Multicode STBC in Frequency Selective Fading

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***Abstract*—Multicode operation in space-time block coded (STBC) multiple input multiple output (MIMO) systems can provide additional degrees of freedom in code domain to achieve high data rates. In such multicode STBC systems, the receiver experiences code domain interference (CDI) in frequency selective fading. In this paper, we propose a linear parallel interference cancellation (LPIC) approach to cancel the CDI in multicode STBC signals in frequency selective fading. The proposed detector first performs LPIC followed by STBC decoding. We present an SINR for the proposed detector. We evaluate the bit error rate (BER) performance of the system, and show that the proposed detector effectively cancels the CDI and achieves improved error performance. Our BER results further illustrate how the combined effect of interference cancellation, transmit diversity, and RAKE diversity affects the performance of the system.**

Keywords – *MIMO, multicode STBC, frequency selective fading, interference cancellation, transmit diversity, RAKE diversity.*

I. Introduction

Multiple input multiple output (MIMO) techniques that employ multiple antennas both at the receiver as well as the transmitter can offer the benefits of high data rates and diversity gain [1]. Diversity-multiplexing gain tradeoff is central to such MIMO systems [2]. It would be desirable to have systems which provide high data rates and at the same time do not compromise on data reliability. Space-time coding using orthogonal space-time block codes (OSTBC) is a well known means to achieve improved reliability, by way of transmit diversity. In order to achieve a good tradeoff between multiplexing gain and diversity gain, an architecture for combining STBC and V-BLAST (referred to as **G**roup **L**ayered **S**pace-**T**ime - GLST architecture) has been proposed in [3]. Efficient detectors for these GLST schemes have also been proposed in the recent literature [4].

An alternate way to achieve flexibility in the diversity-multiplexing gain tradeoff is through the use of MIMO in conjunction with multicode spread spectrum techniques. Multicode operation can provide additional degrees of freedom in the code domain to achieve high data rates. Multicode spread-spectrum techniques in conjunction with V-BLAST have been proposed in 3G standards [5] and ad-hoc networks [6] to achieve high data rates, but without using STBC for achieving transmit diversity. Multicode operation in STBC systems, on the other hand, can provide high data rates using additional degrees of freedom in the code domain, while retaining the transmit diversity gains offered by the STBCs. In such multicode STBC systems, the receiver experiences code domain interference (CDI) in frequency selective fading (delay spread larger than one chip duration), which can degrade the bit error performance. Interference cancelling detectors can alleviate this problem at the cost of receiver complexity.

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In this paper, we propose a linear parallel interference cancellation (LPIC) approach to cancel the CDI caused by the frequency selective nature of the channel in multicode STBC systems. The proposed detector first performs LPIC followed by STBC decoding. We present an SINR analysis for the proposed detector. We evaluate the bit error rate (BER) performance of the system, and show that the proposed detector effectively cancels the CDI and achieves improved error performance. Our BER results further illustrate how the combined effect of interference cancellation, transmit diversity, and RAKE diversity affect the performance of the system.

II. System Model

We consider a space-time block coded MIMO system with M transmit antennas as shown in Fig. 1. The input data stream is demultiplexed into K parallel substreams, where K is the number of spreading codes to be multiplexed on each transmit antenna. Each data substream is fed to a M-transmit antenna STBC encoder. There are K STBC encoders, one for each data substream. The M outputs of the kth STBC encoder, $k = 1, 2, \cdots, K$, are spread using a spreading code $c_k(t)$ assigned to the kth data substream. There are K spreading codes, one for each data substream. The mth spread output of all the K data substreams are multiplexed and mounted on the mth transmit antenna, $m = 1, 2, \cdots, M$. In this multicode STBC architecture, increasing K can increase the data rate and increasing M can increase the order of diversity gain.

In the following, we describe the system model using Alamouti scheme (which sends two symbols in two time slots on two transmit antennas), although the model is applicable to other OSTBCs as well. Let $s_{k,1}$ and $s_{k,2}$ denote, respectively, the 1st and 2nd complex information symbols of the kth data substream. We assume QPSK modulation, and therefore each information symbol can take one of four values in $\{(\pm 1 \pm j)/\sqrt{2}\}$. In the Alamouti scheme, $s_{k,1}$ is mounted on transmit antenna 1 (tx1) during time slot 1 (ts1), and $s^*_{k,1}$ is mounted on the transmit antenna 2 (tx2) during time slot 2 (ts2). Similarly, $s_{k,2}$ is mounted on tx2 during ts1, and $-s^*_{k,2}$ is mounted on tx1 during ts2. The spreading waveform for the kth data substream, $c_k(t)$, is given by

$$c_k(t) \;=\; \sum_{p=0}^{P-1} c_{k,p}\, \psi(t - pT_c), \quad k = 1, 2, \cdots, K, \quad (1)$$

