Performance of Space-Time Block Codes with Finite Geometry LDPC Outer Codes

Raymond A. Brown, Sarah J. Johnson and Steven R. Weller

Abstract—Space-time block codes provide the maximum possible diversity advantage for wireless multiple-input, multiple-output (MIMO) systems when low complexity decoding is required at the receiver. However, space-time block codes do not provide a coding advantage and as such require the use of an outer code to provide such coding gain. In this paper we propose the use of finite geometry low-density parity-check (LDPC) codes as outer codes to the Alamouti space-time block code. We present modifications required to a space-time block decoder to provide a soft-information bit probability statistic output for sum-product decoding of LDPC codes and investigate the performance of space-time block codes with EG-LDPC outer codes in a Rayleigh flat fading environment.

Index Terms—space-time coding, block codes, Low-density parity-check codes, concatenated codes, data communication, diversity reception, fading channels, maximum likelihood decoding, multiple transmit antennas, cyclic codes, iterative decoding, finite-length analysis.

I. INTRODUCTION

High data rate wireless communication environments are subject to multi-path propagation in the form of destructive fading and superposition of signals received from different paths. Traditionally, antenna diversity techniques have been employed to reduce the effects of multi-path by providing the receiver with a less attenuated replica of the transmitted signal. Further, the use of multiple transmit and receive antennas has been shown to significantly increase the capacity of flat Rayleigh fading environments [1].

Receive antenna diversity utilizes simple signal combining to provide a significant performance advantage over multi-path channels with low receiver complexity. However with modern user terminals becoming smaller, simpler and less expensive, receive antenna diversity is almost always utilized at the base station for the uplink. Multiple transmit antennas can provide an advantage through spatial separation of antennas and when combined with coding techniques, can also provide a time delayed variant or replica of transmitted signals to the receiver. The feasibility of multiple antennas at the base station makes transmit diversity a more economical proposition by providing a diversity advantage for the down-link.

Space-time trellis codes [2] are one form of space-time code that provide diversity advantage and coding gain through the use of a trellis structure. This scheme uses coding techniques at the transmitter and signal processing at the receiver for decoding. One disadvantage with space-time trellis codes is that they have a decoding complexity which increases exponentially with transmission rate [2].

Alamouti devised a scheme [3] that provides full rate and full diversity advantage for complex signal sets and a low decoder complexity when compared to trellis codes. Tarokh et al. generalized Alamouti's scheme using orthogonal designs and such schemes are now known as space-time block codes [4]. Space-time block codes, despite their low decoding complexity and full diversity advantage, do not provide the coding advantage of trellis-based codes and require the use of an outer code to provide coding gain.

Low-density parity-check (LDPC) codes were first discovered by Gallager in 1962 [5] and, following their rediscovery and generalization by several researchers in the mid-1990s, have been the subject of intense research interest due to their near-Shannon limit performance in additive white Gaussian noise (AWGN) channels. LDPC codes are linear block codes that are well suited to iterative, soft-decision decoding using the sum-product algorithm [6] in a manner reminiscent of turbo decoders. One disadvantage of LDPCs is their potentially high encoder complexity [7].

Recently, Kou et al. showed how LDPC codes can be constructed based on the points and lines of a finite geometry. The cyclic, or quasi-cyclic, form of these finite geometry LDPCs allows linear-time encoding using only simple feedback shift registers [8].

In this paper we propose the use of finite geometry low-density parity-check (LDPC) codes as outer codes to the Alamouti space-time block code. We provide a review of space-time block codes and describe the system model in Section II. Section III provides an introduction and review of LDPC codes with reference to finite geometry and especially Euclidean geometry LDPC (EG-LDPC) codes. Section IV details the procedure for extracting the probability information from a space-time block decoder and provides details on the use of EG-LDPC codes as an outer code to the Alamouti space-time block code scheme. Simulation results are provided in Section V, and conclusions are presented in Section VI.

II. REVIEW OF SPACE-TIME BLOCK CODES AND SYSTEM MODEL

In this section, we review the space-time block code framework of Tarokh et al. [4], which generalized the remarkable two transmit antenna special case of Alamouti [3]. Space-time
block codes are based on the generalized theory of orthogonal designs [9], where a generalized complex orthogonal design $G$ is a $p \times N$ matrix in the indeterminates $x_1, \ldots, x_k$ satisfying the property

$$G^* G = \left( |x_1|^2 + |x_2|^2 + \cdots + |x_k|^2 \right) I,$$  

(1)

where $G^*$ is the complex conjugate transpose of $G$ and $I$ is an $N \times N$ identity matrix.

