Space-time Linear Dispersion Using Coordinate Interleaving

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***Abstract*— This paper proposes a general coordinate-interleaving method for block-based space-time codes or linear dispersion codes, called space-time coordinate interleaving linear dispersion codes (ST-CILDC), which enables not only symbol-level diversity but also coordinate-level diversity for high rate block-based space-time code design. This paper analyzes the upper bound diversity order and provides the analysis results of the upper bound statistical diversity order and average diversity order for ST-CILDC systems. Compared with conventional ST-LDC systems, ST-CILDC systems may show either almost doubled average diversity order or extra coding advantage in time varying channels. With trivial extra complexity over ST-LDC systems, ST-CILDC systems maintain the diversity performance in quasi-static block fading channels, and significantly improve the diversity performance in rapid fading channels.**

I. Introduction

To support high reliability of space-time multiple input multiple output (MIMO) transmission, space-time coding (STC) may be applied to improve system performance and achieve high capacity potential. Space-time trellis codes [1] have great diversity and coding gain but exponential decoding complexity, which motivates the design of low complexity STC. Due to their attractive complexity, a number of block-based STC have been proposed [2], [3]. Recently, Hassibi and Hochwald have constructed a class of high-rate block-based STC known as linear dispersion codes (LDC) [4], which support arbitrary numbers of transmit and receive antenna channels. We treat LDC as a general framework of complex space-time block code design.

A problem in most existing design criteria of block-based space-time codes, including LDC, is that they do not efficiently exploit additional diversity potential in the real and image parts of coordinates of source data constellation symbols. A technique to utilize the diversity potential of real and image parts of coordinates is called coordinate interleaving or component interleaving (CI), which was first proposed for single stream communications systems [5], [6]. Recently, CI has been applied to multiple antenna systems [7]–[9]. However, current existing approaches to using CI in block-based space-time codes are low-rate designs using orthogonal space-time block codes or their variation [7]–[9].

This paper proposes coordinate-interleaving as a general principle for high-rate block-based space-time code design, i.e., space-time coordinate interleaving linear dispersion codes (ST-CILDC). This paper determines the the upper bound diversity order, statistical diversity order and average diversity order of ST-CILDC. ST-CILDC maintains the same diversity order as conventional ST-LDC. However, ST-CILDC may show either almost doubled average diversity order or extra coding advantage over conventional ST-LDC in time varying channels. Compared with conventional ST-LDC, ST-CILDC maintains the diversity performance in quasi-static block fading channels, and notably improves the diversity performance in rapid fading channels.

The following notation is used: $\mathrm{Re}\left[\cdot\right]$ and $\mathrm{Im}\left[\cdot\right]$ denote the real and image parts, respectively, $(\cdot)^{\mathcal{T}}$ matrix transpose, $(\cdot)^{\mathcal{H}}$ matrix transpose conjugate, $\left[\mathbf{A}\right]_{a,b}$ denote the (a,b) entry (element) of matrix $\mathbf{A}$, $\otimes$ Kronecker product, $\mathbf{I}_K$ denotes $K\times K$ identity matrix, and $C^{m\times n}$ denotes a complex matrix with dimensions $m\times n$.

II. Proposed systems

A. MIMO system model for LDC in time varying channels

In frequency-flat, time non-selective Rayleigh fading channels whose coefficients may vary per channel symbol time slot or channel use, a multi-antenna communication system is assumed with N_T transmit and N_R receive antennas. Assume that an uncorrelated data sequence has been modulated using complex-valued source data symbols chosen from an arbitrary, e.g. D-PSK or D-QAM, constellation. Each LDC codeword of size $T\times N_T$ is transmitted during every T time channel uses from N_T transmit antennas.

1) Component matrices in system equations: We now introduce several component matrices during the k-th space-time LDC codeword transmission.

The received signal vector $\mathbf{x}_{LDC}^{(k)} = \left[\left[\mathbf{x}_{LDC}^{(k,1)}\right]^{\mathcal{T}},...,\left[\mathbf{x}_{LDC}^{(k,T)}\right]^{\mathcal{T}}\right]^{\mathcal{T}}$, where $\mathbf{x}_{LDC}^{(k,t)} \in C^{N_R\times 1}, t = 1...,T$ is the received vector corresponding to the t-th row of the k-th LDC codeword, $\mathbf{S}_{LDC}^{(k)}$.

