New Trend in Space Time Error Correcting Codes

Said M. Elnoubi, Shawky E. Shabaan, Ahmed H. Abd El-Malek
Alexandria University, Egypt, 21544

***Abstract*— A new trend in space time error correcting codes denoted by Block turbo code space time error correcting (BTC-STECC) is proposed. This scheme overcomes the complexity of the detection stage in Space time error correcting code (STECC) by using the simple and familiar Alamouti STBC for transmission and detection. Also it uses the properties of BTC and its decoding procedures to make diversity in space, time and code. This provides a better performance than STECC in Rayleigh fading channels. BTC-STECC uses the Chase decoder algorithm to decode the BTC rows and column data. As MAP decoder is the optimal decoder, Chase decoder is a sub-optimal decoder with a high performance close to the optimal one but with less complexity which it is a drawback in STECC.**

***Index Terms*— Block turbo code, modified Chase algorithm, Multiple-input multiple-output, space-time error correcting codes.**

I. Introduction

Turbo codes [1] were a major development in the field of error control coding. These codes are very attractive due to their outstanding performance, very close to the limits of reliable communication given by Shannon limit. Block Turbo Code [BTC] has more powerful performance in fading channels than the Convolutional Turbo Codes (CTC) [2], [3].

Multiple-input multiple-output (MIMO) was firstly introduced by Alamouti in mid of 90's [4] and developed by Tarokh [5]. Due to the drawbacks of Alamouti Scheme new families were introduced later with good improvement in performance and spectral efficiency like Layered Space time code (LST) and other new schemes [6], [7].

In [8] a new family of space time coding technique was introduced with high performance and high data rate, this family was called Space Time Error Correcting Codes (STECC). One of drawbacks of STECC is the complexity of its receiver which increases linearly with the modulation order [9].

This letter introduces a new family called Block Turbo Code Space Time Error Correcting Codes (BTC-STECC). This family combines the advantages of both of STECC and BTC. Also BTC-STECC overcomes the detection complexity problem as it involves the Alamouti Space Time Block Code in transmission and detection. BTC-STECC is a recent FEC scheme [10], [11]. BTC uses iterative SISO decoder using modified Chase algorithm [12],[13] to make use of complexity reduction in decoding stage provided by Chase algorithm. One of the main advantages of BTC is the interleaving stage. As the BTC performance is almost the same with different interleaving schemes [14] a simple block interleaver can be used.

BTC's are very flexible in terms of performance complexity and code rate. Constituent codes can be mixed and matched to achieve desired code characteristics. They can support any block size, a very wide range of code rates from below rate 1/3 to as high as rate 0.98. Code shortening enhances this flexibility. Turbo product codes provide excellent performance

at high code rates and can offer a wide range of block sizes

and code rates with change in coding strategy. Further, block codes have decoders that can operate at very high speeds

This letter is organized as follows. Part II presents the BTC-STECCs construction. Part III describes the MIMO transmission model. Part IV deals with the corresponding receiver structure. Part V describes the performance on a Rayleigh block flat fading channel while part VI concludes the letter.

II. DEFINITION

A. Construction

A BTC-STECC is a space-time block turbo code able to correct errors due to transmission in fading channels. It is built on the principal of turbo codes and two linear forward error correcting codes.

In this letter, a parallel concatenation block turbo code with uniform interlacing is used. Let $C_1(n_1,k_1)$ and $C_2(n_2,k_2)$ be the two linear block codes used to construct the parallel concatenation block turbo codes where n denotes for the code length and k denotes the data dimension [14]. Each block is written row by row in a matrix of k_1 columns and k_2 rows. Each row is systematically encoded as block by adding $(n_1\text{-}k_1)$ parity bits and each column is encoded as a block by adding $(n_2\text{-}k_2)$ parity bits as shown in equation (1).

$$C=\begin{bmatrix} d_{11} & d_{12} & \cdots & d_{1k_1} & p_{11} & p_{12} & \cdots & p_{1(n_1-k_1)} \\ d_{21} & d_{22} & \cdots & d_{2k_1} & p_{21} & p_{22} & \cdots & p_{2(n_1-k_1)} \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ d_{k_2 1} & d_{k_2 2} & \cdots & d_{k_2 k_1} & p_{k_2 1} & p_{k_2 2} & \cdots & p_{k_2(n_1-k_1)} \\ p_{11} & p_{12} & \cdots & p_{1k_1} & & & & \\ p_{21} & p_{22} & \cdots & p_{2k_1} & & & & \\ \vdots & \vdots & \vdots & \vdots & & & & \\ p_{(n_2-k_2)1} & p_{(n_2-k_2)2} & \cdots & p_{(n_2-k_2)k_1} & & & & \end{bmatrix} \quad (1)$$

The rate of parallel concatenation block turbo code is

$$R_C=\frac{R_1R_2}{R_1+R_2} \qquad (2)$$

Where R_1 is the code rate for code $C_1(n_1,k_1)$ and R_2 is the code rate for code $C_2(n_2,R_2)$.In this letter we assume the two linear error correcting codes have the same parameters $C_1(n_1,k_1)=C_2(n_2,k_2)=C(n,k)$ then the parallel concatenation block turbo code will be denoted by $C^2(n,k)$ and the code rate in (2) and due to double transmission of information data bits it will be

$$R_{C^2} = \frac{R^2}{2R} = \frac{R}{2} \qquad (3)$$

The row encoder and column encoder output codewords are obtained using M-order binary to symbol converter then the transmitted symbols are transmitted to Alamouti STBC to obtain the transmission matrix (4) for $n_t = 2$ where n_t is number of transmitted antennas.

