Multicode MIMO for High Data Rate Mobile Ad-hoc Networks

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***Abstract*—In a mobile ad-hoc network scenario, where communication nodes are mounted on moving platforms (like jeeps, trucks, tanks, etc.), use of V-BLAST requires that the number of receive antennas in a given node must be greater than or equal to the sum of the number of transmit antennas of all its neighbor nodes. This limits the achievable spatial multiplexing gain (data rate) for a given node. In such a scenario, we propose to achieve high data rates per node through multicode direct sequence spread spectrum techniques in conjunction with V-BLAST. In the considered multicode V-BLAST system, the receiver experiences code domain interference (CDI) in frequency selective fading, in addition to space domain interference (SDI) experienced in conventional V-BLAST systems. We propose two interference cancelling receivers that employ a linear parallel interference cancellation approach to handle the CDI, followed by conventional V-BLAST detector to handle the SDI, and then evaluate their bit error rates.**

Keywords – *Mobile ad-hoc networks, multicode V-BLAST, code domain interference, space domain interference, interference cancellation.*

I. Introduction

Recently, mobile ad-hoc networks have been receiving increased attention for both military as well as commercial applications. In a military application context, communication nodes mounted on moving platforms (like jeeps, trucks, tanks, etc. in battlefield environments) must be able to form among themselves, and operate in, dynamic ad-hoc network topologies. In addition, high data rate transmission on the wireless links and low media access control overhead are desired. In such scenarios, MIMO techniques [1],[2] can be employed to achieve high data rates, and we investigate a MIMO system that employs spatial multiplexing (V-BLAST) in conjunction with multicode direct sequence spread spectrum techniques. The motivation for this approach is as follows.

Use of V-BLAST in an ad-hoc network scenario requires that the number of receive antennas in a given node must be greater than or equal to the sum of the number of transmit antennas of all its neighbor nodes. Compared to a single user point-to-point VBLAST system, in multiuser V-BLAST ad-hoc networks, for a given number of receive antennas in a node, the maximum number of transmit antennas allowed gets divided among all the neighboring nodes. That is, the number of spatial degrees of freedom per node gets reduced. This limits the achievable spatial multiplexing gain (i.e., data rate) for a given node. In such a scenario, use of *multiplexing and reuse of multiple orthogonal codes* on each transmit antenna in each node can provide additional degrees of freedom in the code domain, which can be leveraged to achieve increased data rate capability at each node.

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In the considered multicode V-BLAST system, the receiver node experiences code domain interference (CDI) in frequency selective fading, in addition to space domain interference (SDI) experienced in conventional V-BLAST systems. A new contribution in this paper is that we develop novel interference cancelling receiver algorithms for the considered multicode V-BLAST system. We propose a linear parallel interference cancellation (LPIC) approach to handle the CDI, followed by the conventional V-BLAST detector [2][1] to handle the SDI. We propose two detectors, namely, *'CD-LPIC/SD-SIC Detector-I'* and *'CD-LPIC/SD-SIC Detector-II.'* In Detector-I, the CD-LPIC cancels the CDI by generating a vector estimate of the CDI, which involves a pseudo-inverse operation on a modified channel matrix. Because of this, the receiver complexity is high, but it results in very good bit error performance. In Detector-II, the CD-LPIC cancels the CDI by generating a scalar estimate of the interference using less receiver complexity. For Detector-II, we derive closed-form expressions for the optimum weights that maximize the output average signal-to-interference plus noise ratio (SINR).

II. System Model

We consider mobile ad-hoc networks where each node in the network is provided with multiple transmit and multiple receive antennas. Consider a given receive node-of-interest. Let N denote the number of receive antennas in the node-of-interest, I denote the number of neighbor nodes of the node-of-interest, and M_i denote the number of transmit antennas in the ith neighbor node, such that the condition

$$M \;\stackrel{\triangle}{=}\; \sum_{i=1}^{I} M_i \;\leq\; N \tag{1}$$

is satisfied. Let K denote the number of orthogonal codes multiplexed on each transmit antenna in each neighbor node.

The multicode V-BLAST transmitter of the ith neighbor node is shown in Fig. 1. The ith neighbor node's data stream is demultiplexed into KM_i parallel substreams. These substreams are partitioned into M_i groups. Each group consists of K substreams, which are spread using different waveforms $c_k(t)$, $k = 1, 2, \cdots, K$, and are transmitted from the same transmit antenna. These spreading waveforms $c_k(t)$, $k = 1, 2, \cdots, K$ are reused for the other substreams on the other transmit antennas of the ith neighbor node, as shown in Fig. 1. This results in KM_i degrees-of-freedom in the ith neighbor node's transmission. The same spreading waveforms $c_k(t)$, $k = 1, 2, \cdots, K$, are reused in all transmit antennas of the other neighbor nodes as well.

[1] We refer to the conventional V-BLAST detector in [2] as a space domain successive interference cancelling (SD-SIC) detector.

