On The Performance of Space-Time Coded Multiuser MIMO Systems with Iterative Receivers

Pei Xiao[†], Jinsong Wu^{††}, Mathini Sellathurai[†], T. Ratnarajah[†]

[†] Queen's University Belfast, BT3 9DT, United Kingdom

E-mail: {pei.xiao, m.sellathurai, t.ratnarajah}@ecit.qub.ac.uk

^{††} Queen's University, Kingston, Ontario, K7L 3N6, Canada

E-mail: jwu@ee.queensu.ca

Abstract—This paper considers multiuser MIMO CDMA systems with high rate space-time linear dispersion codes (LDC) and orthogonal space-time block codes (O-STBC) in time-varying Rayleigh fading MIMO channels. We propose a multi-function process integrating multi-user detection, space-time decoding and symbol demodulation, which can be coupled with soft channel decoding to improve the system performance in an iterative fashion. We show that the space-time coded CDMA systems approach the single-user bound with only two iterations, and full diversity LDCs enable the systems to utilize the time diversity inherent in fast fading channels. The space-time coded CDMA systems are also compared to the MIMO CDMA system based on spatial multiplexing, some recommendations are made on how to design a practical MIMO CDMA system based on the comparative studies.

I. INTRODUCTION

In this paper, we study orthogonally modulated DS-CDMA systems with an inner space-time block code (STBC). The orthogonal modulation is accomplished by Walsh codes, which combines the advantages of spreading and coding to achieve improved performance for spread spectrum (CDMA) systems. It was shown in [1] that orthogonal signalling improves bandwidth efficiency significantly compared to binary signalling in fading and nonfading channels, and the efficiency improves further as the diversity order increases. In [2], we proposed joint multi-user detection (MUD), channel estimation, demodulation and decoding for the orthogonally modulated and convolutionally coded system over multipath fading channels. Results show that the MUD-assisted iterative soft demodulation and decoding scheme greatly improve the performance of the system under question compared to conventional non-iterative schemes.

Multiple-input multiple-output (MIMO) systems have sparked enormous interests as they promise to significantly increase transmission rate without extra bandwidth and power consumption in wireless communications. Both spatial multiplexing (SM) [3] and space-time coding (STC) [4–6] are popular MIMO communication techniques. The former increases the data rate by simultaneously transmitting symbols from multiple transmit antennas. On the other hand, in STC systems, the same information symbol stream is transmitted from different transmit antennas in an appropriate manner to obtain transmit diversity, therefore provide more reliable communications. The simple transmit diversity scheme developed by Alamouti [4] exploits full spatial diversity with simple linear processing at the receiver, it has been adopted in several wireless standards such as wideband-CDMA and CDMA-2000 [7]. For the purpose of this work, we consider the use of the orthogonal space-time block code (O-STBC) proposed by Alamouti.

A critical issue for high data rate transmission is the space-time coding designs for large MIMO arrays. In particular, full rate O-STBCs cannot be found for complex constellations when the number of transmit antennas is greater than two. Hassibi and Hochwald proposed a high-rate space-time coding framework, linear dispersion codes (LDC) [8], which can support any configuration of transmit and receive antennas, and subsume both SM and STBC as special cases. The principle is to transmit sub-streams of data symbols in linear combinations over space and time. Applications of O-STBC to CDMA systems have been studied, e.g., in [9–11]. However, very limited efforts have been made in investigating the application of LDCs to CDMA systems. In order to support future high data rate CDMA systems, the use of high-rate space-time block codes, e.g. LDCs, may be desirable. In [12], a LDC decoder combined with a blind subspace-based multi-user detector is studied for the downlink of a DS-CDMA system, and a subspace-based sphere decoding algorithm is proposed to further improve the performance. The iterative decoding of LDC codes in a frequency-selective channel is considered in [13], where only a single-user approach is studied, and multi-user scenarios were not investigated. To the best of our knowledge, the issue of iterative multi-user detection and LDC/O-STBC decoding has not been addressed for the orthogonally modulated DS-CDMA systems in the existing literature. In this paper, we propose a joint multi-user detection, space-time decoding, and Q -ary demodulation algorithm for the system under investigation, and the turbo processing principle is applied to improve the system performance, while maintain a reasonable computational complexity.

II. SYSTEM MODEL

Fig. 1 illustrates a block diagram of the baseband received signal due to the k th user, whose l th information bit is denoted as $b_l^k \in \{+1, -1\}$ ($k = 1, \dots, K$, $l = 1, \dots, L_b$, and L_b is the block length). The information bits are convolutionally encoded into code bits $\{u_{n,l}^k\} \in \{+1, -1\}$, where $u_{n,l}^k$ denotes the n th code bit due to b_l^k . For example, in the case of a rate $1/3$ code, b_l^k is encoded into $u_{0,l}^k, u_{1,l}^k, u_{2,l}^k$. Code bits are subsequently interleaved and each block of $\log_2 Q$ coded and interleaved bits $\{u_{n,l}^k\} \in \{+1, -1\}$ is mapped into $\mathbf{w}_k(j) \in \{\mathbf{w}_0, \dots, \mathbf{w}_{Q-1}\}$, which is one of the Q Walsh symbols (codewords). The Q -ary orthogonal modulation is followed by space-time (ST) encoding and spreading. The interleaver and deinterleaver are denoted as Π and Π^{-1} , respectively, in Figs. 1 and 2.

