Super-Orthogonal Space-Time Trellis Codes in Rapid Rayleigh Fading Channels

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Abstract— **Space-time trellis codes have shown to provide a good performance in Rayleigh fading channels. Recently Super-orthogonal space-time trellis codes have shown to outperform these codes, and also provide a systematic design method to maximize diversity and coding gain in quasi-static Rayleigh fading channels. We investigate the performance of these new codes in rapid Rayleigh fading channels. Some simulation results of the two schemes in quasi-static and rapid Rayleigh fading are presented, followed by some analytical results.**

I. INTRODUCTION

With the integration of Internet and multimedia applications in next generation wireless communications, the demand for wide-band high data rate communication services is growing. As the available radio spectrum is limited, higher data rates can be achieved only by designing more efficient signaling techniques [4]. Among many cutting edge wireless technologies, a new class of transmission techniques, known as Multiple-Input Multiple-Output (MIMO) technique, has emerged as an important technology leading to promising link capacity gains of several fold increase in achievable data rates and spectral efficiency.

While the use of MIMO techniques in the third generation (3G) standards is minimal, it is anticipated that these technologies will play an important role in the physical layer of fixed and fourth generation (4G) wireless systems. Multiple antennas, when used at both the transmitter and receiver, create a multiple-input multiple-output propagation channel. Using sophisticated coding at the transmitter and substantial signal processing at the receiver, the MIMO channel can be provisioned for higher data rates, resistance to multipath fading, lower delays, and support for multiple users. However, the key question is how to exploit this new capability of MIMO wireless communications in a computationally efficient manner.

Space-time coding (STC) techniques have been identified as the solution, as it is a set of practical signal design techniques aimed at approaching the information theoretic capacity limit of MIMO channels. Space-time coding is based on introducing joint correlation in transmitted signals in both the space and time domains. Through this approach, simultaneous diversity and coding gains can be obtained, as well as high spectral efficiencies [4]. A few years ago, Tarokh et al. proposed Space-Time Trellis Codes (STTC) [1] by jointly designing the channels coding, modulation, transmit diversity and the optional receive diversity scheme. The performance criteria for designing STTC were derived under the assumption that the channel is fading slowly and that the fading is frequency non-selective (quasi-static). These advances were then also extended to fast fading channels. STTCs perform extremely well at the cost of relatively high complexity. Other design criteria for STTCs have been proposed in [5] and [6], which show relative improvement in performance under the same conditions but only for more than one receive antenna ($N_R > 1$).

In addressing the issue of complexity Alamouti [8], discovered a remarkable scheme for transmission using two transmit antennas. A simple decoding algorithm was devised by Alamouti [8] which can be generalized to an arbitrary number of receive antennas. This proposal motivated Tarokh et al. [9] to generalize Alamouti's scheme to an arbitrary number of transmit antennas, leading to the concept of Space-Time Block Codes (STBC). Although STBC achieved full diversity and had simple decoding methods, they could not provide any coding gain.

Recently, a new class of space-time codes called Super-Orthogonal Space-Time Trellis Codes (SOSTTC) [2] was introduced, that combine set-partitioning with a super set of orthogonal space-time block codes in such a way as to provide full diversity with increased rate and improved coding gain over previous space-time trellis code constructions in quasi-static Rayleigh fading channels. The idea can be basically defined as a combination of space-time block codes with a trellis code to come up with a new structure that guarantees the full diversity and increased rate. The result also provides a systematic method to design space-time trellis codes for any given rate and number of states, unlike the conventional STTCs which had handcrafted code designs.

In this paper we show the performance of various configurations of SOSTTCs in rapid fading Rayleigh channels and compare its performance with conventional STTCs under the same conditions. An exact bound analysis is also provided in terms of their pairwise error probabilities to get a better estimate of the performance of these codes.

Section II reviews the principles behind Space-Time Trellis codes and gives a brief description of the analytical method used for the exact bound. In section III, we look at the design guidelines for the SOSTTCs and an analysis of the coding gain for SOSTTCs. In section IV, we present the simulation results obtained for quasi-static and rapid Rayleigh fading channels and the analytical bounds obtained. Finally, some concluding remarks are provided in section V.