The system channel matrix is

$$\mathbf{H}_{LDC}^{(k)} = \begin{bmatrix} \mathbf{H}_{LDC}^{(k,1)} & \cdots & \mathbf{0} \\ \vdots & \ddots & \vdots \\ \mathbf{0} & \cdots & \mathbf{H}_{LDC}^{(k,T)} \end{bmatrix}$$

where $\mathbf{H}_{LDC}^{(k,t)} \in C^{N_R \times N_T}, t = 1...,T$ with entries $\left[\mathbf{H}_{LDC}^{(k,t)}\right]_{m,n} = h_{m,n}^{(k,t)}, \ m = 1,...,N_T, \ n = 1,...,N_R$, is a complex Gaussian MIMO channel matrix with zero-mean, unit variance entries corresponding to the t-th row of the k-th LDC codeword, $\mathbf{S}_{LDC}^{(k)}$, and $\mathbf{0}$ denotes a zero matrix of size $N_R \times N_T$.

The complex Gaussian noise vector is $\mathbf{v}_{LDC}^{(k)} = \left[\left[\mathbf{v}_{LDC}^{(k,1)}\right]^{\mathcal{T}}, ..., \left[\mathbf{v}_{LDC}^{(k,T)}\right]^{\mathcal{T}}\right]^{\mathcal{T}}$, where $\mathbf{v}_{LDC}^{(k,t)} \in C^{N_R \times 1}, t = 1...,T$ is a complex Gaussian noise vector with zero mean, unit variance entries corresponding to the t-th row of the k-th LDC codeword, $\mathbf{S}_{LDC}^{(k)}$.

The LDC encoded complex symbol vector $\mathbf{s}_{LDC}^{(k)}$ corresponds to the k-th LDC codeword, $\mathbf{S}_{LDC}^{(k)}$, where

$$\mathbf{s}_{LDC}^{(k)} = vec(\left[\mathbf{S}_{LDC}^{(k)}\right]^{\mathcal{T}}). \tag{1}$$

2) System model equation: The system equation for the transmission of the k-th LDC matrix codeword is expressed as

$$\mathbf{x}_{LDC}^{(k)} = \sqrt{\frac{\rho}{N_T}} \mathbf{H}_{LDC}^{(k)} \mathbf{s}_{LDC}^{(k)} + \mathbf{v}_{LDC}^{(k)} \tag{2}$$

where ρ is the signal-to-noise ratio (SNR) at each receive antenna, and independent of N_T.

B. Procedure of space-time inter-LDC coordinate interleaving

We propose a new space-time LDC encoding procedure using inter-LDC CI (ILDC-CI), called space-time coordinate interleaving linear dispersion codes (ST-CILDC) as follows: Consider a pair of source data symbol vectors $\mathbf{s}^{(1)} = \left[s_1^{(1)}, ..., s_Q^{(1)}\right]^{\mathcal{T}}$ and $\mathbf{s}^{(2)} = \left[s_1^{(2)}, ..., s_Q^{(2)}\right]^{\mathcal{T}}$ with the same number, Q of source data symbol symbols, where $s_q^{(i)} = \mathrm{Re}\left(s_q^{(i)}\right) + j\mathrm{Im}\left(s_q^{(i)}\right)$, $i = 1, 2$, and $q = 1, ..., Q$. The transmitter first coordinate-interleaves $\mathbf{s}^{(1)}$ and $\mathbf{s}^{(2)}$ into $\mathbf{s}^{CI(1)} = \left[s_1^{CI(1)}, ..., s_Q^{CI(1)}\right]^{\mathcal{T}}$ and $\mathbf{s}^{CI(2)} = \left[s_1^{CI(2)}, ..., s_Q^{CI(2)}\right]^{\mathcal{T}}$, where

$$s_q^{CI(1)} = \mathrm{Re}\left(s_q^{(1)}\right) + j\mathrm{Im}\left(s_q^{(2)}\right) \tag{3}$$

and

$$s_q^{CI(2)} = \mathrm{Re}\left(s_q^{(2)}\right) + j\mathrm{Im}\left(s_q^{(1)}\right). \tag{4}$$

Then, $\mathbf{s}^{CI(1)}$ and $\mathbf{s}^{CI(2)}$ are encoded into two LDC codewords of size $T \times N_T$, $\mathbf{S}_{LDC}^{CI(1)}$ and $\mathbf{S}_{LDC}^{CI(2)}$, respectively. Then the transmitter successively sends $\mathbf{S}_{LDC}^{CI(1)}$ and $\mathbf{S}_{LDC}^{CI(2)}$ during two interleaved periods such that channels are less correlated.