$$S = \begin{bmatrix} s_1^- & s_2^- & \cdots & s_L^- \\ s_1^| & s_2^| & \cdots & s_L^| \end{bmatrix}$$

$$S_{STBC} = \begin{bmatrix} s_i^- & -s_i^{|*} \\ s_i^| & s_i^{-*} \end{bmatrix} \qquad (4)$$

Where $1 \le i \le L$, s_i^- denotes symbols obtained by M-ary Quadrature Amplitude Modulator (M-QAM) for row codewords and $s_i^|$ denotes symbols obtained by M-QAM modulator for column codewords. The M-QAM modulator converts m bits of codewords into one symbol as $= 2^{m_b}$. The transmitted symbol matrix has dimension of $n_t \times 2L$ where $L = \frac{n \times k}{m_b}$, and if $n \times k$ is not divisible by m we put zero padding at the end of the code matrix C.

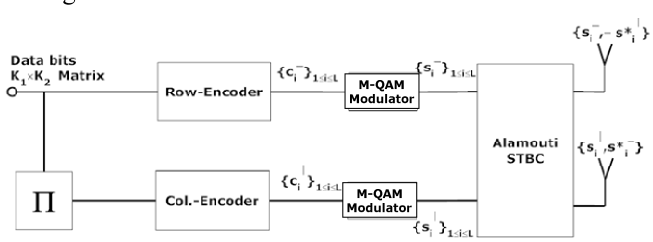

Fig. 1. Transmit Scheme for the $C^2(n,k)$ BTC-STECC with $n_t = 2$

The overall coding rate of BTC-STECC is equal to the coding rate of block turbo code

$$R_{BTC-STECC} = \frac{1}{2} R_{C^2} = \frac{R}{4} \qquad (5)$$

B. Uniform Interleaving Vs Random Interleaving

1) Principal of Interleaving:

Interleaving is a technique commonly used in communication systems to overcome correlated channel noise causing burst error or fading. The interleaver rearranges input data such that consecutive data are split among different blocks. At the receiver end, the interleaved data is arranged back into the original sequence by the de-interleaver. As a result of interleaving, correlated noise introduced in the transmission channel appears to be statistically independent at the receiver and thus allows better error correction

2) *Performance:*

One of the BTC advantages is that the performance of simple row/column interleaver is as good as random interleaving. This allows the use of simpler interleaving structure and reduces the system complexity [14].

III. TRANSMISSION MODEL

We consider a MIMO transmission with $n_t = 2$ transmit antenna and n_r receive antennas. We assume Rayleigh block flat fading channel (i.e. the complex channel matrix H is constant over τ modulation symbols). The received signal is given by:

$$R = HS_{STBC} + N \qquad (6)$$

where R is the $n_r \times 2L$ complex received matrix, S_{STBC} is the $n_t \times 2L$ complex transmit matrix, H is the $n_r \times n_t$ complex channel matrix and N is the $n_r \times 2L$ complex AWGN matrix. Assuming a unitary average gain on each transmit-received antenna link, the signal to noise ratio (SNR) on each receive antenna is given by the following formula [8]

$$SNR = m_b n_t R_{BTC-STECC} \frac{E_b}{N_0} \qquad (7)$$

where E_b is the energy per useful bit and N_0 is the noise.

IV. RECEIVER STRUCTURE

The channel state information is supposed to be perfectly known at the receiver. For the BTC-STECC $C^2(n,k)$, the optimal receiver structure must find the K codewords $\{\hat{c}_1, \hat{c}_2, \ldots, \hat{c}_K\}$ of $C^2(n,K)$ that minimize the distance $\|R - HS_{STBC}\|^2$. By doing so, it performs the detection and the decoding jointly; taking in consideration the advantage of Alamouti detection scheme. The process is divided into two successive operations: a detection stage and a decoding stage as shown in figure 2.

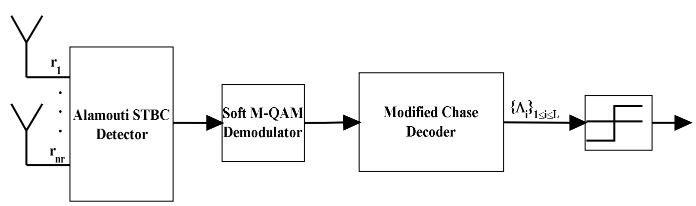

Fig. 2. Receiver structure with optimal detection for BTC $C^2(n,k)$ and n_r receive antennas

A. Space Time Decoding (Maximum Likelihood)

1) Alamouti Soft Optimal detection: Using the familiar Alamouti STBC detector, the modulated symbols $\{s_i^-\}_{1\le i\le L}$ and $\{s_i^|\}_{1\le i\le L}$ can be easily detected. One of the advantages of Alamouti STBC detector is its low-complexity in detecting the data over STECC detector as it detects the transmitted symbols one by one using Alamouti Simple Receiver or Maximum Likelihood (ML) combining. It does not need all the FEC symbols. Alamouti Simple Receiver can be used for detection of the estimated received symbols using detection equations in [4]. i.e. At the receive antenna for 2×2 Alamouti scheme, the received signals over two consecutive symbol periods, denoted by r_1 and r_2 for time t and $t + T$, respectively, can be expressed as

$$r_1 = h_1 s_1 + h_2 s_2 + n_1 \quad (8)$$
$$r_2 = -h_1 s_2^* + h_2 s_1^* + n_2 \quad (9)$$

where n_1 and n_2 are independent complex variables with zero mean and power spectral density $N_0/2$ per dimension, representing additive white Gaussian noise samples at time t and $t + T$, respectively.