II. SPACE-TIME TRELLIS CODES

Space-time coding (STC) schemes combine the channel code design and the use of multiple transmit and receive antennas. The encoded data is split into N_T streams that are simultaneously transmitted using N_T transmit antennas. The received signal is a linear superposition of these simultaneously transmitted symbols corrupted by noise and fading. Space-time decoding algorithms as well as channel estimation techniques are incorporated at the receiver in order to achieve diversity advantage and coding gain.

Space-Time Trellis Codes (STTC) was originally proposed by Tarokh et al [1] from AT&T research labs, which combine the design of channel coding with the symbol mapping onto multiple transmit antennas. These codes are designed to obtain maximum diversity gain and provide the best trade off between constellation size, data rate, trellis complexity and diversity advantage. The encoder is composed of N_T different generator polynomials to determine the simultaneously transmitted symbols. The receiver is based on channel estimation of the fade coefficients and Maximum Likelihood Sequence Estimation (MLSE) decoder, which computes the lowest Euclidean distance metric to extract the most likely transmitted sequence. Literature has shown that STTC achieves greater coding gain than Space-Time Block Codes (STBC) [9], and are shown to outperform layered space-time codes.

Space-Time trellis codes can be considered to function in a similar manner to a convolutional encoder, and the connection weights and output symbols form part of an M-ary alphabet. We can consider the encoder structure shown in Fig. 1 which is described in discrete time using a feedforward implementation.

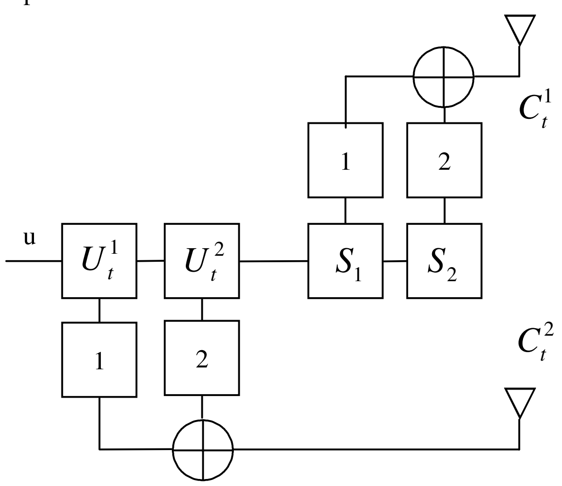
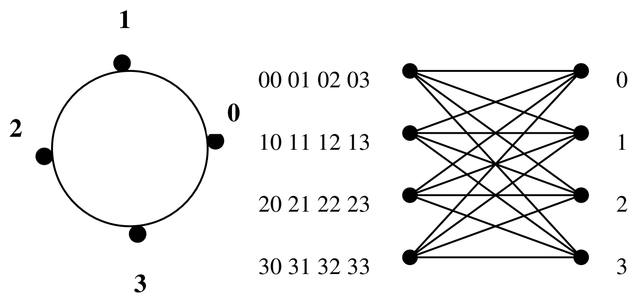


Fig. 1 STTC Encoder

At each instant k input bits ($u_t^1 \cdots\cdots\cdots u_t^k$) are shifted into the STTC encoder. Each M-ary output symbol c_t^i for each of the two transmit antennas is generated by multiplying the input bits and those in shift registers S_1 and S_2 by their corresponding weights and modulo M-addition. Consider the case of QPSK, where m = 2, M=4 ($M = 2^m$) and the number of transmit antennas is two and a single receive antenna. A trellis diagram can be drawn to completely describe the space-time trellis code. The input bits u_t^1 , u_t^2 and the transmitted symbols c_t^1 and c_t^2 are shown as the rows of the trellis.

The different states are represented on the extreme right side of the trellis. The QPSK constellation is the same as Fig 2(a). The trellis is shown in Fig.2 (b),



(a.) 4PSK Constellation (b) 2-Space-Time Code 4-PSK
Fig.2

The decoder consists of a single receive antenna, where the received signal is a noisy super-position of the two transmitted signals. At the receiver, the demodulator computes a decision statistic based on the received signals arriving at the receive antenna. The received signal r_t^j at time t is given by (1),

$$r_t^j = \sum_{i=1}^{n} \alpha_{i,j} C_t^j \sqrt{E_s} + \eta_t^j \qquad (1)$$

where, the noise η_t^j at time t is modelled as independent samples of a zero-mean complex Gaussian random variable with variance No/2 per dimension. The coefficient $\alpha_{i,j}$ is the path gain from transmit antenna i to receive antenna j. The path gain is assumed to be constant during a frame and only vary from frame to frame (quasi-static).

