Linear Dispersion Codes for Asynchronous Cooperative MIMO Systems

Nan Wu
National Institute of Standards and Technology
Gaithersburg, MD 20899, USA
nan.wu@nist.gov

Hamid Gharavi
National Institute of Standards and Technology
Gaithersburg, MD 20899, USA
gharavi@nist.gov

***Abstract*—In this treatise, we propose a novel family of Asynchronous Cooperative Linear Dispersion Codes (ACLDCs), which is capable of maintaining full diversity in cooperative scenarios even at the presence of asynchronous reception. The linear dispersion structure is employed in order to accommodate the dynamic topology of cooperative networks, as well as to achieve higher throughput than conventional space time codes based on orthogonal designs. By introducing guard intervals and block encoding/decoding techniques, the interference signals caused by asynchronous reception can be exploited rather than discarded.**

I. Introduction

The Space Time Block Coding (STBC) [1] [2] [3] techniques provide full spatial diversity in the context of co-located MIMO systems. However, it may not always be practical to accommodate multiple antennas at the mobile nodes in the network, owing to cost, size and other hardware limitations. As a remedy, the concept of 'cooperative diversity' has been proposed in the literature [4] [5] [6] [7], providing diversity using single antennas of other nodes in the network.

Furthermore, it is often the case that propagation delays experienced by the signals from cooperative nodes are different, even if these nodes are scheduled to transmit simultaneously. Thus, the composite pulse seen at the receiver, which is the sum of the pulses from each transmitter shifted by the corresponding propagation delay, will no longer be Nyquist. Hence, Inter Symbol Interference (ISI) is generated after sampling at the receiver, where similar phenomena is observed with frequency selective channels.

In general, there are three classes of techniques in the open literature to deal with the issue of delay ISI, which are time-domain approaches [8] [9], frequency-domain solutions [10] [11] [12], and the use of conventional Equalizer [13].

Firstly, Time Reversed Space-Time Block Codes (TR-STBCs) [8] [9] were proposed in order to protect the Alamouti type scheme [1] [2] [14] from being contaminated by delay ISI. The idea is that every symbol of a STBC codeword is replaced by a block of B symbols, while the conjugate operation requires the corresponding block to be transmitted in a time reversed order. However,the TR-STBCs are unsuitable for high-rate transmission, owing to the embedded orthogonality. Another time-domain approach is called Linear Asynchronous Space-Time Block Codes (LA-STBCs) [11] [15] employing a linear dispersion structure [3] [16]. LA-STBCs are robust, when the propagation delay differences between cooperative nodes are multiple symbol durations ($\tau = nT_s, n = 1, 2, \ldots$), where T_s is the symbol duration. However, when the propagation delay difference is not an integral multiple of T_s, which is often the case in real systems, LA-STBCs degrade significantly. Furthermore, the distributed Threaded Algebraic Space-Time (TAST) codes [17] provide flexible transmission rates, arbitrary antenna support and adjustable complexity for delay-tolerant STBC designs. Unfortunately, distributed-TAST codes remain vulnerable to propagation delays that are not the integral multiplication of a single symbol duration.

Secondly, in frequency domain, OFDM techniques can be employed in order to convert the equivalent frequency selective channel into multiple flat fading channels [18]. Finally, transmission schemes suffering from various propagation delays can be considered as delay diversity schemes [19] and a decision feedback equalizer [13] can be employed in order to achieve spatial diversity.

In this paper, we propose a novel time-domain STBC scheme in order to combat imperfect synchronization in cooperative MIMO Systems, namely Asynchronous Cooperative Linear Dispersion Codes (ACLDCs). The rationale and novelty of the proposed ACLDCs are:

- The proposed scheme is capable of dealing with arbitrary propagation delay difference τ while maintaining full spatial diversity, provided that sufficient guard intervals are appropriately inserted, as opposed to certain delays of LA-STBCs and distributed-TAST codes.
- The proposed scheme features in high-rate transmissions, whereas the TR-STBCs are unable to have a symbol rate higher than unity.
- The ACLDC scheme consists of a space-time encoder to achieve full spatial diversity and a block encoder (or interleaver) to combat the propagation delay, which are designed jointly, rather than separately.

We commence our discourse by providing a detailed description of a linear dispersion structure in Section II. Section III extends the linear dispersion structure by introducing block encoding/decoding techniques, so that the proposed ACLDC scheme is capable of maintaining full spatial diversity. Our simulation results are discussed in Section IV. Finally, we conclude our discourse in Section V.

