediction in realistic deployments. For the asymptotic case, we give a semi-analytical upper bound for the asymptotic worst-case PEP, and reveal the relationship between space-frequency coding, channel coding, channel parameters and the achievable diversity gain.
Simulation results confirmed the diversity gains predicted in our analysis. Future work involves extending the analysis to correlated MIMO channels and various channel power delay profiles.

References