Assuming that r_t^j is the received signal at receive antenna j at time t, the branch metric for a transition labelled q_0, q_1, q_2, q_3 is given by (2),

$$m_i^t = \sum_{j=1}^{m} | r_t^j - \sum_{i=1}^{n} \alpha_{i,j} q_t^i |^2 \qquad (2)$$

We used the Viterbi algorithm to compute the path with the lowest accumulated metric. The result obtained coincided with those results in the paper by Tarokh [1].

Analytical model

In order to compute the upper union bound on the average bit error probability

(i) We add over all error events weighting each information bit errors associated with that event,

(ii) Statistically average this sum over all possible transmitted sequences, and

(iii) Divide by the number of input bits per transmission. This can be expressed as,

$$P_b(e) \le \frac{1}{n_c} x \sum_{x} P(x) \sum_{x \ne \hat{x}} n(x,\hat{x}) P(x,\hat{x}), \tag{3}$$

where $P(x)$ is the probability that the sequence x is transmitted, and $n(x,\hat{x})$ is the number of information bit errors occurring by choosing $\hat{x}$ instead of x. For a large class of trellis codes, a symmetry property exists such that the correct sequence x can always be chosen as the all-zeros sequence, thus avoiding the necessity of averaging over all possible transmitted sequences. Codes of this type are referred to as uniform error probability (UEP) codes. Then we have,

$$P_b(e) \le \frac{1}{n_c} \sum_{x \ne \hat{x}} n(x,\hat{x}) P(x,\hat{x}). \tag{4}$$

Considering only a number of events yields an approximation (i.e. truncation of the upper bound). Thus, we have,

$$P_b(e) \cong \frac{1}{n_c} \sum_{k=1}^{K} n(x,\hat{x}) P(x,\hat{x}), \tag{5}$$

where K is the length of the error event, and the error events are classified into types according to the number of bit errors present in each of them e.g. Type1, Type2, etc.

The transfer function bounding method is an efficient method for computing the weighted sum in the union bound and based on the exact pairwise error probability (PEP) [7] this can be shown as,

$$P_b(e) \le \int_0^{\pi/2} \left[\frac{1}{n_c} \frac{\partial}{\partial I} T(D(\theta), I)\big|_{I=1} \right] d\theta \tag{6}$$

From [7], the differential of the transfer function of a 4-state QPSK STTC approximated for an error event of K=2 can be derived as,

$$\frac{\partial}{\partial I} T(\{D_\lambda(\theta)\}, I)\big|_{I=1} \cong 3D_2^2(\theta) + D_4^2(\theta) \tag{7}$$

Hence,

$$P_b(e) \cong \frac{1}{\pi} \int_0^{\pi/2} \frac{1}{2} \left[3\left(\frac{Sin^2\theta}{Sin^2\theta + \bar{\gamma}/2} \right)^{2L_r} + \left(\frac{Sin^2\theta}{Sin^2\theta + \bar{\gamma}} \right)^{2L_r} \right] d\theta \tag{8}$$

which, can be evaluated in the closed form as,

$$\begin{aligned} P_b(e) = {} & \frac{3}{4}\left[1 - \sqrt{\frac{\bar{\gamma}/2}{1+\bar{\gamma}/2}} \sum_{k=0}^{K} \binom{2k}{k} \left(\frac{1}{4(1+\bar{\gamma}/2)} \right)^k \right] \\ & + \frac{1}{4}\left[1 - \sqrt{\frac{\bar{\gamma}}{1+\bar{\gamma}}} \sum_{k=0}^{K} \binom{2k}{k} \left(\frac{1}{4(1+\bar{\gamma})} \right)^k \right] \end{aligned} \tag{9}$$

III. SUPER-ORTHOGONAL SPACE-TIME TRELLIS CODES

An example of a full-rate full diversity complex space-time block code is the scheme proposed in [9], which is defined by the following transmission matrix,

$$C(x_1, x_2) = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix}. \tag{10}$$

The scheme can be used for N=2 transmit antennas (N_{T1}) and any number of receive antennas (N_R). The scheme transmits 2b bits every two symbol intervals, where the two-dimensional (2-D) constellation size is $L = 2^b$. For each block, 2b bits arrive at the encoder and the encoder chooses two modulation symbols s_1 and s_2. Then, using $C(s_1, s_2)$, the encoder transmits s_1 from N_{T1} and s_2 from N_{T2}, at time t=1. Also, the encoder transmits $-s_2^*$ from N_{T1} and s_1^* from N_{T2} at time t=2. This scheme provides diversity gain, but no additional coding gain. It was clear that there was a rate loss associated with achieving any coding if the constituent 2-D signal constellation size does not increase. This is because these schemes are not using all of the possible 4-D signal constellations.