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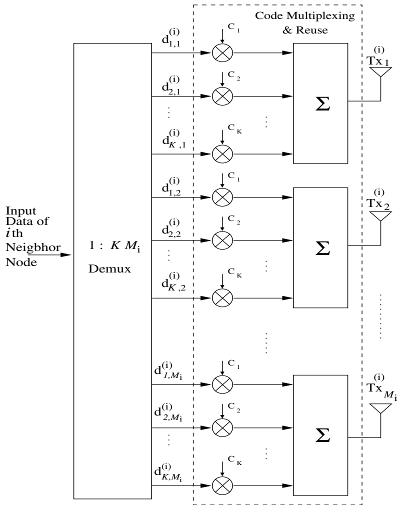

Fig. 1. Multicode V-BLAST transmitter of the ith neighbor node.

Let T_s denote one data symbol duration. The complex baseband transmitted signal from the mth transmit antenna of the ith neighbor node over one symbol interval is given by

$$s_m^{(i)}(t) = \sum_{k=1}^{K} d_{k,m}^{(i)}\, c_k(t), \;\; m = 1, 2 \cdots, M_i, \;\; 0 \leq t \leq T_s, \quad (2)$$

where $d_{k,m}^{(i)}$ denotes the data symbol on the kth code, of the mth transmit antenna, of the ith neighbor node. We assume QPSK modulation. Hence, $d_{k,m}^{(i)}$ takes one of four complex values in $\left\{(\pm 1 \pm j)/\sqrt{2}\right\}$ with equal probability, and $c_k(t)$ is the spreading waveform for the kth code, given by

$$c_k(t) \;=\; \sum_{p=0}^{P-1} c_{k,p}\, \psi(t - pT_c), \quad k = 1, 2, \cdots, K, \quad (3)$$

where T_c is the chip duration, $c_{k,p}$ is the pth chip of the kth 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., unity 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 $c_{k,p} = c_{k,p}^{(Real)} + jc_{k,p}^{(Imag)}$, where $c_{k,p}^{(Real)}$ and $c_{k,p}^{(Imag)}$ take the random values of $+1/\sqrt{2}$ and $-1/\sqrt{2}$ with equal probability. Moreover, the $c_k(t)$ are mutually orthogonal for all k, i.e., $\sum_{p=0}^{P-1} c_{k_1,p}\, c_{k_2,p}^{*} = 0, \;$ for $k_1 \neq k_2$, where $(\cdot)^*$ denotes the complex conjugate operation.

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

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

where L is the number of resolvable multipath components, $h_{n,m,l}^{(i)}$ is the complex fading coefficient from the mth transmit antenna of the ith neighbor node to the nth receive antenna on the lth multipath, and the $\{h_{n,m,l}^{(i)}\}$'s are assumed to be circularly symmetric complex Gaussian random variables with zero mean. It is assumed that the $\{h_{n,m,l}^{(i)}\}$'s are constant over one symbol duration, and are independent for all $n, m, l,$ and i. The second moment of $|h_{n,m,l}^{(i)}|$, Ω_l, is assumed to have an exponential multipath intensity profile, given by

$$\Omega_l = E\left[|h_{n,m,l}^{(i)}|^2\right] = \Omega_0 e^{-l\beta}, \quad l = 0, 1, \cdots, L-1, \quad (5)$$

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: Assuming that all the neighbor nodes' transmissions arrive synchronously at the receiver, the received complex baseband equivalent signal at the nth receive antenna, $n = 1, 2, \cdots, N$, can be expressed as

$$\begin{aligned} r_n(t) \;&=\; \sum_{i=1}^{I} \sum_{m=1}^{M_i} \sum_{l=0}^{L-1} h_{n,m,l}^{(i)}\, s_m^{(i)}(t - lT_c) + w_n(t) \\ &= \sum_{i=1}^{I} \sum_{m=1}^{M_i} \sum_{l=0}^{L-1} \sum_{k=1}^{K} h_{n,m,l}^{(i)}\, d_{k,m}^{(i)}\, c_k(t - lT_c) + w_n(t), \quad (6) \end{aligned}$$

where $w_n(t)$ is the AWGN at the nth receive antenna with one-sided power spectral density σ^2. It is assumed that the channel fade coefficients are perfectly estimated and known at the receiver. We also assume perfect timing at the receiver.

