

ERROR PERFORMANCE OF CONCATENATED SUPER-ORTHOGONAL SPACE-TIME-FREQUENCY TRELLIS CODED MIMO-OFDM SYSTEM

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Abstract: In this paper, we investigate the performance of serially concatenated convolutional code with super-orthogonal space-time trellis code (SOSTTC) in orthogonal frequency division multiplexing (OFDM) over frequency selective fading channels. We consider both recursive systematic convolutional code (RSC) and non-recursive convolutional code (NRC) as the outer code, and 16-state QPSK SOSTTC as the inner code. Employing these, two concatenated schemes consisting of single convolutional outer code and two serially concatenated convolutional outer codes are proposed. We evaluate the performance of the concatenated schemes by means of computer simulations with maximum *a posteriori* (MAP) algorithm based iterative decoding. Simulation results indicate that the performance of the proposed concatenated schemes improved significantly when compared with schemes without concatenation under the same channel condition.

Keywords: Coding gain, convolutional code, frequency selective fading channels, iterative decoding, orthogonal frequency division multiplexing, super-orthogonal space-time trellis code.

1. INTRODUCTION

Space-time coding (STC) has been shown to be an effective method of increasing the capacity of wireless communication channels by combining the benefits of diversity transmission and error correction coding to combat impairments of wireless channels [1-4]. Super-orthogonal space-time trellis code (SOSTTC) is the recently proposed space-time code (STC) that combine set partitioning based on the coding gain distance and a super set of orthogonal space-time block code in a systematic way, to provide full diversity and improved coding gain over the earlier proposed space-time trellis code constructions [5-8].

Orthogonal Frequency Division Multiplexing (OFDM) is a multicarrier modulation scheme used to combat frequency selective fading channels [9]. Although OFDM eliminates the inter-symbol interference (ISI) problem caused by the multipath effect, it does not eliminate errors caused by channel fading and additive white Gaussian noise (AWGN) in wireless channels [10]. Antenna diversity, through space-time coding, is one of the adopted techniques being used to improve the performance of OFDM systems in the presence of channel fading and AWGN impairment. Due to its high bandwidth efficiency and suitability for high data-rate wireless applications, OFDM was chosen as a modulation scheme for the physical layer in several new wireless standards such as digital audio and digital video broadcasting (DAB, DVB) in Europe, the three broadband wireless local area networks (WLAN), European HIPERLAN/2, American IEEE 802.11a and Japanese MMAC [11].

Space-time coded OFDM system was first introduced in [12] where OFDM technique was employed to transform a frequency selective fading channel into many flat fading channels. The initial work of [12] led to many design considerations for space-time coded OFDM system in order to improve its performance [13-17]. It is known that STCs are designed to maximize the diversity gain for a given number of transmit antennas and that the coding gain of STC is low. Increasing the number of states of STC will lead to an increase in the achievable coding gain but the decoding complexity also increases exponentially [10]. Concatenated coding schemes with sub-optimum, yet powerful iterative decoding, have been shown to guarantee improved error performance while the complexity of the decoders is kept comparable to single coding schemes [18, 19]. Several concatenated schemes with constituent codes of STC and convolutional codes were proposed in [20-27], with reported improved coding gain over their un-concatenated counterparts.

In order to improve the coding gain of SOSTTC in frequency selective fading channels, we hereby propose two concatenated schemes consisting of convolutional codes and SOSTTC for OFDM systems. The first involves a serially concatenated convolutional outer code with SOSTTC inner code while the second scheme involves two outer serially concatenated convolutional codes with SOSTTC inner code. The systems have the advantage of achieving diversity gain by exploiting available diversity resources of the frequency selective fading channel. Also, by iterative information exchange, the concatenation schemes achieve additional decoding gain without bandwidth expansion. It is well known that the theoretical evaluation of the exact performance of such concatenated schemes using iterative decoding in

frequency selective fading channel is a very difficult task, and hence we used computer simulations to evaluate the performance of the proposed systems. In particular, we considered rate $\frac{1}{2}$ and rate $\frac{2}{3}$ outer convolutional codes for both recursive and non recursive outer codes. As pointed out in [5-8], SOSTTC has a large number of parallel transitions in their trellis which limit their error performance in frequency selective fading channels. To avoid such parallel transitions, at least a 16-state code for QPSK is required [28, 29]. This is the reason for the choice of the 16-state QPSK SOSTTC as the inner code for the proposed concatenated schemes.

The rest of the paper is organized as follows. In section 2, we describe the system model which includes a brief description of the channel model, the transmitter and the receiver structure. Section 3 describes the outer and the inner code used in this paper while the error performance of the proposed schemes is evaluated by computer simulation in section 4. Finally, section 5 concludes this paper.