A wireless communications system employing space-time block codes consists of $N$ transmit antennas and $M$ receive antennas as shown in Fig. 1. The coefficients $h_{i,j}$ are the path gains from transmit antenna $i$ to receive antenna $j$. At time $t$, $k$ information symbols arrive at the space-time block encoder and each information symbol is mapped to a complex modulation constellation symbol resulting in signals $s_{i}, i = 1, \ldots, k$. Setting $x_{i} = s_{i}$ for $i = 1, \ldots, k$ in the orthogonal design $G$ we arrive at a matrix $C$ where the entries of $C$ correspond to linear combinations of the modulated symbols $s_{i}, i = 1, \ldots, k$ and their conjugates.

If the entries of $C$ are represented by the elements $c_{i}$, where $i$ represents the number of rows and $t$ represents the number of columns, the entries $c_{i}^*$, $i = 1, 2, \ldots, N$ are transmitted from the $N$ transmit antennas at each time slot $t = 1, 2, \ldots, p$ simultaneously.

At time slot $t = 1, 2, \ldots, p$ signal $r_{j}^t$ received at antenna $j$ is given by

$$r_{j}^t = \sum_{i=1}^{N} h_{i,j} c_{i}^t + \eta_{j}^t,$$  

(2)

where $\eta_{j}^t$ are zero-mean independent complex Gaussian random variables with dual-sideband uniform power spectral density (PSD) of $N_0/2$. Note that the average symbol energy is normalized to be $1/N$ for each of the $N$ receive antennas.

The space-time block code of Alamouti [3],

$$G_2 = \begin{pmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{pmatrix},$$  

(3)

is considered further due to its special properties. It has been proven to be the only complex orthogonal design to provide both full diversity and full rate [4]. All other complex orthogonal designs provide full-diversity but at a rate penalty. One set of complex space-time block codes forgo diversity in order to achieve full rate. These codes are known as quasi-orthogonal space-time block codes [10] which are based on designs that incorporate orthogonal subspaces.

In our system model we use the $G_2$ matrix with $N = 2$ transmit and $M$ receive antennas. The channel is assumed to be a quasi-static flat fading channel so that the path gains, $h_{i,j}$, are modelled as samples of independent complex Gaussian random variables with a variance of 0.5 per dimension. The path gains are assumed to be constant over a frame of length $f$ transmissions and vary from one frame to another.

Assuming perfect channel state information is available, decoding can be achieved using maximum likelihood detection in which a hard decision output is formed by minimizing the decision statistic

$$\sum_{j=1}^{M} \left( |r_{1}^t - h_{1,j}s_{1} - h_{2,j}s_{2}|^2 + |r_{2}^t + h_{1,j}s_{1}^* - h_{2,j}s_{2}^*|^2 \right)^2$$  

over all possible symbol values $s_{1}$ and $s_{2}$.

By expanding (4) and removing the terms that are independent of codewords, we can separate the metric into two parts, each depending only on $s_{1}$ and $s_{2}$ separately. For $s_{1}$ the decision metric is

$$\left[ \sum_{j=1}^{M} \left( r_{1}^t h_{1,j}^* + r_{2}^t h_{2,j}^* \right) \right] - s_{1}^2$$  

$$\left[ -1 + \sum_{j=1}^{M} h_{i,j}^2 \right] |s_{1}|^2, \quad (5)$$

and for $s_{2}$ the decision metric is

$$\left[ \sum_{j=1}^{M} \left( r_{1}^t h_{2,j}^* + r_{2}^t h_{1,j}^* \right) \right] - s_{2}^2$$  

$$\left[ -1 + \sum_{j=1}^{M} h_{i,j}^2 \right] |s_{2}|^2. \quad (6)$$

The decision metrics (5) and (6) provide a hard decision output that relies on a decision being made on the minimum Euclidean distance from received symbols to possible transmitted symbols. The minimization of (4) does not produce any performance penalty but at the same time does not provide any coding gain.

