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Iterative Multiuser Detection and Decoding for DS-CDMA System With Space-Time Linear Dispersion

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Abstract—This paper considers a $Q$-ary orthogonal direct-sequence code-division multiple-access (DS-CDMA) system with high-rate space-time linear dispersion codes (LDCs) in time-varying Rayleigh fading multiple-input–multiple-output (MIMO) channels. We propose a joint multiuser detection, LDC decoding, and channel-decoding algorithm and apply the turbo processing principle to improve system performance in an iterative fashion. The proposed iterative scheme demonstrates faster convergence and superior performance compared with the V-BLAST-based DS-CDMA system and is shown to approach the single-user performance bound. We also show that the CDMA system is able to exploit the time diversity offered by the LDCs in rapid-fading channels.

Index Terms—Code-division multiple access (CDMA), iterative detection and decoding, linear dispersion codes (LDCs), multiple-input–multiple-output (MIMO), multiuser detection (MUD).

I. INTRODUCTION

In this paper, we study a $Q$-ary orthogonally modulated orthogonal direct-sequence code-division multiple-access (DS-CDMA) system that employs high-rate space-time linear dispersion codes (LDCs) and an iterative detection and decoding receiver. The orthogonal modulation of the DS-CDMA system is accomplished by Walsh codes, which combines the advantages of spreading and coding to achieve improved performance for spread spectrum [code-division multiple-access (CDMA)] systems. Multiuser detection (MUD) [1] is an effective tool to increase the capacity of interference-limited CDMA systems while reducing some technical requirements such as power control. Several iterative MUD schemes were proposed, e.g., in [2], for un-coded $Q$-ary orthogonal systems with affordable complexity, which is significantly less than that of an optimum receiver. It has been shown that the performance of an iterative MUD receiver is far better than that of the conventional receiver, particularly in high-load networks in which the interference from other users is severe [2].

Channel coding is usually employed in practical systems to improve the error detection and correcting capability and power efficiency. The CDMA systems exhibit their full potential when combined with forward error-correction coding [3]. In this context, the joint MUD and decoding, including soft interference cancellation, linear minimum mean square error (MMSE) filtering, trellis-based Log-MAP multiuser detector, and blind Bayesian MUD are some of the schemes studied in [4]–[7] to reduce the deteriorative effect of interference. In [8], we provide a thorough treatment of joint MUD, channel estimation, demodulation, and decoding for the serially concatenated CDMA system (orthogonal modulation concatenated with outer convolutional code) over multipath fading channels. Decision-directed interference cancellation/suppression and channel estimation were proposed to combat the effect of multiple-access interference (MAI) and to improve the reliability of the demodulation process. For a general reference to iterative methods for joint detection, see [9].

In recent years, multiple-transmit–multiple-receive [multiple-input–multiple-output (MIMO)] antenna systems have attracted extensive interest and research. In particular, space-time block coding (STBC) has emerged as one of the most promising technologies for meeting the high data rate and high service quality requirements for wireless communications [10]–[12]. A critical issue for high-data-rate transmission is the STBC designs for large MIMO arrays. In particular, full-rate 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, called LDCs [13]. The principle is to transmit substrims of data in linear combinations over space and time. LDCs are designed to optimize the mutual information between the transmitted and received signals and, at the same time, retain decoding simplicity due to their linear structure. They provide a powerful means to combat fading by dispersing the transmitted signals over time and space, which is equivalent to creating better effective channels and improving effective signal-to-interference-plus-noise ratio, leading to an improved system performance. It is shown in [13] that LDC may achieve

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a coding rate of up to one and outperform the well-known full-rate uncoded V-BLAST [14] scheme.

Applications of low-rate orthogonal space-time block codes to CDMA systems have been studied, e.g., in [15]–[17]. To support future high-data-rate CDMA systems, the use of high-rate space-time block codes, e.g., LDCs, may be desirable. However, very limited efforts have been made in investigating the application of LDCs to CDMA systems. In [18], an LDC decoder combined with a blind subspace-based multiuser 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 LDCs in a frequency-selective channel is considered in [19], where only a single-user approach is studied, and multiuser scenarios are not investigated. To the best of our knowledge, the issue of iterative MUD and LDC decoding has not been addressed for the orthogonally modulated DS-CDMA systems in the existing literature.

In this paper, we propose a joint MUD, LDC decoding, and Q-ary demodulation algorithm for the system under investigation. The turbo processing principle is employed to improve the system performance in an iterative manner, while maintaining a reasonable computational complexity. The role of LDC in this work is to exploit space and time diversity for multiple users in DS-CDMA systems. The performance of the LDC-coded DS-CDMA system is compared to that of the V-BLAST-based system, and time diversity gains obtained by the LDCs are investigated for the case when the channel gains change within one LDC codeword.

The remainder of this paper is organized as follows. The system model is introduced in Section II. Different LDCs suitable for the orthogonally modulated CDMA system are discussed in Section III. Iterative MUD, demodulation, and decoding schemes are introduced in Section IV. Different algorithms are examined and compared numerically in Section V. Conclusions are drawn in Section VI.