The matched filter (MF) bank for the k_0th spreading code is shown in Fig. 2. The received signal at the nth receive antenna, $r_n(t)$, is passed through L correlators (lth correlator synchronized to lth multipath), each correlated with the complex conjugate of the k_0th spreading waveform, $c_{k_0}(t)$. The same is true on all the receive antennas, as shown in Fig. 2. The output of the lth correlator, at the nth receive antenna, for the k_0th spreading code, $k_0 = 1, 2, \cdots, K$; $n = 1, 2, \cdots, N$; $l = 0, 1, \cdots, L-1$, can be written as

$$\begin{aligned} z_{k_0,n,l} \;&=\; \int_{lT_c}^{T_s + lT_c} r_n(t)\, c_{k_0}^{*}(t - lT_c)\, dt \\ &= \sum_{i=1}^{I} \sum_{m=1}^{M_i} \sum_{q=0}^{L-1} \sum_{k=1}^{K} h_{n,m,q}^{(i)}\, d_{k,m}^{(i)}\, R_{k_0,k}(q - l) + n_{k_0,n,l}, \quad (7) \end{aligned}$$

where $\quad R_{k_0,k}(q-l) \triangleq \int_{-\infty}^{\infty} c_{k_0}^{*}(t)\, c_k(t - (q-l)T_c)\, dt, \quad (8)$

$$n_{k_0,n,l} \triangleq \int_{lT_c}^{T_s + lT_c} w_n(t)\, c_{k_0}^{*}(t - lT_c)\, dt. \quad (9)$$

Considering all the correlator outputs from all the N receive antennas, the $NL \times 1$ MF output vector for the k_0th spreading code, $\mathbf{Z}_{k_0}$, can be written as

$$\begin{aligned} \mathbf{Z}_{k_0} \;&=\; \Big[z_{k_0,1,0}, z_{k_0,1,1}, \cdots, z_{k_0,1,L-1}, z_{k_0,2,0}, \cdots, \\ &\quad z_{k_0,N,L-2}, z_{k_0,N,L-1}\Big]^T. \quad (10) \end{aligned}$$

By rearranging, $\mathbf{Z}_{k_0}$ can be written in the form

$$\mathbf{Z}_{k_0} \;=\; \mathbf{Y}_{k_0,k_0}\mathbf{d}_{k_0} + \sum_{k=1, k \neq k_0}^{K} \mathbf{Y}_{k_0,k}\mathbf{d}_k + \mathbf{n}_{k_0}, \quad (11)$$

where $\mathbf{d}_{k_0}$ is the $M \times 1$ input data vector given by

$$\mathbf{d}_{k_0} = \left[d_{k_0,1}^{(1)}, d_{k_0,2}^{(1)}, \cdots, d_{k_0,M_1}^{(1)}, \cdots, d_{k_0,1}^{(I)}, \cdots, d_{k_0,M_I}^{(I)}\right]^T, \quad (12)$$

$\mathbf{n}_{k_0}$ is the $NL \times 1$ noise vector given by

$$\begin{aligned} \mathbf{n}_{k_0} \;&=\; \Big[n_{k_0,1,0}, n_{k_0,1,1}, \cdots, n_{k_0,1,L-1}, n_{k_0,2,0}, \cdots, \\ &\quad n_{k_0,N,L-2}, n_{k_0,N,L-1}\Big]^T, \quad (13) \end{aligned}$$

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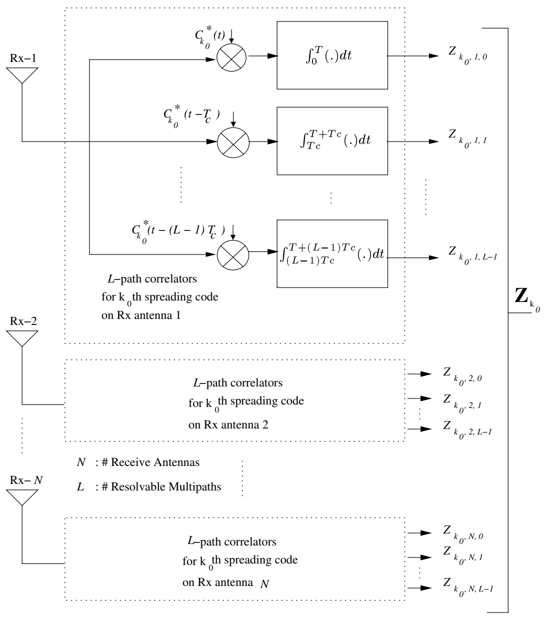

Fig. 2. Matched filter bank for k_0th spreading code.

$\mathbf{Y}_{k_0,k}$ is an $NL \times M$ matrix given by

$$\mathbf{Y}_{k_0,k} \;=\; \left[\mathbf{Y}^T_{k_0,k,1}, \mathbf{Y}^T_{k_0,k,2}, \cdots, \mathbf{Y}^T_{k_0,k,N}\right]^T, \tag{14}$$

and $\mathbf{Y}_{k_0,k,n}$ is an $L \times M$ matrix given by (15) at the bottom of this page [2]. Note that the first term on the RHS of (11) is composed of the desired signal components, multipath interference (MPI), and SDI for the k_0th code. On the other hand, the second term on the RHS of (11) is the CDI for the k_0th code (i.e., interference from other codes to the k_0th code).

