# LINEAR COMPLEX-FIELD CODING FOR COOPERATIVE NETWORKING

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## ABSTRACT

Commonly used protocols involving $J$ cooperating communicators are based on repetition encoding and achieve diversity of order $J$ with bandwidth efficiency $1/J$. We introduce a protocol capable of achieving the same diversity with bandwidth efficiency essentially equal to $1/2$. The protocol is based on linear complex-field coded (LCFC) relay transmissions over orthogonal frequency division multiplexed (OFDM) subcarriers. Cooperators provide diversity by repeating delayed versions of the original packet, thereby generating a frequency-selective multipath channel. The so enabled diversity is collected by standard LCFC-OFDM decoders. Analysis and corroborating simulations establish that the novel protocol achieves diversity order equal to the number of users.

## 1. INTRODUCTION

User cooperation is a promising spatial diversity enabler that has been recently introduced as an alternative to collocated multi-antenna systems. The basic principle is to let users share data packets in order to effect a distributed virtual antenna array [1, 2, 7, 8]. Active users engage in a data sharing phase followed by a cooperation phase [2, 8]. Recently, there has also been a growing interest in cooperation provided by idle users in a setup reminiscent of relay channels which is particularly attractive for fixed multi-access, random access, and ad-hoc networks [6, 7].

Of particular relevance to the present work is the observation that user cooperation may be regarded as a form of multipath [5]. From this vantage point, cooperative protocols become available whereby diversity can be *provided* by a frequency-selective channel created by relay transmissions, and then *collected* at the destination by any of the available decoders that have been designed over the years to deal with multipath. To this end, orthogonal frequency division multiplexing (OFDM) is particularly attractive since it diagonalizes the resultant multipath channel and thus offers a simple means of coping with the emerging inter-symbol interference (ISI); see e.g., [9] and references therein. Although uncoded OFDM cannot enable the underlying multipath diversity, error control coding (ECC) and recent alternatives are known to remedy this limitation. One popular alternative is linear complex-field coding (LCFC), where instead of a single uncoded symbol per subcarrier, one transmits distinct linear combinations of the information symbols [10, 11]. Unlike ECC, LCFC does not sacrifice bandwidth.

Building on these results we introduce a protocol in which cooperators re-transmit properly delayed versions of OFDM symbols. Due to OFDM, transmissions from different users create an easy-to-deal with frequency-selective channel that provides multipath diversity which, thanks to LCF coding/decoding, can be collected at the destination. Besides the fact that the novel protocol relies on well-tested approaches to dealing with multipath, it offers a twofold advantage:

1. Unlike existing repetition protocols that achieve diversity $J+1$ at the price of lowering spectral efficiency to $1/(J+1)$, the LCFC-based one ensures the same diversity at spectral efficiency approximately $1/2$.
2. In contrast to existing protocols which need synchronization either at the bit-level [2] or even at the chip-level [8], the LCFC-based one requires synchronization at packet-level only.

The rest of the paper is organized as follows. In Section 2, we present the system model and introduce the multipath-inducing protocol. In Section 3, we analyze the novel protocol and show how LCFC can achieve a diversity order equal to the number of users. In Section 4, we describe corroborating simulations, and conclude the paper in Section 5.

## 2. LCFC AND USER COOPERATION

Consider a set of $J+1$ users $\{u_j\}_{j=0}^{J}$, where $u_0$ is the active source communicating to a destination $D$ in cooperation with the remaining users $\{u_j\}_{j=1}^{J}$. Let $h_{j1,j2}$ denote the Rayleigh fading channel between $u_{j_1}$ and $u_{j_2}$ with average power $\bar{h}_{j1,j2} := \mathrm{E}[|\bar{h}_{j1,j2}|^2]$. In particular, let $h_{j,D} := h_j$ denote the channel $u_j \to D$. The information stream from $u_0$ is parsed in $N_s \times 1$ blocks $\mathbf{s}(n) := [s(nN_s), \ldots, s(nN_s+N_s-1)]^T$, with the entries of $\mathbf{s}(n)$ drawn from a signal constellation $\mathcal{S}$ so that $\mathbf{s}(n) \in \mathcal{S}^{N_s}$. This block is then modulated onto OFDM subcarriers by applying the inverse (I-)fast Fourier transform (FFT). With $\mathbf{F}_{N_s}$ denoting the $N_s \times N_s$ FFT matrix with entries