II. Cooperative Linear Dispersion Codes (CLDCs)

After introducing the linear dispersion framework, in this section, the power loss caused by the propagation delay ISI is analyzed together with the associated BER performance.

Cooperative schemes in general contain two phases of transmission, namely the broadcast phase and the cooperation phase. During the cooperation phase, relays collaboratively transmit the re-encoded source information, where 'virtual' space-time codewords can be formed. Since the issue of synchronization only involves the cooperation phase, we focus on the cooperation phase.

More explicitly, assume each relay to process the perfect source information vector $\mathbf{K} = [s^1, \ldots, s^Q]^T$ containing Q symbols, which are obtained through the broadcast interval. The k-th relay ($k = 1, \ldots, M$) disperses vector $\mathbf{K}$ by:

$$\mathbf{S}_k = \mathbf{A}_k\mathbf{K}, \tag{1}$$

where the dispersion matrix $\mathbf{A}_k$ having a size of $(T \times Q)$ characterizes how the information is distributed among the T channel uses. By stacking the transmitted signals $\mathbf{S}_k$ from all the relays, a cooperative space-time codeword $\mathbf{C}$ having a size of $(M \times T)$ can be obtained as follows:

$$\mathbf{C} = \begin{pmatrix} \mathbf{S}_1^T \\ \vdots \\ \mathbf{S}_M^T \end{pmatrix} = \begin{pmatrix} (\mathbf{A}_1\mathbf{K})^T \\ \vdots \\ (\mathbf{A}_M\mathbf{K})^T \end{pmatrix}, \tag{2}$$

which should satisfy the overall power constraint of $E\{tr(\mathbf{CC}^H)\} = T$. We can further ensure information vector $\mathbf{K}$ is dispersed with equal power into T channel uses of each relay by restricting $\mathbf{A}_k$ to satisfy

$$\mathbf{A}_k\mathbf{A}_k^H = \frac{1}{M}\mathbf{I}, \tag{3}$$

where $\mathbf{I}$ denotes an identity matrix having a size of $(T \times T)$. Note that when constraint of Equation (3) applied, we should have $Q \geq T$.

At the destination node, the received signal matrix $\mathbf{Y}$ having a size of $(1 \times T)$, becomes

$$\mathbf{Y} = \mathbf{HC} + \mathbf{V}, \tag{4}$$

where $\mathbf{V}$ having a size of $(1 \times T)$ represents realizations of an i.i.d. complex AWGN process with zero-mean and variance σ_0^2 determined by the associated SNR ρ. Each entry of $\mathbf{H}$ represents the Rayleigh fading coefficients between a transmit-receive antenna pair. [1] The entries of the channel matrix are assumed to be known to the destination node, but not to the relays.

Define the $row()$ operation as the vertical stacking of the rows of an arbitrary matrix. Subjecting both sides of Equation (4) to the $row()$ operation gives the equivalent system matrix:

$$\bar{\mathbf{Y}} = \bar{\mathbf{H}}\bar{\chi}\bar{\mathbf{K}} + \bar{\mathbf{V}}. \tag{5}$$

The equivalent channel matrix $\bar{\mathbf{H}}$ of Equation (5) is given by

$$\bar{\mathbf{H}} = \mathbf{H} \otimes \mathbf{I}, \tag{6}$$

where $\otimes$ denotes the Kronecker product. The equivalent dispersion matrix $\bar{\chi}$ of Equation (5) having a size of $(MT \times MQ)$ becomes

$$\chi = \begin{pmatrix} \mathbf{A}_1 & \mathbf{0} & \cdots & \mathbf{0} \\ \mathbf{0} & \mathbf{A}_2 & \cdots & \mathbf{0} \\ & & \ddots & \\ \mathbf{0} & \mathbf{0} & & \mathbf{0} \\ \mathbf{0} & \cdots & \cdots & \mathbf{A}_M \end{pmatrix}, \tag{7}$$

where $\mathbf{0}$ denotes a zero matrix having a size of $(T \times Q)$. Finally, $\bar{\mathbf{K}}$ of Equation (5) is the repetition of the information vector for M times and is given by

$$\bar{\mathbf{K}} = \begin{pmatrix} \mathbf{K} \\ \vdots \\ \mathbf{K} \end{pmatrix}. \tag{8}$$

Thus, conventional Maximize Likelihood (ML) detection can be carried out of in order to recover the original information.