2. SYSTEM MODEL

2.1 Channel model

We considered a MIMO-OFDM system consisting of two transmit antenna and M_r receive antennas. Each transmit antenna employs an OFDM modulator with K subcarriers. We assume no spatial correlation exists between the antennas and that the receiver has perfect knowledge of the channel while the channel is unknown to the transmitter. The channel impulse response (CIR) between the transmit antenna p and receive antenna q with L independent delay paths on each OFDM symbol and an arbitrary power delay profile can be expressed as [30]

$$h_{p,q}(t) = \sum_{l=0}^{L-1} \alpha_{p,q}(l) \delta(t - \tau_l), \quad (1)$$

where τ_l represents the l th path delay and $\alpha_{p,q}(l)$ are the fading coefficients at delay τ_l . Note that each $\alpha_{p,q}(l)$ is a complex Gaussian random variable with zero mean and variance $\frac{\sigma_l^2}{2}$ on each dimension. For normalization purposes we assumed that $\sum_{l=0}^L \sigma_l^2 = 1$ in each of the transmit–receive links.

The channel frequency response (CFR), that is the fading coefficient for the k th subcarrier between transmit antenna p and receive antenna q with a proper cyclic prefix and a perfect sampling time, is given by

$$H_{p,q}(k) = \sum_{l=0}^{L-1} \alpha_{p,q}(l) e^{-j2\pi\Delta_f \tau_l} \quad (2)$$

where Δ_f is the inter-subcarrier spacing, $\tau_l = lT_s$ is the l th path delay and $T_s = \frac{1}{K\Delta_f}$ is the sampling interval of the OFDM system.

A space-frequency codeword for two transmit antennas transmitted at the t th OFDM symbol period can be represented by $\mathbf{C}_{SF}^t = [c_1^t(k) \ c_2^t(k)] \in C^{K \times 2}$, where $c_p^t(k)$ is the complex data transmitted by the p th transmit antenna at the k th subcarrier for, $k = 0, \dots, K-1$. Moreover, \mathbf{C}_{SF}^t satisfies the power constraint $E\|\mathbf{C}_{SF}^t\|_F^2 = K$. A STC codeword has an additional time dimension added to the above space frequency codeword, and can be represented as $\mathbf{C}_{STF} = [\mathbf{C}_{SF}^t \ \mathbf{C}_{SF}^{t+1}] \in C^{K \times 2 \times 2}$. At the receiver, after matched filtering, removal of the cyclic prefix and application of fast Fourier transform (FFT), the signal at the k th subcarrier and antenna q is given by

$$r_q^t(k) = \sum_{i=1}^{n_T=2} c_i^t(k) H_{p,q}^t(k) + N_q^t(k), \quad (3)$$

where $q = 1, \dots, M_r$, and $N_q^t(k)$ is a circularly symmetric Gaussian noise term, with zero-mean and variance N_0 at t th symbol period.

2.2 Encoder structure

Convolutional code with super-orthogonal space-time trellis code (CC-SOSTTC-OFDM): We consider a serially concatenated Multi-Input Multi-Output (MIMO) OFDM communication system that employs $n_T = 2$ antennas at the transmitter, and $n_R = 1$ antenna at the receiver. The transmitting block diagram of the concatenated scheme is shown in Figure 1. The encoder consists of an outer convolutional code concatenated with an inner SOSTTC code. In this system, a block of N independent data bits is encoded by the convolutional outer encoder and the output block of coded bits are interleaved by using a random bit interleaver (π). The interleaved sequence are then passed to the SOSTTC encoder to generate a stream of QPSK symbols. Each of the symbols from the SOSTTC encoder is converted to a parallel output, and an inverse fast Fourier transform (IFFT) is performed on each of the parallel symbols. At the end, cyclic prefix (CP) is added to each of the transformed symbols before transmission from each of the transmit antennas.

Double convolutional code with super orthogonal space-time trellis code (CC-CC-SOSTTC-OFDM): The encoder structure of the double concatenated scheme is shown by Figure 2. Two outer serially concatenated convolutional codes are concatenated with an inner SOSTTC in a bid to improve the overall coding gain of the system. The encoding process is similar to that described above, except for the addition of an extra convolutional outer encoder. In this system, a block of N independent data bits is encoded by the first convolutional outer encoder and then interleaved using a random bit interleaver (π_1). The output stream from the interleaver is then encoded by the second convolutional outer code and thereafter interleaved by the second interleaver (π_2). The permuted sequence is thereafter SOSTTC encoded. The remaining process of encoding follows the description given above. In the two systems, both the recursive systematic (RSC) and non recursive convolutional (NRC) codes were considered and each of the encoders was terminated using appropriate tail bits.

2.3 Decoder structure

In this section, a description of the iterative decoding of the two concatenated schemes is given. The employed decoders operate on bit streams using the Soft-Input Soft-Output (SISO) algorithm [31]. Extrinsic information is exchanged between the component decoders using the soft estimates of their Log Likelihood Ratio (LLR) with the presence of the feedback loop.

CC-SOSTTC-OFDM decoder: The decoder structure of the CC-SOSTTC-OFDM is shown in Figure 3. As shown in the figure, the inserted CP is first removed and thereafter fast Fourier transform (FFT) is performed on each of the symbols. The parallel outputs obtained from this transformed symbol are then converted to serial streams for computation of the coded intrinsic LLR of the SOSTTC SISO module.

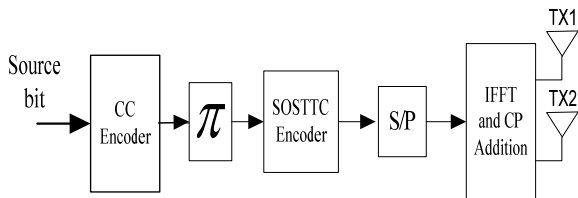


Figure 1: Encoder block diagram of CC-SOSTTC-OFDM system.