We remark that

1) using different permutations, other methods of space-time inter-LDC CI than (3) and (4) are also possible;
2) The LDC encoding matrices for $\mathbf{S}_{LDC}^{CI(1)}$ and $\mathbf{S}_{LDC}^{CI(2)}$ need not be the same.

C. ST-CILDC system structure

The proposed ST-CILDC system structure is shown in Figure 1. The system structure basically consists of three layers : (1) mapping from data bits to constellation points, (2) inter-LDC coordinate interleaving, and (3) LDC coding. Using the proposed layered structure, the only additional complexity compared with a conventional ST-LDC system is the coordinate interleaving operation. Thus, ST-CILDC system is computationally efficient. The motivation of ST-CILDC is to render the fading more independent of each coordinate of the source data signals. Note that due to the superposition effects of signals from multiple transmit antennas at the space-time MIMO receivers, existing LDC designs cannot guarantee fading independence of each coordinate of the source data signals. Compared with ST-LDC, ST-CILDC introduces coordinate fading diversity at the cost of more decoding delay using a pair of LDC codewords of the same size.

III. Diversity analysis

Su and Liu [10] recently analyzed the diversity of space-time modulation over time-correlated Rayleigh fading channels. A modified strategy can be used to investigate the diversity of ST-CILDC systems.

Consider a ST-CILDC block $\mathbf{C}$, which consists of two ST-LDC codewords of size $T \times N_T$, $\mathbf{S}_{LDC}^{(k)}$, where $k = 1, 2$.

The communication model for one ST-CILDC block $\mathbf{C}$ can be rewritten as

$$\mathbf{Y} = \sqrt{\frac{\rho}{N_T}} \mathbf{M}\mathbf{H} + \mathbf{Z}, \tag{5}$$

where

1) the noise vector is $\mathbf{Z}$,
2) the received signal vector $\mathbf{Y} = \left[\left[\mathbf{Y}^{(1)}\right]^{\mathcal{T}}, \left[\mathbf{Y}^{(2)}\right]^{\mathcal{T}}\right]^{\mathcal{T}}$, where $\mathbf{Y}^{(k)} = \left[\mathbf{Y}_1^{(k)}, ..., \mathbf{Y}_{N_R}^{(k)}\right]^{\mathcal{T}}$, $\mathbf{Y}_n^{(k)} = \left[\left[\mathbf{x}_{LDC}^{(k,1)}\right]_{n,1}, ..., \left[\mathbf{x}_{LDC}^{(k,T)}\right]_{n,1}\right]^{\mathcal{T}}$, $k = 1, 2$,
3) $\mathbf{M}$ is the channel symbol matrix corresponding to the block $\mathbf{C}$, $\mathbf{M} = diag\left(\mathbf{M}^{(1)}, \mathbf{M}^{(2)}\right)$, where $\mathbf{M}^{(1)}$ and $\mathbf{M}^{(2)}$ are the matrices corresponding to the LDC codeword $\mathbf{S}_{LDC}^{(1)}$ and $\mathbf{S}_{LDC}^{(2)}$, respectively, $\mathbf{M}^{(k)} = \mathbf{I}_{N_R} \otimes diag\left[\mathbf{M}_1^{(k)}, ..., \mathbf{M}_{N_T}^{(k)}\right]$, $\mathbf{M}_m^{(k)} = diag\left(\left[\mathbf{S}_{LDC}^{(k)}\right]_{1,m}, ..., \left[\mathbf{S}_{LDC}^{(k)}\right]_{T,m}\right)$, $k = 1, 2$, and $m = 1, ..., N_T$,
4) the channel vector $\mathbf{H} = \left[\left[\mathbf{H}^{(1)}\right]^{\mathcal{T}}, \left[\mathbf{H}^{(2)}\right]^{\mathcal{T}}\right]^{\mathcal{T}}$, where

$$\mathbf{H}^{(k)} = \left[\mathbf{h}_{(k)1,1}^{\mathcal{T}}, ..., \mathbf{h}_{(k)N_T,1}^{\mathcal{T}}, ..., \mathbf{h}_{(k)1,N_R}^{\mathcal{T}}, ..., \mathbf{h}_{(k)N_T,N_R}^{\mathcal{T}}\right]^{\mathcal{T}},$$

and $\mathbf{h}_{(k)m,n} = \left[h_{m,n}^{(k,1)}, ..., h_{m,n}^{(k,T)}\right]^{\mathcal{T}}$.