The equivalent system Equation (5) clearly demonstrates that the achievable performance of the CLDC is entirely determined by the Dispersion Character Matrix (DCM) $\bar{\chi}$ of Equation (7). In other words, the challenge of achieving 'cooperative diversity' is equivalent to designing a single DCM $\bar{\chi}$, while obeying the power constraint of Equation (3).

Obviously, this flexible linear dispersion framework can support any number of cooperative nodes M, arbitrary channel use of T as well as arbitrary information vector $\mathbf{K}$ containing Q symbols, since the combination of M, T and Q can be reflected on the design of $\bar{\chi}$. Therefore, we denote such a scheme as CLDC(MTQ).

In Figure 1, the BER performance of the CLDC(222) scheme under various delay differences is characterized, when transmitting over i.i.d. Rayleigh-fading channels having a Normalized Doppler frequency of $f_d = 10^{-2}$.

We assume that the destination node is always synchronized to the first transmitter (Node-1), which means that the sampling process at the receiver will not cause ISI for signals transmitted from Node-1. On the other hand, the second transmitter's (Node-2) signal

[1] In this treatise, the correlated fading process is generated with a filtering of the complex Gaussian process to achieve a given Doppler spectrum and the Jakes' model is applied.

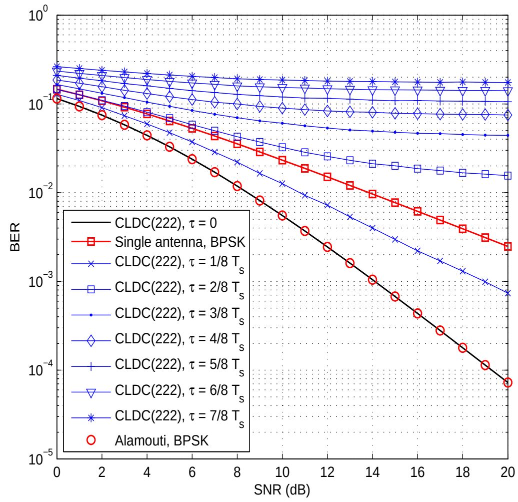

Figure 1. BER of BPSK modulated CLDC(222) scheme obeying the structure of Equation (5) using an ML detector and the propagation delay difference between two-path is characterized as τ, when transmitting over i.i.d. Rayleigh-fading channels having a Normalized Doppler frequency of $f_d = 10^{-2}$.

Table I
THE POWER LOSS CAUSED BY ASYNCHRONOUS RECEPTION AT THE DESTINATION.

Delay different τ	P	P_1	P_2
0 (T_s)	1	0	0
1/8 (T_s)	0.971	0.116	0.08
2/8 (T_s)	0.887	0.263	0.123
3/8 (T_s)	0.759	0.429	0.133
4/8 (T_s)	0.600	0.600	0.120
5/8 (T_s)	0.429	0.759	0.092
6/8 (T_s)	0.263	0.887	0.058
7/8 (T_s)	0.116	0.971	0.025
8/8 (T_s)	0	0	1

Note: P denotes useful signal power; P_1 denotes the signal power of 'Next' Symbol; P_2 denotes the signal power of 'Previous' Symbol;

arrives with a relative delay τ. Given that the signals transmitted are simulated using Raised Cosine Pulses having a roll-off factor of 0.5, we further assume that only one side-lobe contributes to the ISI from Node-2. More explicitly, the sampled signal for Node-2 not only contains the useful symbol information, but also contains the 'previous' transmitted symbol and the 'next' adjacent symbol in, owing to the side-lobe effect of the Raised Cosine pulse. Furthermore, as a result of asynchronous reception, Table I summarizes the power loss of the desired signal and the power increase of ISI signals with respect to the value of τ. The entries in Table I are generated by sampling the received raised cosine waves using the above-mentioned method. Note that the ISI addressed in our scheme is caused by the propagation delay difference after sampling at the receiver, rather than the one caused by multi-path, although both of them have the similar effect of contaminating the received signals. More explicitly, the multi-path ISI can be removed as as long as it is within the length of guard intervals, whereas the propagation delay ISI is exploited by the interleaver and the ML decoder.