II. SYSTEM MODEL

Fig. 1 illustrates a block diagram of the baseband received signal due to the $k$th user. The $k$th user’s $l$th information bit is denoted as $b_k^l \in \{+1, -1\}$ ($k = 1, \ldots, K$, $l = 1, \ldots, L_b$, where $L_b$ 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 $nl$th code bit due to $b_k^l$. For example, in the case of a rate-1/3 code, $b_k^l$ is encoded into $u_{0,l}^k$, $u_{1,l}^k$, and $u_{2,l}^k$. 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 $w_k(j) \in \{w_0, \ldots, w_{Q-1}\}$, which is one of the $Q$ Walsh symbols ($j$ is the symbol index). The interleaver and deinterleaver are denoted as $\Pi$ and $\Pi^{-1}$, respectively, in Figs. 1 and 2. The Walsh codeword $w_k(j) \in \{+1, -1\}^Q$ is then encoded with an LDC as follows:

$$W_k(j) = \sum_{q=1}^{Q} w_k^q(j) A_q$$

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

In a more general case, when the data sequence is modulated using complex-valued symbols $\gamma_q = \alpha_q + i\beta_q$, chosen from an arbitrary constellation (e.g., r-PSK or r-QAM), an LDC $S_{LDC}$ is defined as [13]

$$S_{LDC} = \sum_{q=1}^{Q} (\alpha_q A_q + i\beta_q B_q)$$

where $i = \sqrt{-1}$. Note that $w_k^q$ in (1) is real valued; therefore, $B_q$ becomes irrelevant for the system under investigation.

To facilitate LDC decoding in the receiver, it is desirable to reorder $W_k(j)$ column by column in a vector form. Define the $\text{vec}(\cdot)$ operation of the $m \times n$ matrix $K$ as
vec(K) = [K_1^T, K_2^T, \ldots, K_n^T]^T, where \( T \) denotes the transpose operation, and \( K_i \) is the \( i \)th column of \( K \). Denoting

\[
G_{vec} = [\text{vec}(A_1^T), \text{vec}(A_2^T), \ldots, \text{vec}(A_Q^T)]
\]

\[
w_k(j) = [w^1_k(j), w^2_k(j), \ldots, w^Q_k(j)]^T
\]

we can now express the \( j \)th LDC codeword of the \( k \)th user in vector form as

\[
x_k(j) = \text{vec}(W_k(j))^T = G_{vec}w_k(j)
\]  \hspace{1cm} (3)

where the vectors \( x_k(j) \) and \( w_k(j) \) and the LDC generator matrix \( G_{vec} \) are of sizes \( TN_k \times 1 \), \( Q \times 1 \), and \( TN_k \times Q \), respectively. The LDC codeword \( x_k(j) \) is then repetition encoded into symbol sequence \( s_k(j) = \text{rep}(x_k(j), N_c) \), where \( \text{rep} \{ \cdot \} \) 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, which causes bandwidth expansion (signal spreading in the frequency domain). The spread sequence \( s_k(j) \) is then scrambled (randomized) by a scrambling code unique to each user to form the transmitted symbol sequence \( a_k(j) = C_k(j)s_k(j) \), where \( C_k(j) \in \{ -1, 1 \}^{N_cTN_k \times N_cTN_k} \) 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 \( a_k(j) \) is transmitted over a MIMO channel via multiple transmit antennas. For simplicity, we only consider a flat-fading channel here. The received signal is the sum of \( K \) users’ signals plus the additive white complex Gaussian noise. After descrambling and despreading, the received signal due to the \( k \)th user’s \( j \)th transmitted sequence can be written in a vector form as

\[
y_k(j) = N_cH_k(j)x_k(j) = N_cH_k(j)G_{vec}w_k(j)
\]

Denoting the channel gain of the path between the \( m \)th transmit antenna and the \( n \)th receive antenna for the \( t \)th channel use of \( W_k(j) \) as \( h_{m,n}^{k,t} \), the MIMO channel matrix corresponding to the \( k \)th user’s \( j \)th symbol can be expressed as

\[
H_k(j) = \begin{bmatrix}
H_k^{(1)}(j) & \cdots & 0 \\
\vdots & \ddots & \vdots \\
0 & \cdots & H_k^{(T)}(j)
\end{bmatrix}
\]  \hspace{1cm} (4)

where

\[
H_k^{(l)}(j) = \begin{bmatrix}
h_{1,j}^{k,l}(j) & \cdots & h_{N_c,j}^{k,l}(j) \\
\vdots & \ddots & \vdots \\
h_{N_c,j}^{k,l}(j) & \cdots & h_{N_c,j}^{k,l}(j)
\end{bmatrix}
\]

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

\[
r(j) = \sum_{k=1}^{K} y_k(j) + n(j) = N_cH_k(j)G_{vec}w_k(j) + n(j)
\]

\[
+ \sum_{s=1}^{K-1} (N_c - \tau_k + \tau_s)H_s(j)G_{vec}w_s(j)
\]

\[
+ \sum_{s=k+1}^{K-1} (\tau_k - \tau_s)H_s(j+1)G_{vec}w_s(j+1)
\]

\[
+ \sum_{s=k+1}^{K} (N_c - \tau_s + \tau_k)H_s(j)G_{vec}w_s(j)
\]

\[
+ \sum_{s=k+1}^{K} (\tau_s - \tau_k)H_s(j-1)G_{vec}w_s(j-1)
\]

\[
+ \sum_{s=k+1}^{K} (N_c - \tau_s + \tau_k)H_s(j)G_{vec}w_s(j)
\]

\[
\text{MAI}
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where \( n(j) \in \mathbb{C}^{TN_k} \) is a vector of independent identically distributed complex Gaussian noise samples with zero mean and variance matrix \( N_0 \), i.e., \( n(j) \sim \mathcal{C}N(0, N_0I) \).

