Performance Analysis of MIMO Systems using TCM and Comparison with OSTBC

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Abstract

Emerging demands for high data rate services and high spectral efficiency are the key driving forces for the continued technology evolution in wireless communications. MIMO technology has attracted attention in wireless communications, because it offers significant increases in data throughput and link range without additional bandwidth or transmit power. It achieves this by higher spectral efficiency (more bits per second per hertz of bandwidth) and link reliability or diversity (reduced fading). Orthogonal space-time block codes (STBC's) have received considerable attention in recent open-loop multiple-input-multiple-output (MIMO) wireless communication because they allow low decoding complexity and guarantee full diversity. Trellis-Coded Modulation (TCM) is a kind of channel coding with improving the coding gain. Therefore, concatenation of STBC and TCM is a combined channel coding to achieve both diversity gain and coding gain advantages. This paper presents a detailed study of space-time block coding (STBC) schemes including orthogonal STBC for 3×4 antennas and high-coding rate STBC. In this paper the performance of OSTBC along with TCM is evaluated using QPSK.

Keywords: *Diversity, Multiple Input and Multiple Output (MIMO), orthogonal space- time coding (OSTBC), Channel State information (CSI), Pair-wise Error Probability, Trellis-Coded Modulation (TCM)*

1. Introduction

MIMO wireless systems have captured the attention of international standard organizations. The use of MIMO has been proposed multiple times for use in the high-speed packet data mode of third generation cellular systems (3G) as well as the fourth generation cellular systems (4G). MIMO has also influenced wireless local area networks (WLANs) as the IEEE 802.11n standard exploits the use of MIMO systems to acquire high throughputs. MIMO systems employing space-time coding strategies to support greatly enhanced performance. Space-Time coding [5, 4], uses the advantage of transmitter diversity, is an effective technique to improve the performance of wireless communication systems. In space-time coding, different signals are simultaneously transmitted from different transmit antennas. The signal which is received is the superposition of the different transmitted signals, and the detection process needs estimates of the channel parameters [3]. All these designs were based on the assumption that channel state information is perfectly known at the receiver, but unknown at the transmitter.

The work presented in this paper is motivated by the observation that for the special case of STBC 3×4(3transmitter and 4 receiver) and high code-rate STBC, it is possible to obtain exact closed-form expression for the pair-wise error probability. An exact PEP expression would serve as an attractive alternative to previously derived bounds for evaluating performance [6]. Our expressions are derived from the PDF of the phase of the received signal. Simulated PEP results of STBC 3×4, TCM and concatenation of STBC with TCM using QPSK are presented.

2. System Model: MIMO

When a transmitter and a receiver, with an appropriate channel coding and decoding scheme, are equipped with multiple antennas, the presence of multipath fading can be improved over a Raleigh fading channel. Space-time-coded MIMO systems with N_T transmit antennas and N_R receive antennas is showed in the figure. In the space-time coded MIMO systems, bit stream is mapped into symbol stream $\{\tilde{x}_i\}_{i=1}^{N}$ as depicted in Figure a symbol stream of size N is space-time-encoded into $\{\tilde{x}_i\}_{i=1}^{N}$ t=1, 2, 3….T, where i is the antenna index and t is the symbol time index. Note that the number of symbols in a space-time codeword is N_T .T (*i.e.*, N =N_T × T). In other words $\{\tilde{x}_i\}_{i=1}^{N}$, t=1, 2, 3…T, forms a space-time codeword. As N symbols are transmitted by a codeword over T symbol times, the symbol rate of the space-time-coded system example shown in the figure is given as

$$R = \frac{N}{T} \text{ (Symbols / Channel use)} \quad (1)$$

At the receiver side, the symbol stream $\{\tilde{x}_i\}_{i=1}^{N}$ is estimated by using the receive signal $\{\tilde{y}_i{}^{(t)}\}_{i=1}^{N_R}$, t=1,2,…………T. Let h_{ij}^{t}denotes the Rayleigh-distributed channel gain from the ith transmit antenna to the j^{th} receive antenna over the t^{th} symbol period (i = 1; 2; . . . ; , N_T),(j = 1; 2; . . . ; N_R) and t = 1; 2; . . . ; T). If we assume that the channel gains do not change during T symbol periods, the symbol time index $\{\{h\}_{ij}^{t}$ can be omitted. Furthermore, as long as the transmit antennas and receive antennas are spaced sufficiently apart, N_T ×N_R fading gains can be assumed to be statistically independent [3].