Here, we focus on the description of the LDC coded MIMO CDMA system, the O-STBC coded system will be briefly introduced at the end of this section. The LDC encoding can be

Fig. 1. Block diagram of the transmitter for the STBC coded MIMO DS-CDMA system.

expressed as

$$\mathbf{W}_k(j) = \sum_{q=1}^Q w_k^q(j) \mathbf{A}_q, \quad (1)$$

where the $w_k^q(j)$ denotes the q th bit of the Walsh codeword $\mathbf{w}_k(j)$ (j is the symbol index), and $\mathbf{W}_k(j) \in \mathbb{C}^{T \times N_t}$ is the LDC encoded matrix (T is the number of time slots or channel uses needed to transmit Q LDC symbols, and N_t is the number of transmit antennas, and the symbol $\mathbb{C}$ denotes the complex field). The matrices $\mathbf{A}_q \in \mathbb{C}^{T \times N_t}$, $q = 1, \dots, Q$ are called dispersion matrices, which transform data symbols (Walsh codeword in this context) into a space-time matrix.

In order to facilitate linear LDC decoding in the receiver, it is desirable to reorder $\mathbf{W}_k(j)$ column by column in a vector form. Define $\text{vec}(\cdot)$ operation of $m \times n$ matrix $\mathbf{K}$ as $\text{vec}(\mathbf{K}) = [\mathbf{K}_{\cdot 1}^T \ \mathbf{K}_{\cdot 2}^T \ \dots \ \mathbf{K}_{\cdot n}^T]^T$, where $()^T$ denotes the transpose operation, and $\mathbf{K}_{\cdot i}$ is the i th column of $\mathbf{K}$. Reordering $\mathbf{W}_k(j)$ in (1) and each matrix $\mathbf{A}_q$ into a $TN_t \times 1$ column vector, respectively, we transform (1) into

$$\text{vec}(\mathbf{W}_k(j)) = [\text{vec}(\mathbf{A}_1) \ \dots \ \text{vec}(\mathbf{A}_Q)] [w_k^1(j) \ \dots \ w_k^Q(j)]^T.$$

Denote

$$\mathbf{G}_{\text{vec}} = [\text{vec}(\mathbf{A}_1^T) \ \text{vec}(\mathbf{A}_2^T) \ \dots \ \text{vec}(\mathbf{A}_Q^T)];$$

$$\mathbf{w}_k(j) = [w_k^1(j) \ w_k^2(j) \ \dots \ w_k^Q(j)]^T,$$

we can now express the LDC code due to the transmission of the k th user's j th symbol in vector form as

$$\mathbf{x}_k(j) = \text{vec}(\mathbf{W}_k(j)^T) = \mathbf{G}_{\text{vec}} \mathbf{w}_k(j), \quad (2)$$

which is then repetition encoded into symbol sequence $\mathbf{s}_k(j) = \text{rep}\{\mathbf{x}_k(j), N_c\}$ where $\text{rep}\{\cdot, \cdot\}$ denotes the repetition encoding operation, its first argument is the input bits and the second one is the repetition factor. Therefore, each LDC symbol per channel use is spread (repetition coded) into N_c symbols. The spread sequence $\mathbf{s}_k(j)$ is then scrambled (randomized) by a scrambling code unique to each user to form the transmitted symbol sequence $\mathbf{a}_k(j) = \mathbf{C}_k(j) \mathbf{s}_k(j)$, where $\mathbf{C}_k(j) \in \{-1, +1\}^{N_c T N_t \times N_c T N_t}$ is a diagonal matrix whose diagonal elements correspond to the scrambling code for the k th user's j th symbol. The purpose of scrambling is to separate users. In this paper, we focus on the use of long codes,

e.g., the scrambling code differs from symbol to symbol. The scrambled sequence $\mathbf{a}_k(j)$ is transmitted over multiple-input, multiple-output (MIMO) channel via multiple transmit antennas. The received signal is the sum of K users' signals plus the additive white complex Gaussian noise. After descrambling and despreading, the received signal corresponding to the k th user's j th transmitted sequence can be written in vector form as $\mathbf{y}_k(j) = N_c \mathbf{H}_k(j) \mathbf{x}_k(j) = N_c \mathbf{H}_k(j) \mathbf{G}_{\text{vec}} \mathbf{w}_k(j)$.

Denote the channel gain of the path between the m th transmit antenna and n th receive antenna for the t th channel use of $\mathbf{W}_k(j)$ as $h_{m,n}^{k,t}$, the MIMO channel matrix for the user k can be expressed as

$$\mathbf{H}_k(j) = \begin{bmatrix} \mathbf{H}_k^{(1)} & \dots & \mathbf{0} \\ \vdots & \ddots & \vdots \\ \mathbf{0} & \dots & \mathbf{H}_k^{(T)} \end{bmatrix}, \mathbf{H}_k^{(t)}(j) = \begin{bmatrix} h_{1,1}^{k,t}(j) & \dots & h_{N_t,1}^{k,t}(j) \\ \vdots & \ddots & \vdots \\ h_{1,N_r}^{k,t}(j) & \dots & h_{N_t,N_r}^{k,t}(j) \end{bmatrix} \quad (3)$$