Assume the channel fading coefficients, h_1 and h_2, can be perfectly recovered at the receiver, the decoder will use them as the channel state information (CSI). Assuming that all the signals in the modulation constellation are equiprobable, the maximum likelihood decoder chooses a pair of signals $(\hat{s}_1, \hat{s}_2)$ from the signal modulation constellation to minimize the distance metric

$$d^2(s_1, \ldots, s_L) = \sum_{i=1}^{2} \|R_i - H_i \hat{s}_{STBC}\|^2$$

$$d^2(r_1, h_1\hat{s}_1 + h_2\hat{s}_2) + d^2(r_2, -h_1\hat{s}_2^* + h2\ \hat{s}_1^*)$$
$$= |r_1 - h_1\hat{s}_1 - h_2\hat{s}_2|^2 + |r_2 + h_1\hat{s}_2^* - h2\,\hat{s}_1^*|^2 \quad (10)$$

where R_i is the received vector. H_i is the MIMO channel matrix at the time $t = i$, with $1 \le i \le L$. over all possible values of $\hat{s}_1$ and $\hat{s}_2$. Substituting (8) and (9) into (10), the maximum likelihood decoding can be represented as

$$(\hat{s}_1, \hat{s}_2) = \arg \min_{(\hat{s}_1, \hat{s}_2) \in C} (|h_1|^2 + |h_2|^2 - 1)(|\hat{s}_1|^2 + |\hat{s}_2|^2) + d^2(\tilde{s}_1, \hat{s}_1) + d^2(\tilde{s}_2, \hat{s}_2) \quad (11)$$

where C is the set of all possible modulated symbol pairs $(\hat{s}_1, \hat{s}_2)$, $\tilde{s}_1$ and $\tilde{s}_2$ are two decision statistics constructed by combining the received signals with channel state information. The decision statistics are given by

$$\tilde{s}_1 = h_1^* r_1 + h_2 r_2^* \quad (12)$$
$$\tilde{s}_2 = h_2^* r_1 - h_1 r_2^* \quad (13)$$

Substituting r_1 and r_2 from (8) and (9), respectively, into (12) and (13), the decision statistics can be written as,

$$\tilde{s}_1 = (|h_1|^2 + |h_2|^2)s_1 + h_1^* n_1 + h_2 n_2^* \quad (14)$$

$$\tilde{s}_2 = (|h_1|^2 + |h_2|^2)s_2 - h_1 n_2^* + h_2^* n_1$$

For a given channel realization h_1 and h_2, the decision statistics $\tilde{s}_i$, $i = 1, 2$, is only a function of s_i, $i = 1, 2$. Thus, the maximum likelihood decoding rule (11) can be separated into two independent decoding rules for s_1 and s_2, given by

$$\hat{s}_1 = \arg \min_{\hat{s}_1 \in S} (|h_1|^2 + |h_2|^2 - 1)|\hat{s}_1|^2 + d^2(\tilde{s}_1, \hat{s}_1)$$

and (15)

$$\hat{s}_2 = \arg \min_{\hat{s}_2 \in S} (|h_1|^2 + |h_2|^2 - 1)|\hat{s}_2|^2 + d^2(\tilde{s}_2, \hat{s}_2)$$

Respectively. For M-PSK signal constellations, $(|h_1|^2 + |h_2|^2 - 1)|\hat{s}_i|^2, i = 1, 2$, are constant for all signal points, given the channel fading coefficients. Therefore, the decision rules in (8) can be further simplified to

$$\hat{s}_1 = \arg \min_{\hat{s}_1 \in S} d^2(\tilde{s}_1, \hat{s}_1)$$

and (16)

$$\hat{s}_2 = \arg \min_{\hat{s}_2 \in S} d^2(\tilde{s}_2, \hat{s}_2)$$

This detector is called the maximum Likelihood detector [4]. The Log Likelihood Ratios (LLRs) and their approximations can be easily derived.

2) Soft M-QAM Demodulator: As M-QAM modulator is used in the transmitter, it is needed to reconstruct the received bits using soft M-QAM. It extracts the received bits with its corresponding LLR values. These LLR values are used by SISO decoder to extract the information bits.

For QPSK soft demodulation, the soft output of the decoder represents the log likelihood ratio (LLR) of the probabilities of a bit being zero and one. Therefore;

$$soft_{bit} = \log\left(\frac{p}{1-p}\right) \quad (17)$$

where p is the probability of the bit being one. We can express the probability for estimated symbol $\hat{s}$ using the calculated Euclidean distance

$$P(\hat{s}|\tilde{s}_i) = \frac{1}{2\pi\sigma_N^2} e^{-\left(\frac{d^2(\tilde{s}_i,\hat{s})}{2\sigma_N^2}\right)} \quad (18)$$

where $\hat{s}$ is the estimated symbol depending on the modulation technique. For QPSK modulation $\hat{s} = 0, 1, 2,$ or 3, and $\tilde{s}_i$ is the statistic decision where $i = 1, 2$.