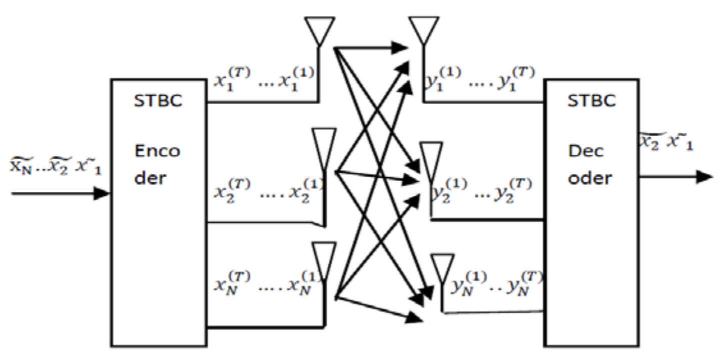

Figure 1. Block Diagram of Space-time Coded MIMO Systems

If x_i^t is the transmitted signal from the i^{th} transmit antenna during t^{th} symbol period, the received signal at the receive antenna during j^{th} symbol period is

$$Y_{j(t)} = \sqrt{\frac{E_X}{N_0 N_T}} \left[h_{j1}^{(t)} h_{j2}^{(t)} \dots h_{j3}^{(t)}\right] + Z_j^{(t)} \quad (2)$$

Where Z_j^t is the noise process at the j^{th} receive antenna during t^{th}symbol period, which is modeled as the ZMCSCG noise of unit variance, and E_X is the average energy of each transmitted signal. Meanwhile, the total transmitted power is constrained as

$$\sum_{i=1}^{N_T} E\{\,|x_i^{(t)}\,|^2\} = N_T, \qquad t = 1,2,\ldots\ldots T \tag{3}$$

Variance is assumed to be 0.5 for real and imaginary parts of h_{ij}.

Considering the relationship i Equation (2) for N_R receive antennas, while assuming quasi-static channel gains (i.e.$h_{ij}^{t}= h_{ij}$, t=1 ,2,….T), the system equation is given as

$$\begin{bmatrix} y_1^{(1)} & y_1^{(2)} \cdots & y_1^{(T)} \\ \vdots & \ddots & \vdots \\ y_{N_R}^{(1)} & y_{N_R}^{(2)} \cdots & y_{N_R}^{(T)} \end{bmatrix} =$$

$$\sqrt{\frac{E_x}{N_O N_T}} \begin{bmatrix} h_{11} & h_{12}.. & h_{1N_T} \\ \vdots & \vdots & \vdots \\ h_{N_R1} & h_{N_R2}.. & h_{N_RN_T} \end{bmatrix} \begin{bmatrix} x_1^{(1)} & x_1^{(2)}.. & x_1^{(T)} \\ \vdots & \vdots & \vdots \\ x_{N_T}^{(1)} & x_{N_T}^{(2)}.. & x_{N_T}^{(T)} \end{bmatrix} + \begin{bmatrix} Z_1^{(1)} & Z_1^{(2)}.. & Z_1^{(T)} \\ \vdots & \vdots & \vdots \\ Z_{N_R}^{(1)} & Z_{N_R}^{(2)}.. & Z_{N_R}^{(T)} \end{bmatrix} \tag{4}$$