We note that detectors in the literature, namely, i) the RAKE receiver for multicode V-BLAST analyzed in [3], which does not do any interference cancellation, ii) the SD-SIC detector in [2], which cancels only SDI and treats CDI as merely noise, and iii) the CD/SD SIC detector in [4], which uses successive cancellation techniques to cancel CDI and SDI, can be used to detect the multicode V-BLAST signals in the above.

III. Proposed Detectors for Multicode V-BLAST

Here, we propose two interference cancelling detectors for the multicode V-BLAST system described above. The proposed detectors employ an LPIC approach for CDI cancellation, followed by the SD-SIC detector for SDI cancellation.

A. Proposed CD-LPIC/SD-SIC Detector-I

In the CD-LPIC/SD-SIC Detector-I, the CD-LPIC cancels the CDI, leaving the MPI and the SDI to be handled by the SD-SIC. The MF bank output vectors, $\mathbf{Z}_{k_0}$, $k_0 = 1, 2, \cdots, K$, are fed as the input to the CD-LPIC. The corresponding CDI cancelled output vectors, $\tilde{\mathbf{Z}}_{k_0}$, $k_0 = 1, 2, \cdots, K$, are fed to the SD-SICs.

2

As pointed out earlier, the 2nd term in (11) is the CDI for the k_0th code, which we would like to cancel using the CD-LPIC. Towards that end, consider the following operation on the MF output vectors, for all codes $\mathbf{Z}_k, k = 1, 2, \cdots, K$:

$$\widehat{CDI}_{k_0} \;=\; \sum_{k=1,k\neq k_0}^{K} \mathbf{Y}_{k_0,k}\mathbf{Y}^{\dagger}_{k,k}\mathbf{Z}_k, \tag{16}$$

where $[.]^{\dagger}$ denotes the pseudo-inverse operation. Using (11) in the above and the fact that $\mathbf{Y}^{\dagger}_{k,k}\mathbf{Y}_{k,k} = \mathbf{I}$, we can write

$$\widehat{CDI}_{k_0} \;=\; \sum_{k=1,k\neq k_0}^{K} \mathbf{Y}_{k_0,k}\mathbf{Y}^{\dagger}_{k,k}\left(\mathbf{Y}_{k,k}\mathbf{d}_k + \sum_{s=1,s\neq k}^{K}\mathbf{Y}_{k,s}\mathbf{d}_s + \mathbf{n}_k\right)$$

$$= \underbrace{\sum_{k=1,k\neq k_0}^{K}\mathbf{Y}_{k_0,k}\mathbf{d}_k}_{T_1} + \underbrace{\sum_{k=1,k\neq k_0}^{K}\mathbf{Y}_{k_0,k}\mathbf{Y}^{\dagger}_{k,k}\sum_{s=1,s\neq k}^{K}\mathbf{Y}_{k,s}\mathbf{d}_s}_{T_2} + \underbrace{\sum_{k=1,k\neq k_0}^{K}\mathbf{Y}_{k_0,k}\mathbf{Y}^{\dagger}_{k,k}\mathbf{n}_k}_{T_3}. \tag{17}$$

Note that the 1st term, T_1, in (17) is the same as the CDI term (i.e., the 2nd term) in (11) which we want to cancel. Hence, we can view (17) as an $NL \times 1$ vector estimate of the CDI for the k_0th code (i.e., estimate of the interference from other codes to code k_0). This CDI estimate is imperfect in the sense that, although it exactly recreates the CDI term in (11), it also generates additional terms (T_2 and T_3 in (17)). Further, the 2nd term, T_2, in (17) can be written as

$$T_2 \;=\; \underbrace{\sum_{k=1,k\neq k_0}^{K}\mathbf{Y}_{k_0,k}\mathbf{Y}^{\dagger}_{k,k}\mathbf{Y}_{k,k_0}\mathbf{d}_{k_0}}_{\text{desired signal leak}} + \sum_{k=1,k\neq k_0}^{K}\mathbf{Y}_{k_0,k}\mathbf{Y}^{\dagger}_{k,k}\sum_{s=1,s\neq k,k_0}^{K}\mathbf{Y}_{k,s}\mathbf{d}_s.$$

Note that the 1st term in the above equation is essentially the desired k_0th code signal component that leaked into the CDI estimate. Consequently, if the CDI estimate in (17) is subtracted from $\mathbf{Z}_{k_0}$, then some amount of the desired signal is also removed in the cancellation process. We propose to use the CDI estimate in (17) for cancellation. We will see later that, because both these additional terms and the desired signal leak term are not dominant, even using this imperfect CDI estimate results in an effective cancellation and thus performance improvement. We obtain the CDI cancelled output vector for k_0th code, $\tilde{\mathbf{Z}}_{k_0}$, as

$$\tilde{\mathbf{Z}}_{k_0} \;=\; \mathbf{Z}_{k_0} - \widehat{CDI}_{k_0} \;=\; \mathbf{Z}_{k_0} - \sum_{k=1,k\neq k_0}^{K}\mathbf{Y}_{k_0,k}\mathbf{Y}^{\dagger}_{k,k}\mathbf{Z}_k$$

