ERROR PERFORMANCE OF SUPER-ORTHOGONAL SPACE-TIME TRELLIS CODES WITH TRANSMIT ANTENNA SELECTION

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ABSTRACT

A space-time coding scheme with transmit antenna selection that employs conventional super-orthogonal space-time trellis and improved super-orthogonal space-time trellis codes is considered. Transmit antenna selection criterion is given and the error performance of the proposed structure is tested by computer simulations for systems that have two active transmit antennas and single receive antenna in quasi-static flat Rayleigh fading channels. It has been shown that proposed scheme outperforms the previous space-time trellis coding structures with transmit antenna selection given in the literature.

1. INTRODUCTION

MIMO (multiple-input multiple-output) technology has become one of the prominent research areas in communication systems since it significantly increases the information capacity of wireless channels [1]. In order to improve the data rate and the error performance of MIMO systems over fading channels, coding techniques appropriate for use with multiple transmit antennas have been developed, namely space-time coding [2]. Space-time coding, first introduced by Tarokh et al. [2] is based on correlating the transmitted signals in both spatial and temporal domains without sacrificing the bandwidth. Two main types of space-time coding techniques are space-time trellis coding and space-time block coding. Space-time trellis codes can provide full diversity and coding gain at the cost of relatively high decoding complexity that increases exponentially with diversity level and data rate. On the other hand, space-time block codes obtained by generalization of Alamouti scheme [3] can provide low decoding complexity with respect to space-time trellis codes and full diversity order due to the orthogonal design method. However they present little or no coding gain. Another class of space-time coding technique exploiting orthogonal design approach is super-orthogonal space-time trellis (SOSTT) coding [4]. This class of space-time coding is based on the concept of concatenating a trellis structure with a space-time block code and allows a considerable increase in coding gain over the conventional space-time trellis codes existing in literature. Continual efforts to build trellis structures with better performance over space-time block codes yielded another concept known as generalized rotations. This new approach leaded to performance improvements over existing SOSTT structures [5].

The main drawback of MIMO systems is the requirement of multiple RF (Radio Frequency) chains consisting of amplifiers, A/D converters etc. that increase system complexity and hardware cost. Therefore, techniques which enable to reduce number of required RF chains while utilizing the benefits of exploiting multiple antennas is considerably important for future communication systems. A promising approach to attain this goal is transmit antenna selection (TAS) [6] technique where some of the transmit antennas are chosen and activated for transmission and all of the other transmit antennas are silent. This technique automatically provides full diversity order as if all transmit antennas were used during transmission. The TAS method does not only reduce the number of RF chains but also decrease the receiver complexity by transferring the antennas from receiver side to transmitter side. In [6], an uncoded transmission scheme combining transmit antenna and maximal-ratio combining is studied. In [7], error performance expressions of a MIMO system that utilizes transmit antenna selection with Alamouti code [3] have been derived. In [8], error performance of space-time trellis codes with transmit antenna selection is analyzed.

In this paper, a scheme coupling SOSTT and improved SOSTT coding with transmit antenna selection is proposed. The antenna selection criterion that minimizes conditional pairwise error probability is obtained. It is shown by computer simulations that the new system provides much better error performance than the previously known space-time trellis codes using transmit antenna selection.

2. BACKGROUND

2.1 Super-Orthogonal Space-Time Trellis Codes

Super-orthogonal space-time trellis coding has been introduced by Jafarkhani and Seshadri [4] to obtain additional coding gain for a given signal constellation and data rate. In this technique, space-time block codes are integrated with a trellis structure. The full diversity level and redundancy providing coding gain are realized with parameterized class of space-time block code structures. The crucial point in design process is expansion of the number of available orthogonal transmission matrices without expanding the signal constellation. An example of

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the orthogonal designs for two transmit antennas can be given by

$$\mathbf{X(x_1,x_2,\theta)} = \begin{bmatrix} x_1 e^{j\theta} & x_2 \\ -x_2^* e^{j\theta} & x_1^* \end{bmatrix} \qquad (1)$$

where the rows correspond to the transmission slots while the columns correspond to the symbols transmitted from each antenna. It is important to determine the possible values of θ in order to prevent constellation expansion. For M-PSK signal constellations signals are represented by $e^{j2\pi m/M}$, m = 0, 1, ... , M - 1, or θ takes values of $\theta = 2\pi m'/M$ $m' = 0,1,..M - 1$. Note that θ = 0 gives the Alamouti code.