$$[\mathbf{F}_{N_s}]_{n,k} = (1/\sqrt{N_s})\exp(-j2\pi nk/N_s) \;, \tag{1}$$

the $n$th block $\mathbf{x}(n) := \mathbf{F}_{N_s}^{\mathcal{H}}\mathbf{s}(n)$ with power $P$ per entry is broadcasted over the flat fading wireless channel which is described by the $N_s \times N_s$ diagonal matrix $\mathbf{D}_{0,j} := \mathrm{diag}[h_{0,j}, \ldots, h_{0,j}]$. In discrete-time equivalent baseband form, the block is received at $u_j$ as

$$\mathbf{y}_{u_j}(n) = \mathbf{D}_{0,j}\mathbf{x}(n) + \mathbf{w}_{u_j}(n) \;, \quad j \in [1, J] \tag{2}$$

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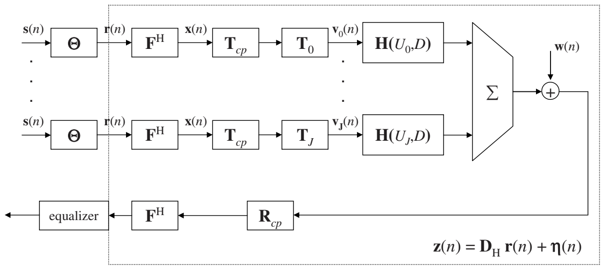


**Fig. 1**. Block diagram for transmission and reception.

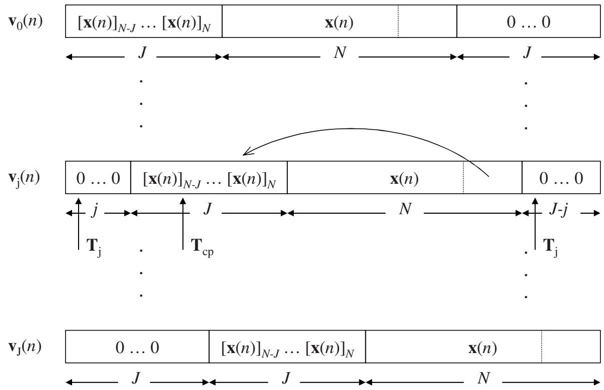


**Fig. 2**. Packets transmitted during the cooperative phase.

where $\mathbf{w}_{u_j}(n)$ denotes additive white Gaussian noise (AWGN) with power $\mathrm{E}[\mathbf{w}_{u_j}^{\mathcal{H}}(n)\mathbf{w}_{u_j}(n)] = \sigma^2\mathbf{I}_{N_s}$ ($\mathbf{I}_{N_s}$ stands for the identity matrix of size $N_s$).

Each block is demodulated by multiplying $\mathbf{y}_{u_j}(n)$ with the FFT matrix to form the decision vector

$$\mathbf{z}_{u_j}(n) := \mathbf{F}_{N_s}\mathbf{y}_{u_j}(n) = \mathbf{D}_{0,j}\mathbf{s}(n) + \quad_{u_j}(n) \tag{3}$$

