On Relay Selection in Frequency Selective Channels

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***Abstract*—We consider relay selection in cooperative relay networks with frequency selective fading, and focus on a system where multiple decode-and-forward relays share a single channel orthogonal to the source. We propose a relay selection scheme and end-to-end transciever architecture which employs a simple decision feedback equalizer, and we show analytically that the proposed system can optimally exploit both spatial diversity and frequency diversity. By simulation, we compare the BER performance of a finite-length MMSE-DFE receiver with that of the MLSE and linear ZF receivers. The numerical results show that the finite-length MMSE-DFE receiver, combined with the relay selection method and zero-padded transmission, can achieve full diversity.**

I. Introduction

Cooperative relay networks have emerged as a powerful technique to combat multipath fading and increase energy efficiency [1], [2]. In high data-rate mobile wireless systems, the coherence bandwidth of the channel tends to be smaller than the bandwidth of the signal, resulting in frequency selective (FS) fading [3]. For such high rate communication in cooperative relay systems, existing techniques for flat channels need to be adapted, or new techniques need to be designed for FS fading channels. While the majority of research in cooperative diversity makes the assumption that the channel are non-frequency selective, there are some exceptions. For example, in [4] and [5], the authors considered a system with amplify-and-forward (AF) relays over FS channels, and proposed distributed space-time codes (DSTCs) or space-frequency codes; the achieved frequency diversity was shown to be equal to the minimum of the source-relay channel length and the relay-destination channel length. The decoding complexity of such DSTC-based schemes may be prohibitive to permit their use in low-cost wireless ad-hoc networks. Furthermore, the non-linearity of most existing RF-frond ends makes distributed coding less desirable. These problems are similar to the shortcomings of DSTC in flat fading channel, which can be overcome by using relay selection. Relay selection generally uses simple repetition coding, very simple scheduling, a single relay channel [6] and can achieve the same diversity-multiplexing tradeoff (DMT) [7] as DSTC.

Relay selection methods have predominantly focused on the frequency flat fading channels. Relay selection in the presence of FS fading has seen some attention in OFDM systems. For example, in [8], [9], uncoded OFDM is studied, and it was shown that if relay selection is done on a per-subcarrier basis, full spatial diversity can be achieved. However, neither of these OFDM-based relay selection methods were able to exploit the frequency diversity of the ISI channel. A linearly precoded OFDM system was proposed in [10] which uses multiple amplify-forward relays with linear transmit precoding; a simulation-based study showed that two relay selection schemes exhibited a coding gain improvement compared to an orthogonal round-robbin relaying scheme.

In this paper, we propose a transmission scheme, a relaying selection strategy, and a receiver architecture for single-carrier transmission with decode-and-forward half-duplex relays in the presence of FS fading. The scheme we propose does not need transmitter precoding, uses simple repetition coding at the relays, and does not require maximum likelihood (ML) detectors. We first analytically prove that the proposed infinite-length zero-forcing (ZF) decision feedback equalizer (DFE) with the relay selection can asymptotically achieve the optimal DMT of the system model [11]. While such a result is attractive for its theoretical simplicity, its practical implication is limited because the equalizer is of infinite-length and the decoding delay is infinite. To reduce the decoding delay and simplify the equalization structure, we switch to consider finite-length MMSE-DFE receiver at the destination. BER analysis is complicated for such finite-length MMSE-DFE receiver as the residual ISI is not Gaussian; hence we show by simulations that the performance of a finite-length MMSE-DFE receiver in relay selection is better than that of linear ZF receiver which is proven to achieve the optimal DMT [11].

II. System Model

A. Channel Model and DMT Bound

We consider the system shown in Fig. 1, comprised of a single source node ($\mathbf{S}$), K relay nodes ($\mathbf{R}_{1,2,\ldots,K}$), and a single destination node ($\mathbf{D}$). We assume that all nodes have equal average power constraint P watts and transmission bandwidth W Hz. The links between the nodes are assumed to be FS quasi-static fading channels, modelled as complex FIR filters with each channel coefficient having i.i.d. Rayleigh fading statistics [3]. The channel coefficients are assumed to be constant over a block, and are independent from one block to the next. In addition, all links have additive noise which is assumed to be mutually independent, zero-mean circularly symmetric complex Gaussian with variance N_0 and the discrete-time signal-to-noise ratio is defined as