where T_c is one chip duration, T_s is one symbol duration, $c_{k,p}$ is the pth chip of the kth spreading code, $P = T_s/T_c$ is the processing gain, and $\psi(t)$ is the chip waveform, which is assumed to be rectangular, i.e., one for $0 \leq t \leq T_c$ and zero otherwise. The chip sequence, $c_{k,p}$ is assumed to be a complex spreading sequence, and is given by $c_{k,p} = c^{(real)}_{k,p} + jc^{(imag)}_{k,p}$, where $c^{(real)}_{k,p}$ and $c^{(imag)}_{k,p}$ take the random values of $+1/\sqrt{2}$ and $-1/\sqrt{2}$ with equal probability. Moreover, $c_k(t)$ are mutually orthogonal for all k, i.e., $\sum_{p=0}^{P-1} c_{k_1,p}\, c^*_{k_2,p} = 0, \;$ for $k_1 \neq$

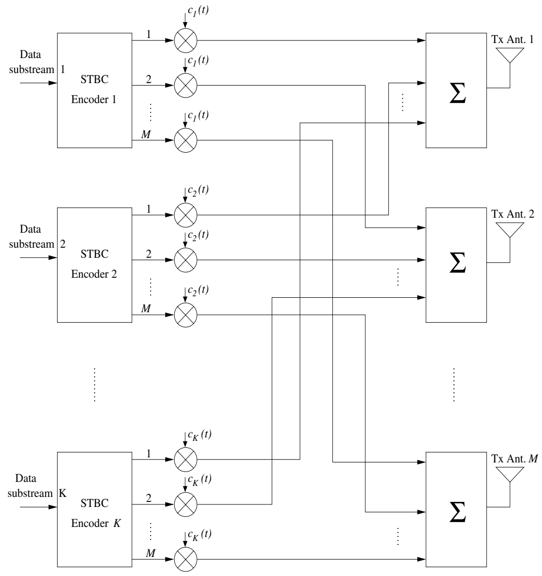

Fig. 1. Multicode STBC transmitter.

k_2. The spread signals sent on transmit antennas tx1 and tx2 during time slots ts1 and ts2, denoted by $S_{tx1}^{ts1}(t)$, $S_{tx2}^{ts1}(t)$, $S_{tx1}^{ts2}(t)$, and $S_{tx2}^{ts2}(t)$, respectively, are then given by

$$S_{tx1}^{ts1}(t) = \sum_{k=1}^{K} s_{k,1}\, c_k(t), \quad S_{tx2}^{ts1}(t) = \sum_{k=1}^{K} s_{k,2}\, c_k(t), \quad (2)$$

$$S_{tx1}^{ts2}(t) = \sum_{k=1}^{K} -s_{k,2}^{*}\, c_k(t), \quad S_{tx2}^{ts2}(t) = \sum_{k=1}^{K} s_{k,1}^{*}\, c_k(t). \quad (3)$$

Channel Model: We consider a frequency selective tapped delay line multipath fading channel. The complex channel impulse response from the mth transmit antenna to the nth receive antenna is expressed as

$$h_{n,m}(t) = \sum_{l=0}^{L-1} h_{n,m,l}\, \delta(t - lT_c), \quad (4)$$

where L is the number of resolvable multipaths, $h_{n,m,l}$ is the complex fading coefficient from the mth transmit antenna to the nth receive antenna on the lth multipath, and $\{h_{n,m,l}\}$'s are assumed to be circularly symmetric complex Gaussian r.v's with zero mean. It is assumed that $\{h_{n,m,l}\}$'s are constant over two time slots duration, and are independent for all n, m and l. The second moment of the channel amplitude $|h_{n,m,l}|$, denoted by Ω_l, is assumed to have an exponential multipath intensity profile (MIP) given by $\Omega_l = E\left[|h_{n,m,l}|^2\right] = \Omega_0 e^{-l\beta}, \ \ l = 0, 1, \cdots, L-1$, where β represents the rate of the exponential decay of the average path power. We assume that the delay spread is small compared to the symbol duration (i.e., $L << P$) so that there is inter-chip interference but no inter-symbol interference.