where $\mathbf{D}_{0,j} := \mathbf{F}_{N_s}\mathbf{D}_{0,j}\mathbf{F}_{N_s}^{\mathcal{H}}$ and the noise $\quad_{u_j}(n) := \mathbf{F}_{N_s}\mathbf{w}_{u_j}(n)$ remains white with power $\mathrm{E}[\ {}^{\mathcal{H}}_{u_j}(n)\ \ _{u_j}(n)] = \mathbf{F}_{N_s}\mathrm{E}[\mathbf{w}_{u_j}^{\mathcal{H}}(n)\mathbf{w}_{u_j}(n)]\mathbf{F}_{N_s}^{\mathcal{H}} = \sigma^2\mathbf{I}_{N_s}$, since the FFT matrix is unitary; i.e., $\mathbf{F}_{N_s}\mathbf{F}_{N_s}^{\mathcal{H}} = \mathbf{I}_{N_s}$.

Likewise, upon defining $\mathbf{D}_0 := \mathrm{diag}[h_0, \ldots, h_0]$ we can write the equivalent input-output relationship at the destination $D$ during this first "reach-out" phase as

$$\mathbf{z}_1(n) := \mathbf{D}_0\mathbf{s}(n) + \quad_1(n) \ . \tag{4}$$

Block $\mathbf{z}_1(n)$ will be combined with LCF coded blocks transmitted from the cooperators during the second "relay" phase.

Demodulating $\mathbf{z}_{u_j}(n)$ in (3), each user performs maximum-likelihood detection to recover the source packet as

$$\hat{\mathbf{s}}_{u_j}(n) = \arg\min_{\hat{\mathbf{s}}_{u_j}(n)\in\mathcal{S}^{N_s}} \|\mathbf{D}_{0,j}\hat{\mathbf{s}}_{u_j}(n) - \mathbf{z}_{u_j}(n)\| \ . \tag{5}$$

The demodulator in (5) is equivalent to a symbol-by-symbol detector since $[\mathbf{z}_{u_j}(n)]_{k_1}$ is independent of $[\mathbf{z}_{u_j}(n)]_{k_2}$ for $k_1 \neq k_2$.

Use of OFDM in this first phase brings no particular advantage and is adopted for consistency with the second phase we describe next, where its instrumental role will become clear.

### 2.1. User cooperation

If the source packet is correctly decoded; i.e., $\hat{\mathbf{s}}_{u_j}(n) = \mathbf{s}(n)$, then $u_j$ proceeds to retransmit it in the next time slot. This transmission entails three steps (see also Fig. 1):

1. *Precoding*: Prior to transmission, $\mathbf{s}(n)$ is left multiplied by the $N \times N_s$ LCF matrix $\boldsymbol{\Theta}$ to yield the $N \times 1$ LCF coded block
$$\mathbf{u}(n) := \boldsymbol{\Theta}\mathbf{s}(n). \tag{6}$$
2. *OFDM modulation*: With $\mathbf{F}_N$ denoting the $N \times N$ FFT matrix in (1) with $N_s = N$, block $\mathbf{u}(n)$ is IFFT processed to obtain
$$\mathbf{x}(n) := \mathbf{F}_N^{\mathcal{H}}\mathbf{u}(n) = \mathbf{F}_N^{\mathcal{H}}\boldsymbol{\Theta}\mathbf{s}(n), \tag{7}$$
where the transmit power per entry is $P := \mathrm{E}[[\mathbf{x}(n)]_k^2]$.
3. *Cyclic prefix insertion and delay*: Define the matrix $\mathbf{T}_{cp} := [\mathbf{I}_{cp}^T, \mathbf{I}_N^T]^T$ formed by the concatenation of the last $L$ rows of the $N \times N$ identity matrix (that we denote as $\mathbf{I}_{cp}$) and the identity matrix itself. Also define a "pure-delay" matrix $\mathbf{T}_j := [\mathbf{0}_{N,j}, \mathbf{I}_N, \mathbf{0}_{N,J-j}]^T$ where $\mathbf{0}_{N,j}$ denotes the $N \times j$ all-zero matrix. Using these two matrices, we form the block transmitted by the cooperators as
$$\mathbf{v}_{u_j}(n) := \mathbf{T}_j\mathbf{T}_{cp}\mathbf{x}(n) = \mathbf{T}_j\mathbf{T}_{cp}\mathbf{F}_N^{\mathcal{H}}\boldsymbol{\Theta}\mathbf{s}(n) \ . \tag{8}$$
Note that $\mathbf{T}_{cp}$ is a cyclic prefix insertion matrix so that when left multiplied by an $N \times 1$ vector $\mathbf{x}(n)$ it yields an $(N+J) \times 1$ vector which places the last $J$ entries of $\mathbf{x}(n)$ on top of $\mathbf{x}(n)$ [9]. Left multiplying $\mathbf{x}(n)$ by $\mathbf{T}_j$ creates its delayed version by $j$-bits and when performed by $J$ cooperators it manifests a $J$th order frequency-selective channel; see also Fig. 2.