To obtain a coding gain an outer code that utilizes an iterative algorithm based on soft information is most suitable. For this we propose the use of LDPC codes, using the sum-product algorithm for decoding. The following section provides an introduction to LDPC codes, with emphasis on a particular class of cyclic LDPC codes known as Euclidean geometry (EG) LDPC codes. Section IV describes the modification to the space-time block decoder described above to provide a soft information output to the LDPC decoder.

III. LOW-DENSITY PARITY-CHECK CODES

Low-density parity-check codes are binary block codes defined by sparse parity-check matrices. The enormous benefit of LDPC codes is their error correction performance when decoded with the sum-product algorithm. Codes with very long blocklengths have been shown to perform within a fraction of a decibel of the relevant Shannon limit in an AWGN channel,
and to do this with reasonable decoding complexity. The sparsity of the parity-check matrix, \( H \), gives a complexity of decoding only linear in the blocklength. The parity-check matrices of LDPC codes have traditionally been defined pseudo-randomly, subject to the requirement that \( H \) be sparse, and code construction of binary LDPC codes involves randomly assigning a small number of the values in an all-zero matrix to 1. See [MacKay - March 1999 IT] for details on several pseudo-random constructions of LDPC codes.

LDPC codes are often described by their Tanner graph, which is a bi-partite graph representing the relationship between the code bits and the code parity-check equations, and so is dependent only on the choice of \( H \). A Tanner graph consists of a set of bit nodes, one for each code bit, and a set of check nodes, one for each parity-check equation. The \( j \)th bit node is joined by an edge to the \( i \)th check node if the \((i,j)\)th entry of \( H \) is non-zero. The edges in the Tanner graph represent the paths of information exchange between bits and parity-checks in the decoding process. An important property of LDPC codes is the length of the cycles in its Tanner graph. A cycle is a sequence of connected nodes which start and end at the same node in the graph and contain no other node more than once. The existence of short cycles in the graph reduces the performance of the sum-product decoding.

The aim of sum-product decoding is to compute the a posteriori probability (APP) for each codeword bit, \( P_i = P\{c_i = 1|N\} \), which is the probability that the \( i \)th codeword bit is a 1 conditional on the event \( N \) that all parity-check constraints are satisfied. The intrinsic or a priori probability, \( p_i \), is the original bit probability independent of knowledge of the code constraints, which depends only on the received signal and knowledge of the channel. The extrinsic probability \( E_i \) represents what has been learnt from the event \( N \).

The sum-product algorithm computes an approximation of the APP value for each code bit by iteratively calculating new extrinsic and estimated APP probabilities at the bit and parity-check nodes based on probabilities passed along the edges. Extrinsic information generated in one iteration is used as a priori information for the subsequent iteration. The extrinsic bit information obtained in one iteration is independent of the a priori values at the start of the iteration. The extrinsic information provided in subsequent iterations remains independent of the original a priori probabilities until that information is returned via a cycle which is why short cycles are avoided in LDPC codes.

The extrinsic probability, from the \( j \)th parity-check equation, that a codeword bit \( i \) is a one is the probability that that parity-check equation is satisfied if bit \( i \) is assumed to be a 1, which is the probability that an odd number of the other codeword bits are a 1.

\[
E_{i,j}^1 = \frac{1}{2} + \frac{1}{2} \prod_{l \in B_j, l \neq i} (1 - 2P_l),
\]  

(7)

and similarly for the probability that bit \( i \) is a 0.

\[
E_{i,j}^0 = \frac{1}{2} - \frac{1}{2} \prod_{l \in B_j, l \neq i} (1 - 2P_l).
\]  

(8)

The notation \( B_j \) represents the set of column locations of the bits in the \( j \)th parity-check equation of the code.

The APP estimate for a bit is the product of the a priori and extrinsic probabilities for that bit:

\[
P_i^1 = \alpha_i p_i \prod_{j' \in A_i} E_{i,j'}^1,
\]  

(9)

\[
P_i^0 = \alpha_i (1 - p_i) \prod_{j' \in A_i} E_{i,j'}^0,
\]  

(10)

where \( \alpha_i \) is chosen so that \( P_i^1 + P_i^0 = 1 \) and the set of row locations of the parity-check equations which check on the \( i \)th bit of the code is represented by \( A_i \).