3. Orthogonal Space Time Block Codes

In higher order STBC in-order to facilitate computationally-efficient ML detection at the receiver, the following property is required:

$$XX_H = c(|\,x_i^1\,|^2 + |\,x_i^2\,|^2 + \ \ldots.\ |\,x_i^T\,|^2) I_{N_T}$$

$$= \ c\,||\,x_i\,||^2\, I_{N_T} \tag{5}$$

Consider transmitting antennas N_T=3 transmitting complex space time block codes in 8 time slots with coding rate of 1/2, while satisfying a full rank condition

$$X_{3,\text{complex}}^{\text{low rate}} = \begin{bmatrix} x_1 & -x_2 & -x_3 & -x_4 & x_1^* & -x_2^* & -x_3^* & -x_4^* \\ x2 & x1 & x4 & -x_3 & x_2^* & x_1^* & x_4^* & -x_3^* \\ x3 & -x4 & x1 & x2 & x_3^* & -x_4^* & x_1^* & x_2^* \end{bmatrix} \tag{6}$$

Space-time block codes can be used for various numbers of receive antennas. However, only a single receive antenna is assumed. We express the received signals from a single receive antenna as

$$[y_1 y_2 y_3 y_4 y_5 y_6 y_7 y_8] = \sqrt{\frac{E_x}{3N_0}}\,[h_1 h_2 h_3] \begin{bmatrix} x_1 & -x_2 & -x_3 & -x_4 & x_1^* & -x_2^* & -x_3^* & -x_4^* \\ x2 & x1 & x4 & -x_3 & x_2^* & x_1^* & x_4^* & -x_3^* \\ x3 & -x4 & x1 & x2 & x_3^* & -x_4^* & x_1^* & x_2^* \end{bmatrix} +$$

$$[Z_1 Z_2 Z_3 Z_4 Z_5 Z_6 Z_7 Z_8] \tag{7}$$

The above input-output relation can be also expressed as

$$\begin{bmatrix} y_1 \\ y_2 \\ y_3 \\ y_4 \\ y_5 \\ y_6 \\ y_7 \\ y_8 \end{bmatrix} = \sqrt{\frac{E_x}{3N_0}} \begin{bmatrix} h_1 & h_2 & h_3 & 0 \\ h_2 & -h_1 & 0 & -h_3 \\ h_3 & 0 & -h_1 & h_2 \\ 0 & h_3 & -h_2 & -h_1 \\ h_1^* & h_2^* & h_3 & 0 \\ h_2^* & -h_1^* & 0 & -h_3^* \\ h_3^* & 0 & -h_1^* & h_2^* \\ 0 & h_3 & -h_2^* & h_1^* \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix} + \begin{bmatrix} Z_1 \\ Z_2 \\ Z_3 \\ Z_4 \\ Z_5 \\ Z_6 \\ Z_7 \\ Z_8 \end{bmatrix} \quad (8)$$

Again, using the orthogonality of the above effective channel matrix, the received signal is modified as

$$y_{eff} = H_{eff}^{H} y_{eff}$$
$$= 2\sqrt{\frac{E_x}{3N_o}} \sum_{j=1}^{3} |h_j|^2 I_4 X_{eff} + Z_{eff} \quad (9)$$

Using the above result, the ML signal detection in which detector searches over the whole set of transmit signals and decides in favor of the transmit signal that minimizes the Euclidian distance to the receive vector is performed as

$$x_{i,ML} = Q\left(\frac{y_{eff,i}}{2\sqrt{\frac{E_X}{3N_0}}\sum_1^3 |h_i|^2}\right) \qquad i = 1,2,3,4 \quad (10)$$

If a decoding complexity at the receiver is compromised, however, higher coding rates can be achieved by the following codes:

$$X_{3,complex}^{high\ rate} = \begin{bmatrix} x_1 & -x_2^* & \frac{x_3^*}{\sqrt{2}} & \frac{x_3^*}{\sqrt{2}} \\ x_2 & x_1^* & \frac{x_3^*}{\sqrt{2}} & \frac{-x_3^*}{\sqrt{2}} \\ \frac{x_3}{\sqrt{2}} & \frac{x_3}{\sqrt{2}} & \frac{(-x_1-x_1^*+x_2-x_2^*)}{2} & \frac{(x_2+x_2^*+x_1-x_1^*)}{2} \end{bmatrix} \quad (11)$$