In the second phase, $D$ receives the superposition of these $J+1$ transmissions (source plus cooperators) in the presence of AWGN given by the $(N+2J) \times 1$ block

$$\mathbf{r}_2(n) = \sum_{j=0}^{J} \tilde{\mathbf{H}}_j\mathbf{v}_j(n) + \mathbf{w}(n) \tag{9}$$

where $\tilde{\mathbf{H}}_j := \mathrm{diag}[\tilde{h}_j, \ldots, \tilde{h}_j]$ denotes the corresponding channel matrix with $\tilde{h}_j = h_j$, if $\hat{\mathbf{s}}_{u_j}(n) = \mathbf{s}(n)$, and $\tilde{h}_j = 0$ otherwise; (Note that $\tilde{h}_j$ accounts both for the underlying propagation medium and also for the probability that cooperator $u_j$ may not participate in the second phase due to decoding errors.)

The first task at the receiver side is to remove the first and last $J$ bits that are affected by interblock interference. This is accomplished by left multiplying $\mathbf{r}_2(n)$ with the matrix $\mathbf{R}_{cp} = [\mathbf{0}_{N,J}, \mathbf{I}_N^T, \mathbf{0}_{N,J}]^T$ to obtain

$$\begin{aligned}\mathbf{y}_2(n) &= \mathbf{R}_{cp}\left[\sum_{j=0}^{J}\mathbf{H}_j\mathbf{T}_j\right]\mathbf{T}_{cp}\mathbf{F}^{\mathcal{H}}\mathbf{x}(n) + \mathbf{R}_{cp}\mathbf{w}(n) \\ &:= \mathbf{H}\mathbf{F}^{\mathcal{H}}\mathbf{x}(n) + \tilde{\mathbf{w}}(n), \end{aligned} \tag{10}$$

where we defined $\tilde{\mathbf{w}}(n) := \mathbf{R}_{cp}\mathbf{w}(n)$ and the equivalent channel matrix $\mathbf{H} := \mathbf{R}_{cp}\left[\sum_{j=0}^{J}\mathbf{H}_j\mathbf{T}_j\right]\mathbf{T}_{cp}$. Interestingly, it is easy to see that $\mathbf{H}$ is a circulant matrix given by

$$\mathbf{H} = \begin{pmatrix} h_0 & 0 & \dots & \dots & \dots & 0 & \tilde{h}_J & \dots & \tilde{h}_1 \\ \vdots & \ddots & \ddots & \dots & \dots & \dots & \ddots & \ddots & \vdots \\ \tilde{h}_{J-1} & \dots & h_0 & 0 & \dots & \dots & \dots & \ddots & \tilde{h}_J \\ \tilde{h}_J & \dots & \tilde{h}_1 & h_0 & 0 & \dots & \dots & \dots & 0 \\ \vdots & \ddots & \vdots & \ddots & \ddots & \ddots & \dots & \dots & \vdots \\ 0 & \dots & \tilde{h}_J & \dots & \tilde{h}_1 & h_0 & 0 & \dots & 0 \end{pmatrix}. \quad (11)$$

Indeed, we can see that upon adopting the convention $h_{[u_{(l-k)} \mod N]} = 0$ for $j > J$, the elements of $\mathbf{H}$ are $[\mathbf{H}]_{lk} = h(u_{(l-k)} \mod N)$, which by definition corresponds to a circulant matrix.