Observe in Figure 1 that our identical throughput CLDC(222) with $\tau = 0$ is able to achieve the same BER performance as the Alamouti scheme, which corresponds to full spatial diversity. Please note that CLDC(222) is not an Alamouti scheme, even though they exhibit identical performance. The proposed CLDC scheme disperses each information symbol into all the spatial and temporal dimensions,

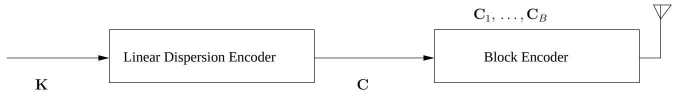

Figure 2. ACLDC encoder for the nodes having cooperative transmission.

whereas the Alamouti scheme exploits only half of the available resources for each symbol. When the propagation delay difference begin to increase, i.e. $\tau = 1/8T_s$, significant BER degradation has been recorded in Figure 1. Furthermore, when we have $\tau = 2/8T_s$, the resultant BER performance of Figure 1 is already worse than the identical throughput single-antenna aided system. Again, Figure 1 explicitly demonstrates that the issue of asynchronous reception is critical for 'cooperative' diversity schemes.

III. Asynchronous Cooperative Linear Dispersion Codes

In this section, a novel family of ACLDCs is proposed in order to combat the severe performance degradation caused by asynchronous reception, which has been characterized in Figure 1. We assume that the propagation delay difference τ_k, which is the difference between Node-1 and Node-k ($k = 2, \ldots, M$) is known to the receiver. For coherently-detected cooperative systems, pilot signals are employed for channel estimations of each relay. From the arrival time of the pilot signals, the knowledge of propagation delay difference can be obtained. Again, the receiver is synchronized to Node-1 and ISI signals are generated for the signals transmitted from all the remaining Nodes.

Figure 2 portrays the encoder of ACLDCs employed at every cooperative node. The 'Linear Dispersion Encoder', as described in Section II generates codeword matrices $\mathbf{C}$ obeying Equation (2). More explicitly, given a block of information vectors $[\mathbf{K}_1, \ldots, \mathbf{K}_B]$, the CLDC encoder generates the corresponding B number of codewords $[\mathbf{C}_1, \ldots, \mathbf{C}_B]$ based on Equation (2). The 'Block Encoder' interleaves the incoming B number codewords into T number of transmissions as follows:

$$\mathbf{F}_i = [L_i(\mathbf{C}_1), \ldots, L_i(\mathbf{C}_B), \mathbf{0}_D], \qquad (i = 1, \ldots, T) \quad (9)$$

where $L_i()$ denotes the i-th column of a matrix and $\mathbf{0}_D$ denotes a zero matrix having D number of columns, serving as guard intervals. We further assume the length of guard intervals to be equal or greater than the maximum delay difference $D \geq \tau_{max}$, which implies the interference between transmission blocks $\mathbf{F}_i$ is removed.

Note that the block encoder can also be viewed as an interleaver, where the interleaving sequence is given in Equation (9). Furthermore, it is the linear dispersion structure of Section II that determines the interleaving sequence, so the receiver can carry out the linear block ML decoding of Equation (16). In other words, if the linear dispersion structure is achieved differently, the interleaver structure should be designed accordingly.

The 'Block Encoder' of Figure 2 is introduced for two reasons. Firstly, effective throughput can be increased, when appropriate channel conditions are available. Since guard intervals are inserted every B block of space-time codewords, the effective symbol rate becomes $\frac{BQ}{T(B+D)}$, which approaches the maximum rate $\frac{Q}{T}$ with an increase of block length B. By contrast, conventional schemes append guard intervals after every codeword. Hence, effective throughput is degraded significantly. Secondly, the B number of codewords are 'interleaved' as seen in Equation (9). This re-arrangement of the codewords is necessary because 'intra-codeword' interference is removed. For example, the second column of a codeword $L_2(\mathbf{C}_i)$ does not interfere with other columns within the same codeword. In other words, the interference only comes from other codewords, not within the same codeword. In this way, we can re-write the equivalent interference signals in a format that can be exploited, as exemplified later in Equation (15). Note that there are other methods to 'interleave' the block of codewords, which could achieve similar effects. In this paper, we only present one as illustrated in Equation (9).