In this work, we focus on asynchronous transmission. Without loss of generality, we assume $\tau_1 < \tau_2 < \dots < \tau_k < \dots < \tau_K$, where τ_k is the propagation delay for user k , and is assumed to be multiple of chip intervals. The maximum delay spread is assumed to be less than or equal to a symbol interval (N_c chip intervals). After descrambling and despreading, the received signal corresponding to the k th user's j th transmitted sequence can now be expressed as

$$\mathbf{r}(j) = \sum_{k=1}^K \mathbf{y}_k(j) + \mathbf{n}(j) = \underbrace{N_c \mathbf{H}_k(j) \mathbf{G}_{\text{vec}} \mathbf{w}_k(j)}_{\text{desired signal}} + \underbrace{\mathbf{n}(j)}_{\text{noise}}$$

$$+ \underbrace{\sum_{s=1}^{k-1} [(N_c - \tau_k + \tau_s) \mathbf{H}_s(j) \mathbf{G}_{\text{vec}} \mathbf{w}_s(j) + (\tau_k - \tau_s) \mathbf{H}_s(j+1) \mathbf{G}_{\text{vec}} \mathbf{w}_s(j+1)]}_{\text{MAI}}$$

$$+ \underbrace{\sum_{s=k+1}^K [(\tau_s - \tau_k) \mathbf{H}_s(j-1) \mathbf{G}_{\text{vec}} \mathbf{w}_s(j-1) + (N_c - \tau_s + \tau_k) \mathbf{H}_s(j) \mathbf{G}_{\text{vec}} \mathbf{w}_s(j)]}_{\text{MAI}},$$

where $\mathbf{n}(j) \in \mathbb{C}^{TN_t}$ is a vector of iid complex Gaussian noise samples with zero mean and variance N_0 , i.e., $\mathbf{n}(j) \sim \mathcal{CN}(\mathbf{0}, N_0 \mathbf{I})$.

The transmitter of the O-STBC system differs from that of the LDC system only in the ST encoder and spreader block. In the original Alamouti scheme [4, 9], two symbols are transmitted from two antennas at a time, which does not suit the system with orthogonal modulation under investigation since the basic

transmission unit is the Walsh sequence. To work around this problem, we propose a block-based encoding and transmission scheme as follows. After performing spreading operation on the transmitted Walsh codeword, we group each Walsh chip sequence $\mathbf{s}_k(j) = \{s_k^0(j), s_k^1(j), \dots, s_k^{(N-1)}(j)\}$ (where $s_k^q(j)$ is the q th chip of $\mathbf{s}_k(j)$) into two data blocks $\mathbf{s}_k^0(j)$ and $\mathbf{s}_k^1(j)$ at each antenna, each containing $N_c/2$ chips, i.e.,

$$\begin{aligned} \mathbf{s}_k^0(j) &= \{s_k^0(j), s_k^1(j), \dots, s_k^{(N/2-1)}(j)\} \\ \mathbf{s}_k^1(j) &= \{s_k^{(N/2)}(j), s_k^{(N/2+1)}(j), \dots, s_k^{(N-1)}(j)\} \\ \mathbf{s}_k^{0*}(j) &= \{s_k^{0*}(j), s_k^{1*}(j), \dots, s_k^{(N/2-1)*}(j)\} \\ -\mathbf{s}_k^{1*}(j) &= \{-s_k^{(N/2)*}(j), -s_k^{(N/2+1)*}(j), \dots, -s_k^{(N-1)*}(j)\}, \end{aligned} \quad (4)$$

where $()^*$ denotes the conjugate operation. The data transmission is divided into two time frames. During the first time frame, $\mathbf{s}_k^0(j)$ is transmitted from the first antenna; and $\mathbf{s}_k^1(j)$ is transmitted from the second antenna. During the second transmission frame, the first antenna transmits a complex conjugated and sign inverted version of $\mathbf{s}_k^1(j)$, denoted as $-\mathbf{s}_k^{1*}(j)$; the second antenna transmits a complex conjugated version of $\mathbf{s}_k^0(j)$, denoted as $\mathbf{s}_k^{0*}(j)$. The rest of the transmission mechanism is the same as in the LDC system.

III. JOINT ST DECODING AND Q -ARY SYMBOL DEMODULATION

The proposed iterative detection and decoding scheme is illustrated in Fig. 2. The soft metric $\lambda(u_{n,l}^k; O)$ from the inner soft-input, soft-output (SISO) block is deinterleaved to $\lambda(u_{n,l}^k; I)$. The k^{th} user's Log-MAP decoder computes an a posteriori log-likelihood ratio (LLR) of each information bit $\lambda(b_l^k; O)$ and each code bit $\lambda(u_{n,l}^k; O)$ based on the soft input $\lambda(u_{n,l}^k; I)$ and the trellis structure of the convolutional code. The former is used to make a decision on the transmitted information bit at the final iteration, while the latter is used for multi-user detection in the inner SISO block at the next iteration. We use the notations $\lambda(\cdot; I)$ and $\lambda(\cdot; O)$ to denote the input and output ports of a SISO device.

The algorithms discussed above require the design of an inner SISO block that can produce soft reliability values for each bit $u_{n,l}^k$ from the received signal in order to enable soft input channel decoding. To this end, we propose an integrated multi-user detection (MUD), ST decoding, and symbol demodulation scheme, which will be described next. To simplify the notation, we sometimes suppress the index j from $\mathbf{s}_k(j)$, $\mathbf{w}_k(j)$ and $\mathbf{H}_k(j)$, etc., whenever no ambiguity arises.