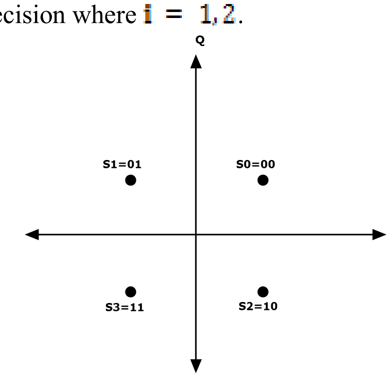

Fig. 3. QPSK modulation signal constellation

The Log Likelihood Ratio (LLR) for the bits in QPSK modulation technique could be written:

$$b_0 = \ln\left(\frac{P(1|s_i) + P(3|s_i)}{P(0|s_i) + P(2|s_i)}\right) \quad (19)$$

and

$$b_1 = \ln\left(\frac{P(2|s_i) + P(3|s_i)}{P(0|s_i) + P(1|s_i)}\right) \qquad (20)$$

from (19) and (20) we can simplify the soft decision rules to be

$$b_0 = \mathrm{Re}\{\hat{s}_i\} \quad \text{and} \quad b_1 = \mathrm{Im}\{\hat{s}_i\} \qquad (21)$$

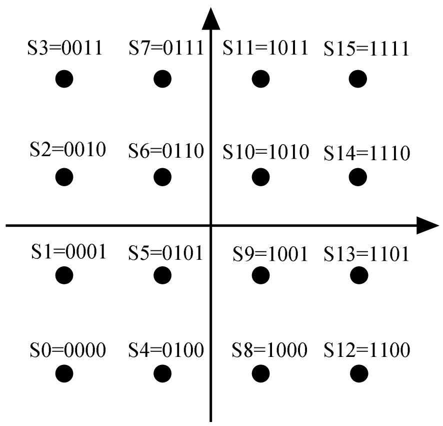

Fig. 4. 16-QAM signal constellation

For 16-QAM modulation $\hat{s} = 0, 1, 2, \text{or } 16$, and $\hat{s}_i$ is the statistic decision where $i = 1, 2$.

$$b_0 = \ln\left(\frac{\sum_{j=8}^{15} P(j|s_i)}{\sum_{j=0}^{7} P(j|s_i)}\right) \qquad (22)$$

$$b_2 = \begin{cases} \ln\left(\frac{P(10|s_i) + P(11|s_i) + P(14|s_i) + P(15|s_i)}{P(8|s_i) + P(9|s_i) + P(12|s_i) + P(13|s_i)}\right), b_0 \geq 0 \\ \ln\left(\frac{P(2|s_i) + P(3|s_i) + P(6|s_i) + P(7|s_i)}{P(0|s_i) + P(1|s_i) + P(4|s_i) + P(5|s_i)}\right), b_0 < 0 \end{cases} \qquad (23)$$

For $b_0 \geq 0$, the two bits b_1 and b_3 are depending on b_2 such as

$$b_1 = \begin{cases} \ln\left(\frac{P(14|s_i) + P(15|s_i)}{P(10|s_i) + P(11|s_i)}\right), b_2 \geq 0 \\ \ln\left(\frac{P(12|s_i) + P(13|s_i)}{P(8|s_i) + P(9|s_i)}\right), b_2 < 0 \end{cases} \qquad (24)$$

and

$$b_3 = \begin{cases} \ln\left(\frac{P(11|s_i) + P(15|s_i)}{P(10|s_i) + P(14|s_i)}\right), b_2 \geq 0 \\ \ln\left(\frac{P(9|s_i) + P(13|s_i)}{P(8|s_i) + P(12|s_i)}\right), b_2 < 0 \end{cases} \qquad (25)$$

While $b_0 < 0$

$$b_1 = \begin{cases} \ln\left(\frac{P(6|s_i) + P(7|s_i)}{P(2|s_i) + P(3|s_i)}\right), b_2 \geq 0 \\ \ln\left(\frac{P(4|s_i) + P(5|s_i)}{P(0|s_i) + P(1|s_i)}\right), b_2 < 0 \end{cases} \qquad (26)$$

and

$$b_3 = \begin{cases} \ln\left(\frac{P(3|s_i) + P(7|s_i)}{P(2|s_i) + P(6|s_i)}\right), b_2 \geq 0 \\ \ln\left(\frac{P(1|s_i) + P(5|s_i)}{P(0|s_i) + P(4|s_i)}\right), b_2 < 0 \end{cases} \qquad (27)$$

Similarly we can use the same method for 64-QAM soft demodulator depending on its signal constellation.

B. Decoding

1) Principle: Decoding stage is the bottleneck of BTC-STECC, as it is responsible for the effect of BTC encoding before the transmission. As we transmit the information bits by two different codes, space and time. We use modified chase algorithm instead of MAP algorithm to reduce complexity of decoding process [14].

2) Chase Decoder: Chase decoder [13],[17] is one of SISO decoder for constituent codes. Upon receiving a soft decision vector R of length n for a given constituent block code, a binary vector Y and a set of test patterns Z_i are formed from the soft input vector in the Chase decoder. The forming of test patterns is performed by a Chase algorithm within the Chase decoder. A hard-input hard-output (HIHO) block decoder is used to decode each binary vector $X_i = Y + Z_i$. If the HIHO decoder is successful the resulting codeword C_i from the hard decoding of X_i is saved in a set Ω. In addition to each codeword in Ω, an associated metric is saved, where the associated metric can be computed from the soft input, the test pattern and the resulting codeword. The modified chase algorithm is shown in Figure 5.