Given the channel CIR matrix $\mathbf{H} = P \cdot [h_1, \ldots, h_M]$ and assume $\mathbf{H}$ to be constant over $(B+D)$ CLDC codewords, the received signal for the i-th transmission block becomes:

$$\mathbf{y}_i = \mathbf{H}\mathbf{F}_i + \mathbf{V}_i. \qquad (i = 1, \ldots, T) \quad (10)$$

In order to re-construct the original codewords $[\mathbf{C}_1, \ldots, \mathbf{C}_B]$, the received signals are firstly sampled, as described in Section II, and 'de-interleaved'. Thus, we have

$$\begin{aligned} \mathbf{Y}_j &= [L_i(\mathbf{y}_1), \ldots, L_i(\mathbf{y}_T)] \qquad (j = 1, \ldots, B) \\ &= \mathbf{H}\mathbf{C}_j + \sum_{k=2}^{M} h_k[P_1, P_2]\mathbf{G}_k + \mathbf{V}_j, \end{aligned} \quad (11)$$

where $\mathbf{G}_k$ denotes the ISI matrix of the k-th cooperative node, caused by the propagation delay after sampling at the receiver. Particularly, the first row of $\mathbf{G}_k$ denotes the interference from the 'Previous' codeword $\mathbf{C}_{j-1}$ and the second row of $\mathbf{G}_k$ denotes interference from the 'Next' codeword $\mathbf{C}_{j+1}$. The power of the interference is denoted by P_1 and P_2, as seen in Table I.

Similar to Equation (5), $row()$ operation is applied to both sides of Equation (11), then we have

$$\bar{\mathbf{Y}}_j = \bar{\mathbf{H}}\bar{\chi}\bar{\mathbf{K}}_j + \sum_{k=2}^{M} \bar{\mathbf{H}}_k\bar{\chi}_k\bar{\mathbf{K}}_{isi} + \bar{\mathbf{V}}_j, \quad (12)$$

where the first item on the right is the desirable signal and the second item is the ISI signals from the k-th node. The equivalent system matrices $\bar{\mathbf{H}}$, $\bar{\chi}$ and $\bar{\mathbf{K}}_j$ have been shown in Equations (6), (7) and (8), respectively. Note that the DCM optimized for CLDCs of Section II is employed, which is optimized for $\tau = 0$. With the help of the interleaver, ACLDCs are expected to maintain the diversity advantage under different values of τ. The equivalent ISI matrix $\bar{\mathbf{H}}_k$ for the k-th node is given by

$$\bar{\mathbf{H}}_k = h_k[P_1, P_2] \otimes \mathbf{I}, \quad (13)$$

where $\mathbf{I}$ denotes an identity matrix. The corresponding DCM $\bar{\chi}_k$ is given by

$$\bar{\chi}_k = \begin{pmatrix} \mathbf{A}_k & \mathbf{0} \\ \mathbf{0} & \mathbf{A}_k \end{pmatrix}, \quad (14)$$

where '$\mathbf{0}$' denotes a zero matrix having a size of $(T \times Q)$. The vector $\bar{\mathbf{K}}_{isi}$ denotes the interference signal from the 'Previous' and 'Next' vectors, which is given by

$$\bar{\mathbf{K}}_{isi} = \begin{pmatrix} \mathbf{K}_{j-1} \\ \mathbf{K}_{j+1} \end{pmatrix}. \quad (15)$$

Note that it is the specific interleaving sequence presented in Equation (9) allows DCM $\bar{\chi}_k$ of Equation (14) and interference signal vector of Equation (15) to be rewritten in such a simplified form. In other words, the design of the interleaver and that of the space-time decoder have to be considered jointly.

Hence, we can recover the information vector block $[\mathbf{K}_1, \ldots, \mathbf{K}_B]$ by calculating

$$\underline{[\mathbf{K}_1, \ldots, \mathbf{K}_B]} = \arg\{\min(||\sum_{j=1}^{B}(\bar{\mathbf{Y}}_j - \bar{\mathbf{H}}\bar{\chi}\bar{\mathbf{K}}_j - \sum_{k=2}^{M} \bar{\mathbf{H}}_k\bar{\chi}_k\bar{\mathbf{K}}_{isi})||^2)\}, \quad (16)$$

when all possible combinations of $[\mathbf{K}_1, \ldots, \mathbf{K}_B]$ are explored. Note that low-complexity Sphere Decoders designed for space-time block codes of multiple antenna systems can be employed to achieve near ML performance with a much lower decoding complexity. The basic

Table II
SYSTEM PARAMETERS FOR ACLDC SCHEMES OF FIGURE 2.

Number of cooperative nodes	M
Number of antenna per node	1
Number of channel uses per block	T
Number of symbols per information vector	Q
Length of a decoding block	B
Length of the guard interval	D
Propagation delay difference	τ_k
Channel constant for	$(B+D)$ blocks
Normalized Doppler Frequency	f_d
Modulation	BPSK
Mapping	Gray mapping
Detector	ML of Equation (16)

idea is that instead of searching through all the constellation points, the decoder only searches the points of the lattice which are found inside a sphere of a given radius centered at the received point. We refer the readers to [20] for more details.