Among different MUD techniques, the multistage parallel interference cancellation (PIC) scheme [14] is known to be simple and effective for mitigating the effect of MAI and increasing the capacity of interference-limited CDMA systems. The MUD-based iterative scheme is shown in Fig. 2. It uses soft information of $\mathbf{w}_k(j)/\mathbf{s}_k(j)$, denoted as $\bar{\mathbf{w}}_k(j)/\bar{\mathbf{s}}_k(j)$, for interference cancellation. To this end, we compute $\bar{\mathbf{w}}_k(j) = [\bar{w}_k^0(j) \ \bar{w}_k^1(j) \ \dots \ \bar{w}_k^{Q-1}(j)]^T$, the soft estimate of the codeword $\mathbf{w}_k(j)$, from its LLR $\lambda(\mathbf{w}_k(j))$, which is derived by feeding $\lambda(u_{n,l}^k) = \Pi\{\lambda(u_{n,l}^k; O)\}$ into a soft modulator [2]. Then we perform ST encoding and spreading to produce an soft estimate of the transmitted chip sequence $\bar{\mathbf{s}}_k(j)$. The initial symbol estimates are obtained by single-user detection schemes, such as matched filtering.

The joint ST decoding and symbol demodulation schemes for the LDC and O-STBC systems are introduced in Section III-A and III-B, respectively.

A. LDC coded system

To demodulate the signal transmitted from user k , we remove the interference by subtracting the estimated signals of the interfering users from the received signal $\mathbf{r}(j)$ to form a new signal vector $\mathbf{r}'_k(j)$, i.e.,

$$\begin{aligned} \mathbf{r}'_k(j) &= \mathbf{r}(j) - \sum_{s=1}^{k-1} [(N_c - \tau_k + \tau_s) \mathbf{H}_s(j) \mathbf{G}_{\text{vec}} \bar{\mathbf{w}}_s(j) \\ &\quad - (\tau_k - \tau_s) \mathbf{H}_s(j+1) \mathbf{G}_{\text{vec}} \bar{\mathbf{w}}_s(j+1)] \\ &\quad - \sum_{s=k+1}^K [(\tau_s - \tau_k) \mathbf{H}_s(j-1) \mathbf{G}_{\text{vec}} \bar{\mathbf{w}}_s(j-1) \\ &\quad - (N_c - \tau_s + \tau_k) \mathbf{H}_s(j) \mathbf{G}_{\text{vec}} \bar{\mathbf{w}}_s(j)] \\ &= N_c \mathbf{H}_k(j) \mathbf{G}_{\text{vec}} \mathbf{w}_k(j) + \mathbf{v}'_k(j), \end{aligned} \quad (5)$$

where $\mathbf{v}'_k(j)$ denotes the combined cancellation residual and noise, and $\bar{\mathbf{w}}_s(j)$ is the soft estimate of $\mathbf{w}_s(j)$. The vector $\mathbf{r}'_k(j)$ is the interference canceled version of $\mathbf{r}(j)$ after subtracting the contributions from all the other users using decision feedback. In case of perfect interference cancellation, the cancellation residual vanishes, and $\mathbf{v}'_k(j) \sim \mathcal{CN}(\mathbf{0}, N_v \mathbf{I})$, where $N_v = N_c N_0$. As will become apparent in Section IV, the assumption of perfect cancellation can be approached by proper design of the iterative receiver.

The Walsh codeword $\mathbf{w}_k$ (or equivalently the j th Walsh symbol for user k) can be estimated by a linear MMSE algorithm, i.e.,

$$\hat{\mathbf{w}}_k = \Phi^H \mathbf{r}'_k = \Phi^H (N_c \mathbf{H}_k \mathbf{G}_{\text{vec}} \mathbf{w}_k + \mathbf{v}_k) = \mathbf{U} \mathbf{w}_k + \xi_k, \quad (6)$$

where the matrix Φ is designed to minimize $E[\|\hat{\mathbf{w}}_k - \mathbf{w}_k\|^2]$, leading to the solution $\Phi = \mathbf{R}^{-1} \mathbf{P}$, where

$$\begin{aligned} \mathbf{R} &= E[\mathbf{r}'_k \mathbf{r}'_k{}^H] = E[(N_c \mathbf{H}_k \mathbf{G}_{\text{vec}} \mathbf{w}_k + \mathbf{v}_k)(N_c \mathbf{w}_k^H \mathbf{G}_{\text{vec}}^H \mathbf{H}_k^H + \mathbf{v}_k^H)] \\ &= N_c^2 \mathbf{H}_k \mathbf{H}_k^H + N_v \mathbf{I}; \end{aligned} \quad (7)$$

$$\begin{aligned} \mathbf{P} &= E[\mathbf{r}'_k \mathbf{w}_k^H] = E[(N_c \mathbf{H}_k \mathbf{G}_{\text{vec}} \mathbf{w}_k + \mathbf{v}_k) \mathbf{w}_k^H] = N_c \mathbf{H}_k \mathbf{G}_{\text{vec}}; \\ \mathbf{U} &= \Phi^H N_c \mathbf{H}_k \mathbf{G}_{\text{vec}} = \Phi^H \mathbf{P}. \end{aligned} \quad (8)$$