The next step in design procedure is the construction of matrix subsets using the classical set partitioning method. This is accomplished in a way similar to Ungerboeck's set partitioning method [9] but utilizing a different criterion namely coding gain distance (CGD) [4] instead of Euclidean distance. CGD between two codewords c_1 and c_2 is defined as the determinant of the following matrix

$$\mathbf{A(c_1, c_2) = B(c_1, c_2).\ B^H(c_1, c_2)} \qquad (2)$$

where $\mathbf{B(c_1, c_2)}$ is the difference of transmission matrices for this codeword pair c_1, c_2 and H is Hermetian operator. The set partitioning process is carried out by maximizing the minimum CGDs over all possible pairs of distinct codewords [4]. The set partitioning for QPSK signal set is given Figure 1. At the each level of the tree structure given in Figure 1, the set that contains all possible codewords is partitioned into two subsets denoted by S. It can be seen that the minimum value of CGD is increased as going lower levels of the tree. The codewords are represented by two numbers each of which takes the values of 0,1,2,3 that correspond to QPSK signal constellation elements 1,j,-1,-j respectively.

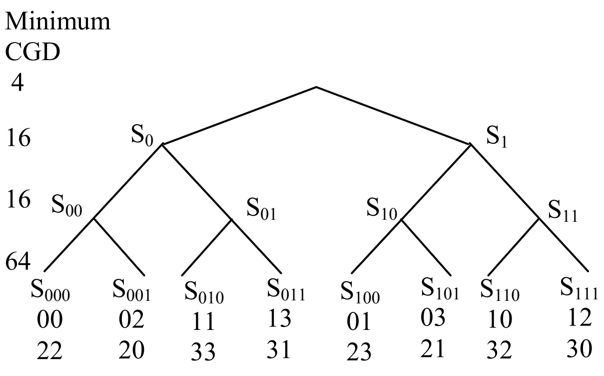

Figure 1 - Set partitioning for QPSK

After the set partitioning, each of constituent orthogonal block codes and codeword sets are assigned to the branches of trellis in order to maximize coding gain. The 4-state SOSTT code designed by Jafarkhani and Seshadri [4] using this systematic way is demonstrated in Figure 2. The parameter that determines the error performance of the trellis is defined by the minimum of CGD values corresponding to parallel paths and error events with path length of three. For the 4-state code in Figure 2, the minimum value of the CGD values of error paths is greater than the minimum value of CGD values of parallel transitions. Thus CGD is dominated by parallel transitions for this structure. The minimum CDG value of the code in Figure 2 is 16 for QPSK signal constellation [4].

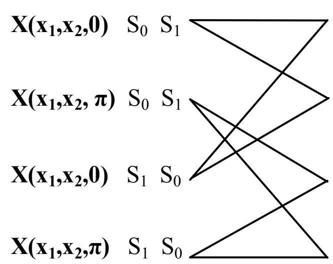

Figure 2 4-state SOSTT code [4]

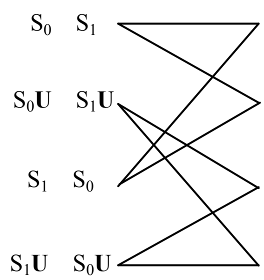

Figure 3 4-state improved SOSTT code [5]