$$\rho \triangleq \frac{P}{WN_0}.$$

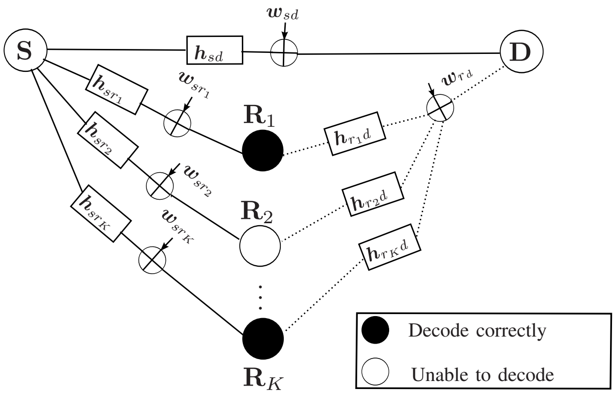

Fig. 1. Channel model.

The source to destination channel coefficients are contained in the vector $\boldsymbol{h}_{sd}$ with length L_{sd} and the AWGN is $\boldsymbol{w}_{sd}$. Similarly, for $i \in 1, 2, \ldots, K$, the source to relay $\mathbf{R}_i$ channels are $\boldsymbol{h}_{sr_i}$ with length L_{sr_i} and AWGN $\boldsymbol{w}_{sr_i}$ at each relay. The channels from each relay $\mathbf{R}_i$ to destination are denoted $\boldsymbol{h}_{r_id}$ with length L_{r_id} and AWGN $\boldsymbol{w}_{rd}$ at the destination. We assume the message sent from the transmitter by the source node is encoded to a block of N source symbols. The relays operate in half-duplex mode, and thus do not transmit and receive at the same time. In addition, the relay nodes and the source use the same transmission bandwidth but employ time division so that the relays transmit on a channel orthogonal to the source. We define the relays which can decode the message from the source correctly as as *decoding relays* and they form a *decoding set* $\mathcal{D}$. The transmission of a complete message is divided into two phases:

1) In phase one, the source broadcasts the message to the destination and the relays, and each relay attempts to decode the message.
2) In phase two, the source is silent. Depending on the relay selection strategy, some subset of relays in the decoding set (possibly a single node) are chosen to forward the message to the destination.

We are concerned with the development of a practical relaying strategy and transceiver design that simultaneously exploit both the maximum available frequency and spatial diversity. Using a combination of the cut-set and matched filter bounds, it was recently shown [11] that the optimal DMT for this system in Fig. 1 is given by

$$d(r) \leq (L_{sd} + \sum_{i=1}^{K} \min(L_{sr_i}, L_{r_id}))(1 - 2r), \qquad (1)$$

which holds regardless of the relaying protocol. As such, we hope that any proposed relaying schemes would be able to attain this DMT in order to fully exploit the available diversity.

In the next three subsections, we propose a specific transmission scheme, relay selection strategy, and receiver architecture. Then, in the following section, we will show that this scheme attains the optimal DMT.

B. Transmission Scheme

In the transmission scheme, we assume zero-padded transmission to eliminate interblock interference, though intersymbol interference (ISI) is still present. In the proposed scheme, we assume that the transmission rate $R = r \log \rho$, and the source transmits $\boldsymbol{x}$, a block of N QAM symbols with $L_{\max} - 1$ trailing zeros, where the QAM symbols are drawn from a constellation of $Q = \rho^{2r'}$ points [12, Equation (2)] with

$$r' = \frac{r}{1 - \frac{L_{\max}-1}{N+L_{\max}-1}} \qquad (2)$$

and $L_{\max} \geq \max_{i \in 1,\ldots,K} \{L_{sr_i}, L_{r_id}, L_{sd}\}$.

The choice of Q or r' here is to make sure the total transmission rate is still R with the guard interval. $L_{\max}$ is essentially an upper bound on the length of all channels in the system. In practice, it is unrealistic for the transmitter to have knowledge of the lengths of all channels in the system. Due to the insertion of guard time between alternating phases of source/relay transmission, we see from (2) that the system incurs a rate penalty that can be made small by increasing the block length N.