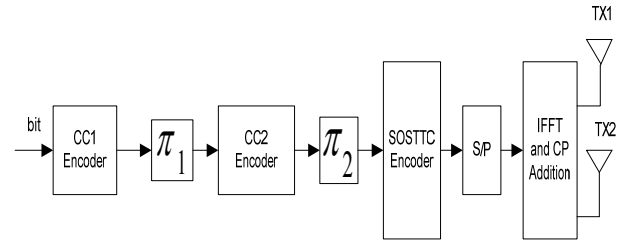


Figure 2: Encoder block diagram of double serially concatenated SOSTTC-OFDM (CC-CC-SOSTTC-OFDM) system.

Given that the received symbol from subcarrier k is

$$r(k) = x(k)H(k) + w(k), \quad (4)$$

the coded intrinsic LLR for the SOSTTC SISO $\lambda(C_{st}, I)$ is computed as [26]

$$\lambda(C_{st}, I) = \log \frac{\Pr[x | r(k)]}{\Pr[x_0 | r(k)]}, \quad (5)$$

where x_0 is a reference symbol. By dropping the subcarrier index k for simplicity, we can express (5) by

$$\lambda(C_{st}, I) = -\frac{1}{2\sigma^2} \sum_{q=1}^{n_R} \left| r - \sum_{p=1}^{n_T} H_{pq} x \right|^2 + \frac{1}{2\sigma^2} \sum_{q=1}^{n_R} \left| r - \sum_{p=1}^{n_T} H_{pq} x_0 \right|^2, \quad (6)$$

where σ^2 is the variance of the independent complex Gaussian noise variable.

The SOSTTC SISO takes $\lambda(C_{st}, I)$ and the *a priori* information from the CC-SISO (initially set to zero) and compute the extrinsic information given by

$$\hat{\lambda}(U_{st}, O) = \lambda(U_{st}, O) - \lambda(U_{st}, I). \quad (7)$$

This extrinsic information is de-interleaved (π^{-1}) and fed to the CC-SISO to become its *a priori* information $\lambda(C_{cc}, I)$. The *a priori* information is then used to compute the extrinsic information for the convolutional code SISO (CC-SISO). The extrinsic information for the CC-SISO is given by

$$\tilde{\lambda}(C_{cc}, O) = \lambda(C_{cc}, O) - \lambda(C_{cc}, I). \quad (8)$$

The extrinsic LLR in (8) is then interleaved to become the *a priori* information $\lambda(U_{st}, I)$ for the SOSTTC SISO for the next iteration. During the first iteration, we set $\lambda(U_{st}, I)$ to zero, as no *a priori* information is available at the SOSTTC-SISO. We assumed that the source symbols transmitted are equally likely. Therefore, the input LLR $\lambda(U_{cc}, I)$ to the CC-SISO is permanently set to zero. We iterated the process several times. On the final iteration, a decision is taken on the extrinsic information $\lambda(C_{cc}, O)$ to obtain the estimate of the original transmitted bit stream.

CC-CC-SOSTTC-OFDM decoder: The decoder structure of the CC-CC-SOSTTC-OFDM is shown in Figure 4. During the first iteration, the LLRs $\lambda(U_{st}, I)$ and $\lambda(U_2, I)$ are set to zero as no *a priori* information is available. Since we assumed that the source symbols transmitted are equally likely, the input LLR $\lambda(U_1, I)$ to the C1-SISO is permanently set to zero. The coded intrinsic LLR for the SOSTTC SISO $\lambda(C_{st}, I)$ is computed using (6).

The SOSTTC SISO takes $\lambda(C_{st}, I)$ and the *a priori* information from the C2-SISO (initially set to zero) for the computation of the extrinsic information given by

$$\hat{\lambda}(U_{st}, O) = \lambda(U_{st}, O) - \lambda(U_{st}, I). \quad (9)$$

The extrinsic LLR $\tilde{\lambda}(U_{st}, O)$ is then de-interleaved through π_2^{-1} to become the input LLR $\lambda(C_2, I)$ of the SISO decoder for the second outer convolutional code. The C2-SISO takes the input LLR $\lambda(C_2, I)$ and the *a priori* information from the C1-SISO to compute the extrinsic information for the C2-SISO of the coded and the un-coded values. The un-coded extrinsic information for the C2-SISO decoder is given by

$$\tilde{\lambda}(U_2, O) = \lambda(U_2, O) - \lambda(U_2, I). \quad (10)$$

The extrinsic information from (10) is then de-interleaved by π_1^{-1} to become the intrinsic LLR information for the C1-SISO decoder $\lambda(C_1, I)$. The intrinsic information is then used by the C1 SISO to compute the extrinsic information for the C1 SISO as

$$\hat{\lambda}(C_1, O) = \lambda(C_1, O) - \lambda(C_1, I). \quad (11)$$

The C1-SISO output LLR is thereafter passed through the interleaver π_1 to obtain the *a priori* information for the C2-SISO. The coded LLR output obtained from the C2-SISO given by

$$\tilde{\lambda}(C_2, O) = \lambda(C_2, O) - \lambda(C_2, I). \quad (12)$$

is also passed through the interleaver π_2 to obtain the *a priori* information $\lambda(U_{st}, I)$ for SOSTTC-SISO for the next iteration. During the final iteration, decision is taken on $\lambda(U_1, O)$ from the C1-SISO output to obtain the estimate of the original transmitted symbol.