Received Signal Model: We will present the received signal model for the case of one receive antenna. Because of this, we will drop the receive antenna index n. It is straightforward to extend the model to more than one Rx antenna. Using (3) and (4), the signals received from tx1 in time slots ts1 and ts2, denoted by $R_{tx1}^{ts1}(t)$ and $R_{tx1}^{ts2}(t)$, respectively, are given by

$$\begin{aligned} R_{tx1}^{ts1}(t) &= \sum_{l=0}^{L-1}\sum_{k=1}^{K} h_{1,l}\, c_k(t - lT_c)\, s_{k,1}, \\ R_{tx1}^{ts2}(t) &= \sum_{l=0}^{L-1}\sum_{k=1}^{K} -h_{1,l}\, c_k(t - lT_c)\, s_{k,2}^{*}. \end{aligned} \quad (5)$$

Likewise, the signals received from tx2 in time slots ts1 and ts2, denoted by $R_{tx2}^{ts1}(t)$ and $R_{tx2}^{ts2}(t)$, respectively, are

$$\begin{aligned} R_{tx2}^{ts1}(t) &= \sum_{l=0}^{L-1}\sum_{k=1}^{K} h_{2,l}\, c_k(t - lT_c)\, s_{k,2}, \\ R_{tx2}^{ts2}(t) &= \sum_{l=0}^{L-1}\sum_{k=1}^{K} h_{2,l}\, c_k(t - lT_c)\, s_{k,1}^{*}. \end{aligned} \quad (6)$$

Using (5) and (6), the total signal received at the receive antenna in time slots ts1 and ts2, denoted by $R^{ts1}(t)$ and $R^{ts2}(t)$, respectively, are given by

$$R^{ts1}(t) = \alpha \sum_{l=0}^{L-1}\sum_{k=1}^{K}\Big(h_{1,l}s_{k,1} + h_{2,l}s_{k,2}\Big)c_k(t - lT_c) + w^{ts1}(t), \quad (7)$$

$$R^{ts2}(t) = \alpha \sum_{l=0}^{L-1}\sum_{k=1}^{K}\Big(-h_{1,l}s_{k,2}^{*} + h_{2,l}s_{k,1}^{*}\Big)c_k(t - lT_c) + w^{ts2}(t), \quad (8)$$

where α in the above is defined as

$$\alpha = \sqrt{\frac{\overline{\gamma}(1 - e^{-\beta})}{2(1 - e^{-\beta L})}}, \quad (9)$$

where $\overline{\gamma}$ is the average SNR at the receiver, and $w^{ts1}(t)$ and $w^{ts2}(t)$ are AWGN at time t for ts1 and ts2, respectively.

III. Detection of Multicode STBC Signals

The proposed receiver for the multicode STBC system consists of a matched filter (MF) bank and a code domain interference (CDI) canceller followed by STBC decoding, as shown in Fig. 2. The MF bank consists of L correlators for each spreading code on each receive antenna. So, there are LKN correlators, where N is the number of receive antennas. As mentioned earlier, we will consider $N = 1$. The output of the correlator on the rth path, $r = 0, 1, \cdots, L-1$, for k_0th data substream, $k_0 = 1, 2, \cdots, K$, in the time slot ts1, denoted by $Z_{k_0,r}^{ts1}$, is given by

$$Z_{k_0,r}^{ts1} = \int_{rT_c}^{T + rT_c} R^{ts1}(t)\, c_{k_0}^{*}(t - rT_c)\, dt, \quad r = 0, \cdots, L-1. \quad (10)$$

A similar equation can be written for time slot ts2 as well. The correlators can be thought of as composed of two operations. In the first operation, the received signal $R^{ts1}(t)$ is matched to the chip waveform, and the corresponding pth chip output, denoted by $R_{mf}^{ts1}(p)$, is given by

$$R_{mf}^{ts1}(p) = \int_{pT_c}^{(p+1)T_c} R^{ts1}(t)\, \psi(t - pT_c)\, dt, \quad p = 0, \cdots, P-1. \quad (11)$$

Using (11) and splitting the integration in (10) into integrands over one chip period, T_c, (10) can be rewritten as

$$Z_{k_0,r}^{ts1} = \sum_{p=r}^{p=P+r-1} R_{mf}^{ts1}(p)\, c_{k_0,(p-r)}^{*}, \quad r = 0, \cdots, L-1. \quad (12)$$

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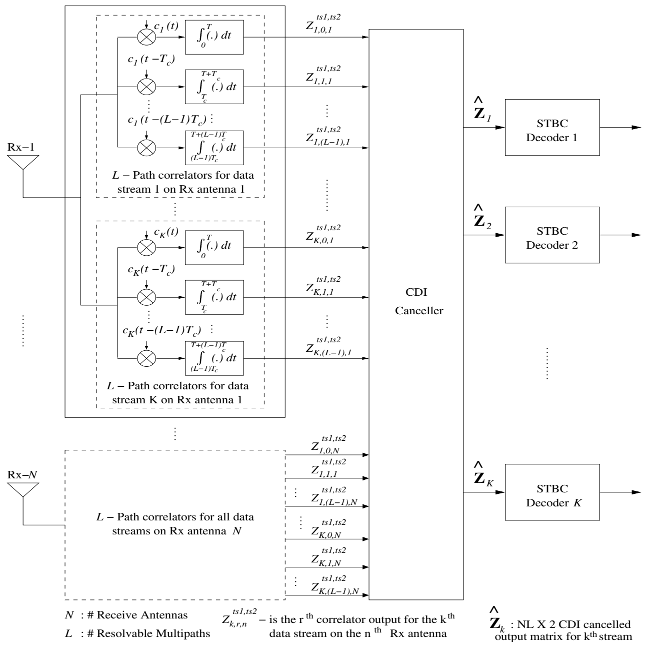

Fig. 2. Receiver architecture for multicode STBC.