Equations (7) and (8) are derived utilizing the fact that for Walsh codewords, $E[\mathbf{w}_k \mathbf{w}_k^H] = \mathbf{I}_Q$, and $\mathbf{G}_{\text{vec}}$ is an unitary matrix for LDC codes ($\mathbf{G}_{\text{vec}} \mathbf{G}_{\text{vec}}^H = \mathbf{I}_Q$). The noise term $\xi_k = \Phi^H \mathbf{v}_k$ is Gaussian, since it is a linear transformation of a Gaussian random vector, with zero mean and covariance $E[\xi_k \xi_k^H] = N_v \Phi^H \Phi$. The probability density function (PDF) of the MMSE filter output $\hat{\mathbf{w}}_k$, conditioned on that the m th Walsh symbol is transmitted, can be expressed as

$$\begin{aligned} f(\hat{\mathbf{w}}_k | \mathbf{w}_m) &= \frac{\exp[-(\hat{\mathbf{w}}_k - \mathbf{U} \mathbf{w}_m)^H (N_v \Phi^H \Phi)^{-1} (\hat{\mathbf{w}}_k - \mathbf{U} \mathbf{w}_m)]}{\pi^Q \det(N_v \Phi^H \Phi)} \\ &= \frac{1}{\pi^Q \det(N_v \Phi^H \Phi)} \exp \left[-\frac{\|\Phi^{-H} \hat{\mathbf{w}}_k - \mathbf{P} \mathbf{w}_m\|^2}{N_v} \right]. \end{aligned}$$

The soft metric for the bit $u_{n,l}^k$ can thus be computed in terms

Fig. 2. Iterative receiver structure.

of LLR as

$$\begin{aligned}
 \lambda(u_{n,l}^k; O) &= \ln \frac{\sum_{m: u_{n,l}^k = +1} f(\hat{\mathbf{w}}_k | \mathbf{w}_m)}{\sum_{m: u_{n,l}^k = -1} f(\hat{\mathbf{w}}_k | \mathbf{w}_m)} \\
 &\approx \ln \frac{\max_{m: u_{n,l}^k = +1} f(\hat{\mathbf{w}}_k | \mathbf{w}_m)}{\max_{m: u_{n,l}^k = -1} f(\hat{\mathbf{w}}_k | \mathbf{w}_m)} \quad (9) \\
 &= \ln \frac{\max_{m: u_{n,l}^k = +1} \exp(-\|\Phi^{-\mathcal{H}} \hat{\mathbf{w}}_k - \mathbf{P} \mathbf{w}_m\|^2 / N_v)}{\max_{m: u_{n,l}^k = -1} \exp(-\|\Phi^{-\mathcal{H}} \hat{\mathbf{w}}_k - \mathbf{P} \mathbf{w}_m\|^2 / N_v)} \\
 &= \frac{1}{N_v} \text{Re}\{[2(\Phi^{-1} \mathbf{P} \mathbf{w}^+)^{\mathcal{H}} \hat{\mathbf{w}}_k - \|\mathbf{P} \mathbf{w}^+\|^2] - \\
 &\quad [2(\Phi^{-1} \mathbf{P} \mathbf{w}^-)^{\mathcal{H}} \hat{\mathbf{w}}_k - \|\mathbf{P} \mathbf{w}^-\|^2]\}, \quad (10)
 \end{aligned}$$

where $m : u_{n,l}^k = \pm 1$ denotes the set of Walsh codewords $\{\mathbf{w}_m\}$ that correspond to the code bit $u_{n,l}^k = \pm 1$, and $\mathbf{w}^+$ denotes the Walsh codeword $\mathbf{w}_m$ that corresponds to $\max_{m: u_{n,l}^k = +1} f(\hat{\mathbf{w}}_k | \mathbf{w}_m)$; $\mathbf{w}^-$ denotes the Walsh codeword $\mathbf{w}_m$ that corresponds to $\max_{m: u_{n,l}^k = -1} f(\hat{\mathbf{w}}_k | \mathbf{w}_m)$.

B. O-STBC coded system

The receiver design for the O-STBC system is simpler than that of the LDC system due to its orthogonal structure. PIC can also be applied in the O-STBC system for interference mitigation. The received signal at the l th receive antenna after despreading and interference cancellation during the first and second time frames can be expressed in vector form as

$$\begin{bmatrix} \mathbf{r}_{k,l}^0 \\ \mathbf{r}_{k,l}^{1*} \end{bmatrix} = \underbrace{\begin{bmatrix} h_{k,l}^0 \mathbf{I} & h_{k,l}^1 \mathbf{I} \\ h_{k,l}^{1*} \mathbf{I} & -h_{k,l}^{0*} \mathbf{I} \end{bmatrix}}_{\mathbf{H}_{k,l}} \underbrace{\begin{bmatrix} \mathbf{s}_k^0 \\ \mathbf{s}_k^1 \end{bmatrix}}_{\mathbf{s}_k} + \underbrace{\begin{bmatrix} \mathbf{v}_{k,l}^0 \\ \mathbf{v}_{k,l}^{1*} \end{bmatrix}}_{\mathbf{v}_{k,l}},$$