A directional pair, denoted as $\mathbf{X} \rightarrow \mathbf{Y}$, means that a system detects $\mathbf{X}$ as $\mathbf{Y}$. Consider the direction pair of matrices $\mathbf{M}$ and

$\tilde{\mathbf{M}}$ corresponding to two different ST-LDC blocks C and $\tilde{C}$. The upper bound pairwise error probability [11] is

$$\Pr\left(\mathbf{M} \to \widetilde{\mathbf{M}}\right) \leqslant \begin{pmatrix} 2r-1 \\ r \end{pmatrix} \left(\prod_{a=1}^{r} \gamma_a\right)^{-1} \left(\frac{\rho}{N_T}\right)^{-r} \quad (6)$$

where r is the rank of $\left(\mathbf{M}-\tilde{\mathbf{M}}\right)\mathbf{R_H}\left(\mathbf{M}-\tilde{\mathbf{M}}\right)^{\mathcal{H}}$, and $\mathbf{R_H} = \mathbb{E}\left\{\mathbf{H}\left[\mathbf{H}\right]^{\mathcal{H}}\right\}$ is the correlation matrix of vector $\mathbf{H}$, $\mathbf{R_H}$ is of size $2N_TN_RT \times 2N_TN_RT$, $\gamma_a, a = 1, ..., r$ are the non-zero eigenvalues of

$$\mathbf{\Lambda} = \left(\mathbf{M}-\tilde{\mathbf{M}}\right)\mathbf{R_H}\left(\mathbf{M}-\tilde{\mathbf{M}}\right)^{\mathcal{H}}.$$

Then the rank and product criteria are

1) Rank criterion: The minimum rank of $\mathbf{\Lambda}$ over all direction pairs of different matrices $\mathbf{M}$ and $\tilde{\mathbf{M}}$ should be as large as possible.
2) Product criterion: the minimum value of the product $\prod_{a=1}^{r} \gamma_a$ over all direction pairs of different $\mathbf{M}$ and $\tilde{\mathbf{M}}$ should be maximized.

To maximize the rank of $\mathbf{\Lambda}$, we need to maximize the ranks of both $\mathbf{R}_H$ and $\left(\mathbf{M}-\widetilde{\mathbf{M}}\right)$. Denote

$$\mathbf{\Omega}^{(k)} = \mathbf{M}^{(k)} - \widetilde{\mathbf{M}^{(k)}},$$

where $k = 1, 2$.

Assume that all the possible $\mathbf{M}^{(k)}$ and $\widetilde{\mathbf{M}^{(k)}}$ are contained in a set $\mathcal{M}^{(k)}$, i.e., $\left\{\mathbf{M}^{(k)}, \widetilde{\mathbf{M}^{(k)}}\right\} \in \mathcal{M}^{(k)}$, where $k = 1, 2$.

Then the diversity order of the ST-CILDC, r_d, is

$$r_d = \min\left\{rank\left(\mathbf{\Lambda}\right), \mathbf{M} \in \mathcal{M}, \widetilde{\mathbf{M}} \in \mathcal{M}, \mathbf{M} \neq \widetilde{\mathbf{M}}\right\}. \quad (7)$$

When $\mathbf{M} \neq \widetilde{\mathbf{M}}$, there are three distinct categories of situations,

1) $\mathbf{M}^{(1)} \neq \widetilde{\mathbf{M}^{(1)}}$ and $\mathbf{M}^{(2)} = \widetilde{\mathbf{M}^{(2)}}$,
2) $\mathbf{M}^{(1)} = \widetilde{\mathbf{M}^{(1)}}$ and $\mathbf{M}^{(2)} \neq \widetilde{\mathbf{M}^{(2)}}$,
3) $\mathbf{M}^{(1)} \neq \widetilde{\mathbf{M}^{(1)}}$ and $\mathbf{M}^{(2)} \neq \widetilde{\mathbf{M}^{(2)}}$.