III. LDCs

In this section, we discuss two classes of LDCs that are well suited for the DS-CDMA system with orthogonal signaling under study.

A. Rectangular U-LDC

As an extension of rate-one square LDC matrices of size \( N_c \times N_c \) defined in [13, eq. (31)], a class of algebraically designed rate-one rectangular LDCs of arbitrary size \( T \times N_t \) called uniform LDCs (U-LDCs), have been reported in [22]. They are particularly suited to the DS-CDMA application under investigation due to their flexible choice of space-time dimensions.

1) Case of \( T \leq N_t \): The \( T \times N_t \) LDC dispersion matrices are defined as

\[
A_{N_t(k-1)+l} = \frac{1}{\sqrt{T}}D^{k-1}\Gamma l^{-1}
\]  \hspace{1cm} (6)

where \( k = 1, \ldots, T \), \( l = 1, \ldots, N_t \), \( D = \text{diag}(1, e^{i(2\pi/T)}, \ldots, e^{i(2\pi(T-1)/T)}), \Pi = \begin{bmatrix} 0_{T \times (N_t-1)} & 1 \\ I_{N_t-1} & 0_{(N_t-1) \times 1} \end{bmatrix} \), and \( \Gamma = [I_T \ 0_{T \times (N_t-T)}] \), where \( \text{diag}(a) \) denotes a diagonal matrix with vector \( a \) on the main diagonal, \( i = \sqrt{-1} \), \( D \) is of size \( T \times T \), \( \Pi \) is of size \( N_t \times N_t \), \( \Gamma \) is of size \( T \times N_t \), \( I_X \) denotes an identity matrix of size \( X \times X \), and \( 0_{X \times Y} \) denotes an all-zero matrix of size \( X \times Y \).
2) Case of \( T > N_t \): The \( T \times N_t \) LDC dispersion matrices are defined as

\[
A_{N_t(k-1)+l} = \frac{1}{\sqrt{N_t}} \Pi^{k-1} \Gamma^{l-1}
\]

where \( k = 1, \ldots, T \), \( l = 1, \ldots, N_t \), \( \Pi = \text{diag}(1, e^{i(2\pi/N_t)}, \ldots, e^{i(2\pi(N_t-1)/N_t)}) \), and \( \Gamma = \begin{bmatrix} I_{N_t} & 0_{(T-N_t) \times N_t} \\ 0_{(T-N_t) \times N_t} & I_{N_t} \end{bmatrix} \), where \( \Pi \) is of size \( N_t \times N_t \), \( \Pi \), and \( \Gamma \) are defined as \( T \times T \), and \( \Gamma \) is of size \( T \times N_t \).

\[ \text{B. FDFR LDC} \]

In general, space-time LDCs [13] do not necessarily reach any guaranteed space diversity order. LDC was applied to space-time CDMA systems in [18]; however, the selected LDC (see [13, eq. (31)]) does not offer full diversity. Damen et al. [13] proposed a class of high-rate full-transmit diversity-based space-time CDMA systems in [23]. However, they actually use space codes instead of space-time codes. Note that in [23], \( N_t \) source data symbols are transformed using a signal space diversity rotation to obtain \( N_t \) coded symbols, and each of the coded symbols is spread through a signature sequence. Although these spread symbols are over time, the signatures are not spread over multiple data symbols. The approach in [23] actually limits the potential to exploit time diversity in signal levels.

To obtain full diversity over multiple-symbol time channel uses, we adopt the full-diversity full-rate (FDFR) complex-field code proposed in [24] and [25] in this work. It is a class of full-space-diversity LDCs with equal real and imaginary dispersion matrices in each source data symbol. The generator matrix of FDFR codes is calculated as

\[
G_{\text{vec}} = \left[ (P_1D_\beta) \otimes [\theta_1]^T \ldots (P_{N_t}D_\beta) \otimes [\theta_{N_t}]^T \right]^T
\]

where \( [\theta_k]^T \) is the \( k \)th row vector of \( \Theta \), and

\[
\Theta = \frac{1}{N_t} [F_{JN_t}]^H \text{diag}(1, \alpha, \ldots, \alpha^{N_t-1})
\]

\[
= [\theta_1, \ldots, \theta_{N_t}]^T
\]

\[
D_\beta = \text{diag}(1, \beta, \ldots, \beta^{N_t-1})
\]

\[
P_m = \begin{bmatrix} 0_{(N_t-m+1) \times (m-1)} & I_{m-1} \\ I_{N_t-m+1} & 0_{(m-1) \times (N_t-m+1)} \end{bmatrix}
\]

where the superscript operator \((\cdot)^H\) denotes the conjugate transpose operation, and \( F_Q \) denotes a discrete Fourier transform matrix of dimension \( Q \times Q \). The constants \( \alpha \) and \( \beta \) for \( N_t = 2^k \) (\( k \) is a positive integer) are specified as

\[
\alpha = \begin{cases} \exp(i\pi/2N_t) & \text{for Design A} \\ \exp(i\pi/N_t) & \text{for Design B} \end{cases}
\]

\[
\beta = \begin{cases} \exp(i\pi/(4N_t)^{3/2}) & \text{for Design A} \\ \alpha^{1/N_t} & \text{for Design B}. \end{cases}
\]