C. Relay Selection

The relay selection occurs during the second phase of the transmission of a message. The selection is based on instantaneous channel state information and is within the decoding set. When the decoding set is not empty, the proposed selection strategy selects the relay which has the largest sum-square of the relay-to-destination channel taps, and that relay forwards the decoded message to the destination. The chosen relay, denoted $\boldsymbol{h}_{r^*d}$, is given by

$$\|\boldsymbol{h}_{r^*d}\|^2 \triangleq \max_{\mathbf{R}_i \in \mathcal{D}} \|\boldsymbol{h}_{r_id}\|^2. \qquad (3)$$

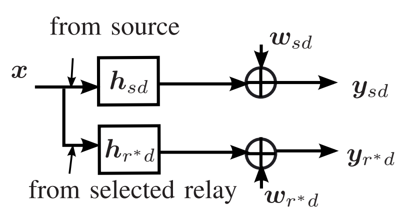

Fig. 2. Received signal at destination

Fig. 2 shows the equivalent channel model for the received signal at destination under the assumption that at least one relay was able to correctly decode the source message. The received signal from the source and the received signal from the selected relay are respectively

$$y_{sd}[k] = \sum_{m=-\infty}^{+\infty} h_{sd}[k-m]x[m] + w_{sd}[k], \qquad (4)$$

$$y_{r^*d}[k] = \sum_{m=-\infty}^{+\infty} h_{r^*d}[k-m]x[m] + w_{rd}[k]. \qquad (5)$$

for $k = 1, \cdots, \max(L_{sd}, L_{r^*d} + N - 1)$, $x[m] = 0$ for $m \notin \{1, \cdots, N\}$, and $\boldsymbol{w}_{sd}$ is the noise at the destination at the first phase.

D. DFE Receiver

We assume that a DFE receiver is used at the destination, as well as at all relays. After the DFE, the decoding is performed by a simple memoryless slicer. We assume that the receiver either uses the infinite-length ZF-DFE (which we adopt for its simplicity in analysis) or alternatively the receiver uses a finite-length MMSE-DFE (which is used more commonly in practice). If no relay is able to decode the transmitted signal, the receiver at the destination follows the standard DFE design just as the receiver at the relays, since the channel is a single point-to-point channel. In the case where a relay is able to successfully decode, and after one is chosen as described above, the channel is equivalent to a single-input multi-output (SIMO) channel as shown in Fig. 2. Fig. 3 shows the basic structure of DFE receiver for SIMO channel with the two outputs. The feed forward filters (FFFs) are $F_1(z)$ and $F_2(z)$ and the feedback filter (FBF) is $G(z)$. $\mathcal{Q}$ represents the memoryless decision device with input $\tilde{\boldsymbol{x}}$ and output $\hat{\boldsymbol{x}}$.

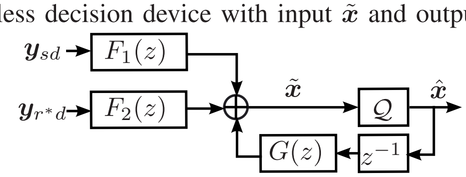

Fig. 3. DFE receiver

Next, we analyze the probability of error for the system with infinite-length ZF-DFE for SIMO channel.

III. Analysis of the Probability of Error with Infinite-length ZF-DFE receiver

We now analyze the probability of error of the system with an infinite-length ZF-DFE receiver to show that the relay selection method can exploit both frequency diversity and spatial diversity as given by the optimal DMT. We first calculate the probability of a given decoding set, then we present the conditional probability of error at the destination conditioned on a certain decoding set, and finally we combine the above two probabilities by total probability theorem to arrive at the final probability of error.

As each DF relay uses an infinite-length ZF-DFE to decode the received signal from the source, the error probability at each relay is equivalent to that of a ZF-DFE in a point-to-point channel. Using the error probability bound in [12, Theorem III.10] for the case of a point-to-point transmission with an infinite-length ZF-DFE, we have

$$P_{e,i} \dot{\leq} \rho^{-L_{sr_i}(1-r')}.$$

We assume that a relay is in the decoding set only when all N symbols are decoded correctly, so the probability that a relay is in the decoding set becomes

$$\Pr[\mathbf{R}_i \in \mathcal{D}] \dot{\leq} (1-P_{e,i})^N.$$

and the probability of a given decoding set is bounded as

$$\begin{aligned} \Pr[\mathcal{D}] &\dot{\leq} \prod_{\mathbf{R}_i \notin \mathcal{D}} (1-(1-P_{e,i})^N)) \prod_{\mathbf{R}_i \in \mathcal{D}} (1-P_{e,i})^N \\ &\doteq \rho^{-(1-2r')\sum_{\mathbf{R}_i \notin \mathcal{D}} L_{sr_i}} \end{aligned} \tag{6}$$

where asymptotic equality in (6) follows from the binomial theorem.