Coding rate is $R = \frac{3}{4}$

Decoding of higher coding rates can be achieved by the following equations

$$[y_1 y_2 y_3 y_4] = \sqrt{\frac{E_x}{3N_0}}\,[h_1 h_2 h_3] \begin{bmatrix} x_1 & -x_2^* & \frac{x_3^*}{\sqrt{2}} & \frac{x_3^*}{\sqrt{2}} \\ x_2 & x_1^* & \frac{x_3^*}{\sqrt{2}} & \frac{-x_3^*}{\sqrt{2}} \\ \frac{x_3}{\sqrt{2}} & \frac{x_3}{\sqrt{2}} & \frac{(-x_1-x_1^*+x_2-x_2^*)}{2} & \frac{(x_2+x_2^*+x_1-x_1^*)}{2} \end{bmatrix} + [Z_1 Z_2 Z_3 Z_4] \quad (12)$$

Using the equation (12), the ML signal detection is performed as

$$x_{i,ML} = Q\left(\frac{y_{eff,i}}{2\sqrt{\frac{E_X}{3N_0}}\sum_1^3 |h_i|^2}\right) \qquad i = 1,2,3,4 \quad (13)$$

Effective channel construction for $X_{3,complex}^{high\ rate}$ is rather more complex than the previous coding.

4. Trellis-Coded Modulation

TCM is a modulation scheme which allows highly efficient transmission of information over band-limited channels such as telephone lines. The functions of a TCM consist of a Trellis Code and a Constellation Mapper as shown in Figure 2 TCM combines the functions of the convolutional coder of rate: R=k/(k+1) and a M-ary signal mapper that maps M=2k input points into a larger constellation of M=2k+1 constellation points. TCM is a convolutional coding. Unlike a true Convolutional code, not all incoming bits are coded and only 1 extra bit is always added. Increasing the constellation size reduces Euclidean distances between the constellation points but sequence coding offers a coding gain that overcomes the power disadvantage of going to the higher constellation. The decoding metric is the Euclidean distance and not the Hamming distance. TCM uses set-partitioning and small number of states.

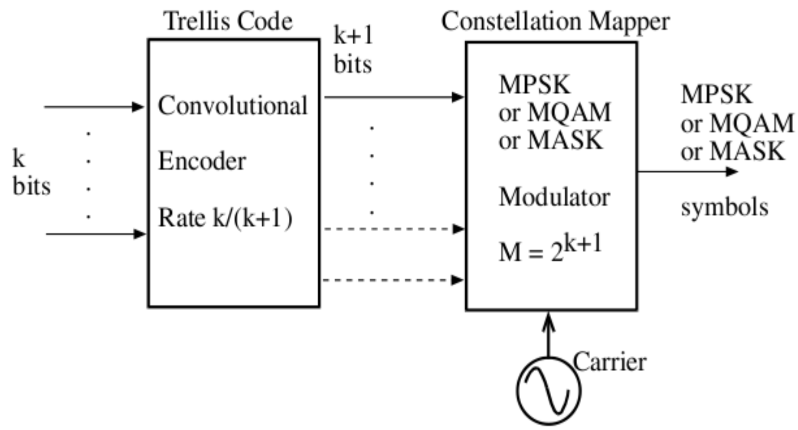

Figure 2. General Trellis Coded Modulation

Assuming that $x_i(k)$ is the transmitted signal from antenna *i* at time *k*, the received signal at the *j*th antenna corresponding to this time interval is given by

$$y(k)=\sqrt{\rho}\sum_{i=1}^{N_t} h_{i,j}x_i(k)+z_i(k) \quad (14)$$

where i = 1, 2, . . .,Nt , j = 1, 2, . . .,Nr , t = 1, 2, . . .,N, and N is the frame length; $h_{i,j}$ denotes the complex Gaussian channel coefficient between the ith transmit and jth receive antennas. ρ denotes the signal-to-noise ratio at each receive antenna.