In [24], FDFR codes of size \( N_t \times N_t \) were proposed as space-time codes; FDFR codes of size \( JN_t \times N_t \) (\( J \) is an integer number) were proposed as space-frequency codes and space-Doppler codes. In this paper, we consider FDFR codes of size \( JN_t \times N_t \) as a class of full-space-diversity rectangular space-time LDCs.

Defining the LDC coding rate as \( R_{\text{LDC}} = Q/(N_tT) \) [20], one can see that both U-LDC and FDFR codes mentioned above have a rate of one \( (Q = N_tT) \). This is in contrast to all orthogonal space-time codes, for which the rate is less than one (even the so-called full-rate Alamouti code [11] has a rate of 1/2 by this definition).

\[ \text{IV. JOINT LDC DECODING AND Q-ARY SYMBOL DEMODULATION} \]

The proposed iterative detection and decoding scheme is illustrated in Fig. 2. The soft metric \( \lambda(u^k_{n,l};O) \) from the inner soft-input–soft-output (SISO) block is deinterleaved to \( \lambda(u^k_{n,l};I) \). The \( k \)th user’s Log-MAP decoder computes an \textit{a posteriori} log-likelihood ratio (LLR) of each information bit \( \lambda(b^k_{n,l};O) \) and each code bit \( \lambda(u^k_{n,l};O) \) based on the soft input \( \lambda(u^k_{n,l};I) \) 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 MUD in the inner SISO block in 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 \( u^k_{n,l} \) from the received signal to enable soft-input channel decoding. To this end, we propose an integrated MUD, LDC decoding, and symbol demodulation scheme, which will be described next. To simplify the notation, we sometimes suppress the index \( j \) from \( x_k(j), w_k(j), H_k(j), \text{etc.} \), whenever no ambiguity arises.

A. Single-User Approach

Let \( r_k \) denote the delay-aligned version of the received vector due to the transmission of the \( k \)th user’s \( j \)th symbol. The single-user detection (SUD) approach is based on the assumption that different users’ scrambling codes are (nearly) orthogonal to each other (their cross correlations are approximately zero) and their autocorrelation approximates the delta function. The received vector corresponding to the \( k \)th user’s \( j \)th symbol after descrambling and despreading can be expressed as

\[
z_k = N_e H_k x_k + v_k = N_e H_k G_{\text{vec}} w_k + v_k
\]

where \( z_k \) is the descrambled and despread version of \( r_k \), and

\( v_k \) denotes the combined MAI and noise, which is a complex Gaussian random vector, i.e., \( v_k \sim CN(0, N_e I) \) [26], where \( N_e \) is the combined MAI and noise variance.

The Walsh codeword \( w_k \) (or equivalently the \( j \)th Walsh symbol for user \( k \)) can be estimated by a linear MMSE algorithm, i.e.,

\[
w_k = \Phi^H z_k = \Phi^H (N_e H_k G_{\text{vec}} w_k + v_k) = U w_k + \xi_k
\]
where the matrix $\Phi$ is designed to minimize $\mathbb{E}[\|\hat{w}_k - w_k\|^2]$, leading to the solution $\Phi = R^{-1}P$, where

$$R = E[zz^H]_{\gamma}$$

$$= E\left( (N_c H_k G_{\text{vec}} w_k + v_k) (N_c w_k^H G_{\text{vec}}^H H_k^H + v_k^H) \right)$$

$$= N_c^2 H_k H_k^H + N_c I$$

Equations (13) and (14) are derived utilizing the fact that for Walsh codewords, $E[w_k w_k^H] = I_n$, and $G_{\text{vec}}$ is a unitary matrix for both U-LDC and FDFR codes (the proof in [22], and a proof can be similarly derived for FDFR). Therefore, $G_{\text{vec}} G_{\text{vec}}^H = I_n$. The noise term $\xi_k$ is $\Phi N_k H_k$, where the linear transformation of a Gaussian random vector, with zero mean and covariance $E[\xi_k^H \xi_k] = N_c \Phi^2 \Phi$. The probability density function of the MMSE filter output $\hat{w}_k$, conditioned on that the $m$th Walsh symbol is transmitted, can be expressed as

$$f(\hat{w}_k|w_m) = \frac{1}{\pi^Q \det(N_c \Phi^2 \Phi)} \exp \left[ -\|\Phi^2 \Phi \hat{w}_k - Pw_m\|^2 / N_v \right].$$