where $\mathbf{r}_{k,l}^0$ ($\mathbf{r}_{k,l}^1$) and $\mathbf{v}_{k,l}^0$ ($\mathbf{v}_{k,l}^1$) are the received signal and noise plus interference cancellation residual corresponding to the k th user's j th transmitted sequence, respectively, during the first (second) time frame; $\mathbf{r}_{k,l}^{1*}$ is the complex conjugated version of $\mathbf{r}_{k,l}^1$. The transmitted chip sequences $\mathbf{s}_k^0$, $\mathbf{s}_k^1$ are defined in (4). The channel coefficients from the k th user's first and second transmit antenna to the l th receive antenna are denoted as $h_{k,l}^0$ and $h_{k,l}^1$, respectively. An explicit requirement for the O-STBC system is that the channel coefficients have to remain constant during two consecutive time frames, while the LDC system does not have this restriction, and is able to exploit the time diversity in fast fading channels.

Assuming perfect interference cancellation, $\mathbf{v}_{k,l}$ only contains the contribution from the noise, i.e., $\mathbf{v}_{k,l} \sim \mathcal{CN}(\mathbf{0}, N_0 \mathbf{I})$,

the conditional PDF of the received vector $\mathbf{v}_{k,l}$ can be expressed as

$$f(\mathbf{r}_{k,l} | \mathbf{s}_m) = \frac{1}{(\pi N_0)^{N_c}} \exp\left(-\frac{\|\mathbf{r}_{k,l} - \mathbf{H}_{k,l} \mathbf{s}_m\|^2}{N_0}\right).$$

The receiver combines the signals received from different antennas, i.e., $\mathbf{r}_k = \sum_{n=0}^{N_r-1} \mathbf{r}_{k,n}$ where N_r is the number of receive antennas. Since $\mathbf{v}_{k,0}$, $\mathbf{v}_{k,1}$, $\mathbf{v}_{k,2} \dots$ are uncorrelated, thus the conditional PDF for $\mathbf{r}_k$ is

$$\begin{aligned}
 f(\mathbf{r}_k | \mathbf{s}_m) &= \prod_{l=0}^{N_r-1} f(\mathbf{r}_{k,l} | \mathbf{s}_m) \\
 &= \prod_{l=0}^{N_r-1} \frac{1}{(\pi N_0)^{N_c}} \exp\left(-\frac{\|\mathbf{r}_{k,l} - \mathbf{H}_{k,l} \mathbf{s}_m\|^2}{N_0}\right)
 \end{aligned}$$

The LLR for the bit $u_{n,l}^k$ can be derived as

$$\begin{aligned}
 \lambda(u_{n,l}^k; O) &\approx \ln \frac{\max_{m: u_{n,l}^k = +1} \prod_{l=0}^{N_r-1} f(\mathbf{r}_{k,l} | \mathbf{s}_m)}{\max_{m: u_{n,l}^k = -1} \prod_{l=0}^{N_r-1} f(\mathbf{r}_{k,l} | \mathbf{s}_m)} \\
 &= \ln \frac{\prod_{l=0}^{N_r-1} \exp(-\|\mathbf{r}_{k,l} - \mathbf{H}_{k,l} \mathbf{s}^+\|^2 / N_0)}{\prod_{l=0}^{N_r-1} \exp(-\|\mathbf{r}_{k,l} - \mathbf{H}_{k,l} \mathbf{s}^-\|^2 / N_0)} \\
 &= \frac{2}{N_0} \sum_{l=0}^{N_r-1} \text{Re}\{(\mathbf{H}_{k,l} \mathbf{s}^+)^{\mathcal{H}} \mathbf{r}_{k,l} - (\mathbf{H}_{k,l} \mathbf{s}^-)^{\mathcal{H}} \mathbf{r}_{k,l}\}
 \end{aligned}$$

where $\mathbf{s}^+$ denotes the Walsh sequence $\mathbf{s}_m$ that corresponds to $\max_{m: u_{n,l}^k = +1} f(\mathbf{r}_{k,l} | \mathbf{s}_m)$, and $\mathbf{s}^-$ is defined similarly. Due to the orthogonality of the channel matrix $\mathbf{H}_{k,l}$, we have

$$\sum_{l=0}^{N_r-1} (\mathbf{H}_{k,l} \mathbf{s}_m)^{\mathcal{H}} \mathbf{r}_{k,l} = \sum_{l=0}^{N_r-1} (|h_{k,l}^0|^2 + |h_{k,l}^1|^2) \mathbf{s}_m^{\mathcal{H}} \mathbf{s}_m + (\mathbf{H}_{k,l} \mathbf{s}_m)^{\mathcal{H}} \mathbf{v}_{k,l},$$

which shows that the above scheme achieves full spatial diversity.

IV. NUMERICAL RESULTS

In the simulations, we employ the following convolutional codes: 1) a weak code with code rate $R_c = 1/2$, constraint length $L_c = 3$, generator polynomials $(5, 7)_8$; 2) a strong code with $R_c = 1/3$, $L_c = 5$, generator polynomials $(25, 33, 37)_8$.

Fig. 3. Comparison of different schemes. All the curves represent the performance of different systems upon reaching convergence at the 4th iteration.

Fig. 4. Comparison of different schemes. All the curves are plotted at the 3rd iteration.