We next analyze the error probability at the destination conditioned on the decoding set. When the decoding set is not empty, the forwarding relay is selected according to (3). The input-output relation between the original message sent from the source and the received signals at the destination during the two phases are

$$\boldsymbol{y}[k] \triangleq \begin{bmatrix} y_{sd}[k] \\ y_{r^*d}[k] \end{bmatrix} = \sum_{m=-\infty}^{+\infty} \begin{bmatrix} h_{sd}[k-m] \\ h_{r^*d}[k-m] \end{bmatrix} x[m] + \begin{bmatrix} w_{sd}[k] \\ w_{rd}[k] \end{bmatrix} \tag{7}$$

for $k = 1, \cdots, \max(L_{sd}, L_{r^*d} + N - 1)$, $x[m] = 0$ for $m \notin \{1, \cdots, N\}$. Thus the transfer function for (7)

$$\boldsymbol{H}(z) = \begin{bmatrix} H_{sd}(z) \\ H_{r^*d}(z) \end{bmatrix} = \begin{bmatrix} \sum_{l=0}^{L_{sd}-1} h_{sd}[l] z^{-l} \\ \sum_{l=0}^{L_{r^*d}-1} h_{r^*d}[l] z^{-l} \end{bmatrix}.$$

According to the minimum-phase spectral factorization

$$\boldsymbol{H}^H(1/z^*)\boldsymbol{H}(z) = V^*(1/z^*)\Gamma^2 V(z),$$

the infinite-length ZF-DFE receiver consists of a precursor equalizer is

$$\boldsymbol{F}(z) = \begin{bmatrix} F_1(z) \\ F_2(z) \end{bmatrix} = \boldsymbol{H}^H(1/z^*) \frac{1}{\gamma^2 V^*(1/z^*)}$$

and a postcursor equalizer is $G(z) = (V(z) - 1)z$.

Assuming correct decisions, the effective SNR at the output of this equalizer is

$$\rho_{\text{eff}}^{\text{ZF-DFE}} = e^{\frac{1}{2\pi}\int_0^{2\pi} \log(\rho \boldsymbol{H}^H(\omega)\boldsymbol{H}(\omega))}.$$

Let $X : [0, 2\pi) \to \mathbb{R}$ and define the set

$$\mathcal{U}_\epsilon \triangleq \{\omega \in [0, 2\pi) | X(\omega) \leq \epsilon\}.$$

Define $\bar{H}_{sd}(\omega) \triangleq H_{sd}(\omega)/\|\boldsymbol{h}_{sd}\|^2$ and $\bar{H}_{r_id}(\omega) \triangleq H_{r_id}(\omega)/\|\boldsymbol{h}_{r_id}\|^2$.

The error probability at the destination conditioned on $\mathcal{D}$ [12, Lemma VII.6] is

$$\begin{aligned} P_{e|\mathcal{D}} &\doteq \Pr[\rho_{\text{eff}}^{\text{DFE-ZF}} < \rho^{2r'} | \mathcal{D}] \\ &\leq \Pr[e^{\frac{1}{2\pi}\int_0^{2\pi}\log(\rho\boldsymbol{H}^H(\omega)\boldsymbol{H}(\omega))} < \rho^{2r'}] \\ &= \Pr[e^{\frac{1}{2\pi}\int_0^{2\pi}\log(\rho(|H_{sd}(\omega)|^2 + |H_{r^*d}(\omega)|^2))} < \rho^{2r'}] \\ &< \Pr[e^{\frac{1}{2\pi}\int_0^{2\pi}\log(\rho\max(|H_{sd}(\omega)|^2, |H_{r^*d}(\omega)|^2))} < \rho^{2r'}] \\ &= \Pr[e^{\frac{1}{2\pi}\int_0^{2\pi}\log(\rho|H_{sd}(\omega)|^2)} < \rho^{2r'}] \cdot \\ &\quad \prod_{\mathbf{R}_i \in \mathcal{D}} \Pr[e^{\frac{1}{2\pi}\int_0^{2\pi}\log(\rho|H_{r_id}(\omega)|^2))} < \rho^{2r'}] \\ &\dot{\leq} \frac{1}{\epsilon^{L_{sd}}}\rho^{-L_{sd}(1-\frac{r'}{1-\sup_{\boldsymbol{h}_{sd}\in\mathbb{C}^{L_{sd}}}|\mathcal{U}_\epsilon(\bar{\boldsymbol{H}}_{sd})|})} \cdot \\ &\quad \prod_{\mathbf{R}_i \in \mathcal{D}} \frac{1}{\epsilon^{L_{r_id}}}\rho^{-L_{r_id}(1-\frac{r'}{1-\sup_{\boldsymbol{h}_{r_id}\in\mathbb{C}^{L_{r_id}}}|\mathcal{U}_\epsilon(\bar{\boldsymbol{H}}_{r_id})|})} && (8) \\ &\dot{\leq} \rho^{(2r'-1)(L_{sd} + \sum_{\mathbf{R}_i \in \mathcal{D}} L_{r_id})} && (9) \end{aligned}$$