3. COMPONENT CODES

3.1 Inner code

The SOSTTC code with the transmission matrix given by (13) is considered as the inner codes.

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

where for M-PSK signal constellation, the signals x_1 and x_2 which are selected by input bits can be represented by $e^{j\frac{2\pi l}{M}}$, where $l = 0, 1, \dots, M-1$ and θ which is the rotation angle can take on the values $\theta = 2\pi l'/M$, where $l' = 0, 1, \dots, M-1$. As noted in [28, 29] and in [32], SOSTTC have parallel transitions that limit its error performance in a frequency selective fading channel. To exploit the multipath diversity of frequency selective fading channels, at least 16-state SOSTTC is needed for QPSK constellation. We considered 16-state QPSK SOSTTC with the trellis diagram shown by Figure 5 [33]. The SOSTTC was designed using the set partitioning principle of [5] and design rules outlined in [7, 8]. For comparison, we also considered a 16-state QPPK STTC presented in [1] as inner code.

3.2 Outer code

Convolutional codes are considered as the outer codes. We considered both recursive systematic and non recursive codes. The generator matrices of the outer codes considered in our investigation are given in Table 1.

4. RESULTS

In this section, we provide the simulation results illustrating the performance of the proposed concatenated schemes for multiple-antenna OFDM systems. The performance is presented in terms of frame error rate (FER) versus the received SNR for QPSK constellation. To properly model a frequency selective fading channel, we considered the typical urban six-path (TU6) COST 207 model reported in [34]. The OFDM modulator utilizes 64 subcarriers with a total system bandwidth of 800 kHz and FFT duration of 100 μ s.

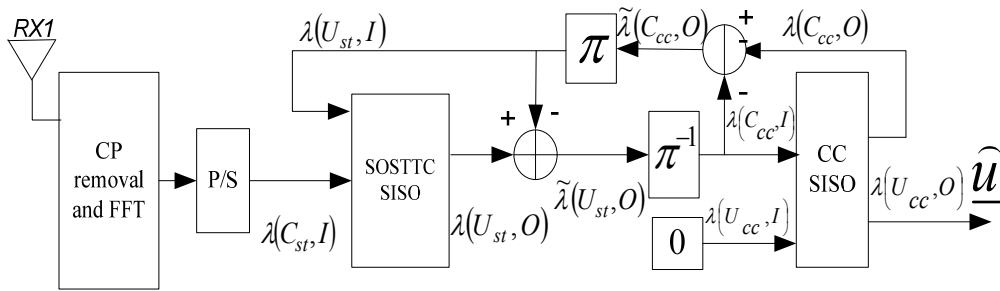


Figure 3: Decoder block diagram of CC-SOSTTC-OFDM.

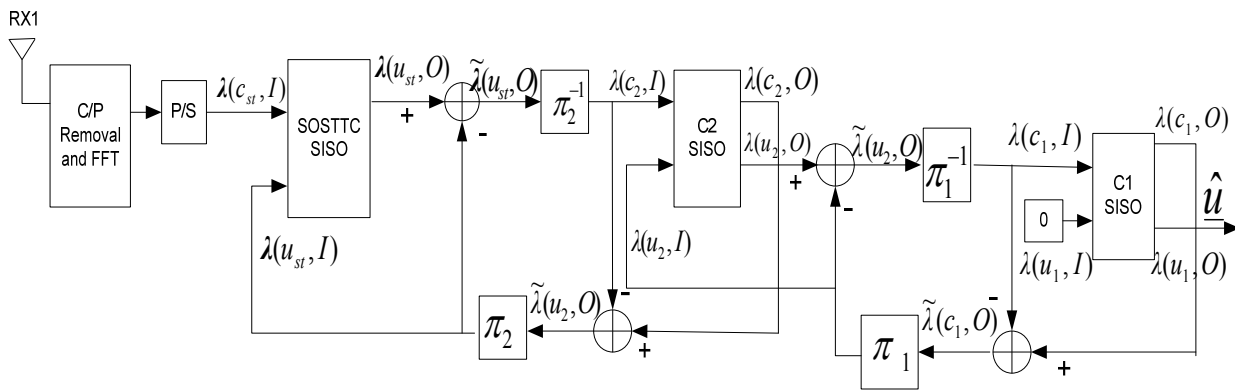


Figure 4: Decoder block diagram of CC-CC-SOSTTC-OFDM.

The system's subcarrier spacing is 12.5 kHz with symbol duration of 80 μ s while the guard band interval is 20 μ s. The channel is assumed to be static during one OFDM symbol duration and the receiver is assumed to have the full knowledge of the channel (perfect channel estimation). We also assume perfect time and frequency synchronization between the transmitter and the receiver.