For example, consider other codes which provide behaviour similar to those of (10) for the same rate and number of transmit antennas. The set of all such codes which use x_1, x_2 and their conjugates with positive or negative signs are listed as follows,

$$\begin{aligned} & \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix}, \begin{bmatrix} -x_1 & x_2 \\ x_2^* & x_1^* \end{bmatrix}, \begin{bmatrix} x_1 & -x_2 \\ x_2^* & x_1^* \end{bmatrix}, \\ & \begin{bmatrix} x_1 & x_2 \\ x_2^* & -x_1^* \end{bmatrix}, \begin{bmatrix} -x_1 & -x_2 \\ x_2^* & -x_1^* \end{bmatrix}, \begin{bmatrix} -x_1 & x_2 \\ -x_2^* & -x_1^* \end{bmatrix}, \\ & \begin{bmatrix} x_1 & -x_2 \\ -x_2^* & x_1^* \end{bmatrix}, \begin{bmatrix} -x_1 & -x_2 \\ -x_2^* & x_1^* \end{bmatrix}, \end{aligned} \tag{11}$$

The union of all these codes is referred to as "super-orthogonal code" set C. Using just one of the constituent codes from C, e.g. the code in (10), one cannot create all possible orthogonal 2x2 matrices for a given constellation. By using two of the codes in C, it was shown that it could be possible to build all possible 2x2 orthogonal matrices [2]. Together with the code from (10), one can create additional matrices from the code,

$$\begin{bmatrix} -x_1 & x_2 \\ x_2^* & x_1^* \end{bmatrix} \tag{12}$$

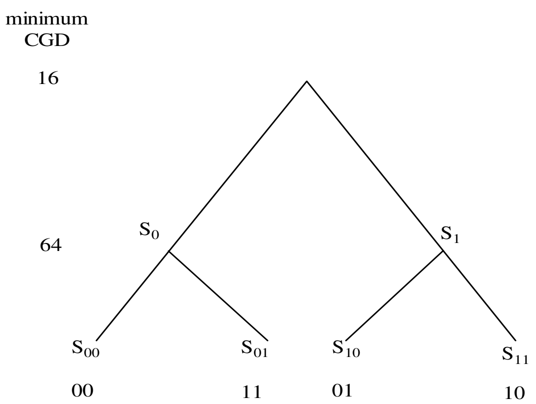
This represents a phase shift of the signals transmitted from Tx1 by π. We denote a set including all 2x2 orthogonal matrices from (10) and (12) as O_2. By using more than one code from set C, we can create all possible 2x2 orthogonal matrices from O_2. Therefore, the scheme provides a sufficient number of constellation matrices to design a trellis codes with the highest possible rate.

Hence for the SOSTTC scheme, we assign a STBC with specific constellation symbols to transitions originating from a state. Therefore, in general, for a TxN STBC, picking a trellis branch emanating from a state is equivalent to transmitting NT symbols from N transmit antennas in T time intervals. By doing so, it is guaranteed that we get the diversity of the corresponding STBC and the design of trellis code for the highest possible rate, as was done in [STTC], to get maximum coding gain as well. In this paper, we use the following class of orthogonal designs as transmission matrices similar to [2],

$$C(x_1, x_2, \theta) = \begin{bmatrix} x_1 e^{j\theta} & x_2 \\ -x_2 e^{j\theta} & x_1^* \end{bmatrix}. \tag{13}$$

Then the codes obtained for $\theta = 0$ is $C(x_1, x_2, 0)$ and when $\theta = \pi$ is $C(x_1, x_2, \pi)$. So by using $C(x_1, x_2, 0)$ and $C(x_1, x_2, \pi)$ for the constellations, one can obtain all 2x2 orthogonal matrices in O_2. Therefore, the set of orthogonal codes is a subset of the set of super-orthogonal codes. Also, while super-orthogonal code does not extend the constellation alphabet of the transmitted signals, it does expand the number of available orthogonal matrices. This is of great benefit and crucial in the design of full-rate, full-diversity trellis codes. Another advantage is that the codes are parameterized.