The input-output relationship can be written in the matrix form as

$$Y=\sqrt{\rho}XH+N \quad (15)$$

In decoding of space Time Trellis Codes, assuming that the receiver has access to the channel coefficients, the optimal decision rule minimizing the probability of error is given by

$$\hat{X}=\arg_X \min P(X \mid Y, H). \quad (16)$$

This is the maximum a posteriori (MAP) decoding rule. If all the symbols are equally likely, then it is equivalent to the ML decoding rule, and it is given by

$$\hat{X}=\arg_X \min P(Y \mid X, H). \quad (17)$$

which is easier to manipulate. Given the transmitted signal matrix and the set of channel coefficients, the elements of the received matrix are jointly Gaussian since the additive noise is Gaussian, and temporally and spatially white. Therefore, this likelihood function is proportional to the negative of the squared Euclidean distance between the received matrix and the transmitted matrix (multiplied by the channel coefficient matrix), resulting in the optimal decoding rule

$$\hat{X} = \arg_X \min \parallel Y - \sqrt{\rho} XH \parallel^2 \quad (18)$$

Where $\parallel . \parallel^2$ denotes the sum of the norm squares of the elements of its matrix argument (*i.e.*, the square of the Frobenius norm of a matrix). Clearly, the resulting decision rule can also be written as

$$\hat{X} = \arg_X \min \sum_{k=1}^{N} \sum_{j=1}^{N_r} \left| y_j(k) - \sqrt{\rho} \sum_{i=1}^{N_t} h_{i,j} x_i(k) \right|^2 \quad (19)$$

By way of construction, the space-time trellis codewords are paths through the code trellis. Therefore, the minimization above is nothing but the process of finding a path through the space-time code trellis with the minimum (Euclidean) distance from the received signal. Also, observing that the metric is additive at each time step, it is clear that one can use the well-known Viterbi algorithm to perform the decoding very efficiently.

Let us describe the Viterbi decoding algorithm in this context in a bit more detail. At time k = 0, we assume that the encoder is in state S0. We extend the paths emanating from this state, and record the value of the path metric computed using

$$\sum_{j=1}^{N_r} \left| y_j(k) - \sqrt{\rho} \sum_{i=1}^{N_t} h_{i,j} x_i(k) \right|^2 \quad (20)$$

At time k, we have one path through the trellis for each state of the encoder together with the corresponding value of the accumulated path metric. To extend each of these paths by one more step, for each state, at time k + 1, all the paths that merge with that particular state are considered as candidates. The possible path metrics are computed by adding

$$\sum_{j=1}^{N_r} \left| y_j(k+1) - \sqrt{\rho} \sum_{i=1}^{N_t} h_{i,j} x_i(k+1) \right|^2 \quad (21)$$

to the current path metrics. All of these extensions except the one with the minimum accumulated path metric are discarded, and the time index is incremented. Therefore, for these steps, only the extensions of the paths that lead to trellis termination are considered, and at the end of the frame, the path through the trellis that is closest to the received vector is declared as the maximum likelihood codeword.

5. Pair-wise Error Probability

Pair-wise error probability is defined as probability of transmitting c^1 and detecting it as c^2, when there are no other code-words. It is represented as P ($c^1 \rightarrow c^2$).

Conditional probability is written as, refer to [6].

$$P\left(c^1 \rightarrow \frac{c^2}{H}\right) = Q\left(\sqrt{\frac{\gamma}{2} Tr[H^H (C^2 - C^1)^H . ((C^2 - C^1)^H]}\right) \quad (22)$$

According to the Orthogonality conditional PEP is written as

$$P(C^1 \rightarrow C^2/H) = Q\left(\sqrt{\frac{\gamma}{2} K \sum_{K=1}^{K} |S_K^2 - S_K^1|^2 \ T_r \ [H^H, H]}\right)$$

$$= Q\left(\sqrt{\frac{Y}{2}\ K\sum_{K=1}^{K}|S_K^2 - S_K^1|^2\ \sum_{n=1}^{N}\sum_{m=1}^{M}\left|\alpha_{n,m}\right|^2}\right) \quad (23)$$

Euclidian distance between T_x and detected symbol is given by

$$d_E = \sqrt{\sum_{k=1}^{k}\left|s_k^2 - s_k^1\right|^2}$$

$$P(C^1 \rightarrow C^2/H) = Q\left(\sqrt{\frac{Y}{2}\ Kd_E^2\sum_{n=1}^{N}\sum_{m=1}^{M}\left|\alpha_{n,m}\right|^2}\right) \quad (24)$$