4.1 CC-SOSTTC-OFDM

Figure 6 shows the FER simulation results for the first concatenated scheme for various numbers of iterations. The amount of improvement resulting from iteration is seen to become smaller as the number of iteration increases. This is especially evident after the 4th iteration. In Figure 7 the FER performance is shown for normalized Doppler frequencies f_{dt} of 0.05, 0.02, 0.01 and 0.005, corresponding to mobile speed of 150 m/s, 60 m/s, 30 m/s and 15 m/s respectively. Rate $\frac{1}{2}$, 4-state recursive convolutional code is considered in this investigation. As it can be observed, the scheme's FER performance improves as the mobile speed decreases. This is due to the fact that the channels are more correlated at high normalized Doppler frequencies. In other words, channel correlation may cause long burst errors during deep fades, which degrades the FER performance.

Figure 8 illustrates the effect of the number of the states of the outer codes on the performance of the proposed

schemes. Here, the systems have 2 transmit antennas and 1 receive antenna. We evaluate the system using the 4-state, 8-state and 16-state rate $\frac{1}{2}$ non recursive convolutional codes as the outer code. As can be observed from Figure 8, the system with 4-state exhibits the best performance. This shows that increasing the number of the states of the outer code results in degradation of the system's performance. This is a typical phenomenon in iterative decoding [20, 35].

4.2 CC-CC-SOSTTC-OFDM

For the CC-CC-SOSTTC-OFDM scheme, we considered the rate $\frac{2}{3}$ outer codes for the case of both recursive and non recursive code. Figure 9 shows the FER variation with the number of decoding iterations. The performance improvement produced by each iteration is observed to converge with increase in the number of iterations. Figure 10 also shows the effect of the mobile speed on the FER performance of the proposed scheme for various normalized Doppler frequencies. In Figure 11 we show the FER results for four various combinations of the outer channel codes. From the FER performance curve, it is observed that having RSC as the middle code achieves the best performance irrespective of having either RSC or NRC as the outer code.

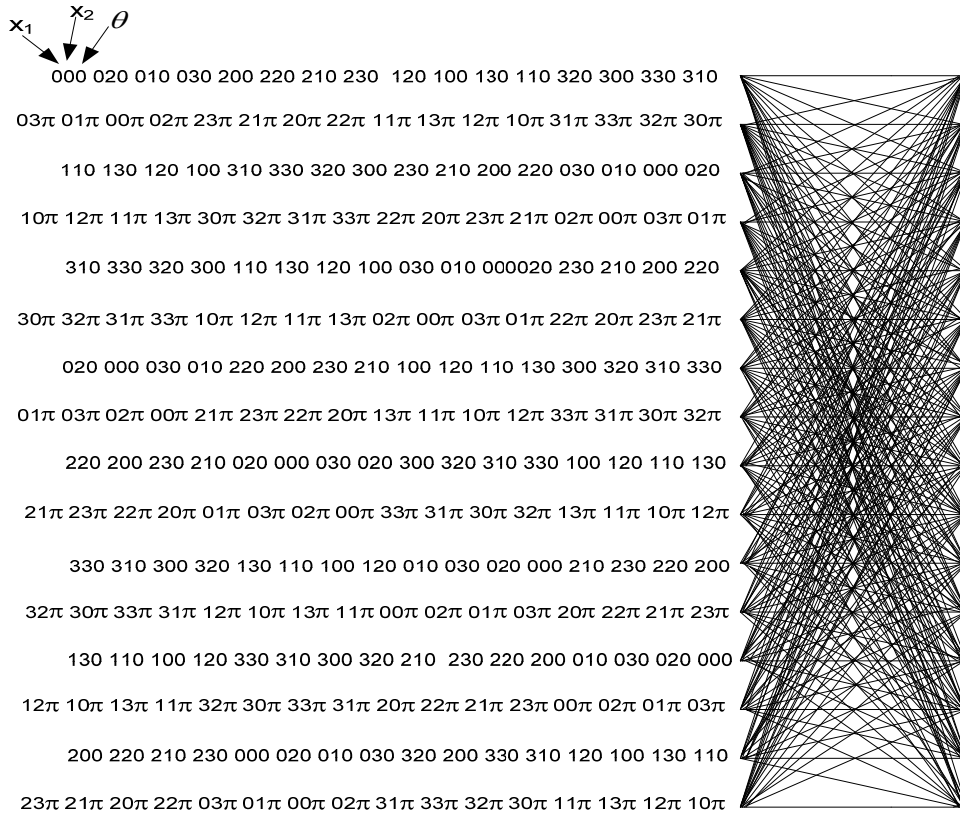


Figure 5: Trellis diagram of 16 states SOSTTC at rate 2 bits/s/Hz [33].