Note that when $\mathbf{R_H}$ is full rank,

1) in the above Situations (1) and (2), the upper bound of $rank(\mathbf{\Lambda})$ is N_RT,
2) in the above Situation (3), the upper bound of $rank(\mathbf{\Lambda})$ is $2N_RT$.

Thus ST-CILDC does not further increase the diversity order over ST-LDC in terms of the conventional definition (7). However, ST-CILDC does increase r over ST-LDC for the above-mentioned third situation, which is not the conventional diversity order of the STC and may significantly impact system performance. It is necessary to introduce a new concept to quantify this effect as follows,

Definition 1: Statistical diversity order, r_{sd}, is the rank of $\mathbf{\Lambda}$ achieved with a certain probability α, mathematically written as

$$\Pr\left\{\begin{array}{l} rank\left(\mathbf{\Lambda}\right) \geqslant r_{sd}, \\ \mathbf{M} \neq \widetilde{\mathbf{M}}, \\ \left\{\mathbf{M}, \widetilde{\mathbf{M}}\right\} \in \mathcal{M}, \end{array}\right\} = \alpha. \quad (8)$$

Then, we have the following theorem.

Theorem 1: A ST-CILDC is constructed through coordinate interleaving across a pair of component LDC codewords. Both component LDC encoders are able to generate different codewords for different input sequences. The diversity orders of the component LDCs are $r_d^{(1)}$ and $r_d^{(2)}$, respectively. Suppose that $\mathbf{R_H}$ is full rank. The codebook sizes of the two component LDCs are the same value, N_a.

1) *The diversity order of this ST-CILDC, r_d, is $\min\left\{r_d^{(1)}, r_d^{(2)}\right\}$.*
2) *Assuming that all directional pairs $\mathbf{M}$ and $\tilde{\mathbf{M}}$ are equally probable, the statistical diversity order of this ST-CILDC, r_{sd}, is $r_d^{(1)} + r_d^{(2)}$ with probability*

$$\alpha = \frac{\begin{pmatrix} N_a \\ 2 \end{pmatrix}\begin{pmatrix} N_a \\ 2 \end{pmatrix}}{\begin{pmatrix} N_a \\ 2 \end{pmatrix}\begin{pmatrix} N_a \\ 2 \end{pmatrix} + N_a \begin{pmatrix} N_a \\ 2 \end{pmatrix}}$$

The proof of Theorem 1 is omitted due to space limitations.

A problem of the above discussion is that the analysis is purely based on pairwise error probability. However, system performance is normally expressed as average error probability (AEP). We further introduce a diversity concept based on AEP.

Definition 2: Denote AEP of the communications system with the codeword block set $\{\mathbf{M}\}$ at average receive SNR ρ as AEP $\{\mathbf{M}, \rho\}$. Assume that AEP $\{\mathbf{M}, \rho\}$ is differentiable at ρ.

Denote

$$f(\rho) = \log_{10}\left(AEP\left\{\mathbf{M}, \rho\right\}\right)$$

and

$$g(\rho) = \log_{10}\rho.$$

The average diversity order, r_{ad}, at the average signal-to-noise ratio (SNR) of each receive antenna, ρ, is defined as a differential

$$r_{ad} = -\frac{\partial f(\rho)}{\partial g(\rho)}. \quad (9)$$

Note that AEP cannot be generally derived. Thus we propose an analysis of the diversity performance of CI-STLDC based on the error union bound. EUB, an upper bound on the average error probability, is an average of the pairwise error probabilities between all direction pairs of codewords. Due to space limitations, the EUB based analysis is not provided in detail. The result of this analysis is that the average diversity order of CI-STLDC can be approximated as either $\min\left\{r_d^{(1)}, r_d^{(2)}\right\}$ or $r_d^{(1)} + r_d^{(2)}$, the choice of which depends on the value of SNR ρ and the codebook size N_a. In the case of $r_{ad} = \min\left\{r_d^{(1)}, r_d^{(2)}\right\}$, the merit of CI appears as an extra coding advantage.