Next, let us recall that circulant matrices are diagonalized by the (I)FFT processing. Thus, after OFDM demodulation we obtain the decision vector

$$\mathbf{z}_2(n) := \mathbf{F}_N\mathbf{y}_2(n) = \mathbf{D}_H\boldsymbol{\Theta}\mathbf{s}(n) + \ {}_2(n), \quad (12)$$

where ${}_2(n) := \mathbf{F}_N\tilde{\mathbf{w}}(n)$ is AWGN with power $\mathrm{E}[\ {}_2^{\mathcal{H}}(n)\ {}_2(n)] = \sigma^2\mathbf{I}_N$ (once more because the FFT matrix is unitary), and $\mathbf{D}_H := \mathbf{F}_N\mathbf{H}\mathbf{F}_N^{\mathcal{H}}$ is a diagonal matrix.

To combine the blocks received by $D$ in both phases we define $\mathbf{z}(n) := [\mathbf{z}_1^T(n), \mathbf{z}_2^T(n)]^T$ and $\mathbf{H}_{eq} := [\mathbf{D}_0^T, (\mathbf{D}_H\boldsymbol{\Theta})^T]^T$. The maximum likelihood estimate of $\mathbf{s}(n)$ based on received signals from the first and second phases is finally obtained as

$$\hat{\mathbf{s}}(n) = \arg\min_{\hat{\mathbf{s}}(n)\in\mathcal{S}^{N_s}} \|\mathbf{H}_{eq}\hat{\mathbf{s}}(n) - \mathbf{z}(n)\|. \quad (13)$$

This second phase transmission entails creating a multipath channel with $J+1$ taps by means of properly delayed transmissions from each of the cooperating users. For transmission over this "manually created" multipath channel we use OFDM modulation which diagonalizes the channel and thus simplifies equalization at the receiver. Unfortunately, uncoded OFDM does not enable the diversity provided by the multipath channel. Effecting this diversity, is the task of the LCF precoding matrix $\boldsymbol{\Theta}$ as we outline in the next section.

## 3. ERROR PERFORMANCE ANALYSIS

The protocol outlined in Section 2 generates the frequency selective channel described by the matrix $\mathbf{H}$ in (11). This circulant matrix is in turn uniquely specified by the impulse response vector

$$\mathbf{h} := [h_0, \tilde{h}_1, \dots, \tilde{h}_J]^T, \quad (14)$$

a relation that we denote as $\mathbf{H} = \mathrm{circ}(\mathbf{h}, N)$. If entries of $[\mathbf{h}]_j = \mathcal{N}[0, \bar{h}_j]$ of $\mathbf{h}$ are uncorrelated Gaussian distributed, then it is well known that for large enough signal to noise ratio (SNR), the error probability is $P_e \le G_c(P/\sigma^2)^{-(J+1)}$, where $G_c$ is a constant (a.k.a. coding gain) not dependent on the SNR$=(P/\sigma^2)$. In the case at hand though, $[\mathbf{h}]_j$ is non-Gaussian distributed. Nonetheless, it is possible to ensure diversity advantage of order $(J+1)$, as we show in the following theorem.