We now continue by offering a few remarks concerning the equivalent ACLDC system model of Equation (12).

1) Spatial diversity: The fundamental idea of achieving diversity is to have independent copies of the same information. Alamouti-type schemes achieve this objective by transmitting redundant information from the extra antenna. However, redundancy can be reduced by employing the proposed linear dispersion structure, where each transmitted signal is the weighted sum of all the information symbols, as seen in Equation (1). Since this structure remains in the equivalent system model of Equation (12), the proposed system is capable of achieving diversity, regardless of block length B.
2) The effect of delays: Ultimately, delays affect power loss and power distribution of the transmitted information and is summarized in Table I. In the case of perfect synchronization, the power of the desired signal is concentrated within one sampled value. In case of asynchronous reception, the desired signal power is spread into current and adjacent samples having the power of P, P_1 and P_2, respectively. Our powerful block detector of Equation (16) is capable of exploiting all this information so that full diversity is maintained. However, there will be some power loss during the sampling process, which would degrade the achievable performance slightly and is demonstrated in Section IV.
3) Recall that the channel is assumed to be constant over $(B+D)$ STBC blocks in order to facilitate coherent detection. When this condition is violated, the system's achievable performance will degrade.

IV. SIMULATION RESULTS

This section presents the simulation results for a number of ACLDC(MTQ) schemes. The channel is assumed to be constant for $(B+D)$ blocks, then faded to another value governed by the Normalized Doppler frequency f_d. All the system parameters are listed in Table II, unless otherwise stated. For simplicity, we assume $\tau < T_s$. Thus, only one guard interval is necessary. However, our system is capable of supporting arbitrary delay difference values. More explicitly, for any fixed value $\alpha \in (0,1)$, i.e. $\alpha = 0.5$, the BER performance for ACLDC(MTQ)s having $\tau = \alpha T_s + nT_s$ would be identical, where $n = 0, 1, 2, \ldots$. This is because we can still exploit the delayed version of the signals thanks to our flexible system architecture.

Figure 3 characterizes the BER performance of the ACLDC(222) scheme using a block length of $B = 2$, while experiencing different propagation delay difference τ. Compared with the non-interleaving

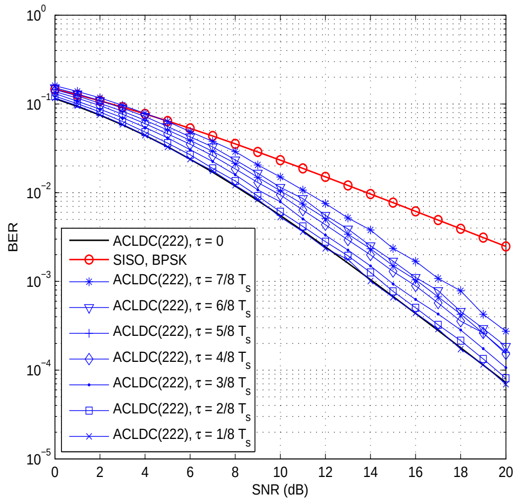

Figure 3. BER comparison of a group of ACLDCs having $M = 2$, $T = 2$, $Q = 2$ and $B = 2$ while experiencing different propagation delay difference τ, when transmitting over i.i.d. Rayleigh-fading channels having $f_d = 10^{-2}$. All the system parameters were summarized in Table II.

CLDC(222) counterpart of Figure 1 having identical τ values, the proposed ACLDC scheme demonstrates a substantial gain thanks to the introduction of the 'interleaver'. For example, in the case of perfect synchronization (i.e. $\tau = 0$), the best achievable performance is recorded. As τ increases, we have (i.e. $0 < \tau < T_s$), the BER performance begins to degrade slowly, owing to the power loss in sampling, as illustrated in Table I.

However, the ACLDC scheme remains capable of maintaining full spatial diversity. Note that the spatial diversity of CLDCs is guaranteed by optimizing the DCM of Equation (7) using the Rank and determinant criteria detailed in [2]. Since the ACLDCs inherit the CLDCs' encoder, full spatial diversity is achieved when we have $\tau = 0$. When we have other τ values, the diversity advantage is proven in terms of BER's decay with the increase of SNR. More explicitly, the decay of BER having diversity of two is much faster than that of a single-antenna aided system. In other words, the slopes of the array of ACLDC's curves are the same. Again, our proposed scheme is capable of supporting arbitrary delay difference values. For example, when $\tau = 2\frac{1}{2}T_s$, the BER performance would be identical to $\tau = \frac{1}{2}T_s$ recorded in Figure 3, provided that $D = 3$ guard intervals are inserted.