2.2 Improved Super-Orthogonal Space-Time Trellis Codes Through Generalized Rotations

The research activities on building trellises over space-time block codes yield a new concept called generalized rotations [5]. This approach provides better error performance over existing SOSTT codes presented in [4]. In this technique, the expansion of the orthogonal transmission matrix is accomplished via transformation matrices $\mathbf{U_i}$. The code is constructed by allocating one rotation of the orthogonal transmission matrix $\mathbf{X_i = X_0 U_i}$ to each trellis state. Then within each state, the rotation matrix $\mathbf{X_i}$ is partitioned into subsets ($\mathbf{X_{i,0},\ldots,X_{i,n}}$) where n equals to the number of the connected states in the trellis [5]. If a transition between state i and j does not have any parallel branches, the subset $X_{i,j}$ has one codeword, otherwise it has more than one codeword [5].

The key point in design procedure is finding good U_i transformation matrices that generate orthogonal rotation matrix set. It has been proven that transformation matrices must be not only unitary but also either diagonal or anti-diagonal [5]. The transformation matrix of the 4-state QPSK code for two transmit antennas given in Figure 3 is

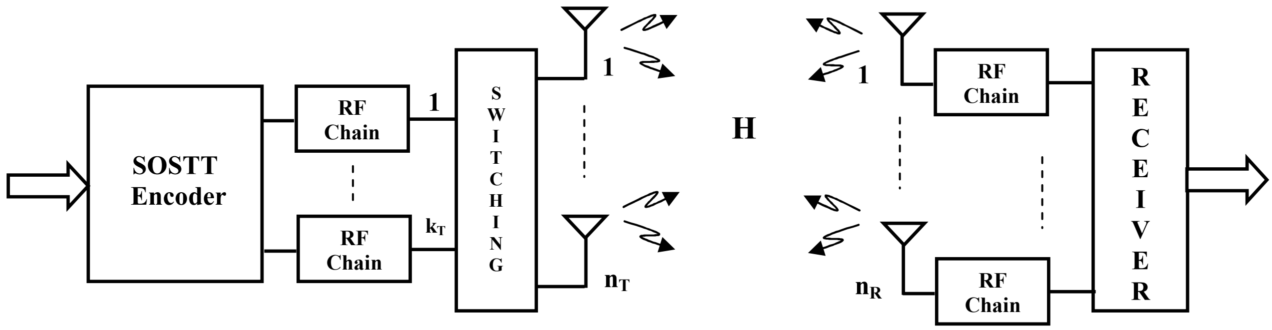

Figure 4 - (n_T, k_T; n_R, n_R) MIMO system with transmit antenna selection employing super-orthogonal coding

$\mathbf{U}$=diag($e^{j3\pi/4}$,$e^{-j3\pi/4}$). Note that the transformation matrix for the code in Figure 2 is $\mathbf{U}$=diag($e^{j\pi}$,1). This transformation provides a gain of approximately 0.3 dB with respect to the code designed by Jafarkhani and Seshadri [4].

Unlike conventional super-orthogonal space-time trellis coding, the generalized rotations approach generates modulation symbols that may not be in the original signal constellation. However, this situation does not cause the detection problems which take place because of constellation expansion, since the symbols out of signal set are a rotation of the original signal set. A match filter can easily eliminate unfavorable effects of the rotations that occur at each trellis transition [5].

3. SOSTT AND IMPROVED SOSTT CODING WITH TRANSMIT ANTENNA SELECTION

The block diagram of MIMO scheme proposed in this paper that couples TAS with SOSTT coding and improved SOSTT coding is demonstrated in Figure 4. In this scheme, k_T of n_T transmit antennas are chosen and activated for a certain time interval for transmission of the baseline code designed for k_T transmit antennas, while all other transmit antennas are silent. All the receive antennas are used without selection. We will refer to this construction as an (n_T,k_T;n_R,n_R) system.