$$= \left[\mathbf{Y}_{k_0,k_0} - \sum_{k=1,k\neq k_0}^{K}\mathbf{Y}_{k_0,k}\mathbf{Y}^{\dagger}_{k,k}\mathbf{Y}_{k,k_0}\right]\mathbf{d}_{k_0} + \mathbf{I}_{k_0} + \mathbf{u}_{k_0}, \tag{18}$$

where

$$\mathbf{I}_{k_0} \;=\; -\sum_{k=1}^{K}\mathbf{Y}_{k_0,k}\mathbf{Y}^{\dagger}_{k,k}\left(\sum_{s=1,s\neq k_0,k}^{K}\mathbf{Y}_{k,s}\mathbf{d}_s\right), \tag{19}$$

$$\mathbf{u}_{k_0} \;=\; \mathbf{n}_{k_0} - \sum_{k=1,k\neq k_0}^{K}\mathbf{Y}_{k_0,k}\mathbf{Y}^{\dagger}_{k,k}\mathbf{n}_k. \tag{20}$$

This cancellation operation is done for all codes in parallel. The CDI cancelled outputs $\tilde{\mathbf{Z}}_{k_0}, k_0 = 1, 2, \cdots, K$, are fed to SD-SIC blocks which cancel the SDI and estimate the data substreams.

$$\mathbf{Y}_{k_0,k,n} = \sum_{l=0}^{L-1}\begin{pmatrix} h^{(1)}_{n,1,l}R_{k_0,k}(l) & h^{(1)}_{n,2,l}R_{k_0,k}(l) & \cdots & h^{(1)}_{n,M_1,l}R_{k_0,k}(l) & h^{(2)}_{n,1,l}R_{k_0,k}(l) & \cdots & h^{(I)}_{n,M_I,l}R_{k_0,k}(l) \\ h^{(1)}_{n,1,l}R_{k_0,k}(l-1) & h^{(1)}_{n,2,l}R_{k_0,k}(l-1) & \cdots & h^{(1)}_{n,M_1,l}R_{k_0,k}(l-1) & h^{(2)}_{n,1,l}R_{k_0,k}(l-1) & \cdots & h^{(I)}_{n,M_I,l}R_{k_0,k}(l-1) \\ \vdots & \vdots & \ddots & \vdots & & & \\ h^{(1)}_{n,1,l}R_{k_0,k}(l-L+1) & h^{(1)}_{n,2,l}R_{k_0,k}(l-L+1) & \cdots & h^{(1)}_{n,M_1,l}R_{k_0,k}(l-L+1) & h^{(2)}_{n,1,l}R_{k_0,k}(l-L+1) & \cdots & h^{(I)}_{n,M_I,l}R_{k_0,k}(l-L+1) \end{pmatrix}. \tag{15}$$

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B. Proposed CD-LPIC/SD-SIC Detector-II

It is noted that the CDI estimate in Detector-I is a vector estimate, which requires a pseudo inverse operation of the $\mathbf{Y}_{k,k}$ matrix. A less complex detector can be realized by obtaining a scalar estimate of the CDI as follows. The MF output on the lth path, at the nth receive antenna for the k_0th code, $z_{k_0,n,l}$, given by (7), can be written as

$$\begin{aligned} z_{k_0,n,l} &= \sum_{i=1}^{I}\sum_{m=1}^{M_i}\sum_{q=0}^{L-1}\sum_{k=1}^{K} h_{n,m,q}^{(i)} d_{k,m}^{(i)} R_{k_0,k}(q-l) + n_{k_0,n,l} \\ &= \sum_{i=1}^{I}\sum_{m=1}^{M_i} h_{n,m,l}^{(i)} d_{k_0,m}^{(i)} \\ &+ \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1}\sum_{i=1}^{I}\sum_{m=1}^{M_i} h_{n,m,q}^{(i)} d_{k,m}^{(i)} R_{k_0,k}(q-l) + n_{k_0,n,l}. \end{aligned} \quad (21)$$

Note that the 2nd term in (21) is composed of CDI and MPI. We propose to estimate and cancel this 2nd term from $z_{k_0,n,l}$. Towards that end, consider the following operation:

$$\mathbf{I}_{k_0,n,l}^{(est)} = \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1} z_{k,n,q} R_{k_0,k}(q-l). \quad (22)$$