Figure 4 demonstrates the BER performance of the ACLDC(222) scheme with $\tau = 3/4T_s$, while having a block length of $B = 1, 2, 4, 6$. Since a guard interval is inserted every block length of B, the resultant system symbol rate becomes $R = 0.5, 0.67, 0.8, 0.86$, respectively. Another advantage of increasing the value B is that there is slightly BER performance gain, as recorded in Figure 4, since the ISI information of Equation (12) has been explored. However, there are two drawbacks associated with the increase the B. Firstly, the decoding complexity will increase exponentially, owing to B number of codewords that are jointly decoded. Secondly, the channel has to be constant over $(B+D)$ codeword blocks in order to carry out coherent detection of Equation (16). In other words, the choice of parameter B involves a fundamental trade-off between the system throughput and decoding complexity.

V. CONCLUSION

In this paper, we first proposed a family of CLDCs for cooperative networks and demonstrated its ability to achieve full spatial diversity,

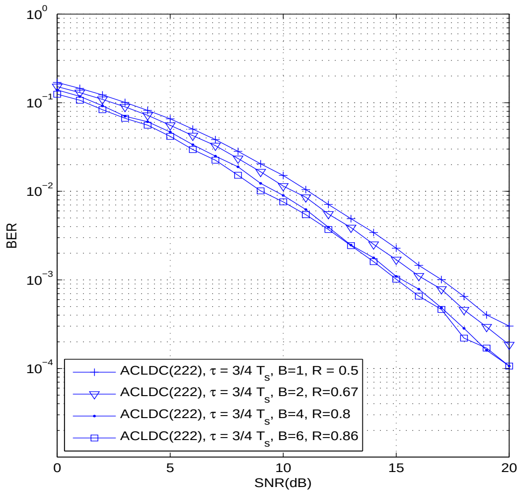

Figure 4. BER comparison of a group of ACLDCs having $M = 2$, $T = 2$, $Q = 2$ and $\tau = 3/4T_s$ while using the decoding block length of $B = 1, 2, 4, 6$, when transmitting over i.i.d. Rayleigh-fading channels having $f_d = 10^{-2}$. All the system parameters were summarized in Table II.

as well as its vulnerability under the situation of asynchronous reception. Later, we proposed a novel time-domain delay-tolerance ACLDC and demonstrated that the desirable cooperative diversity can be maintained, even if severe propagation delay differences exist. The insertion of guard intervals mitigates interference between transmission blocks and the associated throughput loss can be addressed using a flexible linear dispersion structure.

Appendix A
Example of Dispersion Matrices

Since the performance of CLDCs is characterized by the equivalent Dispersion Character Matrix (DCM) $\bar{\chi}$ defined in Equation (7), In this section, some Dispersion Character Matrix (DCM) $\bar{\chi}$ of Equation (7) are given. Note that the optimized linear dispersion code is suitable for any relative delay difference values, because the 'interleaver' coupled with the guard intervals that is responsible for combating the ISI. We restrict ourselves to a discussion of the linear dispersion framework itself, rather than emphasizing the issue of designing dispersion matrices. However, we'd like to point out that there are a number of criteria that can be used to optimize $\bar{\chi}$, such as maximizing mutual information [3], having a non-vanishing determinant [21] or minimizing the maximum Pairwise Symbol Error Probability (PSEP) [2]. In this paper, the BER-oriented PSEP criterion is chosen.

At the bottom of this page, the DCMs for BPSK-modulated CLDC(222) scheme is illustrated. Since the Alamouti scheme can be rewritten using the proposed linear dispersion framework of Equations (2) and (5), the corresponding DCM is also given.

$$\bar{\chi}_{Alamouti} = \begin{pmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & -1 \\ 0 & 0 & 1 & 0 \end{pmatrix}$$

$$\bar{\chi}_{CLDC(222)} = \begin{pmatrix} -0.4651 + 0.4952i & -0.1788 + 0.0807i & 0 & 0 \\ 0.0639 - 0.1855i & -0.4502 + 0.5088i & 0 & 0 \\ 0 & 0 & -0.1080 - 0.2144i & 0.4542 + 0.4859i \\ 0 & 0 & -0.3846 - 0.5427i & -0.1883 - 0.1489i \end{pmatrix}$$

References

[1] S. M. Alamouti. A simple transmit diversity technique for wireless communications. *IEEE Journal on Selected Areas in Communications*, 16(8):1451–1458, 1998.