Using (7) into (11), and normalizing by T_c, we get

$$R_{mf}^{ts1}(p) = \alpha \sum_{l=0}^{L-1} \sum_{k=1}^{K} \Big(h_{1,l} s_{k,1} \,+\, h_{2,l} s_{k,2} \Big) c_{k,(p-l)} \,+\, W_p^{ts1}, \quad (13)$$

$$W_p^{ts1} = (1/T_c) \int_{pT_c}^{(p+1)T_c} w^{ts1}(t)dt, \quad p = 0, \cdots, P-1, \quad (14)$$

where W_p^{ts1} is a complex Gaussian r.v. with zero mean and 0.5 variance per dimension, Using (13) into (12) and rearranging the summation order, (12) can be written as

$$Z_{k_0,r}^{ts1} \;=\; \sum_{k=1}^{K} \alpha \sum_{l=0}^{L-1} \Big(h_{1,l} s_{k,1} + h_{2,l} s_{k,2} \Big) C_{k_0,k,l}^{r} + W_{k_0,r}^{ts1}, \quad (15)$$

where

$$C_{k_0,k,l}^{r} \;=\; \sum_{p=r}^{p=P+r-1} c_{k_0,p-r}^{*}\, c_{k,p-l}, \quad (16)$$

$$W_{k_0,r}^{ts1} \;=\; \sum_{p=r}^{p=P+r-1} W_p^{ts1}\, c_{k_0,p-r}^{*}. \quad (17)$$

Equation (15) can be written in the following form[1]

$$Z_{k_0,r}^{ts1} \;=\; \alpha \sum_{k=1}^{K} \mathbf{h}_{k_0,k,r}\, \mathbf{s}_k^{ts1} \,+\, W_{k_0,r}^{ts1}, \quad (18)$$

where

$$\mathbf{s}_k^{ts1} \;=\; [s_{k,1} \;\; s_{k,2}]^T, \quad \mathbf{h}_{k_0,k,r} \;=\; \sum_{l=0}^{L-1} [h_{1,l} \;\; h_{2,l}]\, C_{k_0,k,l}^{r}. \quad (19)$$

We arrange the L values for $Z_{k_0,r}^{ts1}$ into a L-length column vector. We will have a similar L-length vector for ts2 as well.

Finally, we arrange these two L-length column vectors into a $L \times 2$ matrix $\mathbf{Z}_{k_0}$, whose first column is the L-length vector for ts1, and the second column is for ts2. Note that $\mathbf{h}_{k_0,k,r}$ is a row vector of length 2, and it is same for both ts1 and ts2. Since r takes all values from 0 to $L-1$, we can arrange all the L row vectors into a $L \times 2$ matrix $\mathbf{H}_{k_0,k}$ whose rth row is $\mathbf{h}_{k_0,k,r}$. Similarly, the noise variables $W_{k_0,r}^{ts1}$ can also be arranged into a L-length column vector for ts1 (same can be done for the ts2 noise variables as well). We can then form a $L \times 2$ noise matrix, $\mathbf{W}_{k_0}$. The $L \times 2$ sized $\mathbf{Z}_{k_0}$ matrix can then be written as

$$\mathbf{Z}_{k_0} \;=\; \sum_{k=1}^{K} \alpha \mathbf{H}_{k_0,k}\, \mathbf{S}_k + \mathbf{W}_{k_0}, \quad (20)$$

where

$$\mathbf{S}_k \;=\; \Big[\mathbf{s}_k^{ts1} \;\; \mathbf{s}_k^{ts2} \Big] \;=\; \begin{pmatrix} s_{k,1} & -s_{k,2}^{*} \\ s_{k,2} & s_{k,1}^{*} \end{pmatrix}. \quad (21)$$

For N receive antennas, we will have a $L \times 2$ sized $\mathbf{Z}_{k_0}$ matrix for each receive antenna. These N matrices can be stacked to form a $NL \times 2$ matrix $\mathbf{Z}_{k_0}^{N}$. Similarly, the $\mathbf{H}_{k_0,k}$ and $\mathbf{W}_{k_0}$ matrices for each receiver antenna could be stacked up to form $NL \times 2$ matrices, $\mathbf{H}_{k_0,k}^{N}$ and $\mathbf{W}_{k_0}^{N}$. Therefore, for N receive antennas, the received signal matrix will be given by

$$\mathbf{Z}_{k_0}^{N} \;=\; \sum_{k=1}^{K} \alpha \mathbf{H}_{k_0,k}^{N}\, \mathbf{S}_k + \mathbf{W}_{k_0}^{N}. \quad (22)$$