Let $\mathbf{H}$ be n_R x n_T channel matrix. The elements of this matrix represented by h_{ij} (i = 0, 1,…,n_T and j = 0,1,...,n_R) are fading coefficients which are modeled as independent samples of complex Gaussian random variables with zero mean and the variance of 0.5 per dimension. Then we have the following received signal model:

$$\mathbf{r_t} = \mathbf{x_t H_s} + \mathbf{n_t} \tag{3}$$

where r_t (1 x n_R) is the received vector at time t, x_t (1 x k_T) is the transmitted signal vector, n_t (1 x n_R) is the additive white Gaussian noise (AWGN) vector and $\mathbf{H_s}$ (k_T x n_R) is the sub-channel matrix between the selected transmit antenna subset and the receive antenna elements.

The transmit antenna selection is accomplished according to the channel state information which is assumed to be perfectly available at receiver [6-8]. There are $\binom{n_T}{k_T}$ possible subsets of transmit antennas that can be used antenna subset, k_T transmit antennas are chosen to maximize the received SNR of the data stream that is given by

$$\gamma_s = \gamma \sum_{j=1}^{n_R} \sum_{u=1}^{k_T} \left| h_{d_u j} \right|^2 \tag{4}$$

where d_u is the label of the selected antennas; $\gamma = E_s/N_0$ where E_s is the energy per symbol at each transmit antenna and N_0 is the one-side power spectral density of complex AWGN per receive antenna. It is clearly seen from (4) that the transmit antennas which maximize the Frobenious norm of $\mathbf{H_s}$ sub-channel matrix should be used for transmission. It can be shown that this selection criterion also minimizes the conditional pairwise error probability (PEP) defined as the probability that receiver selects erroneously $\hat{\mathbf{x}} = \hat{x}_1^1 \hat{x}_1^2 ... \hat{x}_1^{k_T} \hat{x}_2^1 \hat{x}_2^2 ... \hat{x}_2^{k_T} .. \hat{x}_L^1 \hat{x}_L^2 ... \hat{x}_L^{k_T}$ as the estimate of transmitted sequence $\mathbf{x} = x_1^1 x_1^2 ... x_1^{k_T} x_2^1 x_2^2 ... x_2^{k_T} ... x_L^1 x_L^2 ... x_L^{k_T}$ where L is the frame length. The conditional PEP is upper bounded by [8]

$$P(\mathbf{x}, \hat{\mathbf{x}} \mid \mathbf{H_s}) \le \frac{1}{2} \exp\left(d_h^2(\mathbf{x}, \hat{\mathbf{x}}) E_s/4N_0 \right) \tag{5}$$

where $d_h^2(\mathbf{x}, \hat{\mathbf{x}})$ is modified Euclidean distance between $\mathbf{x}$ and $\hat{\mathbf{x}}$, and is given as

$$d_h^2(\mathbf{x}, \hat{\mathbf{x}}) = \sum_{t=1}^{L} \sum_{j=1}^{n_R} \left| \sum_{u=1}^{k_T} h_{d_u j} \left(x_t^{d_u} - \hat{x}_t^{d_u} \right) \right|^2 . \tag{6}$$

Let $\mathbf{B}(\mathbf{x}, \hat{\mathbf{x}})$ be a given difference matrix given as

$$\mathbf{B}(\mathbf{x}, \hat{\mathbf{x}}) = \begin{bmatrix} (x_1^1 - \hat{x}_1^1) & (x_2^1 - \hat{x}_2^1) & ... & (x_L^1 - \hat{x}_L^1) \\ (x_1^2 - \hat{x}_1^2) & (x_2^2 - \hat{x}_2^2) & ... & (x_L^2 - \hat{x}_L^2) \\ . & . & & . \\ . & . & & . \\ (x_1^{k_T} - \hat{x}_1^{k_T}) & (x_2^{k_T} - \hat{x}_2^{k_T}) & ... & (x_L^{k_T} - \hat{x}_L^{k_T}) \end{bmatrix} \tag{7}$$

and a $\mathbf{A}(\mathbf{x}, \hat{\mathbf{x}})$ be a nonnegative definite Hermitian distance matrix given as

$$\mathbf{A}(\mathbf{x}, \hat{\mathbf{x}}) = \mathbf{B}(\mathbf{x}, \hat{\mathbf{x}}).\, \mathbf{B}^{\mathbf{H}}(\mathbf{x}, \hat{\mathbf{x}}). \tag{8}$$