Sum-product decoding begins with the estimated APP probabilities initialized to the a priori bit probabilities after which the equations (7)-(10) are iterated until a hard decision on the estimated APP probabilities produces a valid codeword, or until the maximum number of iterations have been reached. A modification of the above replaces equations (9) and (10) by \(|B_j|\) separate equations each one omitting the extrinsic probabilities for one of the parity-check equations in the product. These new values are then used in equations (7) and (8) so that the calculation of each new extrinsic probability is independent of its previous value; See [6] and [11] for a complete presentation of sum-product decoding.

From the above it is evident that for good decoding performances an LDPC code with a sparse parity-check matrix and without small cycles is required and there are many such codes available. However, one of the main concerns with LDPC codes is encoding complexity as for the most part LDPC codes are encoded via matrix multiplication for which complexity increases with the square of the blocklength. For this reason we consider a class of LDPC codes from finite geometries, called Euclidean geometry (EG) codes, which are cyclic and so are very easy to encode. The EG-LDPC codes have also shown excellent error correction performance over a wide range of blocklengths [8]. The EG-LDPC codes are derived from Euclidean geometries, and were first presented as one step majority-logic decodable codes in [12].

A two dimensional Euclidean geometry, \( \text{EG}(2,2^2) \), is a finite set of \( 2^{2^2} - 1 \) points and \( 2^{2^2} - 1 \) lines, with the conditions that each point is incident in exactly \( 2^4 \) lines, each line contains exactly \( 2^4 \) points, and every pair of points is contained in at most one line together. The incidence matrix of an \( \text{EG}(2,2^2) \) is a binary \( 2^{2^2} - 1 \times 2^{2^2} - 1 \) matrix with the \((i,j)\)th entry non zero if the \( j \)th point of the geometry is contained in the \( i \)th line. The incidence matrix of a Euclidean geometry describes the parity-check matrix of an EG-LDPC code. The properties of the \( \text{EG}(2,2^2) \) ensure that the shortest cycles in the resulting Tanner graph are of length six, as a pair of points can be contained in at most one line together. Further, the codes have a minimum distance of \( 2^4 + 1 \), they are regular with exactly \( 2^3 \) checks per bit and bits in each check, and finally they are cyclic which allows for simple encoding via a shift register circuit [8].

One of the reasons for the excellent performance of the EG-LDPC codes over other LDPC codes of the same size and rate is that their parity-check matrices have low ranks over \( GF(2) \). This means that they include an extra set of parity-check equations without reducing the rate of the code. The rank of the
$2^{2s} - 1 \times 2^{2s} - 1$ parity-check matrix from an EG$(2, 2^s)$ is $3^s - 1$ [8], and thus the code rate is

$$\frac{2^{2s} - 3^s}{2^{2s} - 1}.$$  

Choosing EG-LDPC codes as the outer codes for the Alamouti space-time block code scheme provides excellent error correction performance with linear complexity for both encoding and decoding. Encoding is achieved through a simple shift register circuit as in [13], while the inherent parallelism of the sum-product decoding algorithm on the Tanner graph structure is readily exploited for low complexity hardware implementations as in [14].

**IV. FINITE GEOMETRY LDPC WITH STBC**

The proposed use of EG-LDPC codes as an outer code to a space-time block coded wireless system is shown in Fig. 1. The system encodes information data as described in the previous section and then further encodes via a space-time encoder before transmission on multiple transmit antennas. In the conventional Alamouti scheme, received signals are combined and processed by the space-time block decoder as described in Section II. However this only provides a hard decision output using maximum-likelihood estimation. To provide APP information to the iterative sum-product decoder we require a modified structure to provide a soft information statistic. The sum-product algorithm in the LDPC decoder requires a-priori probability statistics dependant on knowledge of the received signal and channel.

To modify the space-time decoder to provide an APP soft output we consider as our starting point the work of Bauch [15], who utilized binary turbo codes as an outer code to the Alamouti $Q_2$ space-time block code scheme. Bauch derived a MAP decoding rule that delivered a logarithmic symbol probability statistic for each symbol, the derivation as follows: Using Bayes rule,

$$P(c_1, c_2 | r_1, r_2) = \text{const} \cdot p(r_1, r_2 | c_1, c_2) \cdot P(c_1, c_2),$$

where

$$p(r_1, r_2 | c_1, c_2) = \text{const} \cdot \exp \left(-\frac{1}{2\sigma^2} \sum_{j=1}^{M} \left( r_1^2 - \sum_{i=1}^{2} h_{1,j} g_{1,i} \right)^2 \right),$$

and, since $c_i$ are statistically independent,

$$P(c_1, c_2) = \prod_{i=1}^{2} P(c_i).$$

$P(c_1, \ldots, c_K)$ is a-priori information obtained from source statistics.