Note that except for the trivial extra computational load of coordinate interleaving, for the same size of LDC encoding matrices, the complexity per LDC codeword of the ST-CILDC system is almost the same as that of conventional LDC systems. However, the upper bound achievable average diversity order of a ST-CILDC system is almost twice that of conventional block-based space-time code (BSTC) systems if the two component LDCs in the ST-CILDC have similar diversity features. It is worth mentioning that using nonlinear sphere or ML decoding, the conventional BSTC systems need much higher complexity to reach an average diversity order comparable to ST-CILDC.

We remark the scope of this approach is not limited to LDC. Other block-based space-time code designs may also be improved using the proposed space-time inter-LDC coordinate interleaving approach. Further, the pair of LDC codewords used in ST-CILDC could be viewed as a single specially designed LDC codeword of size $2T \times N_T$. Thus ST-CILDC systems could be viewed as extensions of LDC systems using different design criteria.

IV. Performance

A. Simulation setup

Perfect channel knowledge (amplitude and phase) is assumed at the receiver but not at the transmitter. Assume the number of receive antennas is equal to the number of transmit antennas. Channel symbols are estimated using MMSE estimation. Data symbols use QPSK modulation in all simulations. The signal-to-noise-ratio (SNR) reported in all figures is the average symbol SNR per receive antenna. The matrix channel is assumed to be constant over different integer numbers of channel uses or symbol time slots, and i.i.d. between blocks. We denote this interval as the channel change interval (CCI).

Three space-time block codes, Code A, Code B, and Code C, are used as component LDC coding matrices of ST-CILDC systems in the simulations. Code A is chosen from Eq. (31) of [4], a class of rate-one square LDC of arbitrary size proposed by Hassibi and Hochwald. Code B is chosen from Design A of full diversity full rate (FDFR) codes proposed by Ma and Giannakis [12]. Code C is a non-rate-one high rate code for the configuration of $N_T = 4, T = 6, Q = 12$, proposed by Hassibi and Hochwald [4].

B. Performance comparison

The performance comparison of code A is shown in Figures 2, 3 and 4. The performance comparison of code B is shown in Figure 5. The performance comparison of code C is shown in Figure 6.

In block fading channels, i.e., when the 4×4 MIMO channels are constant over the pair of ST-LDC codewords and code A is used, ST-CILDC obtains the same performance as that of ST-LDC as shown in Figure 3.

However, as shown in Figures 2, 4 5, and 6, ST-CILDC significantly outperforms ST-LDC at high SNRs in rapid fading channels. Thus, the ST-CILDC procedure may be applied to both rate-one and slightly lower rate codes. Observing Figures 2 and 5, the performances of code A and code B are similar in rapid fading channels. Thus, even though code A is not designed under a diversity criterion, code A appears to possess good diversity properties.

V. Conclusion

This paper has proposed a general space-time inter-LDC coordinate interleaving procedure, ST-CILDC, which may be applied to either rate-one (information lossless) or slightly lower rate block-based space-time coding systems. This enables not only symbol-level diversity but also coordinate-level diversity. The upper bound diversity order of ST-CILDC is analyzed, and the analysis results of the upper bound statistical diversity order and average diversity order of ST-CILDC are provided. Compared with conventional ST-LDC , ST-CILDC show either much higher average diversity order or extra coding advantage in time varying channels. Compared with conventional block-based STCs, ST-CILDC maintains diversity performance in quasi-static block fading channels, and significantly improves the diversity performance in the high SNR regions of rapid fading channels.

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References

[1] V.Tarokh, N.Seshadri, and A.Calderbank, "Space-time codes for high data rate wireless communications: performance criterion and code construction," *IEEE Trans.Inform.Theory*, vol. 44, pp. 744–765, Mar. 1998.

[2] S.Alamouti, "A simple transmitter diversity scheme for wireless communications," *IEEE J.Select.Areas Commun.*, pp. 1451–1458, Oct. 1998.

[3] V.Tarokh, H.Jafarkhani, and A.R.Calderbank, "Space-time block code from orthogonal designs 3," *IEEE Trans.Inform.Theory*, vol. 45, pp. 1456–1467, July 1999.

[4] B. Hassibi and B. M. Hochwald, "High-rate codes that are linear in space and time," *IEEE Trans.Inform.Theory*, vol. 48, no. 7, pp. 1804–1824, July 2002.