For $N = 1$, taking the desired signal term for the k_0th data substream out of the summation in (20), we have

$$\mathbf{Z}_{k_0} \;=\; \underbrace{\alpha\, \mathbf{H}_{k_0,k_0}\, \mathbf{S}_{k_0}}_{\text{desired signal}} + \underbrace{\sum_{k=1,k\neq k_0}^{K} \alpha \mathbf{H}_{k_0,k} \mathbf{S}_k}_{\text{CDI}} + \underbrace{\mathbf{W}_{k_0}}_{\text{noise}}. \quad (23)$$

STBC decoding can be carried out directly on the output matrix $\mathbf{Z}_{k_0}$ to detect the transmitted data symbols on the k_0th data stream . The performance in that case, however, will be degraded by the CDI term in (23). Improved performance can be achieved if the CDI components in $\mathbf{Z}_{k_0}$ can be estimated and cancelled, and STBC decoding is performed on the CDI cancelled output matrix $\widehat{\mathbf{Z}}_{k_0}$ as shown in Fig. 2.

A. CDI Estimation and Cancellation

In order to alleviate the effect of CDI on the receiver performance, we seek to estimate and cancel the CDI term (i.e., 2nd term) in (23). Towards that, consider the following operation

$$\sum_{k=1,k\neq k_0}^{K} \mathbf{H}_{k_0,k} \mathbf{H}_{k,k}^{\dagger} \mathbf{Z}_k, \quad (24)$$

which when expanded using (20) can be written as

$$\begin{aligned} \sum_{k=1,k\neq k_0}^{K} \mathbf{H}_{k_0,k} \mathbf{H}_{k,k}^{\dagger} \mathbf{Z}_k \;=\; & \sum_{k=1,k\neq k_0}^{K} \alpha\, \mathbf{H}_{k_0,k} \mathbf{S}_k \\ & + \sum_{k=1,k\neq k_0}^{K} \alpha\, \mathbf{H}_{k_0,k} \mathbf{H}_{k,k}^{\dagger} \mathbf{H}_{k,k_0} \mathbf{S}_{k_0} \\ & + \sum_{k=1,k\neq k_0}^{K} \alpha \left(\mathbf{H}_{k_0,k} \mathbf{H}_{k,k}^{\dagger} \left(\sum_{q=1,q\neq k,k_0}^{K} \mathbf{H}_{k,q} \mathbf{s}_q \right) \right) \\ & + \sum_{k=1,k\neq k_0}^{K} \mathbf{H}_{k_0,k} \mathbf{H}_{k,k}^{\dagger} \mathbf{W}_k. \quad (25) \end{aligned}$$

Note that the first term on the RHS of the above Eqn. (25) is equal to the CDI for k_0th data substream (i.e., same as the CDI term in Eqn. (23)). So, if we subtract $\sum_{k\neq k_0} \mathbf{H}_{k_0,k} \mathbf{H}_{k,k}^{\dagger} \mathbf{Z}_k$ (i.e., Eqn. (24)) from $\mathbf{Z}_{k_0}$ (i.e., Eqn. (23)), then the CDI term

[1] Vectors are denoted by boldface lowercase letters, and matrices are denoted by boldface uppercase letters. Superscripts T and H denote transpose and conjugate transpose operations, respectively. Superscript † denotes matrix pseudo-inverse operation.

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is (23) is completely removed. Accordingly, we propose the CDI cancellation operation as follows

$$\widehat{\mathbf{Z}}_{k_0} = \mathbf{Z}_{k_0} - \underbrace{\sum_{k\neq k_0} \mathbf{H}_{k_0,k}\mathbf{H}^{\dagger}_{k,k}\mathbf{Z}_k,}_{\text{CDI Estimate}} \tag{26}$$

where $\widehat{\mathbf{Z}}_{k_0}$ is the CDI cancelled output matrix for the k_0th stream. It is noted that, the CDI estimate $\sum_{k\neq k_0} \mathbf{H}_{k_0,k}\mathbf{H}^{\dagger}_{k,k}\mathbf{Z}_k$ has a desired signal component as well (in the 2nd term in Eqn. (25)). So, upon subtracting $\sum_{k\neq k_0} \mathbf{H}_{k_0,k}\mathbf{H}^{\dagger}_{k,k}\mathbf{Z}_k$, some part of the desired signal also gets subtracted, and we call it as desired signal leakage. Also, additional interference terms (terms other than the 1st term on the RHS of Eqn. (25)) are generated in the cancellation process. However, the cancellation benefit can outweigh the effects of relatively small desired signal leakage and additional interference terms (as we will see in the Sec. IV). Further, the CDI estimate in (24) can be scaled by a weight w and the scaled CDI estimate can be subtracted from $\mathbf{Z}_{k_0}$. The SINR at the cancelled output will then vary with w, and hence can be optimized w.r.t w.