Modified Chase algorithm:
Start
Loading the observation data $R = (r_1, r_2, \ldots, r_n)$.
Calculating the vector $Y = (y_1, y_2, \ldots, y_n)$ with $y_i = sgn(r_i)$.
Searching for the positions of the least reliable component of Y.
Determining the codeword in Ω.
For $i = 0$ with $i < N_{dec}$

$$Z^i = (z_1^i, z_2^i, \ldots, z_n^i): \begin{cases} z_j^i = -y_j \text{ if } (j = i \text{ and } i \neq 0) \\ z_j^i = y_j \quad \text{otherwise} \end{cases}$$

X^i = Binary decoding (Z^i)
If X^i is a codeword, then $\begin{cases} \Omega = \Omega \cup X^i \\ met(i) = d_E(R, X^i) \end{cases}$
$D = (d_1, d_2, \ldots, d_n)$: codeword in Ω associated with the smallest Euclidean distance.
End

Fig. 5. Chase decoding algorithm

Figure 6 shows the SISO Chase decoder which is used to decode both of row codewords or column codewords. The soft output Chase decoder calculates the soft output for each bit j in the received codeword based on two or more codewords in Ω and the associated metrics. One codeword is the best estimated D, which is practically found unless the set Ω is empty

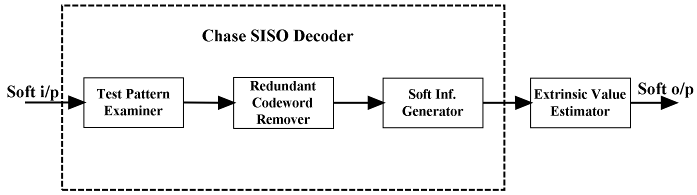

Fig. 6. SISO Chase Decoder

Figure 7 is a flow diagram of the process of examining the test patterns of chase decoder. Then the output of the examiner is applied to redundant codeword remover in order to reduce the number of probable codewords as shown in Figure 8. Figure 9 shows how the soft information is generated according to the minimum Likelihood distance. The soft output from the generator is applied to the extrinsic value estimator to evaluate the soft output value from the rule

$$\Lambda_i = R + L_e \qquad (28)$$

Figure 10 shows the iterative SISO decoding using Modified Chase Algorithm.

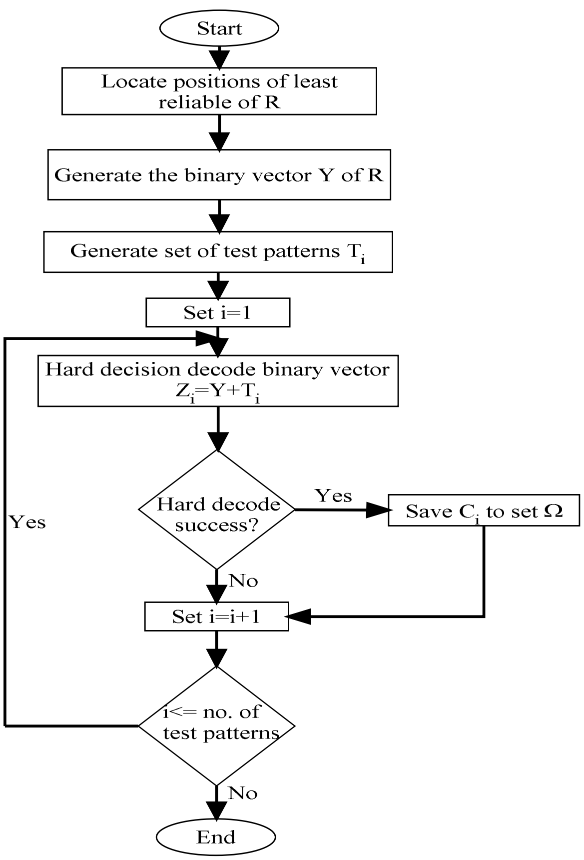

Fig. 7. Test pattern Examiner Flow Diagram

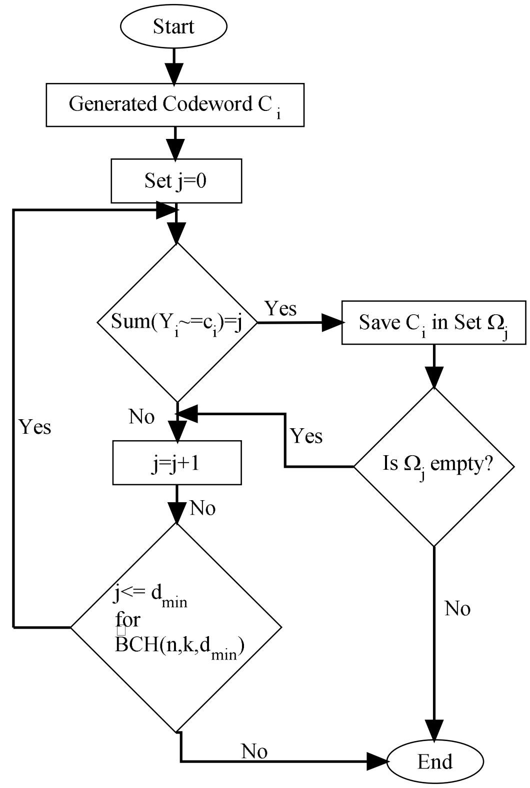

Fig. 8. Redundant codeword remover flow diagram.