The code is chosen to be the same for all the users. The coded and interleaved bits (random interleaver is used) from each user is converted into one of $Q = 8$ Walsh symbols, followed by ST encoding and spreading. For the LDC system, we use the full-diversity, full-rate (FDFR) code (design A) [15]. The parameter setting is chosen to be $N_t = 2$, $T = 4$, and $N_r = 2$ or 4. In this case, 8-bit Walsh codeword is dispersed into 2 transmit antennas, each accommodates 4 LDC symbols ($T = 4$ channel uses before spreading). Each LDC symbol is spread (repetition encoded) to $N_c = 8$ symbols. In order to achieve the same data rate, the spreading factor has to be reduced to $N_c = 4$ in the O-STBC system. The number of users is chosen to be $K = 18$, which represents a fairly heavy loaded system. The proposed LDC/O-STBC system is compared to the SM system. In the latter case, information bearing signals are divided into multiple substreams, each encoded and modulated independently; on the receiver side, we use the iterative soft demodulation and decoding algorithm presented in [2]. In all the systems, the single-user detection is used in the beginning of the iterative process to obtain an initial estimate of data, which are needed for MUD at subsequent iterations. To transmit one Walsh symbol, the SM system needs $Q = 8$ channel uses (time slots), whereas the LDC system only needs $T = 4$ channel uses at both transmit antennas. However, due to simultaneous transmission from two transmit antennas in the SM system, all the systems have the same data rate.

The long scrambling codes C_k are generated randomly. The noise variance N_0 , and C_k as well as path delays are assumed to be known to the receiver. The channel gain $h_{m,n}^{k,t}$ is a complex circular Gaussian process with autocorrelation function $E[h_{m,n}^{k,t} h_{m,n}^{k,t+\tau}] = P_{m,n}^k J_0(2\pi f_D \tau)$ where f_D is the maximum Doppler frequency, $J_0(x)$ is the zero-th order Bessel function of the first kind, and $P_{m,n}^k$ is the power of $h_{m,n}^{k,t}$. The amplitude of $h_{m,n}^{k,t}$ follows a Rayleigh distribution. The Doppler shifts are due to the relative motion between the base station and mobile units. Perfect slow power control is assumed in the sense that $P_k = \sum_{m,n} E[|h_{m,n}^{k,t}|^2]$, the average received power, is equal for all users, and it is normalized such that $P_k = 1$ for all k . However, the instantaneous power $|h_{m,n}^{k,t}|^2$ may vary from one user to another. Perfect knowledge of the channel state information (CSI) is assumed in our simulations. In Figs. 3 – 5, we as-

sume channel gains remain constant during the transmission of one LDC codeword in the LDC system, two data frames in the O-STBC system, and one Walsh chip sequence in the SM system, i.e., $h_{m,n}^{k,t}(j) = h_{m,n}^{k,t+1}(j) = h_{m,n}^{k,t+2}(j) = h_{m,n}^{k,t+3}(j)$. The normalized Doppler frequency is assumed to be $f_D T_s = 0.01$, where T_s is the symbol duration. The simulation results are averaged over random fading, noise, delays, and scrambling codes with minimum of 50 blocks of data transmitted and at least 100 bit errors generated.

The performance of different systems are compared in Fig. 3 when the weak code ($R_c = 1/2$, $L_c = 3$) is used. The number of receive antenna is $N_r = 2$. One can see from the figure that the SM system has the worst performance. The O-STBC system shows clear superiority over other systems. It outperforms the LDC system and the SM system by 0.7 dB and 1.3 dB, respectively, at the target BER= 10^{-4} . In Fig. 4, we examine the behavior of different systems with a stronger channel code ($R_c = 1/3$, $L_c = 5$) and a higher receive diversity order ($N_r = 4$). In this case, the advantage of applying STC becomes much smaller. For example, the performance gain obtained by the O-STBC system is only up to 0.3 dB compared to the SM system.

Fig. 5 shows the performance of the iterative receivers at different iterations. We notice that the SM system has to iterate 3 times before reaching convergence, while the O-STBC system only needs to iterate 2 times to converge. Therefore, the use of STC leads to faster convergence for the iterative receiver. The single user bound for the O-STBC coded system is also shown in Fig. 5. It is obtained by the proposed scheme in single user environment. No interference mitigation is needed in this case, and it gives a lower bound on the best performance achievable by applying MUD technique. It can be seen that the performance of the proposed iterative MUD approach upon reaching convergence is very close to the single user bound, meaning that MAI can be effectively eliminated by proper design of iterative MUD schemes.

Note that the proposed LDC system does not require constant channel during the transmission of one LDC codeword, channel gains can vary from one time slot to another, i.e., $h_{m,n}^{k,t}(j) \neq h_{m,n}^{k,t+1}(j) \neq h_{m,n}^{k,t+2}(j) \neq h_{m,n}^{k,t+3}(j)$. This is in contrast to the previous work on LDC for CDMA systems, which were designed for block fading channel over the whole LDC

Fig. 5. Comparison of the iterative schemes. Curves with the same marker represent the performance of the same scheme at different iterations (including the first single user detection stage). The parameter setting is the same as in Fig. 4.