To calculate PEP, one needs to integrate above equation weighted by density of path gains

$$p\left(c^1 \rightarrow {c^2}/_{H}\right) = \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \exp\left(\frac{-k\gamma d_E^2 \sum_{n=1}^{M}\sum_{m=1}^{M}|a_{n,m}|^2}{4\sin\Phi^2}\right)\ d\Phi$$

$$= \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \prod_{m=1}^{M}\prod_{N=1}^{N}. \exp\left(\frac{-k\gamma d^2 E|\alpha_{n,m}|}{4\sin\emptyset^2}\right) d\emptyset \quad (25)$$

Let the path gains are independent from each other. The integral over the distribution of the path gains is same as product of MN equal integrals *i.e.*, $\int_0^{\frac{\pi}{2}}\prod_{m=1}^{M}\prod_{N=1}^{N}. = \int_0^{\infty}.$

$$p\left(c^1 \rightarrow {c^2}/_{H}\right) = \frac{1}{\pi}\int_0^{\frac{\pi}{2}}\left[\int_0^{\infty}\exp\left(\frac{-k\gamma d_E^2 x}{4\sin\Phi^2}\right)\ \delta_{\chi(x)}\ dx\right]^{MN} d\Phi$$

Where $f_x(x) = e^{-x}$, x>0 is the pdf of $|\alpha_{n,m}|^2$Moment Generating Function of exponential distribution for μ<1 is given by

$$M_x(\mu) = E[e^{\mu x}] = \int_0^{\infty} e^{\mu x} f_x(x) dx$$

$$= \int_0^{\infty} e^{\mu x}\, e^{-x} dx$$

$$= \frac{1}{1-\mu} \quad (26)$$

Since $\mu = \frac{-k\gamma d^2 Ex}{4\sin\emptyset^2}$

$$P(c^1 \rightarrow c^2) = \frac{1}{\pi}\int_0^{\frac{\pi}{2}} \frac{1}{1 + \frac{-k\gamma d_E^2}{4\sin\Phi^2}}$$

$$= \frac{1}{\pi}\int_0^{\frac{\pi}{2}}\left[\frac{\sin\Phi^2}{\sin\Phi^2 + \frac{k\gamma d_E^2}{4}}\right]^{MN} d\Phi \quad (27)$$

The generalized expression of PEP for STBC is given by

$$P(c^1 \rightarrow c^2) = \frac{1}{2}\left\{1 - \sqrt{\frac{a}{1+a}}\sum_{i=0}^{MN-1}\binom{2i}{i}\left[\frac{1}{4(1+a)}\right]^i\right\} \quad (28)$$

where $a = k\frac{\gamma}{4}d^2E$

N= Number of Transmitters
M= Number of receivers

6. Comparison between TCM and OSTBC

It is of interest to compare and contrast the two approaches for space-time coding in this, space-time block coding versus space-time trellis coding.

Both space-time block codes and space-time trellis codes are designed to achieve full diversity advantage over the MIMO wireless channels. However, there are some basic differences. For instance, space-time block codes are very easy to encode and decode (which is achieved using the simple linear processing receivers), whereas space time trellis codes require more complicated trellis-based decoders. Also, it is relatively easy to find and employ space-time block codes for more than two transmit antennas (although there may be a rate loss penalty), this is not the case for space-time trellis codes – they are most widely used for the case of two transmit antennas. These are clear advantages for space-time block coding. On the other hand, the resulting error rates of the space-time trellis codes are generally better than those of the space-time block codes. . We observe that the space-time trellis codes outperform the Alamouti scheme, particularly when the number of states is increased. This is because space-time trellis codes provide a coding advantage in addition to providing full diversity when properly designed.

7. Concatenation of OSTBC and TCM

Concatenated codes were first introduced by Forney (1966), where he proposed a scheme that involves concatenating two single codes in a serial fashion. The inner code is a convolutional code and the outer code is a high-rate algebraic Reed–Solomon (RS) code which has a powerful error correction capability. The performance improvements achieved by this concatenated coding scheme were very promising and opened the door for further developments in this area.