**Theorem 1** *If any $N-J$ rows of $\boldsymbol{\Theta}$ span $\mathbf{C}^{N_s}$, then the error probability for the protocol described in Section 2 is upper bounded as*

$$P_e \le G_c\gamma^{-(J+1)} \quad (15)$$

*where $\gamma := P/\sigma^2$ denotes input SNR; and $G_c$ is a function of $\boldsymbol{\Theta}$, the channel powers $\bar{h}_{j1,j2}$, the constellation $\mathcal{S}$ and the number of users $J$, but does not depend on $\gamma$.*

**Proof:** Let $\mathcal{C} := \{u_j|\hat{\mathbf{s}}_{u_j}(n) = \mathbf{s}(n)\}$ be the set of users that successfully decoded $\mathbf{s}(n)$. (Note that $u_0 \in \mathcal{C}$ always.) Define the channel vector

$$\mathbf{h}(\mathcal{C}) := [h_0, \tilde{h}_{1|\mathcal{C}}, \dots, \tilde{h}_{J|\mathcal{C}}]^T, \quad (16)$$

comprising the coefficients of the matrix $\mathbf{H}$ when the set of cooperators is $\mathcal{C}$. Letting $|\mathcal{C}|$ denote the cardinality of $\mathcal{C}$, we observe that $\mathbf{h}(\mathcal{C})$ contains $(J+1)-|\mathcal{C}|$ zero elements and $\mathcal{C}$ independent Gaussian distributed elements. Thus, $\mathbf{h}(\mathcal{C})$ is a Rayleigh fading FIR channel with $|\mathcal{C}|$ taps and consequently the error probability is bounded as [10, Theorem 3]

$$P_e(\mathcal{C}) \le G_{c,\mathcal{C}}\gamma^{-|\mathcal{C}|}. \quad (17)$$

On the other hand, the probability that a given user does not belong to $\mathcal{C}$ can also be bounded since [4, Section 14.3]

$$\Pr\{u_j \notin \mathcal{C}\} = \Pr\{\hat{\mathbf{s}}_{u_j}(n) \ne \mathbf{s}(n)\} \le G_{c,j}\bar{h}_{0,j}\gamma^{-1}, \quad (18)$$

for some constant $G_{c,j}$. Using (18), we can further bound the probability of a particular set of cooperators as

$$\Pr\{\mathcal{C}\} \le \prod_{u_j\notin\mathcal{C}} G_{c,j}\bar{h}_{0,j}\gamma^{-1} := G'_{c,\mathcal{C}}\gamma^{-(J+1-|\mathcal{C}|)}. \quad (19)$$

Finally, relying upon the theorem of total probability we can write [c.f (17) and (19)]

$$\begin{aligned} P_e &= \sum_{\mathcal{C}} P_e(\mathcal{C})\Pr\{\mathcal{C}\} \quad (20) \\ &\le \sum_{\mathcal{C}} G_{c,\mathcal{C}}\gamma^{-|\mathcal{C}|}G'_{c,\mathcal{C}}\gamma^{-(J+1-|\mathcal{C}|)} \\ &= \left[\sum_{\mathcal{C}} G_{c,\mathcal{C}}G'_{c,\mathcal{C}}\right]\gamma^{-(J+1)} \end{aligned}$$

Defining $G_c := \left[\sum_{\mathcal{C}} G_{c,\mathcal{C}}G'_{c,\mathcal{C}}\right]$, eq. (15) follows readily. □

Theorem 1 establishes that the protocol described in Section 2 achieves $J+1$-order diversity with spectral efficiency $N_s/(N_s+N+2J) \approx 1/2$, where in asserting the last approximation we used that $N_s \approx N \gg J$. As for the existence of choices $\boldsymbol{\Theta}$ satisfying the conditions of Theorem 1, there are families of matrices known to satisfy them. Among these, the most commonly used are the Vandermonde and Cosine matrices [10].

**Remark 1** The treatment here has considered a single user $u_0$ communicating with $D$; but our protocol can be readily tuned to a multi-source setting by invoking orthogonal frequency division multiple access (OFDMA). The diversity order asserted by Theorem 1 holds true with LCF coded OFDMA. Furthermore, relying on OFDMA with pseudo-random subcarrier hopping, and following arguments in [5] it is possible to reach spectral efficiency as high as $N_s/(N+J) \approx 1$.