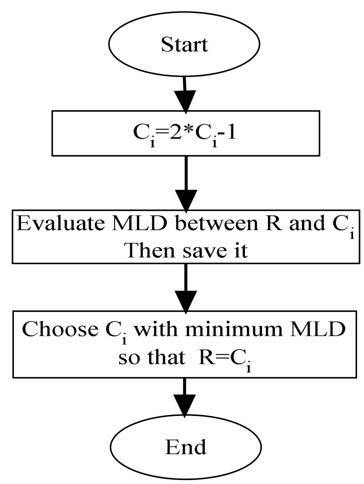

Fig. 9. Soft information generator flow diagram

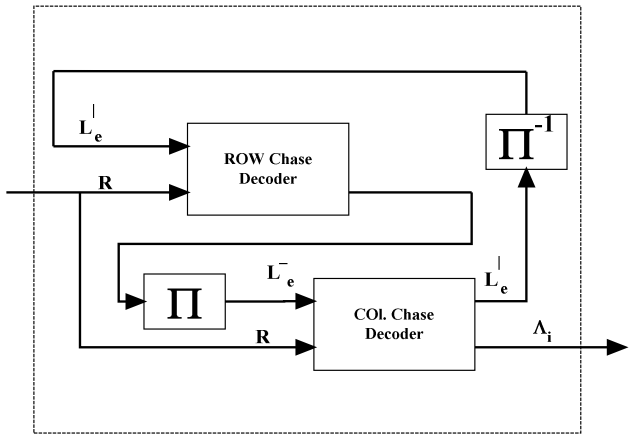

Fig. 10. Iterative SISO decoding using Modified Chase Algorithm

$$\Lambda_i = R + L_e^- + L_e^{|} \qquad (29)$$

Where Λ_i is the LLR value of SISO decoder, R the M-QAM demodulator soft output, L_e^- is the extrinsic information form row decoder, and $L_e^{|}$ is the extrinsic information from column decoder.

C. Complexity

The main disadvantage of STECC is the M^K complexity of its receiver. The detection complexity increases linearly with the modulation order M, where K is the number of transmitted codewords [8]. While in BTC-STECC, because of using Alamouti simple receiver the complexity of the receiver equals to M^2 and reduced to be 2M as we decode the two received signals separately.

V. Simulation Results

Consider the BTC-STECC $C^{BCH(15,11)}$ with Rayleigh block flat fading channel constant over τ modulation symbols $= 2$. The BER versus E_b/N_0 at the output of the combiner stage is plotted is Figure 11. For $E_b/N_0 \leq 1\,dB$ the performance of BTC-STECC is almost as the STECC performance [8]. For $E_b/N_0 \geq 1\,dB$, the performance gets better with different K of STECC. At a BER of 10^{-3} the improvement is 2.25 dB. At a BER of 2×10^{-4} the improvement is 1.25 dB between the BTC-STECC and the STECC with $K = 6$ curves. STECC can not increase K as the complexity of the receiver gets higher and becomes not practical.

Figure 12 shows the performance of BTC-STECC in different fading channels, as τ increases the performance of BTC-STECC becomes worst.

Figure 13 shows that the performance of BTC-STECC in Rayleigh fading channels with different τ assuming the BTC codewords length is 256. Figure 13 shows that the performance of BTC-STECC is better than the STECC performance for different Rayleigh fading channels.

Figure 14 shows the performance of BTC-STECC with different number of iterations for Modified Chase Algorithm. Performance improves with increasing decoding iteration order. However, improvement in performance decreases with increasing number of iteration order and becomes insignificant for number of iteration higher than 5 for a BER of 10^{-4}.

Figure 15 shows a comparison between 2x2 Alamouti scheme BTC-STECC with different modulation techniques. At BER equal to 10^{-2} QPSK modulation provides better performance over Rayleigh fading channel than 16 QAM modulation. The improvement provided QPSK modulation over 16 QAM modulation is about 3.5 dB.

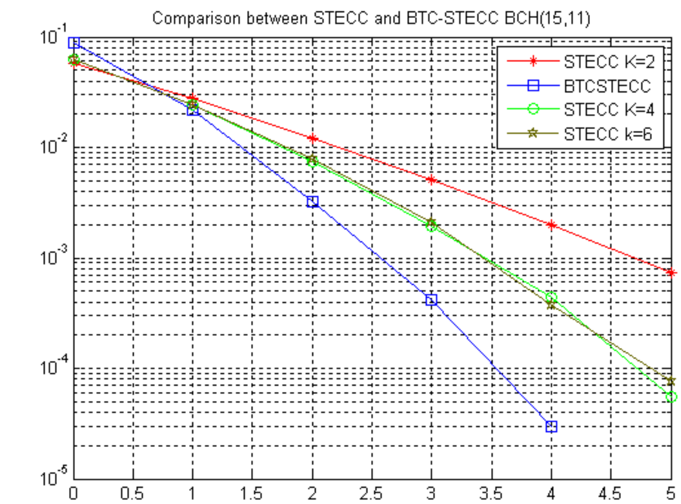

Fig. 11. Comparison between STECC and BTC-STECC, QPSK, $n_t = 2, n_r = 2, \tau = 2$