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

$$\lambda(u_{n,l}^k; O) = \ln \frac{\sum_{m:u_{n,l}^k=+1} f(w_k|w_m)}{\sum_{m:u_{n,l}^k=-1} f(w_k|w_m)}$$

$$\approx \ln \frac{\max_{m:u_{n,l}^k=+1} f(w_k|w_m)}{\max_{m:u_{n,l}^k=-1} f(w_k|w_m)}$$

$$\approx \ln \frac{\max_{m:u_{n,l}^k=+1} \exp \left( -\|\Phi^2 \Phi \hat{w}_k - Pw_m\|^2 / N_v \right)}{\max_{m:u_{n,l}^k=-1} \exp \left( -\|\Phi^2 \Phi \hat{w}_k - Pw_m\|^2 / N_v \right)}$$

$$= \frac{1}{N_v} \text{Re} \left\{ \left[ 2(\Phi^2 \Phi w^+) \mathcal{N}_k - \|Pw^+\|^2 \right] \right\}$$

$$- \left[ 2(\Phi^2 \Phi w^-) \mathcal{N}_k - \|Pw^-\|^2 \right] \right\}$$

where $m: v_{n,l}^k = \pm 1$ denotes the set of Walsh codewords $\{w_m\}$ that correspond to the code bit $u_{n,l}^k = \pm 1$, $w^+$ denotes the Walsh codeword $w_m$ that corresponds to $\max_{m:u_{n,l}^k=-1} f(\hat{w}_k|w_m)$, and $w^-$ denotes the Walsh codeword $w_m$ that corresponds to $\max_{m:u_{n,l}^k=+1} f(\hat{w}_k|w_m)$. In the case where $Q = 8$, the $j$th user's $j$th Walsh codeword $w_k(j)$ corresponds to three coded and interleaved bits: $u_{1,j}^k$, $u_{2,j}^k$, and $u_{3,j}^k$. We know from Table I that $u_{0,j}^k = +1$ holds for $m = 0, 1, 2, 3$ and $u_{0,j}^k = -1$ holds for $m = 4, 5, 6, 7$. According to (15)

$$\lambda(u_{n,l}^k; O) \approx \max \left\{ z_k(0), z_k(1), z_k(2), z_k(3) \right\}$$

$$- \max \left\{ z_k(4), z_k(5), z_k(6), z_k(7) \right\}$$

$$= \frac{(1/N_v) \text{Re} \left\{ 2(\Phi^2 \Phi Pw_m)^H \mathcal{N}_k - \|Pw_m\|^2 \right\} \right\}$$

Similarly

$$\lambda(u_{1,l}^k; O) \approx \max \left\{ z_k(0), z_k(1), z_k(2), z_k(3) \right\}$$

$$- \max \left\{ z_k(4), z_k(5), z_k(6), z_k(7) \right\}$$

$$\lambda(u_{2,l}^k; O) \approx \max \left\{ z_k(0), z_k(2), z_k(4), z_k(6) \right\}$$

$$- \max \left\{ z_k(1), z_k(3), z_k(5), z_k(7) \right\}.$$
introduce an MUD technique to increase the capacity of interference-limited CDMA systems. Among different MUD techniques, the multistage parallel interference cancellation (PIC) scheme [26] is known to be simple and effective for mitigation of MAI in long-code DS-CDMA systems. In what follows, we shall explain how this PIC-based MUD technique can be incorporated into the demodulation and decoding process to mitigate the effect of MAI.

B. Multiuser Approach

Once the transmitted signals are estimated for all the users at the previous iteration, interference can be removed by subtracting the estimated signals of the interfering users from the received signal \( r \). To form a new signal vector \( r'_k \) for demodulating the signal transmitted from user \( k \). The descrambled and despread signal corresponding to the \( k \)-th user’s \( j \)-th symbol after interference cancellation can now be expressed as

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

where \( \hat{\mathbf{w}}_s(j) \) is an estimate of \( \mathbf{w}_s(j) \) using hard decisions, and \( \mathbf{v}'_k(j) \) denotes the combined cancellation residual and noise. 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. The symbol index \( j \) is sometimes omitted for simplicity. With the interference cancellation technique, \( \mathbf{v}'_k(j) \) contains much less MAI compared with \( \mathbf{v}_k \) in (11), leading to a significant performance improvement. In case of perfect interference cancellation, the cancellation residual vanishes, and \( \mathbf{v}'_k \sim \mathcal{CN}(0, N_c \mathbf{N}_0) \). As will become apparent in Section V, the assumption of perfect cancellation can be approached by proper design of the iterative receiver and by proper choice of full-diversity LDCs. The rest of LDC decoding and derivation of LLR values for \( u_n^k \) is the same as described in the previous section, except that \( \mathbf{z}_k \) is replaced by \( \mathbf{r}'_k \), and \( N_c \) is replaced by \( N_c \mathbf{N}_0 \) in (12)-(15).

The conventional interference cancellation is subject to performance degradation due to incorrect decisions on interference that are subtracted from the received signal. To prevent error propagation from the decision feedback, soft interference cancellation was proposed, e.g., in [27], for uncoded \( Q \)-ary DS-CDMA systems. The rationale is that the cancellation with hard decisions tends to propagate errors and increase the interference with incorrect decision feedback, while with soft cancellation, an erroneously estimated symbol usually has a small LLR and does not make much contribution to the feedback, and therefore, the error propagation problem is alleviated. In our case, the interference cancellation scheme using soft symbol estimates can be reformed as (16) with \( \hat{\mathbf{w}}_s(j) \) replaced by \( \overline{\mathbf{w}}_s(j) \) (the soft estimate of \( \mathbf{w}_s(j) \)).