In this paper the coding gain distance (CGD) is used instead of Euclidean distance to define set partitioning similar to Ungerboeck's set partitioning [10]. Consider a four-way partitioning of the orthogonal code as shown in Fig. 3 for the BPSK case.



At the root of the tree, the minimum determinant is 16. At the first level of partitioning, the highest determinant that can be obtained is 64. This is obtained by a set partitioning in which subsets S_0 and S_1 use different transmitted signal elements for different transmit antennas. At the next level of partitioning, we have four sets S_{00}, S_{11}, S_{01} *and* S_{10} with only one element per set. Thus, it is important to ensure that we do the set partitioning such that the CGD is maximized at each level of partitioning. Therefore, a fundamental rule in set partitioning is to choose the codeword that contain signal elements with highest maximum Euclidean distance from each other as the leaves of the set-partitioning tree.

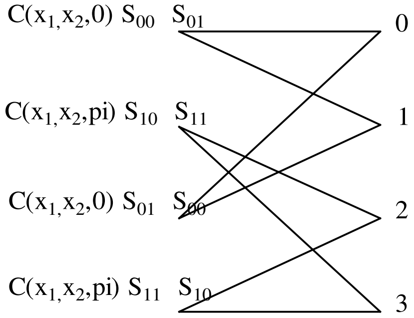


Fig. 4(a)

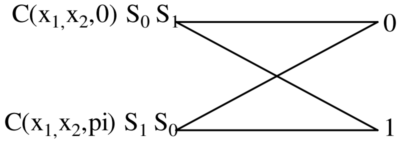


Fig. 4(b)

Fig. 4(a) and 4(b) trellis diagrams of SOSTTCs

Fig. 4 shows an example of the SOSTTC. In this example, we use BPSK and corresponding set partitioning of Fig. 3, and the code rate is 1. We use $C(x_1, x_2, 0)$ when departing from states zero and two and use $C(x_1, x_2, \pi)$ when departing from sates one and three. So with this structure, we have eight possible orthogonal 2x2 matrices instead of four which allows us to design a full-rate code. If we use QPSK constellation the corresponding set partitioning in Fig. 3, the result is a four state SOSTTC code at rate 2b/s/Hz. The minimum CGD for this 2b/s/Hz code is equal to 16 which is greater than 4, the CGD of the corresponding STTC [1].

IV. ANALYTICAL BOUNDS

First we consider the rate $r = 1$ BPSK 2-state code in Fig. 4(b), where two sets, each containing two pairs of BPSK symbols, are assigned to each state, i.e. taking into account the parallel paths between each pair of states. Building on the previous discussion in section II, we consider an error event of length N=1, for a fast fading Rayleigh channel the PEP associated with parallel paths is from [3] given by,

$$P(X \to \widehat{X}) = \frac{1}{\pi}\int_0^{\pi/2}\left(\frac{\sin^2\theta}{\sin^2\theta + 2\overline{\gamma}_s}\right)^{2L_r} d\theta \tag{14}$$

which, can be evaluated in the closed form as,

$$P(X \to \widehat{X}) = \frac{1}{2}\left\{1 - \sqrt{\frac{2\overline{\gamma}_s}{1+2\overline{\gamma}_s}}\sum_{k=0}^{2L_r-1}\binom{2k}{k}\left[\frac{1}{4(1+2\overline{\gamma}_s)}\right]^k\right\} \tag{15}$$

where, $L_r = 1$ γ_s is the signal to noise ratio.

For an error event path of length N=2, we have,

$$P(X \to \widehat{X}) = \frac{1}{\pi}\int_0^{\pi/2}\left(\frac{\sin^2\theta}{\sin^2\theta + \overline{\gamma}_s}\right)^{2L_r}\left(\frac{\sin^2\theta}{\sin^2\theta + 2\overline{\gamma}_s}\right)^{L_r} d\theta \quad , (16)$$

and its closed form is,

$$P(X \to \widehat{X}) = \frac{1}{2}\left[1 - 4\sqrt{\frac{2\overline{\gamma}_s}{1+2\overline{\gamma}_s}} + \sqrt{\frac{\overline{\gamma}_s}{1+\overline{\gamma}_s}}\left(3 + \frac{1}{2(1+\overline{\gamma}_s)}\right)\right]. \tag{17}$$