where (8) and (9) follow the very similar argument as that in proof of [12, Theorem III.10].

Combining (6) and (9) by the total probability theorem, we conclude that the proposed transmission scheme and infinite-length ZF-DFE receiver result in the following upper bound on the error probability:

$$\begin{aligned} P_e &\doteq \sum_{\mathcal{D}} P_{e|\mathcal{D}} \Pr[\mathcal{D}] \\ &\dot{\leq} \sum_{\mathcal{D}} \rho^{(2r'-1)(L_{sd}+\sum_{\mathbf{R}_i\in\mathcal{D}} L_{r_i d}+\sum_{\mathbf{R}_i\notin\mathcal{D}} L_{sr_i})} \\ &\doteq \rho^{(2r'-1)(L_{sd}+\sum_{i=1}^{K}\min(L_{r_i d},L_{sr_i}))}. \end{aligned}$$

Combining [12, Lemma III.1] and the result in [11, equation (24)], we can conclude that

$$P_e \doteq \rho^{(2r'-1)(L_{sd}+\sum_{i=1}^{K}\min(L_{r_i d},L_{sr_i}))}$$

which shows that the proposed scheme can asymptotically achieve the optimal DMT.

IV. Finite-length MMSE-DFE Receiver

In the previous section, we proved that by selecting the decoding relay with the largest sum-square of the relay-destination channel, the infinite-length ZF-DFE receiver can asymptotically achieve the optimal DMT. In [11], we proved that with the same relay selection method, linear zero-forcing equalizer (ZFE) can also asymptotically achieve the optimal DMT. While these theoretical results are encouraging, zero-forcing receivers are rarely used in practice. For example, the block linear ZF equalizer in [11] requires inversion of a matrix of dimension N-by-N, where N is the block length; thus, the equalizer is effectively a bank of equalizers. In addition, the decoding delay is as large as the block length. The infinite-length equalizer is not realizable, and can only be approximated, not to mention the fact that it has infinite decoding delay. Here we propose a realizable finite-length MMSE-DFE receiver which results in a decoding delay that can be chosen to be much shorter than the block length. As MMSE-DFE receivers minimize the mean square error (MSE), however, there is not only noise but residual ISI at the input the decision device. Hence, calculating the exact probability of error is difficult since the symbols are corrupted by residual interference which is not purely Gaussian.

The MMSE-DFE receiver structure which follows the receiver design in [13] is shown in Fig. 3. The feed forward filters can be designed as the finite impulse response filters $\boldsymbol{f}_1 \in \mathbb{C}^{L_{f1}}$ and $\boldsymbol{f}_2 \in \mathbb{C}^{L_{f2}}$ where $F_1(z) = \sum_{l=0}^{L_{f1}} f_1[l]z^{-l}$ and $F_2(z) = \sum_{l=0}^{L_{f2}} f_2[l]z^{-l}$. The feedback filter can be specified by $\boldsymbol{g} \in \mathbb{C}^{L_g}$ where $G(z) = \sum_{l=0}^{L_g} g[l]z^{-l}$.