This MUD-based iterative scheme is shown in Fig. 2. It uses the soft information of \( \mathbf{x}_k \), denoted as \( \overline{\mathbf{x}}_k \), for interference cancellation to reduce the error propagation. To this end, we compute \( \mathbf{w}_k = [w^0_k(j), w^2_k(j), \ldots, w^Q(j)]^T \), the soft estimate of the codeword \( \mathbf{w}_k(j) \), from its LLR \( \lambda(\mathbf{w}_k(j)) \), which is derived by feeding \( \lambda(u^k_{n,l}) = \Pi\{\lambda(u^k_{n,l}; O) \} \) into a soft modulator. The soft estimate \( \overline{\mathbf{w}}_k(j) \) is then LDC encoded to produce the LDC codeword \( \overline{\mathbf{x}}_k(j) \).

When soft information is to be used for interference cancellation, a serially concatenated system would be rather different from the uncoded or nonconcatenated systems in that the soft values are not directly available for all the inner code bits from the outer decoder. In particular, in our case, only the soft information can be extracted for the systematic bits of the Walsh codewords from a SISO channel decoder. To tackle this problem, we design a soft modulator to derive the soft estimates for parity bits, which will be explained next.

In the case where \( Q = 8 \), the code bits to the Walsh codeword (Walsh symbol) mapping rule is given in Table I. The three systematic bits are \( w^1_k(j), w^2_k(j), \) and \( w^4_k(j) \), where \( w^1_k(j) \) denotes the \( j \)-th bit of the codeword. The columns corresponding to the systematic bits are highlighted in the table. For ease of understanding, we use \( Q = 8 \) as an example. However, the extension of the proposed algorithms to other values of \( Q \) is straightforward. In Table I, we can see that the parity bits are formed by systematic bits \( w^3_k(j), w^5_k(j), \) and \( w^7_k(j) \) as

\[
\begin{align*}
\lambda(w^1_k(j)) &= \lambda(u^1_{n,l}) \\
\lambda(w^3_k(j)) &= \lambda(u^3_{n,l}) \\
\lambda(w^5_k(j)) &= \lambda(u^5_{n,l}) \\
\lambda(w^7_k(j)) &= \lambda(u^7_{n,l})
\end{align*}
\]

The LLRs for systematic bits are

\[
\begin{align*}
w^0_k(j) &= +1 \\
w^2_k(j) &= w^1_k(j) + w^3_k(j) \\
w^4_k(j) &= w^1_k(j) + w^5_k(j) \\
w^6_k(j) &= w^2_k(j) + w^4_k(j) \\
w^7_k(j) &= w^3_k(j) + w^5_k(j) + w^7_k(j).
\end{align*}
\]

Considering the fact that the interleaver breaks the memory of the convolutional encoding process, the bits \( u^1_{n,l}, u^3_{n,l}, \) and \( u^5_{n,l} \) can be modeled as statistically independent random variables. Additionally, assuming that they are independently
conditioned on the received signal, the LLRs for parity bits can thus be computed according to [28] by
\[
\lambda (w_k^p(j)) = \lambda (w_k^p(j) \oplus w_k^q(j)) = 2\text{arc tanh}\left\{ \text{tanh}\left( \frac{\lambda (w_k^p(j))/2}{2} \right) \cdot \text{tanh}\left( \frac{\lambda (w_k^q(j))/2}{2} \right) \right\}.
\]
\[
\lambda (w_k^q(j)) = \lambda (w_k^p(j) \oplus w_k^p(j)) = 2\text{arc tanh}\left\{ \text{tanh}\left( \frac{\lambda (w_k^p(j))/2}{2} \right) \cdot \text{tanh}\left( \frac{\lambda (w_k^q(j))/2}{2} \right) \right\}.
\]
\[
\lambda (w_k^p(j)) = \lambda (w_k^p(j) \oplus w_k^q(j) \oplus w_k^r(j)) = 2\text{arc tanh}\left\{ \prod_{n=0}^{2} \text{tanh}\left( \frac{\lambda (w_n^k(j))/2}{2} \right) \right\}.
\]

Finally, the soft estimate (expected value given the received observation) for each bit of the Walsh codeword \(w_k^p(j)\) is computed based on \(\lambda(w_k^p(j))\) as
\[
\hat{w}_k^p(j) = E\{w_k^p(j)|z_k\} = (+1) \times P\{w_k^p(j) = +1|z_k\} + (-1) \times P\{w_k^p(j) = -1|z_k\}
\]
\[
= (+1) \frac{e^{\lambda(w_k^p(j))}}{1 + e^{\lambda(w_k^p(j))}} + (-1) \frac{e^{-\lambda(w_k^p(j))}}{1 + e^{-\lambda(w_k^p(j))}} = \text{tanh}\left( \frac{\lambda (w_k^p(j))/2}{2} \right).
\]