Using (21) in the above, we can write

$$\begin{aligned} \mathbf{I}_{k_0,n,l}^{(est)} &= \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1} \Bigg(\sum_{i=1}^{I}\sum_{m=1}^{M_i} h_{n,m,q}^{(i)} d_{k,m}^{(i)} \\ &+ \sum_{s=1}^{K}\sum_{r=0,r\neq q}^{L-1}\sum_{i=1}^{I}\sum_{m=1}^{M_i} h_{n,m,r}^{(i)} d_{s,m}^{(i)} R_{k,s}(r-q) + n_{k,n,q} \Bigg) R_{k_0,k}(q-l) \\ &= \underbrace{\sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1}\sum_{i=1}^{I}\sum_{m=1}^{M_i} h_{n,m,q}^{(i)} d_{k,m}^{(i)} R_{k_0,k}(q-l)}_{T_4} + T_5 + T_6, \end{aligned} \quad (23)$$

where

$$T_5 = \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1} \left(\sum_{s=1}^{K}\sum_{r=0,r\neq q}^{L-1}\sum_{i=1}^{I}\sum_{m=1}^{M_i} h_{n,m,r}^{(i)} d_{s,m}^{(i)} R_{k,s}(r-q) \right) R_{k_0,k}(q-l),$$

$$T_6 = \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1} n_{k,n,q} R_{k_0,k}(q-l).$$

Note that the 1st term, T_4, in (23) is the same as the scalar interference term (i.e., the 2nd term) in (21) which we want to cancel. Hence, we can view (23) as a scalar estimate of the CDI and MPI for the MF bank output of the lth path, at the nth receive antenna for the k_0th code. We propose to subtract this interference estimate $\mathbf{I}_{k_0,n,l}^{(est)}$ scaled by a scalar weight, $w_l, l = 0, 1, \cdots, L-1$, to obtain the interference cancelled output on the lth path, at the nth receive antenna, for the k_0th code, $\bar{z}_{k_0,n,l}$, as follows:

$$\begin{aligned} \bar{z}_{k_0,n,l} &= z_{k_0,n,l} - w_l \mathbf{I}_{k_0,n,l}^{(est)} \\ &= \sum_{i=1}^{I}\sum_{m=1}^{M_i} h_{n,m,l}^{(i)} d_{k_0,m}^{(i)} \left(1 - w_l \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1} R_{k_0,k}(q-l) R_{k,k_0}(l-q) \right) \\ &+ J_{k_0,n,l} + V_{k_0,n,l}, \end{aligned} \quad (24)$$

where

$$\begin{aligned} J_{k_0,n,l} &= \sum_{i=1}^{I}\sum_{m=1}^{M_i}\sum_{q=0,q\neq l}^{L-1} h_{n,m,q}^{(i)} \sum_{k=1}^{K} d_{k,m}^{(i)} \Bigg(R_{k_0,k}(q-l)(1-w_l) \\ &- w_l \sum_{s=1}^{K}\sum_{r=0,r\neq l,q}^{L-1} R_{k_0,s}(r-l) R_{s,k}(q-r) \Bigg) \\ &- w_l \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1}\sum_{s=1,s\neq k_0}^{K}\sum_{i=1}^{I}\sum_{m=1}^{M_i} h_{n,m,l}^{(i)} d_{s,m}^{(i)} R_{k_0,k}(q-l) R_{k,s}(l-q), \end{aligned} \quad (25)$$

$$V_{k_0,n,l} = n_{k_0,n,l} - w_l \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1} n_{k,n,q} R_{k_0,k}(q-l). \quad (26)$$

This cancellation operation is done for all $z_{k_0,n,l}$ for $k_0 = 1, 2, \cdots, K$, $n = 1, 2, \cdots, N$, and $l = 0, 1, \cdots, L-1$ in parallel. The NL-dimensional interference cancelled output vector for k_0th code, $\overline{\mathbf{Z}}_{k_0}$, can then be written as

$$\begin{aligned} \overline{\mathbf{Z}}_{k_0} &= \Big[\bar{z}_{k_0,1,0}, \bar{z}_{k_0,1,1}, \cdots, \bar{z}_{k_0,1,L-1}, \bar{z}_{k_0,2,0}, \cdots, \\ &\quad \bar{z}_{k_0,N,L-2}, \bar{z}_{k_0,N,L-1} \Big]^T \quad (27) \\ &= \overline{\mathbf{Y}} \mathbf{d}_{k_0} + \mathbf{J}_{k_0} + \mathbf{V}_{k_0}, \quad (28) \end{aligned}$$

where

$$\begin{aligned} \mathbf{J}_{k_0} &= \Big[J_{k_0,1,0}, J_{k_0,1,1}, \cdots, J_{k_0,1,L-1}, J_{k_0,2,0}, \cdots, \\ &\quad J_{k_0,N,L-2}, J_{k_0,N,L-1} \Big]^T, \quad (29) \end{aligned}$$

$$\begin{aligned} \mathbf{V}_{k_0} &= \Big[V_{k_0,1,0}, V_{k_0,1,1}, \cdots, V_{k_0,1,L-1}, V_{k_0,2,0}, \cdots, \\ &\quad V_{k_0,N,L-2}, V_{k_0,N,L-1} \Big]^T, \quad (30) \end{aligned}$$

$$\overline{\mathbf{Y}} = \left[\overline{\mathbf{Y}}_1^T, \overline{\mathbf{Y}}_2^T, \cdots, \overline{\mathbf{Y}}_N^T \right]^T, \quad (31)$$