[5] K.Boulle and J.C.Belfiore, "Modulation schemes designed for the Rayleigh channel," in *Proc. CISS 1992*, 1992, pp. 288–293.

[6] B.D.Jelicic and S.Roy, "Cutoff rates for coordinate interleaved QAM over Rayleigh fading channels," *IEEE Trans.Commun.*, vol. 44, no. 10, pp. 1231–1233, Oct. 1996.

[7] Y.-H. Kim and M.Kaveh, "Coordinate-interleaved space-time coding with rotated constellation," in *Proc. IEEE VTC*, vol. 1, Apr. 2003, pp. 732–735.

[8] M.Z.A.Khan and B.S.Rajan, "Space-time block codes from co-ordinate interleaved orthogonal designs," in *Proc. IEEE ISIT 2002*, 2002, pp. 275–275.

[9] M.Z.A.Khan, B.S.Rajan, and M. H. Lee, "Rectangular co-ordinate interleaved orthogonal designs," in *Proc. IEEE Globecom 2003*, vol. 4, Dec. 2003, pp. 2003–2009.

[10] W. Su, Z.Safar, and K.J.R.Liu, "Diversity analysis of space-time modulation over time-correlated Rayleigh-fading channels," *IEEE Trans.Inform.Theory*, vol. 50, no. 8, pp. 1832–1840, Aug. 2004.

[11] S.Siwamogsatham, M.P.Fitz, and J.H.Grimm, "A new view of performance analysis of transmit diversity schemes in correlated Rayleigh fading," *IEEE Trans.Inform.Theory*, vol. 48, no. 4, pp. 950–956, Apr. 2002.

[12] X. Ma and G.B.Giannakis, "Full-diversity full-rate complex-field space-time coding," *IEEE Trans.on Sig.Proc.*, vol. 51, no. 11, pp. 2917–2930, Nov. 2003.

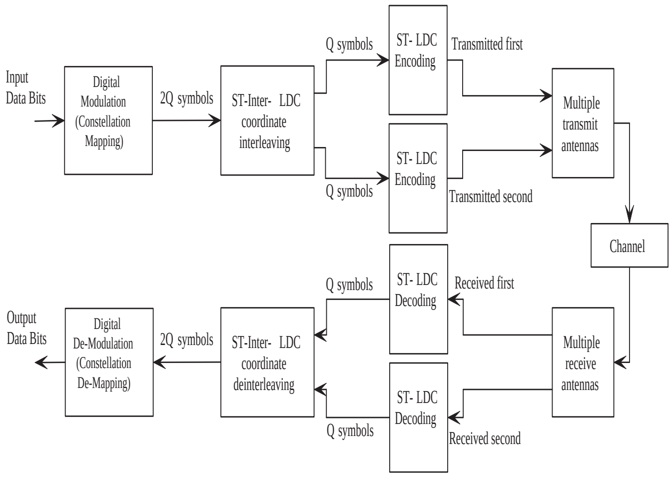

Fig. 1. Space-time inter-LDC coordinate interleaving system structure

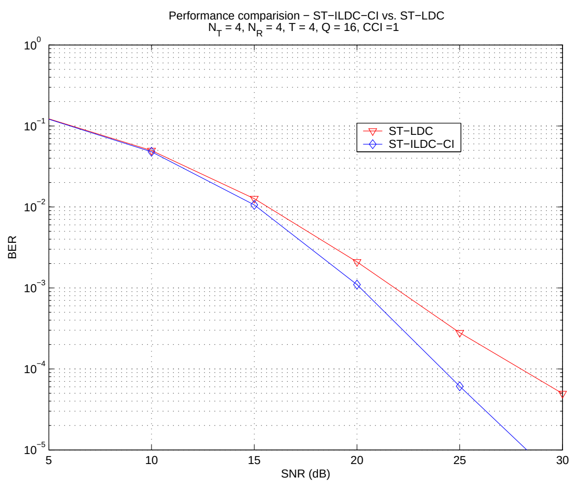

Fig. 2. BER performance comparison of ST-CILDC vs. LDC using code A, $CCI = 1$, $N_T = 4$, $N_R = 4$, $T = 4$, $Q = 16$