B. SINR at the CDI Canceller Output

The received signal matrix, after weighted CDI cancellation, is given by

$$\widehat{\mathbf{Z}}_{k_0} = \mathbf{Z}_{k_0} - w\sum_{k=1,k\neq k_0}^{K} \mathbf{H}_{k_0,k}\mathbf{H}^{\dagger}_{k,k}\mathbf{Z}_k, \tag{27}$$

$$= \mathbf{T}_{S,k_0} + \underbrace{\mathbf{T}_{I,k_0} + \mathbf{T}_{N,k_0}}_{\triangleq\, \mathbf{G}_{k_0}}, \tag{28}$$

where $\mathbf{T}_{S,k_0}$, $\mathbf{T}_{I,k_0}$, and $\mathbf{T}_{N,k_0}$ are the signal, interference and noise terms, respectively, in $\widehat{\mathbf{Z}}_{k_0}$, which are given by

$$\mathbf{T}_{S,k_0} = \alpha \underbrace{\left(\mathbf{H}_{k_0,k_0} - w\sum_{k=1,k\neq k_0}^{K}\mathbf{H}_{k_0,k}\mathbf{H}^{\dagger}_{k,k}\mathbf{H}_{k,k_0}\right)}_{\triangleq\, \mathbf{F}_{k_0}}\mathbf{S}_{k_0}, \tag{29}$$

$$\mathbf{T}_{I,k_0} = \alpha(1-w)\sum_{k=1,k\neq k_0}^{K}\mathbf{H}_{k_0,k}\mathbf{S}_k - w\sum_{k=1,k\neq k_0}^{K}\left(\mathbf{H}_{k_0,k}\mathbf{H}^{\dagger}_{k,k}\left(\sum_{q=1,q\neq k,k_0}^{K}\alpha\mathbf{H}_{k,q}\mathbf{S}_q\right)\right), \tag{30}$$

$$\mathbf{T}_{N,k_0} = \mathbf{W}_{k_0} - w\sum_{k=1,k\neq k_0}^{K}\mathbf{H}_{k_0,k}\mathbf{H}^{\dagger}_{k,k}\mathbf{W}_k. \tag{31}$$

The interference, signal, and noise powers for the k_0th stream, denoted by $\mathcal{P}_{I,k_0}$, $\mathcal{P}_{S,k_0}$, and $\mathcal{P}_{N,k_0}$, respectively, are given by

$$\begin{aligned}\mathcal{P}_{I,k_0} = {} & 2\alpha^2(1-w)^2\sum_{k=1,k\neq k_0}^{K}\|\mathbf{H}_{k_0,k}\|_F^2 \\ & + 2\alpha^2 w^2 \sum_{\substack{k_1=1\\k_1\neq k_0}}^{K}\sum_{\substack{k_2=1\\k_2\neq k_0}}^{K}\sum_{\substack{q=1\\q\neq k_0,k_1,k_2}}^{K} trace\left(\mathbf{H}_{k_0,k_1}\mathbf{H}^{\dagger}_{k_1,k_1}\mathbf{H}_{k_1,q}\left(\mathbf{H}_{k_0,k_2}\mathbf{H}^{\dagger}_{k_2,k_2}\mathbf{H}_{k_2,q}\right)^H\right) \\ & - 4\alpha^2 w(1-w)\,\Re\left(\sum_{\substack{k=1\\k\neq k_0}}^{K}\sum_{\substack{q=1\\q\neq k_0,k}}^{K} trace\left(\mathbf{H}_{k_0,q}(\mathbf{H}_{k_0,k}\mathbf{H}^{\dagger}_{k,k}\mathbf{H}_{k,q})^H\right)\right).\end{aligned} \tag{32}$$

$$\mathcal{P}_{S,k_0} = 2\alpha^2\|\mathbf{H}_{k_0,k_0} - w\sum_{k=1,k\neq k_0}^{K}\mathbf{H}_{k_0,k}\mathbf{H}^{\dagger}_{k,k}\mathbf{H}_{k,k_0}\|_F^2, \tag{33}$$

$$\mathcal{P}_{N,k_0} = 2P\left(L + w^2\sum_{k=1,k\neq k_0}^{K}\|\mathbf{H}_{k_0,k}\mathbf{H}^{\dagger}_{k,k}\|_F^2\right). \tag{34}$$