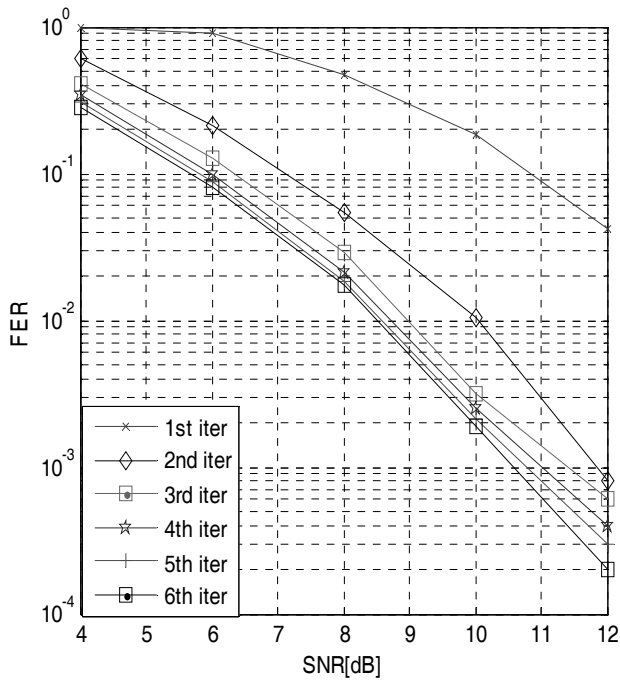


Figure 6: FER performance of CC-SOSTTC-OFDM for various numbers of iterations.

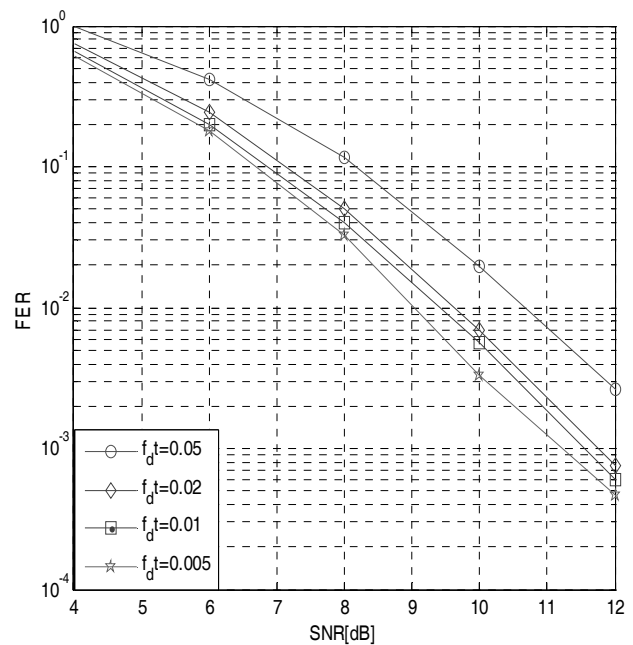


Figure 7: FER performance of the CC-SOSTTC-OFDM with different normalized Doppler frequency.

Table1: Generating matrices for the outer convolutional codes.

No	Code Description	G(D)
1	4-state rate $\frac{1}{2}$ NRC	$[1 + D + D^2, 1 + D^2]$
2	4-state rate $\frac{1}{2}$ RSC	$\left[1, \frac{1 + D^2}{1 + D + D^2} \right]$
3	8-state rate $\frac{1}{2}$ NRC	$[1 + D + D^2 + D^3, 1 + D^2 + D^3]$
4	16-state rate $\frac{1}{2}$ NRC	$[1 + D^2 + D^3 + D^4, 1 + D + D^4]$
5	4-state rate $\frac{2}{3}$ NRC	$\begin{bmatrix} 1 + D, & D, & 1 \\ 1 + D, & 1, & 1 + D \end{bmatrix}$
6	4-state rate $\frac{2}{3}$ RSC	$\left[1, 0, \frac{1 + D^2}{1 + D + D^2} \right]$ $\left[0, 1, \frac{1 + D}{1 + D + D^2} \right]$

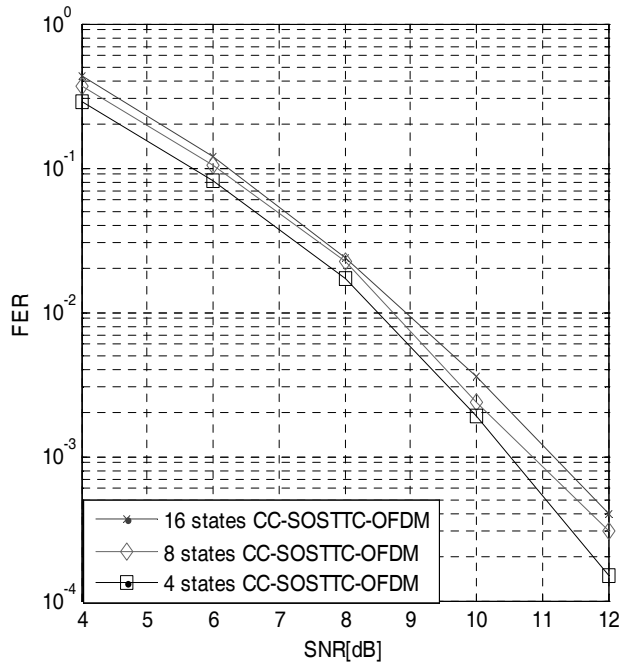


Figure 8: FER performance of CC-SOSTTC-OFDM for different states of outer code.