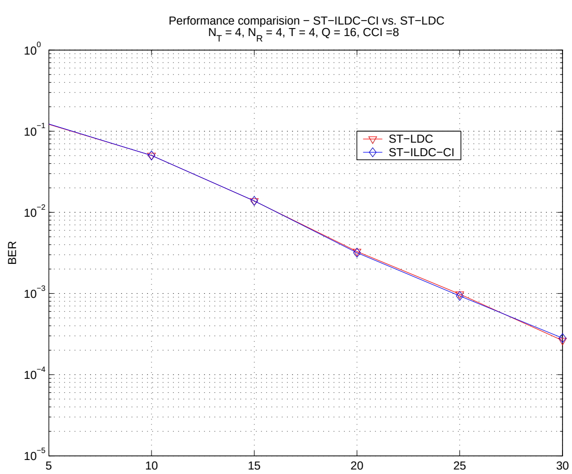

Fig. 3. BER performance comparison of ST-CILDC vs. LDC using code A, $CCI = 8$, $N_T = 4$, $N_R = 4$, $T = 4$, $Q = 16$

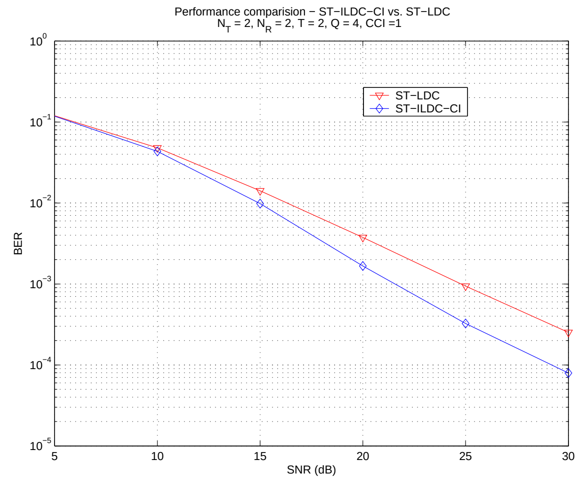

Fig. 4. BER performance comparison of ST-CILDC vs. LDC using code A, $CCI = 1$, $N_T = 2$, $N_R = 2$, $T = 2$, $Q = 4$

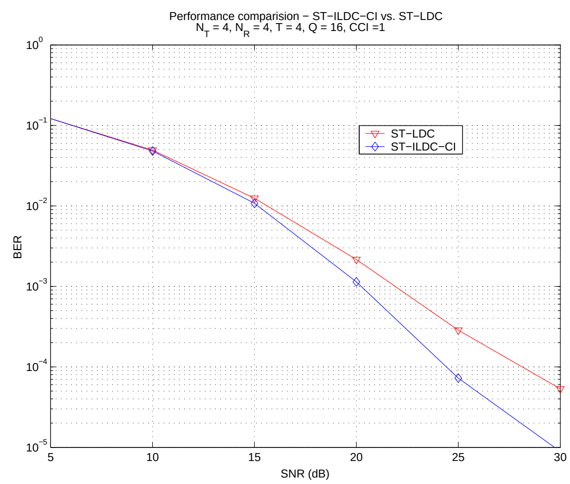

Fig. 5. BER performance comparison of ST-CILDC vs. LDC using code B, $CCI = 1$, $N_T = 4$, $N_R = 4$, $T = 4$, $Q = 16$

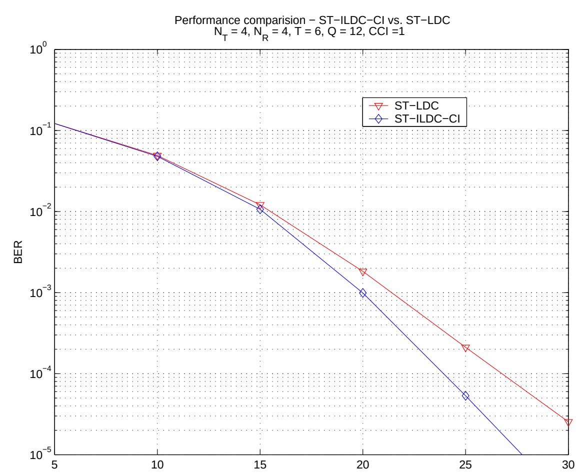

Fig. 6. BER performance comparison of ST-CILDC vs. LDC using code C, $CCI = 1$, $N_T = 4$, $N_R = 4$, $T = 6$, $Q = 12$