Wireless communication systems are being designed to integrate features that include high data rates as well as high quality of service in the existing communication framework. For this an orthogonal space-time block code (OSTBC) technique improves error performance of synchronous data links without sacrificing data rate or requiring more bandwidth. Trellis coded modulation enables efficient transmission scheme and to achieve high coding gain by integrating coding and modulation. In this work an OSTBC concatenated with TCM is implemented for information transmission over different antenna configurations, from Single-Input Single-Output (SISO) to Multi-Input Multi-Output (MIMO) channels.

The TCM scheme, as mentioned previously, encompasses a wide variety of concatenation schemes since there is no restriction on its mapper. Among these schemes is the concatenation of an outer channel code and an inner orthogonal STBC, where in this case the STBC simply replaces the mapper. This scheme was originally proposed by Bauch (1999). It is shown by Liew and Hanzo (2002) that the coded STBC scheme gives the best performance–complexity trade-off among other concatenation schemes when the outer code is a convolutional code or a turbo code.

7.1. Encoder Structure

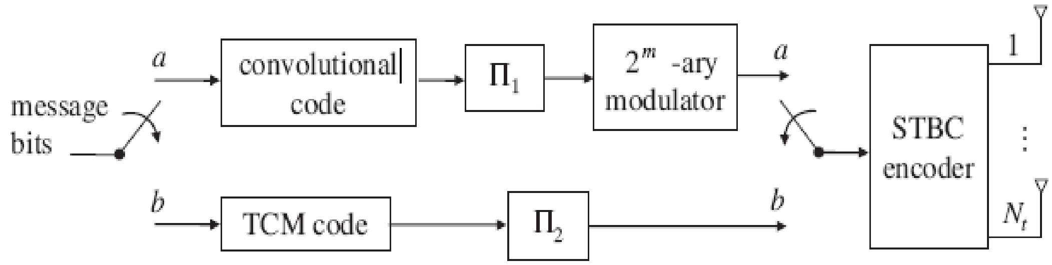

Figure 3. Encoder Structure for the Coded STBC Scheme

The encoder of the coded STBC scheme is shown in Figure 3. In the figure, the primitive data stream is first encoded by an outer channel encoder. There is no restriction on the type of channel code employed, but the focus here is on convolutional codes, turbo codes and TCM codes because they are widely used. In addition, we only consider binary codes. The coded sequence is interleaved and demultiplexed. Each set of m bits at the output of the demultiplexer is mapped onto a symbol taken from a 2^m-ary signal constellation.

The output of the modulator is then fed into the STBC encoder, which groups every N_t consecutive symbols and transmits them from the available N_t antennas according to the STBC encoding principles, assuming that the underlying STBC is full rate. As such, the transmission rate achieved is R = mRc bits per channel use, where Rc is the rate of the outer channel code.

7.2. Decoder Structure

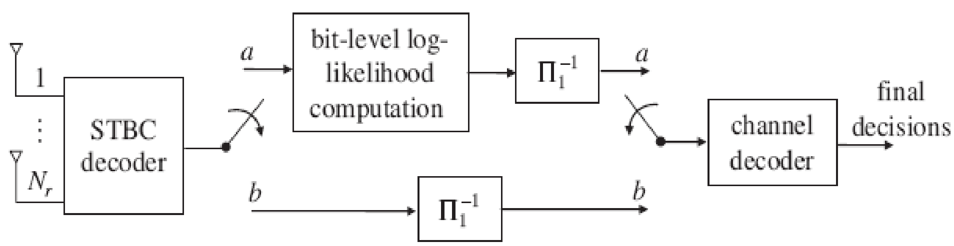

Figure 4. Decoder Structure for the Coded STBC Scheme

The decoder structure for the coded STBC scheme is depicted in Figure 4. In the figure, when the switch is in position '*a*', the resulting decoder corresponds to the convolutional (or turbo) coded system. In this case, the output of the STBC decoder is fed into the log likelihood computation module, which computes the log-likelihoods for the bits comprising the corresponding symbols. These log-likelihoods are then deinterleaved and passed to the channel decoder. When a convolutional code is used, the corresponding decoder would be the Viterbi decoder.