[2] V. Tarokh, H. Jafarkhani, and A. R. Calderbank. Space-time block codes from orthogonal designs. *IEEE Transactions on Information Theory*, 45(5):1456–1467, 1999.

[3] B. Hassibi and B. M. Hochwald. High-rate codes that are linear in space and time. *IEEE Transactions on Information Theory*, 48(7):1804–1824, July 2002.

[4] K. C. Liang, X. D. Wang, and I. Berenguer. Minimum error-rate linear dispersion codes for cooperative relays. *IEEE Transactions on Vehicular Technology*, 56:2143–2157, July 2007.

[5] J. N. Laneman and G. W. Wornell. Distributed space-time-coded protocols for exploiting cooperative diversity in wireless networks. *IEEE Transactions on Information Theory*, 49(10):2415–2425, October 2003.

[6] J. Boyer, D.D. Falconer, and H. Yanikomeroglu. Multihop diversity in wireless relaying channels. *IEEE Transactions on Communications*, 52(10):1820–1830, October 2004.

[7] Y.D. Jing and B. Hassibi. Distributed space-time coding in wireless relay networks. *IEEE Transactions on Wireless Communications*, 5(12):3524–3536, December 2006.

[8] E. Lindskog and A. Paulraj. A transmit diversity scheme for channels with intersymbol interference. In *IEEE International Conference on Communications*, volume 1, pages 307–311, New Orleans, LA, June 2000.

[9] P. Stoica and E. Lindskog. Space-time block coding for channels with intersymbol interference. In *Thirty-Fifth Asilomar Conference on Signals, Systems and Computers*, volume 1, pages 252–256, Pacific Grove, CA, November 2001.

[10] S.G. Rajan and S.B. Rajan. OFDM based distributed space time coding for asynchronous relay networks. In *IEEE International Conference on Communications*, pages 1118–1122, Beijing, May 2008.

[11] Z. Li and X.G. Xia. A simple alamouti space time transmission scheme for asynchronous cooperative systems. *IEEE Signal Processing Letters*, 14(11):804–807, November 2007.

[12] O.S. Shin, A. M. Chanand, H. T. Kung, and V. Tarokh. Design of an OFDM cooperative space-time diversity system. *IEEE Transactions on Vehicular Technology*, 56:2203–2215, July 2007.

[13] S. Q. Wei, D. L. Goeckel, and M. C. Valenti. Asynchronous cooperative diversity. *IEEE Transactions on Wireless Communications*, 5(6):1547–1557, June 2006.

[14] H. Jafarkhani. A quasi-orthogonal space-time block code. *IEEE Transactions on Communications*, 49(1):1–4, 2001.

[15] Y. Shang and X. G. Xia. Limited-shift-full-rank matrices with applications in asynchronous cooperative communications. *IEEE Transactions on Information Theory*, 53(11):4119–4126, November 2007.

[16] R. W. Jr. Heath and A. Paulraj. Linear dispersion codes for MIMO systems based on frame theory. *IEEE Transactions on Signal Processing*, 50(10):2429–2441, October 2002.

[17] M. O. Damen and A. R. Hammons. Delay-tolerant distributed-TAST codes for cooperative diversity. *IEEE Transactions on Information Theory*, 53(10):3755–3773, October 2007.

[18] H. Zhang, X. G. Xia, L. J. Cimini Jr., and P.C. Ching. Synchronization techniques and guard-band-configuration scheme for single-antenna vector-OFDM systems. *IEEE Transactions on Wireless Communications*, 4(5):2454–2464, September 2005.

[19] A. Wittneben. Base station modulation diversity for digital simulcast. In *IEEE Vehicular Technology Conference*, pages 848–853, St. Louis, May 1991.

[20] E. Viterbo and J. Boutros. A universal lattice code decoder for fading channels. *IEEE Transactions on Information Theory*, 45(5):1639–1642, July 1999.

[21] J. C. Belfiore, G. Rekaya, and E. Viterbo. The golden code: a 2×2 full-rate space-time code with nonvanishing determinants. *IEEE Transactions on Information Theory*, 51(4):1432–1436, April 2005.