$\overline{\mathbf{Y}}_n$ is given by Eqn. (41) at the bottom of the next page, and

$$\gamma_l = \left(1 - w_l \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1} R_{k_0,k}(q-l) R_{k,k_0}(l-q) \right). \quad (32)$$

The interference cancelled outputs $\overline{\mathbf{Z}}_{k_0}, k_0 = 1, 2, \cdots, K$, are fed to SD-SIC blocks which cancel the SDI and estimate the data substreams. Note that the conventional SD-SIC detector in [2] can be viewed as a special case of Detector-II for $w_l = 0, \forall l$. In the following subsection, we obtain the optimum weights for Detector-II, w_l^{opt}, by maximizing the average SINR at the output of the CD-LPIC.

C. Derivation of Optimum weights for Detector-II

The desired signal power at the CD-LPIC output on the lth path received at the nth receive antenna for the k_0th code , $S_{k_0,n,l}$, is given by

$$S_{k_0,n,l} = M\Omega_l(1 - w_l a)^2, \quad (33)$$

where

$$a = \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1} R_{k_0,k}(q-l) R_{k,k_0}(l-q), \quad (34)$$

and Ω_l is defined in (5). The interference power, $\sigma_{J_{k_0,n,l}}^2$, and the noise power, $\sigma_{V_{k_0,n,l}}^2$, at the CD-LPIC output on lth path, at the nth receive antenna for the k_0th code, are given by

$$\begin{aligned} \sigma_{J_{k_0,n,l}}^2 &= M \sum_{q=0,q\neq l}^{L-1} \Omega_q \sum_{k=1}^{K} \left(\left| R_{k_0,k}(q-l)(1-w_l) - w_l b \right|^2 \right) \\ &+ c\, w_l^2 \Omega_l, \end{aligned} \quad (35)$$

where

$$b = \sum_{s=1}^{K}\sum_{r=0,r\neq q,l}^{L-1} R_{k_0,s}(r-l) R_{s,k}(q-r), \quad (36)$$

$$c = M \sum_{s=1,s\neq k_0}^{K} \left| \sum_{k=1}^{K}\sum_{q=0,q\neq l}^{L-1} R_{k_0,k}(q-l) R_{k,s}(l-q) \right|^2, \quad (37)$$

and

$$\sigma_{V_{k_0,n,l}}^2 = \sigma^2(1 + d w_l^2 - 2a w_l), \quad (38)$$

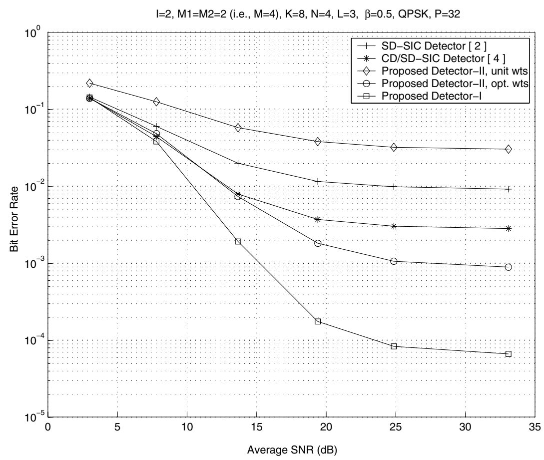

Fig. 3. BER performance of the proposed CD-LPIC/SD-SIC Detectors I and II for multicode V-BLAST. $I = 2$, $M_1 = M_2 = 2$, $K = 8$, $N = 4$, $L = 3$, $\beta = 0.5$, QPSK. $P = 32$.

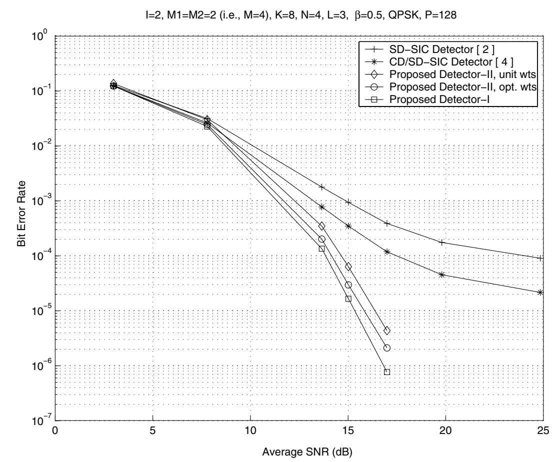

Fig. 4. BER performance of the proposed CD-LPIC/SD-SIC Detectors I and II for multicode V-BLAST. $I = 2$, $M_1 = M_2 = 2$, $K = 8$, $N = 4$, $L = 3$, $\beta = 0.5$, QPSK. $P = 128$.