For a slow Rayleigh fading, the PEP for parallel path is still given by (15) and for an error event of length N=2, we have,

$$P(X \to \widehat{X}) = \frac{1}{\pi}\int_0^{\pi/2}\left(\frac{\sin^2\theta}{\sin^2\theta + 3\overline{\gamma}_s}\right)^{L_r}\left(\frac{\sin^2\theta}{\sin^2\theta + \overline{\gamma}_s}\right)^{L_r} d\theta \tag{18}$$

This, in the closed form is evaluated by,

$$P(X \to \widehat{X}) = \frac{1}{2}\left[1 - \frac{3}{2}\sqrt{\frac{3\overline{\gamma}_s}{1+3\overline{\gamma}_s}} + \frac{1}{2}\sqrt{\frac{\overline{\gamma}_s}{1+\overline{\gamma}_s}}\right]. \tag{19}$$

The above derived PEPs can be used to evaluate in closed form an approximation to the average bit-error probability (BEP) based on accounting only for error events of lengths N less than or equal to H. Hence, if we were to choose to approximate the average BEP by considering only error event paths to H=1 and H=2 then,

$$P_b(E) \cong \frac{1}{2} P\left(X \to \widehat{X}\right) \tag{20}$$

and

$$P_b(E) \cong \frac{1}{2}\left[P(X \to \widehat{X})_I + 12P(X \to \widehat{X})_{II}\right] \quad , (21)$$

where, for the fast Rayleigh channel, $P(X \to \widehat{X})_I$ and $P(X \to \widehat{X})_{II}$ are given by the closed-form expressions in (15) and (17), respectively.

V. RESULTS

In this section, we provide simulation results for STTC and SOSTTCs using two transmit and one receive antenna systems. All results are obtained using frame lengths of 130 transmissions. In Fig. 5, simulation results for 4-state QPSK modulation and 2-state QPSK and BPSK constellations are provided. The codes provide a 2 dB gain over the standard STTC [2], in quasi-static fading channels. In our work we provide simulation results for 4-state QPSK and 2-state SOSTTC codes for rapid Rayleigh fading channels and compare the performance of STTCs in similar conditions. Although the codes show degradation in performance in rapid fading channels compared to the quasi-static case, the SOSTTC codes still maintain a 2 dB gain over the STTC case as shown in Fig. 6. Analytical bounds help provide an understanding of the performance of these codes to justify the obtained simulation results and also provide means for obtaining results where simulation results are not feasible to obtain. We provide analytical results for SOSTTC 2-state codes for error events of length N=1 and N=2 in terms of their PEPs in Fig. 7. The average BEP for 2-state SOSTTC codes shown in Fig. 8, for N=1 and N=2 error events for both quasi-static and rapid Rayleigh fading channels, and for N=2 STTC code in quasi-static fading channels are provided. As the bounds are truncated we obtain the lower bounding performance of the codes due to the short length of the error events considered.

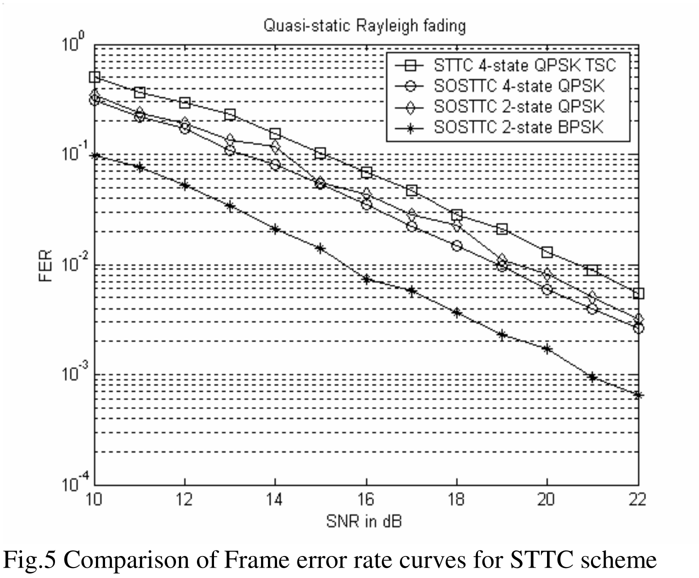


Fig.5 Comparison of Frame error rate curves for STTC scheme and the SOSTTC scheme in quasi-static Rayleigh fading channels.