To simplify the derivation, we assume that $L_{f1} + L_{sd} = L_{f2} + L_{r^*d}$. We can write the received signals from the source and from the selected relay as

$$\boldsymbol{y}_{sd}[k] = \boldsymbol{H}_{sd}\boldsymbol{x}[k] + \boldsymbol{w}_{sd}[k]$$

$$\boldsymbol{y}_{r^*d}[k] = \boldsymbol{H}_{r^*d}\boldsymbol{x}[k] + \boldsymbol{w}_{r^*d}[k]$$

where $\boldsymbol{H}_{sd} \in \mathbb{C}^{L_{f1}\times(L_{f1}+L_{sd}-1)}$ and $\boldsymbol{H}_{r^*d} \in \mathbb{C}^{L_{f2}\times(L_{f2}+L_{r^*d}-1)}$ are the Tœplitz channel convolution matrices defined, for example, as $[\boldsymbol{H}_{sd}]_{i,j} = h_{sd}[j-i]$ and $\boldsymbol{x}[k] = [x[k], x[k-1], \cdots, x[k-L_{f1}-L_{sd}+2]]^T \in \mathbb{C}^{L_{f1}+L_{sd}-1}$. Assume the desired output delay is δ, the coefficients of the FFF $\boldsymbol{f}_1$ and $\boldsymbol{f}_2$ and FBF $\boldsymbol{g}$ can be obtained by assuming correct past decisions and minimizing MSE $= \|x[k-\delta] - \tilde{x}[k]\|_2^2$ where $\tilde{x}[k] = \boldsymbol{u}^T[k]\boldsymbol{c}$ is the equalized output with the filter input $\boldsymbol{u}_k^T$ defined as

$$\boldsymbol{u}^T[k] \triangleq \left[\boldsymbol{y}_{sd}[k]^T, \boldsymbol{y}_{r^*d}[k]^T, \boldsymbol{x}^T[k]\boldsymbol{M}\right]$$

where

$$\boldsymbol{M}^T \triangleq \begin{bmatrix}\boldsymbol{0}_{\delta\times L_g} & \boldsymbol{I}_{L_g\times L_g} & \boldsymbol{0}_{L_g\times(L_{f1}+L_{sd}-L_g-\delta-1)}\end{bmatrix}$$

and the filter coefficient $\boldsymbol{c}$ defined as

$$\boldsymbol{c} \triangleq [\boldsymbol{f}_1^T\ \boldsymbol{f}_2^T\ \boldsymbol{g}^T]^T.$$

By applying the orthogonality principle, the equalizer coefficients are

$$\boldsymbol{c} = (E[\boldsymbol{u}^*[k]\boldsymbol{u}^T[k]])^{-1}E[\boldsymbol{u}^*[k]x[k-\delta]]. \tag{10}$$

When the input and noise processes are mutually uncorrelated, we can compute the correlation

$$E[\boldsymbol{u}^*[k]x[k-\delta]] = \begin{bmatrix}(\boldsymbol{H}_{sd}^*\boldsymbol{e}_\delta)^T & (\boldsymbol{H}_{r^*d}^*\boldsymbol{e}_\delta)^T & \boldsymbol{0}_{1\times L_g}\end{bmatrix}^T \tag{11}$$

where $e_\delta[i] = 0$ except $e_\delta[\delta] = 1$ with $1 \le i \le L_{f1} + L_{sd} - 1$. To complete the computation of the coefficient $\boldsymbol{c}$ in (10), we can further compute the autocorrelation matrix

$$E[\boldsymbol{u}^*[k]\boldsymbol{u}^T[k]] = \begin{bmatrix}\boldsymbol{H}^*\boldsymbol{H}^T + N_0\boldsymbol{I}_{L_{f1}+L_{f2}} & \boldsymbol{H}^*\boldsymbol{M} \\ \boldsymbol{M}^T\boldsymbol{H}^T & \boldsymbol{I}_{L_g}\end{bmatrix} \tag{12}$$

where $\boldsymbol{H}^T = \begin{bmatrix}\boldsymbol{H}_{sd}^T & \boldsymbol{H}_{r^*d}^T\end{bmatrix}$ is the composite SIMO channel. Applying the inversion of a block matrix to (12) and substituting the results and (11) to the cross-correlation (10), we have the feed-forward filter coefficients

$$\boldsymbol{f} = \begin{bmatrix}\boldsymbol{f}_1 \\ \boldsymbol{f}_2\end{bmatrix} = (\boldsymbol{H}^*(\boldsymbol{I} - \boldsymbol{M}\boldsymbol{M}^T)\boldsymbol{H}^T + N_0\boldsymbol{I})^{-1}\boldsymbol{H}^*\boldsymbol{e}_\delta \tag{13}$$

and the feedback filter coefficients

$$\boldsymbol{g} = -\boldsymbol{M}^T\boldsymbol{H}^T\boldsymbol{f}. \tag{14}$$

V. Numerical Results

This section presents numerical examples for the performance of the proposed relay selection method. In evaluating performance over finite SNRs, the diversity measured as the negative slope of each outage curve often does not always coincide exactly with the predicted maximal diversity [14], [15] since the predicted diversity assumes that the SNR grows arbitrarily large to permit the analysis to be mathematically tractable.