The proposed model takes the advantages of the concatenation scheme: the spatial diversity gain offered by OSTBC and the coding gain offered by TCM.

8. Simulation

Results In this work, MATLAB is used to test the BER performance of the Rayleigh fading channel model for STBC with transmitters (NT=3) and receivers (NR=4) for different code rates. Results are shown below. It illustrates the advantages of a STBC and TCM concatenation scheme, the spatial diversity gain offered by STBC, and the coding gain offered by TCM. By using the complex space time block code shown in the equation (6) with low code rate and equation (11) with high code rate, it is also observed that nearly same probability of error can be achieved.

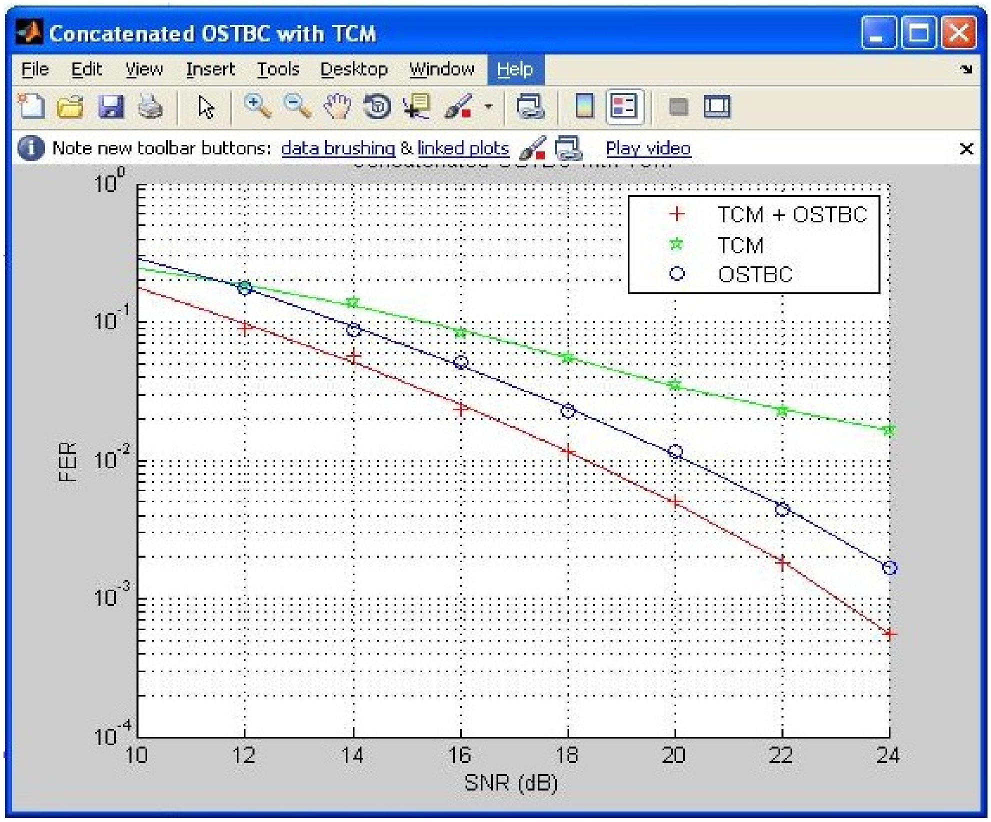

Figure 5. FER versus SNR of MIMO using QPSK

9. Conclusion

This paper provides need and advantages of MIMO systems. A basic introduction to Space-Time Coding was provided by presenting STBC system model. We then discussed block codes schemes for the cases of 3 transmit antennas and 4 receive antennas. High data rate code Scheme also discussed .Generalized pair-wise error probability (PEP) for the STBC was presented. From the simulation results we conclude that data rate can be increased by using high rate code in STBC. Based on the simulation results above, it is shown that, the optimal performance can be achieved by using OSTBC along with TCM.

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