Expanding (11) and eliminating terms not associated with codewords, as shown in Section II, an expression is obtained that is comprised of terms dependant only upon each symbol.

Equation (11) can therefore be decoupled to form two expressions

$$\ln P(c_1 = x_m | r_1, r_2) = \text{const}$$

$$- \frac{1}{2\sigma^2} \left( \sum_{j=1}^{M} \left( r_1^* h_{1,j}^* + r_2^* h_{2,j} \right) - x_m \right)^2$$

$$+ \left( -1 + \sum_{j=1}^{M} \sum_{i=1}^{2} |h_{i,j}|^2 \right) |x_m|^2 + \ln P(c_1),$$

and

$$\ln P(c_2 = x_m | r_1, r_2) = \text{const}$$

$$- \frac{1}{2\sigma^2} \left( \sum_{j=1}^{M} \left( r_1^* h_{2,j}^* - r_2^* h_{1,j} \right) - x_m \right)^2$$

$$+ \left( -1 + \sum_{j=1}^{M} \sum_{i=1}^{2} |h_{i,j}|^2 \right) |x_m|^2 + \ln P(c_2),$$

for each transmitted symbol $x_m$, $m = 1, \ldots, K$, where $K$ is the number of symbols of the constellation in use.

Equations (14) and (15) only provide the logarithmic symbol probabilities and as such need modification to provide the a-priori bit probabilities required for the sum-product algorithm. Rewriting without logarithms we obtain

$$P(c_i = x_m | r_1, r_2) = \text{const} \cdot \exp \left(f_{c_i}(x_m)\right) \cdot P(c_i),$$

for $i = 1, \ldots, k$, where

$$f_{c_1}(x_m) = - \frac{1}{2\sigma^2} \left( \sum_{j=1}^{M} \left( r_1^* h_{1,j}^* + r_2^* h_{2,j} \right) - x_m \right)^2$$

$$+ \left( -1 + \sum_{j=1}^{M} \sum_{i=1}^{2} |h_{i,j}|^2 \right) |x_m|^2,$$

and

$$f_{c_2}(x_m) = - \frac{1}{2\sigma^2} \left( \sum_{j=1}^{M} \left( r_1^* h_{2,j}^* - r_2^* h_{1,j} \right) - x_m \right)^2$$

$$+ \left( -1 + \sum_{j=1}^{M} \sum_{i=1}^{2} |h_{i,j}|^2 \right) |x_m|^2.$$  

Noting that

$$\sum_{m=1}^{K} P(c_i = x_m | r_1, r_2) = 1, \ i = 1, \ldots, k,$$

it is easy to show that

$$P(c_i = x_m | r_1, r_2) = \frac{1}{1 + \sum_{l=1, l \neq i}^{K} \exp \left(f_{c_1}(x_l) - f_{c_i}(x_m)\right)},$$

(20)
Equation (20) provides symbol probabilities given the received signals and provides inherent normalization of the constants. Bit probabilities can simply be obtained by addition of the relevant symbol probabilities for the constellation in use. The bit probability ratios are then obtained by division in the usual manner.

\[
P(b_{i0} = 1) = P(c_1 = 1 | r_1, r_2) + P(c_1 = 3 | r_1, r_2) \]
\[
P(b_{i0} = 0) = P(c_1 = 0 | r_1, r_2) + P(c_1 = 2 | r_1, r_2) \]
\[
P(b_{i1} = 1) = P(c_1 = 2 | r_1, r_2) + P(c_1 = 3 | r_1, r_2) \]
\[
P(b_{i1} = 0) = P(c_1 = 0 | r_1, r_2) + P(c_1 = 1 | r_1, r_2) \]
\[
P(b_{i2} = 1) = P(c_2 = 1 | r_1, r_2) + P(c_2 = 3 | r_1, r_2) \]
\[
P(b_{i2} = 0) = P(c_2 = 0 | r_1, r_2) + P(c_2 = 2 | r_1, r_2) \]
\[
P(b_{i3} = 1) = P(c_2 = 2 | r_1, r_2) + P(c_2 = 3 | r_1, r_2) \]
\[
P(b_{i3} = 0) = P(c_2 = 0 | r_1, r_2) + P(c_2 = 1 | r_1, r_2) \]

The method described above allows for an a-priori probability statistic to be provided to the sum-product algorithm in a bitwise form while decoding by symbols. No performance penalty or gain is associated with using the soft-output decoding method just described.