where

$$d = \sum_{k=1}^{K} \sum_{q=0, q\neq l}^{L-1} \sum_{\bar{k}=1}^{K} \sum_{\bar{q}=0,\bar{q}\neq l}^{L-1} R_{k_0,k}(q-l) R_{k,\bar{k}}(\bar{q}-q) R^*_{k_0,\bar{k}}(\bar{q}-l). \quad (39)$$

The average SINR at the CD-LPIC output on lth path, at the nth receive antenna, for the k_0th code, is then given by

$$\overline{\text{SINR}}_{k_0,n,l} = \frac{S_{k_0,n,l}}{\sigma^2_{J_{k_0,n,l}} + \sigma^2_{V_{k_0,n,l}}}. \quad (40)$$

The optimum weight for the lth path, w_l^{opt}, is the one that maximizes the above SINR, which can be obtained, in closed-form, by differentiating (40) w.r.t w_l and equating the result to zero.

IV. Results and Discussion

In this section, we present the BER performance of the proposed detectors for multicode V-BLAST. In Figs. 3 and 4, we plot the BER performance of various detectors as a function of SNR. The various detectors include a) SD-SIC detector in [2] (i.e., conventional V-BLAST detector), b) CD/SD-SIC detector in [4], c) 'CD-LPIC/SD-SIC Detector-I,' d) 'CD-LPIC/SD-SIC Detector-II' with unit weights, and e) 'CD-LPIC/SD-SIC Detector-II' with optimum weights. We used Walsh codes multiplied with a common complex random binary sequence as the spreading codes. We considered $I = 2$ (2 neighbor nodes), $M_1 = M_2 = 2$ (2 transmit antennas in each neighbor node, i.e., $M = 4$), $K = 8$ (8 spreading codes on each transmit antenna), $N = 4$ (4 receive antennas), $L = 3$ (3 paths), and $\beta = 0.5$ (rate of exponential decay of average path power $= 0.5$). For the above system parameters, Fig. 3 shows the BER performance for a processing gain of $P = 32$ and Fig. 4 shows the BER performance for $P = 128$.

From Figs. 3 and 4, it can be seen that the SD-SIC detector performs poorly. This is expected, because the SD-SIC cancels only the SDI and ignores the CDI, which hurts performance. The CD/SD-SIC detector in [4] performs better than SD-SIC, due to the successive cancellation of the CDI. This observation corroborates with the results in [4].

The performance of Detector-I is much superior compared to both SD-SIC and CD/SD-SIC. Also, Detector-I performs better than Detector-II with optimum weights. This is because the CDI estimate of Detector-I in Eqn. (17) has a larger dimension (i.e., $NL \times 1$), which results in only a small desired signal leakage in the CDI estimate. The CDI estimate of Detector-II with optimum weights, on the other hand, has a smaller dimension (i.e., a scalar estimate) which results in a larger desired signal leakage in the CDI estimate. Because of this, Detector-II performs poorer than Detector-I, but has less complexity compared to Detector-I. As seen in Fig. 3, when the CDI is large (e.g., low processing gain P for a given number of codes K), Detector-II with unit weights can perform worse than the CD/SD-SIC and the SD-SIC detectors, due to an inaccurate estimate of the CDI. However, when optimum weights are used, Detector-II clearly outperforms both CD/SD-SIC and SD-SIC.

V. Conclusions

We investigated a multicode V-BLAST system for achieving high data rates in mobile ad-hoc networks, where communication nodes are mounted on moving platforms. High data rates are realized through degrees-of-freedom in the code domain, in addition to the degrees-of-freedom in the space domain (V-BLAST). In order to handle the CDI arising in frequency selective fading, we proposed two interference cancelling receivers that adopted an LPIC approach.

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$$\overline{\mathbf{Y}}_n = \begin{pmatrix} \gamma_0 h^{(1)}_{n,1,0} & \gamma_0 h^{(1)}_{n,2,0} & \cdots & \gamma_0 h^{(1)}_{n,M_1,0} & \gamma_0 h^{(2)}_{n,1,0} & \cdots & \gamma_0 h^{(I)}_{n,M_I,0} \\ \gamma_1 h^{(1)}_{n,1,1} & \gamma_1 h^{(1)}_{n,2,1} & \cdots & \gamma_1 h^{(1)}_{n,M_1,1} & \gamma_1 h^{(2)}_{n,1,1} & \cdots & \gamma_1 h^{(I)}_{n,M_I,1} \\ \vdots & \vdots & \ddots & \vdots & & & \\ \gamma_{L-1} h^{(1)}_{n,1,L-1} & \gamma_{L-1} h^{(1)}_{n,2,L-1} & \cdots & \gamma_{L-1} h^{(1)}_{n,M_1,L-1} & \gamma_{L-1} h^{(2)}_{n,1,L-1} & \cdots & \gamma_{L-1} h^{(I)}_{n,M_I,L-1} \end{pmatrix}. \quad (41)$$

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