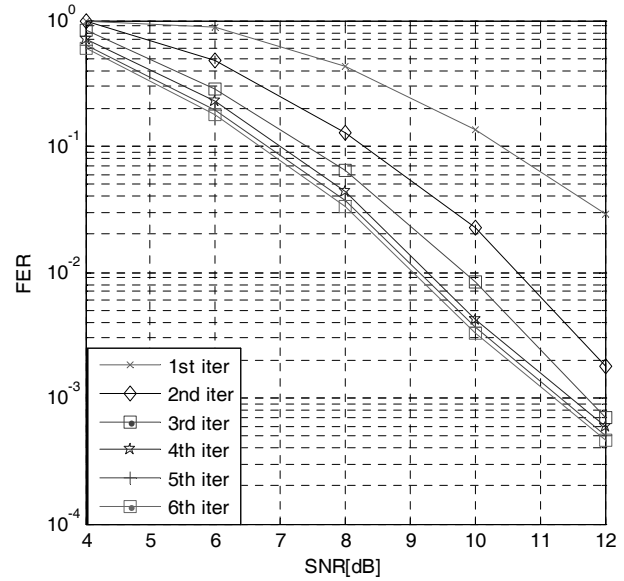


Figure 9: FER performance of CC-CC-SOSTTC-OFDM for various number of iteration.

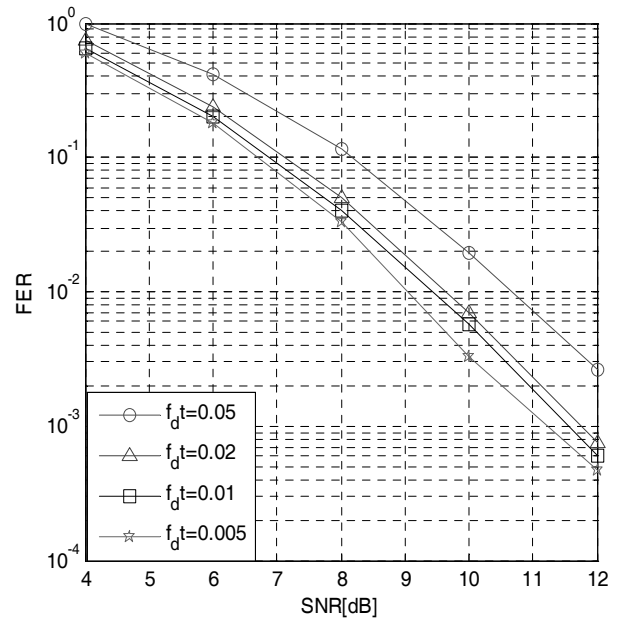


Figure 10: FER performance of CC-CC-SOSTTC-OFDM over various normalized Doppler frequencies.

For the purpose of performance comparison we show the FER of 16-state SOSTTC-OFDM, CC-STTC-OFDM, CC-SOSTTC-OFDM, CC-CC-STTC-OFDM, and the CC-CC-SOSTTC-OFDM schemes in Figure 12. The CC-SOSTTC-OFDM is seen to outperform the SOSTTC-OFDM by about 5 dB at FER of 10^{-3} and also outperform the CC-STTC-OFDM by about 2.5 dB at the same FER. The CC-CC-SOSTTC-OFDM outperforms the SOSTTC-OFDM by about 7 dB at FER of 10^{-3} and also outperforms the CC-CC-STTC-OFDM with about 1.4 dB at the same FER. The CC-CC-SOSTTC-OFDM scheme also outperforms the CC-SOSTTC-OFDM counterpart by about 2 dB at FER 10^{-3} albeit with loss in bandwidth efficiency.

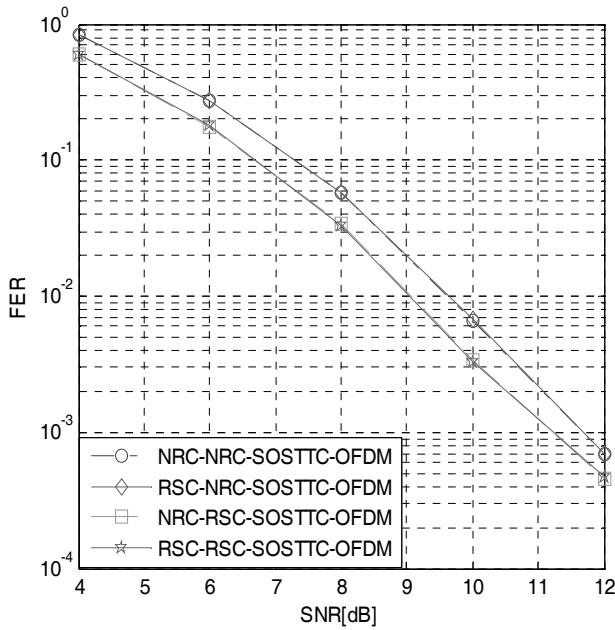


Figure 11: FER performance of CC-CC-SOSTTC-OFDM for different outer code combinations for 2 transmits and 1 receive antennas.