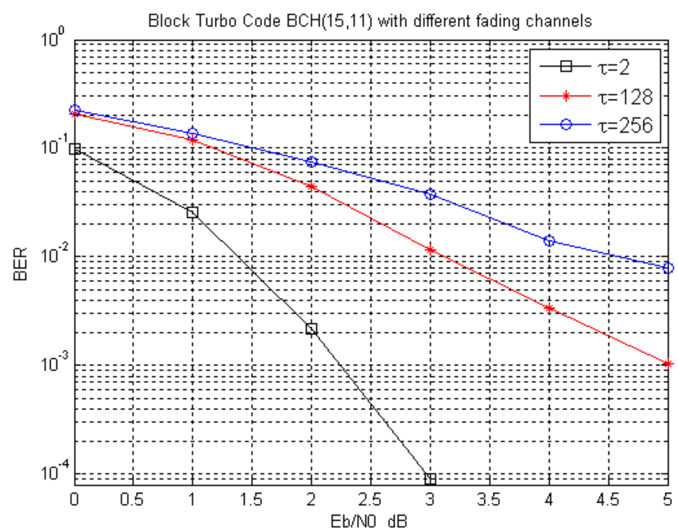

Fig. 12. BTC-STECC BCH (15, 11) with different Rayleigh Fading Channel

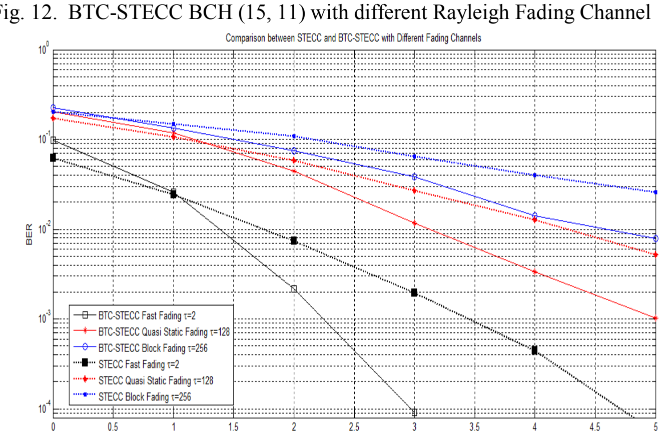

Fig. 13. Comparison between STECC and BTC-STECC with Different Fading Channels

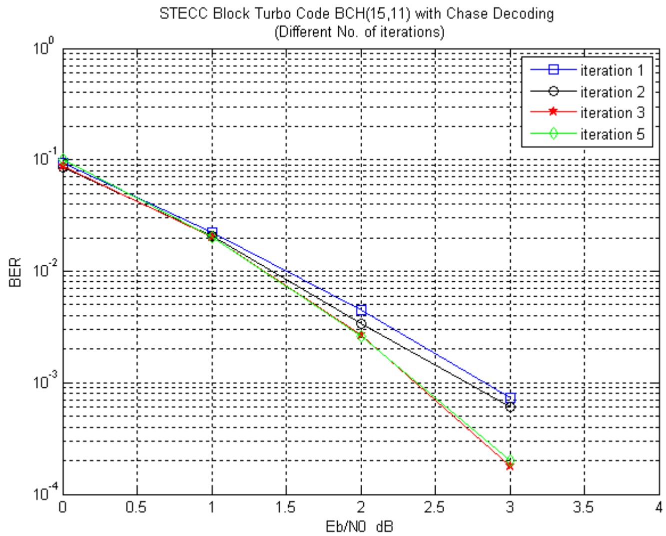

Fig. 14. BTC-STECC [BCH (15, 7, 3)2] with Chase Decoding algorithm with different number iterations

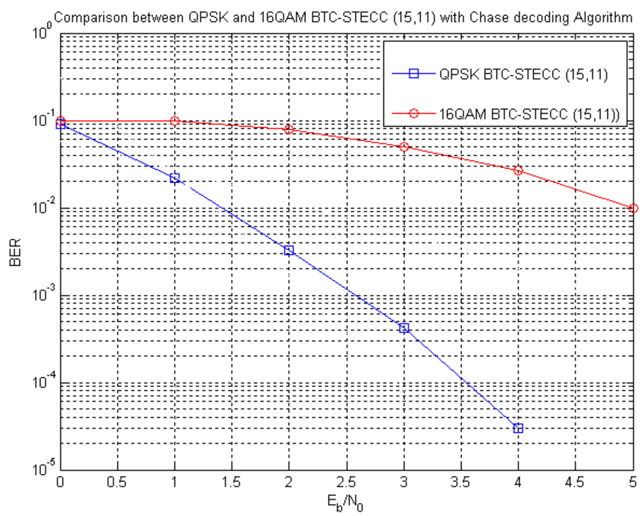

Fig. 15. Comparison BTC-STECC [BCH (15, 11, 3)2] with QPSK and 16QAM modulation techniques

Figure 16 shows a comparison between Alamouti simple receiver and maximum Likelihood (ML) detector for 2x2 Alamouti scheme BTC-STECC. The performance of ML is slightly better than the performance of Alamouti simple receiver. Therefore, we can use Alamouti simple receiver in space time decoding process in order to reduce the receiver complexity.