The SINR of the k_0th data stream at the cancelled output is then given by

$$SINR_{k_0} = \frac{\mathcal{P}_{S,k_0}}{\mathcal{P}_{I,k_0} + \mathcal{P}_{N,k_0}}. \tag{35}$$

C. STBC Decoding

Noting that the $L\times 2$ CDI cancelled output matrix $\widehat{\mathbf{Z}}_{k_0}$ is of the form (from Eqns. (28) and (29))

$$\widehat{\mathbf{Z}}_{k_0} = \alpha\mathbf{F}_{k_0}\mathbf{S}_{k_0} + \mathbf{G}_{k_0}, \tag{36}$$

the STBC decoding on $\widehat{\mathbf{Z}}_{k_0}$ can be performed as follows. Let $s_{k_0,1} = x_{k_0,1} + jy_{k_0,1}$, $s_{k_0,2} = x_{k_0,2} + jy_{k_0,2}$, $j = \sqrt{-1}$,

$$\mathbf{A}_1 = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix}, \quad \mathbf{B}_1 = \begin{pmatrix} 1 & 0 \\ 0 & -1 \end{pmatrix},$$

$$\mathbf{A}_2 = \begin{pmatrix} 0 & -1 \\ 1 & 0 \end{pmatrix}, \quad \mathbf{B}_2 = \begin{pmatrix} 0 & 1 \\ 1 & 0 \end{pmatrix}.$$

Using the knowledge of $\mathbf{F}_{k_0}$, obtain the following

$$\tilde{x}_{k_0,1} = \Re Trace\left(\mathbf{A}_1^H\mathbf{F}_{k_0}^H\widehat{\mathbf{Z}}_{k_0}\right), \tag{37}$$

$$\tilde{y}_{k_0,1} = \Re Trace\left(-j\mathbf{B}_1^H\mathbf{F}_{k_0}^H\widehat{\mathbf{Z}}_{k_0}\right), \tag{38}$$

$$\tilde{x}_{k_0,2} = \Re Trace\left(\mathbf{A}_2^H\mathbf{F}_{k_0}^H\widehat{\mathbf{Z}}_{k_0}\right), \tag{39}$$

$$\tilde{y}_{k_0,2} = \Re Trace(-j\mathbf{B}_2^H\mathbf{F}_{k_0}^H\widehat{\mathbf{Z}}_{k_0}). \tag{40}$$

The decoded symbol for the transmitted symbol $s_{k_0,1}$ is then obtained as

$$\widehat{s}_{k_0,1} = \widehat{x}_{k_0,1} + j\,\widehat{y}_{k_0,1}, \tag{41}$$

where

$$\widehat{x}_{k_0,1} = \begin{cases} 1/\sqrt{2} & \text{if } \tilde{x}_{k_0,1} \geq 0 \\ -1/\sqrt{2} & \text{if } \tilde{x}_{k_0,1} < 0, \end{cases} \tag{42}$$

$$\widehat{y}_{k_0,1} = \begin{cases} 1/\sqrt{2} & \text{if } \tilde{y}_{k_0,1} \geq 0 \\ -1/\sqrt{2} & \text{if } \tilde{y}_{k_0,1} < 0. \end{cases} \tag{43}$$

The transmitted symbol $s_{k_0,2}$ is decoded in a similar manner.

IV. Results And Discussion

We evaluated the bit error performance of the proposed CDI cancelling receiver for the multicode STBC system considered. We point out that the overall error performance of this system is influenced by various factors including $i)$ transmit diversity benefit from the STBC code used (see the STBC decoding operation in Eqns. (37)-(40)), $ii)$ RAKE diversity benefit due to multipath ($L > 1$) for the desired data substream (see the $\mathbf{F}_{k_0}^H\widehat{\mathbf{Z}}_{k_0}$ operation in Eqns. (37)-(40), which essentially amounts to RAKE combining operation), $iii)$ degradation due to multipath (MP) induced CDI from other data substreams (see the CDI terms in Eqn. (23)), and $iv)$ benefit due to CDI cancellation (see the weighted CDI cancellation operation in Eqn. (27)). In the simulation results here, we focus on the interplay between the above factors.

In Fig. 3, we present the BER performance of a system with $K = 16$ substreams/codes, processing gain $P = 32$ chips per bit, number of multipaths $L = 1, 2, 3$, and $\beta = 0$ (i.e., L equal-energy paths). The figure shows plots for different

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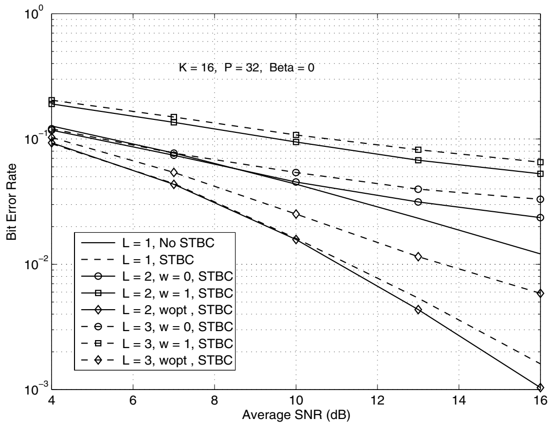

Fig. 3. Bit error performance of the proposed CDI cancelling receiver for multicode STBC in frequency selective fading. $K = 16$, $P = 32$, $\beta = 0$, Alamouti code, $M = 2$, $N = 1$.