V. SIMULATION RESULTS

In this section, we present the simulated performance of the EG-LDPC codes when used as an outer code to the modified Alamouti space-time block code scheme in a Rayleigh flat fading environment. We analyze the performance of the sum-product decoder with respect to the number of iterations required. We also make comparisons to the simple Alamouti space-time block code scheme without an outer code for the one and two receive antenna case and to a random construction LDPC code used as an outer code. The EG-LDPC codes used in these simulations consist of rate 0.46, 0.58 and 0.68 with block lengths 15, 63 and 255 respectively. The random code used in comparisons is a [63,37,4] code.

We present results for the case of two transmit antennas and one receive antenna and also for the case of two transmit and two receive antennas. The modulation scheme used throughout these simulations was QPSK using Gray mapping at each antenna. In each instance a number of bits equal to the least common multiple of the parity-check block length and the space-time frame size sent eliminating the need for data padding.

For the results shown without the use of an outer code, the Alamouti space-time block code, \(G_2\), was used with QPSK modulation and one or two receive antennas to achieve a spectral efficiency of 2 or 4 bits/s/Hz respectively. The results shown with the EG-LDPC codes used as outer codes were completed using the same Alamouti scheme, \(G_2\), with QPSK modulation, and one or two receive antennas. The spectral efficiencies of the 15, 63 and 255 block length EG-LDPC codes using QPSK modulation are 1.87, 2.34 and 2.75 bits/s/Hz respectively. The spectral efficiencies of the 15, 63 and 255 block length EG-LDPC codes using 16-QAM modulation are 3.73, 4.69 and 5.49 bits/s/Hz respectively.

Fig. 2 shows the performance of the \(G_2\) space-time block code using QPSK modulation with 2 receive antennas and the EG-LDPC[63,37,9] outer code and applying different iteration constraints to the sum-product algorithm. The channel was assumed to be constant over a frame size of \(l = 130\) transmissions. The results demonstrate that for a block length 63 code using QPSK modulation there is little to be achieved by iterating more than four times. Further simulations were completed using longer block lengths and with other modulation schemes. It was observed that the number of iterations required was proportional to the size of the code block length and was not dependent upon the constellation size.

Fig. 3 shows the bit-error rate of the \(G_2\) space-time block code using QPSK modulation with one receive antenna. Also shown is the performance of the EG-LDPC codes [15,7.5], [63,37,9] and [255,175,19] when used as outer codes with the same Alamouti scheme and QPSK modulation. In this case the channel coefficients were constant over \(l = 2\) transmissions (one space-time block) but varied between blocks. The results
show a significant gain advantage by using EG-LDPC codes, although there is a slight loss in spectral efficiency. It is further noted that even the use of a very short block length 15 code provides a gain of 10dB over the Alamouti scheme at a bit error rate of 10^{-4}. Fig. 3 also shows a comparison with a randomly generated LDPC code of blocklength 63 as an outer code; the significant error floor at high SNRs is attributed to the poor minimum distance of this code. Note also the poor performance of this randomly generated code in comparison with the same blocklength EG-LDPC code.

Fig. 4 shows the bit-error rate of the G2 space-time block code using QPSK modulation with two receive antennas. Also shown is the performance of the EG-LDPC codes [63,37,9] and [255,175,19] used as outer codes with the same Alamouti scheme and QPSK modulation. In this case the channel coefficients were constant for a frame of l = 130 transmissions (65 space-time blocks) and varied between frames. It can be seen that the performance has been degraded slightly but the EG-LDPC[255,175,19] code still achieves an almost 3dB improvement in performance over the Alamouti scheme at a bit error rate of 10^{-5}. The degraded performance is likely to be due to deep block fades and would benefit from the use of interleavers and/or larger block length codes.

VI. Conclusion

We have have studied the performance of LDPC codes, especially EG-LDPC codes, used as outer codes to a modified Alamouti space-time block code scheme. We have shown that large performance gains are possible using short codes with feasible low decoder complexity and with linear time encoding. Further work will investigate further the performance of LDPC codes as outer codes to space-time block codes in different channel environments, and with comparisons to other modulation formats and outer coding schemes.

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