VI. CONCLUSION

BTC-STECC introduced in this letter provides a good performance in different fading channels in comparison with STECC with less complexity detection and decoding stages. BTC-STECC uses the advantages of simplicity in detection of Alamouti STBC beside the advantages of block turbo codes good performance and less complex decoding. STECC depends on the FEC linear properties and correlation between received symbols to minimize the BER. BTC-STECC makes a simple decoder and shifts the process of detection and correction to the BTC-decoder which is less complex than STECC decoder. BTC-STECC decoder uses Chase decoding algorithm to detect and correct errors. Chase decoder is a sub-optimal decoder, less complex than the MAP decoder and gives a good output performance nearly to the performance of optimal decoder. One of the drawbacks of BTC-STECC is its low rate compared with STECC.

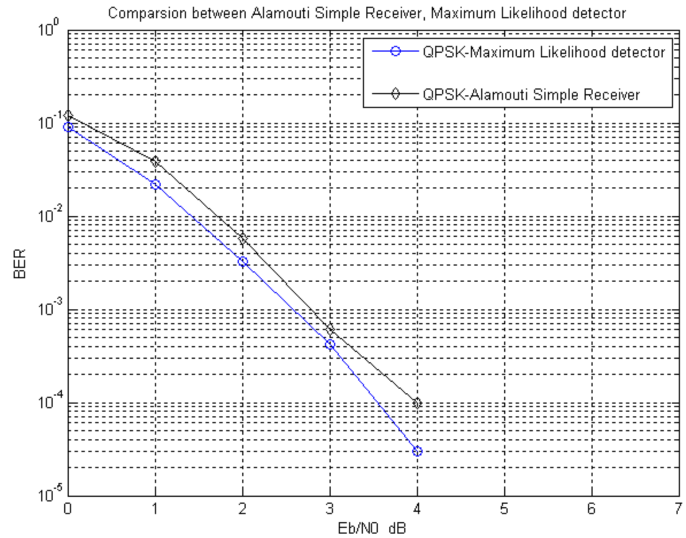

Figure 16: Comparison between Alamouti simple receiver and maximum Likelihood detector.

REFERENCES

[1] C. Berrou, A. Glavieux, and P. Thitimajshima, "*Near Shannon Limit Error Correcting Coding and Decoding: Turbocodes (1)*", IEEE Int. Conf. on Comm. ICC'93, pp. 1064-1071, vol. 2/3, May 1993.

[2] K. Ramasamy, M.U. Siddiq, M.Y. Alias, and A. Arunagiri, "*Performance comparison of convolutional and block turbo code*", IEICE Electronics Express, Vol.2 , No. 13, pp. 322-327, 2006.

[3] Lajos Hanzo, T. H. Liew, and B. L. Yeap, "*Turbo Coding, Turbo Equalization and Space -Time Coding for Transmission over Fading Channels*", July 2002, Wiley-IEEE Press.

[4] S. M. Alamouti, "*A simple transmit diversity technique for wireless communications*," IEEE J. Select. Areas Commun., pp. 1451-1458, Oct.1998.

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

[6] G. J. Foschini, "*Layered space time architecture for wireless communication in a fading environment when using multi-element antennas*," Bell Labs Techn. J., Autumn 1996.

[7] H. Gamal and A. Hammons, "*A new approach to layered space-time coding and signal processing*," IEEE Trans. Inform. Theory, vol. 47, pp. 2321–2334, Sept. 2001.

[8] M. Lalam, K. Amis, and D. Leroux, "*Space-time error correcting codes"*, IEEE Transactions on Wireless Communications, pp.1472-1476 May 2008.

[9] M. Lalam, K. Amis, and D. Leroux, "*Iterative decoding of space-time error correcting codes,*" in Proc. IEEE 18th International Symposium on Personal Indoor and Mobile Radio Communications, PIMRC'07, pp. 1–5, Sept. 2007

[10] R. C. Bose and D. K. Ray Chaudhuri. "*On a Class of Error Correcting Binary Group Codes*", Inf. Control, 3, pp. 68-79, March 1960

[11] Robert H. Morelos-Zaragoza, "*The Art of Error Correcting Coding*", 2nd edition, Wiley,2006

[12] D. Chase, "*A Class of Algorithms for Decoding Block Codes with Channel Measurement Information*", IEEE Trans. Inform. Theory, vol. IT-18, pp. 170-182, January 1972.

[13] S. Dave, J. Kim, and S. C. Kwatra "*An Efficient Decoding Algorithm for Block Turbo Codes*", IEEE Trans. Communications, Vol.49, No.1, pp 41-46, January 2001.

[14] A. Glavieux, "*Channel Coding in Communication Networks From Theory to Turbocodes*", ISTE Ltd, 2007

[15] H. Zerrouki, and M. Feham, "*High Throughput of WiMAX MIMO-OFDM Including Adaptive Modulation and Coding",* International Journal of Computer Science and Information Security, Vol.7, No.1,pp. 86- 90, 2010.

[16] B. Vucetic and J. Yuan, "*Space-Time Coding*", Willey, 2003.

[17] R. M. Pyndiah, "*Near-optimum decoding of product codes: Block turbo codes*", IEEE Trans. Communications, Vol.46, pp.1003-1010, Aug. 1998

[18] M. Surendra Raju, A. Ramesh, and A. Chockalingam , "BER Analysis of QAM with Transmit Diversity in Rayleigh Fading Channels", in Proc. IEEE Global Telecommunications Conference, 2003 (GLOBECOM '03), vol. 2, 1-5 Dec. 2003, pp. 641-645.