Fig. 6. Performance of LDC coded system in fast fading channels. The parameter setting is the same as in Fig. 3.

codeword. In Fig. 6, we examine the performance of the LDC coded system using the proposed iterative MUD approach in faster fading channels when the channel gains keep changing at each time slot within one LDC codeword. To this end, we re-define the normalized Doppler frequency as $f_d T$, where T is the duration of one time slot (channel use). The SNR is set to be $E_b/N_0 = 8$ dB. Slow power control is assumed so that the average received power is equal for all users. The BER curve is plotted at the 3rd stage when the system reaches convergence. Fig. 6 shows that the system performance improves as the Doppler frequency increases, which clearly indicates the time diversity obtained by the LDC codes. However, this time diversity is not exploited for slow fading channels as in the previous cases, where the normalized Doppler frequency is set to be $f_d T_s = 0.01$ and code level channel stationarity is assumed. It was shown in [2], that for non-LDC systems, the performance degrades as the normalized Doppler frequency increases.

V. CONCLUSIONS

Iterative detection for orthogonally modulated and STBC coded MIMO DS-CDMA system has been investigated in this paper. We propose an integrated design of MUD, ST decoding, symbol demodulation, and symbol-to-bit LLR mapping in order to reduce the complexity of the turbo receiver. Results show that in comparison to the SM system with the same transmission rate, the STBC coded systems have superior performance and faster convergence. Furthermore, by exploiting time diversity, LDCs also provide us a powerful means to combat impairment caused by fast fading channels. However, we have observed that the advantages of applying STBC become smaller when a strong channel code is used and/or when the receive diversity increases. Considering the fact that strong channel codes are usually employed in practical communication systems, and a high receive diversity order can be readily fulfilled at the base station, we can conclude that the simple SM scheme would suffice for MIMO CDMA systems in slow fading channels, whereas LDCs are more suitable for fast fading channels. The results from this work could provide a valuable source of information for future versions of the 3GPP standard.

ACKNOWLEDGEMENT

The financial support of the EPSRC (under grant number EP/D07827X/1) is gratefully acknowledged.

REFERENCES

- [1] K. Pahlavan, M. Chase. "Spread-spectrum multiple-access performance of orthogonal codes for indoor radio communications". *IEEE Trans. on Commun.*, vol. 38, no. 5, pp. 574–577, May 1990.
- [2] P. Xiao. *Iterative Detection, Decoding and Channel Parameter Estimation for Orthogonally Modulated DS-CDMA Systems*. PhD thesis, Department of Signals and Systems, Chalmers University of Technology, Gothenburg, Sweden. Jan. 2004. Available at <http://publications.lib.chalmers.se/records/fulltext/243.pdf>
- [3] G. Foschini. "Layered space-time architecture for wireless communication in a fading environment when using multiple antennas". *Bell Labs. Tech. J.*, vol. 1, no. 2, pp. 41–59, 1996.
- [4] A. Alamouti. "A simple transmit diversity technique for wireless communications". *IEEE J. Sel. Areas Commun.*, vol. 16, no. 8, pp. 1451–1458, Oct. 1998.
- [5] V. Tarokh, N. Seshadri, A. Calderbank. "Space-time codes for high data rates wireless communications: performance criterion and code construction". *IEEE Trans. on Inform. Theory*, vol. 44, pp. 744–765, March 1998.
- [6] V. Tarokh, N. Seshadri, A. Calderbank. "Space-time block coding from orthogonal designs". *IEEE Transactions on Information Theory*, vol. 45, pp. 1456–1467, July 1999.
- [7] 3GPP. "Physical channels and mapping of transport channels onto physical channels". Technical Specification TS 25.11 (V3.3.0), Technical Specification Group Radio Access Network, June 2000.
- [8] B. Hassibi, B. Hochwald. "High-rate codes that are linear in space and time". *IEEE Trans. Inform. Theory*, vol. 48, no. 7, pp. 1804–1824, July 2002.
- [9] B. Lu, X. Wang. "Iterative receivers for multiuser space-time coding systems". *IEEE Journal on Selected Areas in Communications*, vol. 18, no. 11, pp. 2322–2335, Nov. 2000.
- [10] X. Yue, H. Fan. "Near-far resistance of Alamouti space-time coded CDMA communication systems". *Sig. Proc.*, vol. 84, no. 3, pp. 669–674, March 2004.
- [11] G. Leus, F. Petré, M. Moonen. "Space-time chip equalization for maximum diversity space-time block coded DS-CDMA downlink transmission". *EURASIP J. Applied Sig. Proc.*, no. 5, pp. 740–750, 2004.
- [12] J. Liu, A. Host-Madsen. "Blind decoders for linear dispersion codes in downlink DS-CDMA systems". *IEEE Commun. Letters*, vol. 6, no. 7, pp. 285–287, July 2002.
- [13] A. Wiesel, X. Mestre, A. Pages, J. Fonollosa. "Turbo equalization and demodulation of multicode space time codes". *Proc. IEEE ICC*, vol. 4, pp. 2296–2300, May 2003.
- [14] P. Xiao, E. Ström. "A theoretical evaluation of parallel interference cancellation in M-ary orthogonal modulated asynchronous DS-CDMA system over multipath rayleigh fading channels". *IEEE Trans. on Veh. Technol.*, vol. 54, no. 4, pp. 1400–1414, July, 2005.
- [15] X. Ma and G. Giannakis. "Full-diversity full-rate complex-field space-time coding". *IEEE Trans. on Sig. Proc.*, vol. 51, no. 11, pp. 2917–2930, Nov. 2003.