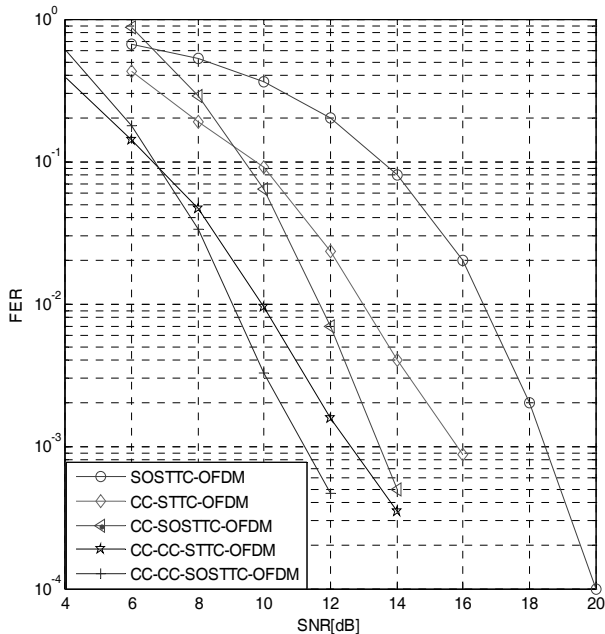


Figure 12: FER comparisons of SOSTTC-OFDM, CC-STTC-OFDM, CC-SOSTTC-OFDM and CC-CC-SOSTTC-OFDM for 2 transmit and 1 receive antenna.

5. COMPARATIVE DECODING COMPLEXITY AND BANDWIDTH EFFICIENCY

In this section, the relative estimated complexity and bandwidth efficiency of the proposed concatenated

schemes in MIMO-OFDM systems are presented. The approach used in [36] is herewith adopted in analyzing the complexity of the various proposed schemes. As stated in [36], the complexity of the channel decoders depends directly on the number of trellis transition per information data bits. This will be used as the basis for our comparison. For SOSTTC, the number of trellis leaving each state is equivalent to 2^{BPS} , where BPS is the number of transmitted bits per modulation symbols. For QPSK SOSTTC, four bits are used for every modulation symbol. The approximate complexity of the SOSTTC - OFDM system decoder using Viterbi algorithm can therefore be calculated by

$$\text{Comp}\{\text{SOSTTC}\} = \frac{2^{BPS} \times \text{No of states}}{BPS} \quad (14)$$

For the concatenated schemes, we applied the Log-MAP decoding algorithm for iterative decoding. Since the Log-MAP algorithm has to perform forward as well as backward recursion and soft output calculation, the number of trellis in Log-MAP decoding algorithm is assumed to be three times higher than that of a conventional Viterbi algorithm. For the rate 1/2 4-state convolutional code (CC) decoder, the complexity is approximated as

$$\text{Comp}\{\text{CC}(2,1,K)\} = 2 \times 2^{K-1} \times 3 \times \text{no of iterations}, \quad (15)$$

where K is the constraint length of the convolutional code. For the rate 2/3 4-state CC decoder, the complexity is estimated as

$$\text{Comp}\{\text{CC}(3,2,K)\} = 3 \times 2^{K-1} \times 3 \times \text{no of iterations}. \quad (16)$$

For SOSTTC with six iterative decoding, the complexity is estimated as

$$\text{Comp}\{\text{SOSTTCiter}\} = \frac{3 \times 6 \times 2^{BPS} \times \text{no of states}}{BPS}. \quad (17)$$

During each OFDM frame, 256 information bits are encoded to generate coded QPSK sequence which is OFDM modulated on 64 subcarriers. Considering the tail bits of the encoders and the guard interval of the OFDM modulation, the bandwidth efficiency η of the concatenated scheme is given by

$$\eta = 2 \times \frac{K}{K + \text{guard band}} \times \frac{\text{no of bits}}{\text{no of bit} + \text{tail bits}} \times \text{over all code rate}. \quad (18)$$

Table 2: Estimated decoder complexity and bandwidth of the proposed schemes.

S/N	SCHEME	Decoding Algorithm	CC	STC states	Complexity	Bandwidth Efficiency (b/s/Hz)	SNR at FER of 10^{-2}
1	SOSTTC-OFDM	VA		16 states	64	1.575	16.8 dB
2	CC-STTC-OFDM	Log-MAP	rate 2/3	16 states	792	1.04	13 dB
3	CC-CC-STTC-OFDM	Log-MAP	rate 2/3	16 states	1008	0.68	10 dB
4	CC-SOSTTC-OFDM	Log-MAP	rate 2/3	16 states	1368	1.04	11.6 dB
5	CC-CC-SOSTTC-OFDM	Log-MAP	rate 2/3	16 states	1584	0.68	9 dB

Applying equations (14) to (18) and considering six iterations for all the concatenated schemes, we summarize the estimated complexity and bandwidth efficiency for the proposed schemes in OFDM systems in Table 2. From Table 2, the CC-CC-SOSTTC-OFDM have the highest decoding complexity but with the best FER performance.

6. CONCLUSIONS

In this paper, investigation of concatenated SOSTTC over time varying frequency selective fading channel for MIMO-OFDM systems was carried out. In the investigation, two concatenated schemes were proposed for OFDM systems. We compared the error performance of the schemes with the use of QPSK STTC inner code combination. The system performance was observed to depend on the time selectivity of the channels, exhibiting high performance over slow varying channel and poor performance over fast varying channels. Results showed performance degradation when outer code with higher number of states and iterative decoding was used. The concatenated SOSTTC-OFDM systems have the advantage of achieving high diversity gain by exploiting available diversity resources of frequency selective fading channels and high coding gain was provided by serial concatenation coding scheme. The decoding complexity of the proposed schemes was also evaluated and compared with the reference systems.

7. ACKNOWLEDGMENT

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