## 4. SIMULATIONS

In this section, we present the results of simulating the protocol in Section 2 under different assumptions. We first consider the case of error-free $u_0 \to u_j$ links which serves as a benchmark for the more realistic case considered later. For simplicity, we choose a tall precoder such that $N - N_s = J$, although other precoders (square or fat) are also capable of achieving full diversity [10, 11].

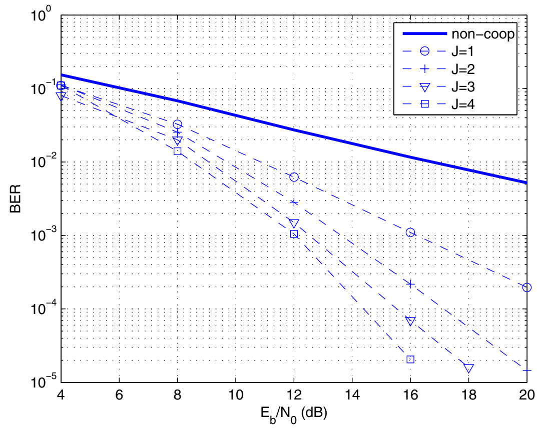


**Fig. 3**. BER for different SNRs between user-user pairs

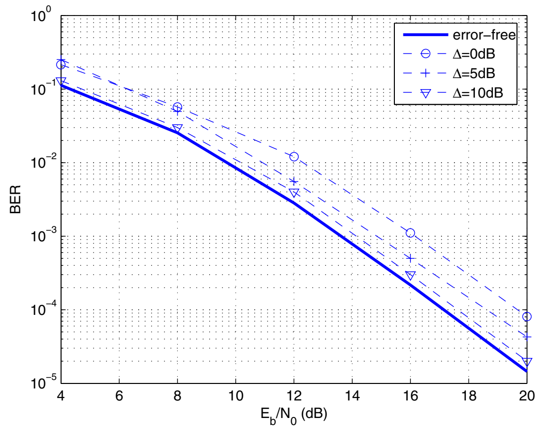


**Fig. 4**. BER for different SNRs between user-user pairs

### 4.1. Error-free links among users

Fig. 3 demonstrates how bit error rate (BER) performance varies with the number of cooperators $J$. We use $N = 16$-point FFT with the number of information symbols therefore given by $N_s = 16 - J$. For near maximum likelihood (ML) demodulation, we used the sphere decoding algorithm [3]. As predicted by Theorem 1, the diversity order achieved is $J + 1$.

### 4.2. Fading user-user channels

To account for fading in the $u_0 \rightarrow u_j$ links we generated the channel coefficients as independent identically distributed (i.i.d.) Rayleigh random variates. Fig. 4 depicts the performance of our protocol for different values of $\Delta = \gamma_1 - \gamma_2$, where $\gamma_1$ and $\gamma_2$ are the average SNRs of user-user and user-destination pairs respectively. We use $N = 16$, $J = 2$, and therefore $N_s = 14$. The number of information bits per block is 200 with perfect error detection assumed. From Fig. 4, we confirm that for different values of $\Delta$, the diversity order is $J + 1 = 3$, as predicted by Theorem 1. Note also that the performance improves as $\Delta$ increases; and when $\Delta = 10$, it comes very close to that of error-free user-user links.

## 5. CONCLUSIONS

We introduced a novel wireless cooperative protocol which hinges on the idea of generating a frequency-selective multipath channel through cooperating relay transmissions. The diversity provided by this "manually" (as opposed to physically) created multipath channel is effected by a linear complex-field coder (LCFC) used in conjunction with orthogonal frequency division multiplexing (OFDM). With $J$ cooperators, the novel protocol offers diversity of order $J + 1$ with bandwidth efficiency essentially equal to 1/2, while commonly used cooperative protocols based on repetition encoding operate at 1/J bandwidth efficiency[1].

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