V. NUMERICAL RESULTS

In the simulations, we employ a rate \(R_c = 1/3\) convolutional code with constraint length \(L_c = 5\) and generator polynomials \((25, 33, 37)\) in octal form for all the users, unless otherwise stated. Each block of three interleaved bits from each user is then converted into one of \(Q = 8\) Walsh symbols, which is subsequently encoded with an LDC. In this paper, we use the U-LDC expressed by \((7)\) and the FDFR code (design A) expressed by \((8)\)–\((10)\). The parameter setting is chosen to be \(N_t = 2, N_r = 2\) and \(T = 4\). In this case, an 8-bit Walsh codeword is dispersed into two transmit antennas, each of which accommodates four LDC symbols (\(T = 4\) channel uses before spreading). Each LDC symbol is spread (repetition encoded) to \(N_c = 8\) symbols. The effective spreading factor of the system is given as the reciprocal of the overall code rate divided by \(N_t\), i.e., \(N_c Q/(N_t R_c \log_2 Q) = 32\) LDC symbols/information bit/antenna. Hence, the spectral efficiency of a single link is equal to 1/32 times the spectral efficiency of an uncoded unspread system. The number of users is chosen to be \(K = 18\), which represents a fairly heavily loaded system, as a \(K = 32\)-user system would be a fully loaded system. The long scrambling codes \(C_k\) are randomly generated, and all the transmit antennas of a specific user are assigned the same scrambling code. The noise variance \(N_0\) and \(C_k\), as well as path delays \(\tau_1, \tau_2, \ldots, \tau_K\), are assumed to be known to the receiver, and the different paths of the MIMO channels for the same user have the same delay. We assume uplink asynchronous transmission, and the delays are therefore different for different users.

We compare the performance of the proposed LDC iterative scheme with that of the soft demodulation and decoding algorithm [8] for the Q-ary orthogonally modulated DS-CDMA system without LDCs in a \(2 \times 2\) or \(2 \times 4\) V-BLAST system. In this case, information-bearing signals are divided into multiple substreams, each encoded and modulated independently. Unlike in the LDC systems, each user’s two transmit antennas are assigned with different scrambling codes and simultaneously transmit the data. The employed convolution code and spreading factor are the same as in the LDC system. To transmit one Walsh symbol, the V-BLAST 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 V-BLAST system, the two systems have the same data rate.

The channel gain \(h_{m,n}^{k,\tau}\) is a complex circular Gaussian process with autocorrelation function
\[
E[h_{m,n}^{k,\tau} h_{m,n}^{k,\tau+\tau}^*] = P_{m,n} J_0(2\pi f_D \tau),
\]
where \(f_D\) is the maximum Doppler frequency, \(J_0(x)\) is the zeroth-order Bessel function of the first kind, and \(P_{m,n}\) is the average power of \(h_{m,n}^{k,\tau}\). The amplitude of \(h_{m,n}^{k,\tau}\) 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 \(P_k = \sum_{m,n} E[|h_{m,n}^{k,\tau}|^2]\), 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 \(|h_{m,n}^{k,\tau}|^2\) may vary from one user to another. Perfect knowledge of the channel state information (CSI) is assumed in our simulations. In Figs. 3–7, we assume that channel gains remain
constant during the transmission of one LDC codeword, i.e.,
\[ h_{m,n}^{k,t}(j) = h_{m,n}^{k,t+1}(j) = h_{m,n}^{k,t+2}(j) = h_{m,n}^{k,t+3}(j), \]
and vary from one codeword to another. The normalized Doppler frequency is assumed to be \( f_D T_s = 0.01 \), where \( T_s \) is the symbol (LDC codeword) duration. During each Monte Carlo run, the block size is set to 1526 information bits followed by 4 tail bits to terminate the trellis. The coded bits are passed through a random interleaver. The simulation results are averaged over random fading, noise, delays, and scrambling codes with a minimum of 50 blocks of data transmitted and at least 100 bit errors generated.

In Fig. 3, we compare the performance of the iterative MUD algorithms at different iterations. The PIC-based soft demodulation scheme [8] is used for the V-BLAST system. Here, \( E_b \) is defined as the received bit energy and is normalized with the number of receive antennas. The single-user scheme is used in the beginning of the iterative process to obtain an initial estimate of the data, which is needed for MUD at subsequent iterations. We notice that the V-BLAST system has to iterate three times before reaching convergence (excluding the SUD stage), while the LDC system only needs to iterate two times to converge. The use of LDC leads to faster convergence for the iterative receiver, as well as superior performance, compared with the V-BLAST system. By comparison, the performance of the U-LDC is not as good as that of the FDFR codes, especially at high SNRs (the curves for the U-LDC are not included to conserve space). To fully exploit the space diversity, FDFR codes should be used.