Equation (6) can be rewritten as

$$d_h^2(\mathbf{x}, \hat{\mathbf{x}}) = \sum_{j=1}^{n_R} \mathbf{\Omega_j}\, \mathbf{A}(\mathbf{x}, \hat{\mathbf{x}})\, \mathbf{\Omega_j}^{\mathbf{H}} \tag{9}$$

where

$$\mathbf{\Omega_j} = [\, h_{j1} \quad .... \quad h_{jk_T} \,] \qquad (j = 1,2...,n_R). \tag{10}$$

When the relations between transmitted signals due to structure of transmission matrix are considered, the difference matrix for a (n_T,2;n_R, n_R) system employing super-orthogonal space-time trellis coding can be given by

$$\mathbf{B}(\mathbf{x}, \hat{\mathbf{x}}) = \begin{bmatrix} (e^{jk_1w} - e^{ja_1w})e^{j\theta_1} & (e^{jl_1w} - e^{jb_1w})e^{j\theta_2} \\ (-e^{-jl_1w} + e^{-jb_1w})e^{j\theta_1} & (e^{-jk_1w} + e^{-ja_1w})e^{j\theta_2} \\ . & . \\ . & . \\ (e^{jk_{L/2}w} - e^{ja_{L/2}w})e^{j\theta_1} & (e^{jl_{L/2}w} - e^{jb_{L/2}w})e^{j\theta_2} \\ (-e^{-jl_{L/2}w} + e^{-jb_{L/2}w})e^{j\theta_1} & (e^{-jk_{L/2}w} + e^{-ja_{L/2}w})e^{j\theta_2} \end{bmatrix}^T \tag{11}$$

where $(e^{jk_1w}, e^{jl_1w}), (e^{jk_2w}, e^{jl_2w}), \dots (e^{jk_{L/2}w}, e^{jl_{L/2}w})$ are the transmitted codeword pairs; (e^{ja_1w}, e^{jb_1w}), $(e^{ja_2w}, e^{jb_2w}), \dots, (e^{ja_{L/2}w}, e^{jb_{L/2}w})$ are the erroneously decided codeword sequence; $k_d, l_d, a_d, b_d \in 0,1.,M-1$ are the M-PSK signal constellation elements (d = 0, 1,..,L/2) and $w = 2\pi/M$.

If we rearrange the distance matrix considering that CGD is dominated by parallel transitions, we obtain following expression

$$\mathbf{A}(\mathbf{x}, \hat{\mathbf{x}}) = \left[\sum_{z=1}^{L/2} [4 - 2\cos[w(k_z - a_z)] - 2\cos[w(l_z - b_z)]]^2 \right] \mathbf{I}$$

where $\mathbf{I}$ is a (2 x 2) identity matrix. Thus, modified Euclidean distance for a (n_T,2;n_R, n_R) super-orthogonal space-time trellis coding system with transmit antenna selection will be

$$d_h^2(\mathbf{x}, \hat{\mathbf{x}}) =$$

$$\sum_{j=1}^{n_R} \sum_{u=1}^{2} \sum_{z=1}^{L/2} \left| h_{d_u j} \right|^2 [4 - 2\cos[w(k_z - a_z)] - 2\cos[w(l_z - b_z)]]^2 \tag{12}$$

where d_1 and d_2 are the labels of the selected two transmit antennas. Expression (12) has to be maximized to improve conditional PEP given in (5). This is accomplished by maximizing the term

$$\sum_{u=1}^{2} \left| h_{d_u j} \right|^2 \tag{13}$$

in Equation (12) which represents the Frobenious norm of $\mathbf{H_s}$ sub-channel matrix.