In the simulations, we assume the source transmits blocks of length $N = 32$ Gray-mapped QPSK symbols. We further assume that each channel in the system has uniform power delay profile, i.e. each tap of each channel experiences i.i.d. fading with variance $1/L$ where L is the channel length.

We compare the BER performance of three receivers: finite-length MMSE-DFE equalizer as in Section IV, linear ZFE, and MLSE equalization.

Fig. 4 shows the BER performance comparison of the three schemes when there are 2 relays in the system. In Fig. 4, all

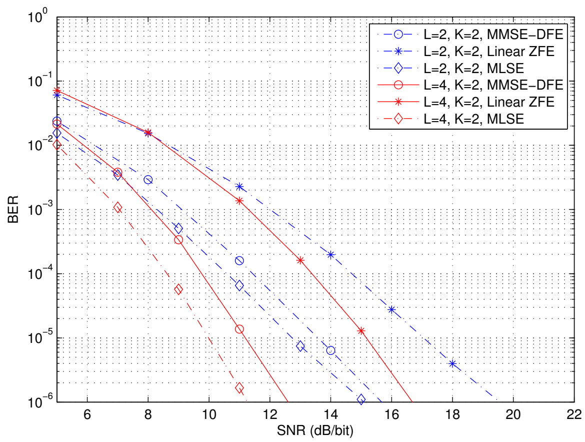

Fig. 4. BER performance with 2 relays.

BER curves of $L = 4$ have approximately equal negative slope of 6.3, and all BER curves with $L = 2$ have approximately equal negative slope of 4, which shows that all of the three schemes can exploit the FS diversity. Fig. 5 shows the BER performance of the three schemes with 10 relays in the system. We found that in Fig. 5 the negative slope is approximately 5 for all BER curves of $L = 2$ and 7.5 for all BER curves of $L = 4$. On both Fig. 4 and Fig. 5, while the diversity orders achieved by the three receivers are approximately the same, the receiver with finite-length MMSE-DFE equalizers and receiver with MLSE have much larger coding gain than the receiver with linear ZF equalizer. Comparing the two figures, we found that with increased number of relays, the power gain is larger, and the diversity order is increasing though not as predicted by theoretical analysis due to finite SNR. This shows that the relay selection method is able to exploit the cooperative diversity offered by the relays.

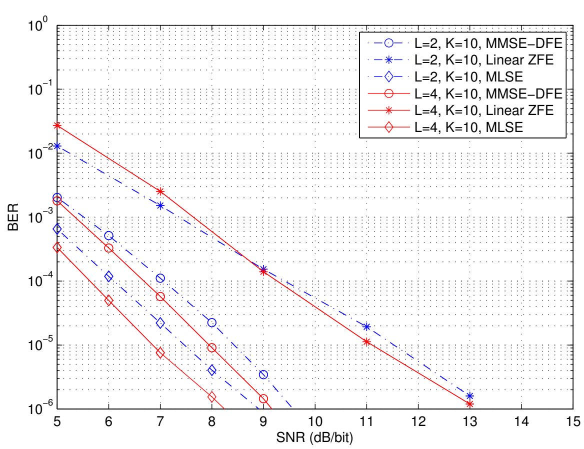

Fig. 5. BER performance with 10 relays.

VI. Conclusion

In this paper, we considered opportunistic relaying with FS fading for the decode-and-forward protocol. Our proposed relay selection method coincides with the relay selection method in [11] which selects the relay with the largest norm of the relay-destination channel. We analyzed the relay selection method with the infinite-length MMSE-DFE receiver, and proved that it can achieve the optimal DMT. Simulation results show that such a relay selection method combined with more practical finite-length MMSE-DFE receivers achieves full diversity, as well.

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