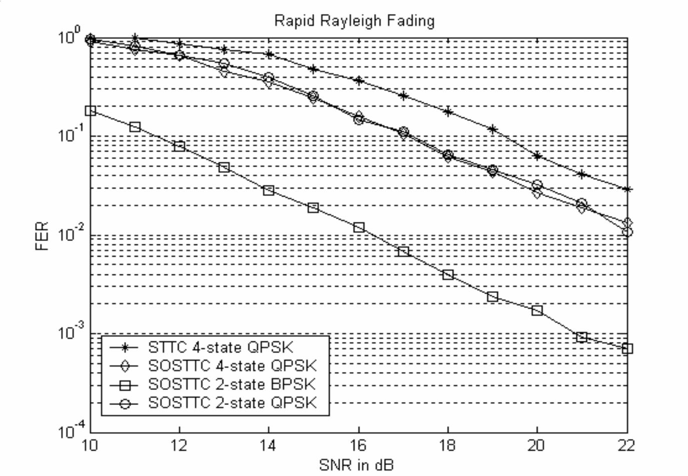


Fig. 6 Comparison of Frame error rate curves for STTC scheme and the SOSTTC scheme in rapid Rayleigh fading channels

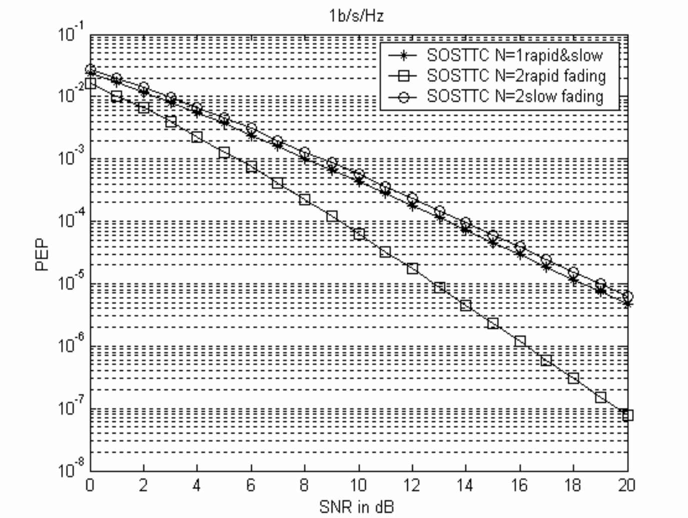


Fig. 7 Comparison of PEP curves for N=1 and N=2 error events of 2-state SOSTTC in both rapid and quasi-static Rayleigh fading channels

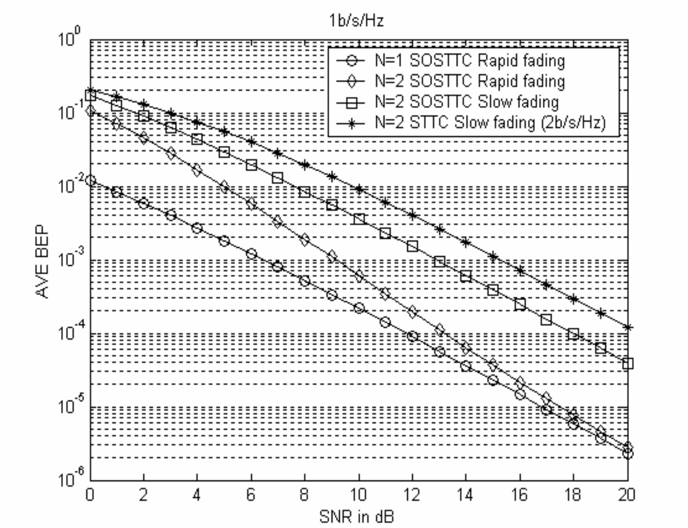


Fig. 8 Comparison of average BEP curves for N=1 and N=2 error events of 2-state SOSTTC and 4-state STTC in both rapid and quasi-static Rayleigh fading channels.

VI. Conclusion

The paper presents a very comprehensive study of SOSTTCs in both quasi-static and rapid Rayleigh fading channels in terms of both simulation and analytical results. These codes provide a systematic design methodology to provide full diversity and improved coding gain over STTC for single receive antennas in varying channel conditions and also provide a platform to design new codes in a more systematic manner.

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