The decision metric is obtained using the same method in [2] and transmission matrix structures developed in [4 - 5]. The decoding is carried out in two steps in case of trellises with parallel branches. At the first step, the decoder finds the branch with the smallest metric. This process eliminates the other parallel branches and provides a reduced trellis with only one branch representing per transition. At the second step, the path with the minimum accumulated metric is determined by employing Viterbi algorithm. The number of the calculations required in Viterbi algorithm is decreased via the elimination process performed in the first step.

The receiver sends optimal transmit antenna subset information to the transmitter through the feedback link. This feedback transmission supplies the transmitter not only the optimal transmit antenna subset but also partial knowledge of the channel state. The receiver must also be able to compute Frobenious norms of total $\binom{n_T}{k_T}$ possible choices and determine the maximum norm. This computation is not a problem in practical systems when the number of transmit and receive antennas is less than four or five.

4. SIMULATIONS

In order to measure the error performance of the proposed scheme, computer simulations have been accomplished for quasi-static flat Rayleigh fading channels where channel state information is constant within a frame and varies independently from one frame to another. All performance curves are obtained for 130 QPSK transmission symbols per transmit antenna. The total radiated signal power from active transmit antennas is assumed to be same for all situations with and without transmit antenna selection. A single antenna is assumed at receiver. Performance comparisons have been carried out at frame error probability of 10^{-1}.

Figure 5 shows the frame error rate results of the 4-state SOSTT and improved SOSTT codes for QPSK. (2,2;1,1) corresponds to the system without antenna selection. It can be clearly seen for both conventional and improved SOSTT codes that the frame error rate decreases as the total number of the transmit antennas is increased although the number of activated transmit antennas is constant. The error performance of the optimum 4-state QPSK space-time trellis code with two transmit antennas given in [10] is also included for comparison purposes. It is seen from the Figure 5 that the proposed scheme provides much better error performance. The gain supplied by the proposed

(3,2;1,1) scheme over the (3,2;1,1) system employing the space-time trellis code is 1.5 dB and 1.8 dB for conventional and improved SOSTT codes respectively. Similarly, the frame error performance improvement over the (4,2;1,1) system using space-time trellis code is 1.8 dB and 2.1 dB for conventional and improved SOSTT codes respectively.

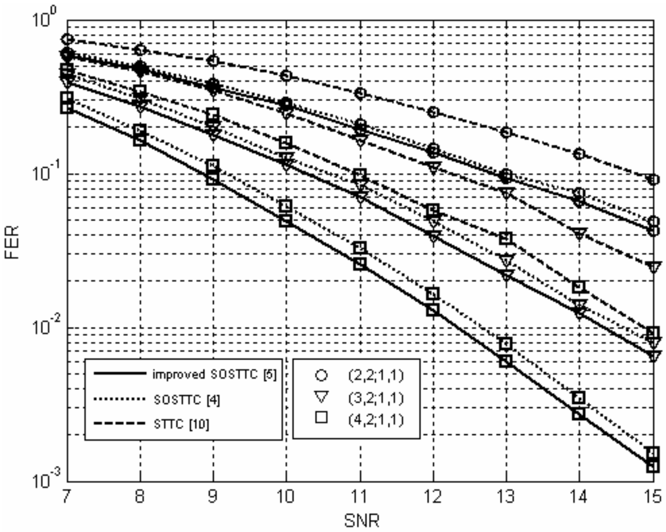

Figure 5 - Error performance curves of the 4-state QPSK space-time trellis structures with and without transmit antenna selection

5. CONCLUSION

A MIMO structure that couples transmit antenna selection technique with conventional and improved super-orthogonal space-time coding is considered. The modified Euclidean distance of the proposed scheme has been derived for the systems employing two active transmit antennas and the antenna selection criterion has been obtained. It has been shown by simulation results that this scheme achieves better error performance with respect to the previously known space-time trellis codes using transmit antenna selection.

ACKNOWLEDGEMENT

The authors wish to thank Coordinatorship of Selcuk University Scientific Research Projects for the sponsorship of this project (Project No: 06701100).

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