Their convergence behavior is further examined at \( E_b/N_0 = 7 \) dB in Fig. 4 using the extrinsic information transfer (EXIT) chart, which traces the evolution of the mutual information \( I_i^M/I_o^M \in [0, 1] \) between input/output LLR and \( u_{k,n}^i \) for the multiuser detector and the mutual information \( I_D^M/I_o^M \) between input/output LLR and \( u_{k,n}^i \) for the Log-MAP decoder. See [29] for a detailed discussion of this analytical method and to [30]...
and [31] for its application to the analysis of iterative schemes in CDMA and MIMO systems. It should be noted that in our case, $I^M > 0$ at the starting point. This is due to the SUD stage applied in the beginning of the iterative process to yield an initial estimate of data; therefore, the $a$ priori information is not zero when the MUD process starts. The output LLR of the detector $I^M_o$ is forwarded to the decoder as input, i.e., $I^D = I^M_o$; the output LLR of the decoder $I^D_o$ is fed back to the detector, i.e., $I^M_o = I^D_o$ and so on. As indicated by the transfer curves in Fig. 4, the output LLR $I^M_o$ becomes more reliable (its value increases) as the input LLR $I^M$ becomes more reliable in the detector. The iterative detection and decoding process is depicted by a staircase trace between the transfer curves of the detector and decoder. The trace shows that two (three) iterations of detection/decoding are needed for the LDC (V-BLAST) system to converge (reach the maximum $I^D$). This is in close agreement with the results presented in Fig. 3. The initial value of $I^M_o$ is obtained through simulations for both LDC and V-BLAST systems.

In Figs. 5–7, we compare the two systems with different channel codes and diversity orders. Two convolutional codes are tested: 1) a weak code with rate $R_c = 1/2$, constraint length $L_c = 3$, and generator polynomials (5, 7) and 2) a strong code with rate $R_c = 1/3$, constraint length $L_c = 9$, and generator polynomials (575, 623, 727), which is an optimum distance spectrum code [32]. Comparing Fig. 5 with Fig. 6, one can see that the advantages of LDCs over V-BLAST become smaller when a stronger code is used. For example, with the (5, 7) code, the LDC system outperforms the V-BLAST system by 1.0 dB upon reaching convergence at a target bit error rate (BER) of $10^{-4}$, whereas the difference is only 0.5 dB when the (575, 623, 727) code is used. Obviously, it is more advantageous to apply LDC with a weak code. The single-user bounds for the FDFR-LDC-coded systems are shown in Figs. 3 and 5. They are obtained by the proposed scheme in a single-user environment, no interference mitigation is needed in this case, and they give lower bounds on the best achievable performance by applying the 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.

Fig. 7 shows the performance comparison between the two systems with the (5, 7) code in a $2 \times 4$ antenna setup. Comparing with Fig. 5, it is obvious that the V-BLAST system converges faster, and the performance gap between the two systems becomes smaller as the number of receive antennas (spatial diversity order) increases.

Note that the proposed system does not require a constant channel during the transmission of one LDC codeword, channel gains can vary from one time slot to another, i.e., $h_{m,n}(j) \neq h_{m,n}(j) \neq h_{m,n}(j) \neq h_{m,n}(j)$. This is in contrast with previous work on LDC for CDMA systems, which were designed for static channel over the whole LDC codeword. In Fig. 8, 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 redefine 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 $E_b/N_0 = 5$ dB. Slow power control is assumed so that the average received power is equal for all users. The BER curve is plotted at the fourth iteration when the system reaches convergence. Fig. 8 shows that the system performance improves as the Doppler frequency increases, which clearly indicates the time diversity obtained by the LDCs, and the diversity gain is more obvious by applying FDFR LDC rather than U-LDC. 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 $f_D T = 0.01$, and code-level channel stationarity is assumed. It was shown in [8] that for non-LDC systems, the performance degrades as the normalized Doppler frequency increases.

The complexity of the iterative MUD approach for the LDC-coded system and the V-BLAST system is compared in Table II, which shows the required number of complex multiplications/divisions and additions/subtractions for one user's one symbol estimate corresponding to the calculation of LLRs for $\log_2 Q$ code bits. One can see in the table that the proposed LDC system increases the complexity from $O(Q^3)$ to $O(Q^2)$ compared to the V-BLAST system, mainly due to the matrix inverse operation [see (15)] at symbol rate in the LDC decoding process. However, the complexity increase is partly compensated by the faster convergence achieved by the LDC system. The time diversity shown in Fig. 8 also justifies the use of LDCs in fast-fading channels.

VI. CONCLUSION

Iterative detection and decoding for an orthogonally modulated and LDC-coded MIMO DS-CDMA system has been investigated in this paper. We have proposed an integrated design of MUD, LDC decoding, symbol demodulation, and symbol-to-LLR mapping to reduce the complexity of the iterative receiver. Numerical results show that the choice of LDCs is important for system performance, and the use of FDFR LDCs leads to high transmission bandwidth efficiency and significantly
improved BER performance. Under the assumption of perfect CSI, it enables the system to approach the single-user bound even in heavily loaded systems. Compared with the V-BLAST system with the same transmission rate, the proposed LDC system shows superior performance and faster convergence. However, we have observed that the advantages of applying LDC become smaller when a strong channel code is used or the diversity order increases. Furthermore, by exploiting time diversity, LDCs also provide us with a powerful means to combat impairment caused by rapid-fading channels. The extension of the current work to frequency-selective channels by incorporating the orthogonal frequency-division multiplexing technique or using the turbo equalization method is a topic for future research.

REFERENCES

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