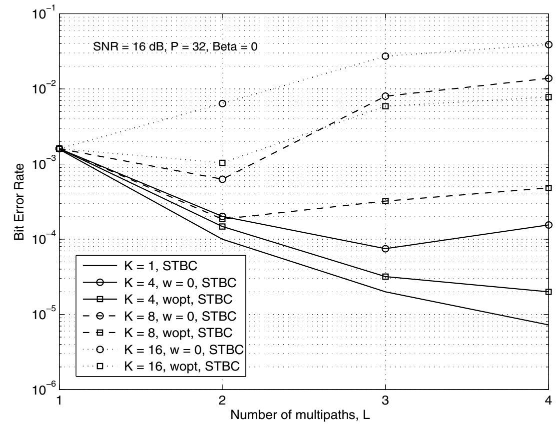

Fig. 4. BER performance of the proposed CDI cancelling receiver as a function of number of multipaths, L, for multicode STBC. $K = 1, 4, 8, 16$, $P = 32$, $\beta = 0$, SNR = 16 dB, Alamouti code, $M = 2$, $N = 1$.

weights used in the CDI canceller, i.e., for $w = 0, 1, w_{opt}$. Note that $w = 0$ corresponds to the case where no CDI cancellation is done and STBC decoding is directly done on the MF outputs. In the case of w_{opt}, these optimum weights were obtained by maximizing the SINR in Eqn. (35). The following interesting observations are made from Fig. 3:

- When $L = 1$, there is no MP induced CDI (and hence no performance loss due to CDI), but there is no RAKE diversity benefit as well. However, there is 2nd order Tx diversity benefit because of the Alamouti code used.
- When L is increased to 2, MP induced CDI occurs. If CDI cancellation is not done (i.e., $w = 0$), the performance degrades significantly, because the interference effect dominates the diversity effects of STBC and RAKE. Hence the high error floor for $w = 0, L = 2$ in Fig. 3. Also, cancelling the CDI with non-optimum weights can make things worse compared to no cancellation, as can be observed for the case of $w = 1, L = 2$. Such a behaviour of cancellation performing worse than no cancellation is known to occur in poor channel conditions (e.g., low SNR, high interference) [7]. However, when optimum weights w_{opt} are used, the cancellation becomes very effective, alleviating the interference effect and restoring the diversity benefits. In fact, we see that the performance for $w = w_{opt}, L = 2$ is even better than $L = 1$ (i.e., no CDI case). This is because in addition to the cancellation benefit, there is RAKE diversity benefit in $L = 2$ compared to $L = 1$.
- When L is further increased to 3, the performance behaviour gets interesting. First, the performance for $L = 3, w = w_{opt}$ is significantly better than no cancellation or non-optimum weight cancellation (i.e., $w = 0, 1$, $L = 3$). Second, even with CDI cancellation using w_{opt} weights, the performance for $L = 3$ is worse than that of $L = 1$ as well as $L = 2, w = w_{opt}$. This is because, for $L = 3$, the CDI becomes larger to an extent that the interference effect outweighs the RAKE diversity benefit achieved with $L = 3$. We point out that this situation can be improved by using more number of cancellation stages so that the STBC and RAKE diversity benefits can outweigh the interference effect.

Figure 4 shows BER as a function of number of multipaths, L, for $K = 1, 4, 8, 16$ and SNR=16 dB. The plots in Fig. 4 further reinforce the observations made in Fig. 3. For $K = 1$, in the absence of other stream interference, increasing L results in only RAKE diversity gain. For $K > 1$, increasing L beyond certain value degrades performance since other stream interference from multipaths dominates RAKE diversity gain. This degradation at large L is alleviated by the proposed CDI canceller using optimum weights. Further improvement can be achieved using multiple cancellation stages.

V. Conclusion

We proposed a parallel interference cancelling (PIC) detector for cancelling frequency selectivity induced code domain interference (CDI) in multicode STBC systems. We evaluated the performance of the proposed PIC detector in terms of BER, and highlighted the interplay between various factors that influence the performance; these factors included the degrading effect of CDI and the beneficial effects of STBC and RAKE diversities and CDI cancellation. The proposed detector was shown to effectively cancel the frequency selectivity induced CDI that resulted in improved error performance.

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