Universal Space-Time Serially Concatenated Trellis Coded Modulations

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Abstract— **In this paper, we propose serially concatenated trellis coded modulations (SCTCMs) that perform consistently close to the available mutual information for the 2-by-2 compound matrix channel. The proposed SCTCMs use universal SCTCMs for the period-2 periodic fading channel in order to deliver consistent performance over eigenvalue skew. Within the family of channels having the same eigenvalue skew, a time-varying linear transformation (TVLT) is used to mitigate the performance variation over different eigenvectors. The proposed SCTCMs of 1, 2 and 3 bits per transmission require excess mutual information in the ranges 0.11-0.15, 0.23-0.26 and 0.35-0.53 bits per antenna, respectively. Because of their consistent performance over all channels, the proposed codes will have good frame-error-rate (FER) performance over any quasi-static fading distribution. In particular, the codes provide competitive FER performance in quasi-static Rayleigh fading.**

I. Introduction

Often, design of channel codes focuses on the optimization of performance on a specific channel such as the additive white Gaussian noise (AWGN) channel, or on average performance under a specific channel probability distribution, such as the Rayleigh fading channel. Numerous turbo TCMs [1][2][3][4] have been designed to optimize average performance for space-time Rayleigh fading channels. Typically, recursive space-time trellis codes that satisfy the slow-fading criteria are used as constituent codes. These codes perform well on the aimed channel or distribution. However, the performance can degrade significantly over some specific channel realizations.

Root and Varaiya's compound channel coding theory for linear Gaussian channels indicates that a single code can reliably transmit information at R bits/symbol on each channel in the ensemble of linear Gaussian channels with mutual information (MI) larger than the attempted rate [5]. In related work, Sutskover and Shamai [6] recently proposed decoding of Low-Density Parity-Check (LDPC) codes jointly with channel estimation for transmission over memoryless compound channels.

For the linear Gaussian vector channels (space-time channels)

$$\mathbf{y} = \mathbf{H}x + \mathbf{n}. \tag{1}$$

where $\mathbf{H}$ denotes the $N_r \times N_t$ channel matrix, $\mathbf{x}$ is the $N_t \times 1$ vector of transmitted symbols, $\mathbf{n} \sim \mathcal{N}(0, N_o I_{N_r})$ denotes the additive white Gaussian noise vector, and $\mathbf{y}$ is the $N_r \times 1$ vector of received symbols. N_t and N_r are the number of transmit antennas and receive antennas respectively in the Multiple-Input Multiple-Output (MIMO) system.

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In [7], [8], the term *universal* was used to describe a channel code that has a good bit-error-rate (BER) or FER for every channel in a family, $\mathcal{H}$, of channels. The code is said to be universal over $\mathcal{H}$. These papers presented trellis codes that approach universal behavior on space-time channels with a proximity similar to that with which trellis codes approach capacity on the AWGN channel.

In [9], [10], Zheng, Tse, and Viswanath examine the trade-off between diversity and multiplexing in MIMO systems. This trade-off turns out to be related to universal behavior. In [11], Tivildar and Viswanath give a precise characterization of approximately universal codes. These universal codes are also universally optimal in their trade-off between diversity and multiplexing. Tse and Viswanath provide an excellent overview of universal codes and their role in the diversity-multiplexing tradeoff in [12].

In this paper, we present SCTCMs for 1, 2, and 3 bits per transmission over the 2×2 space-time channel that provide approximately universal performance. Our technique de-multiplexes universal SCTCMs for periodic fading (as designed in [13], [14]) across the antennas. This structure can deliver consistent performance over a wide range of possible eigenvalue skews. However, nearly singular channels still display a variation due to the particular eigenvectors. The performance difference over eigenvectors can be largely mitigated, but not completely eliminated, by the time-varying linear transformation (TVLT) technique introduced in [15]. The current paper employs a generalized form of TVLT, which uniformly sweeps three phase parameters over $[0, 2\pi)$.

The rest of the paper is organized as follows. Section II defines the matrix channel and its parameters. Section III describes the structure of the space-time SCTCM system . Section IV presents the code design criteria and Section V analyzes the TVLT technique. Section VI presents simulation results of universal SCTCMs transmitting at 1, 2 and 3 bits per channel use and their performance on the quasi-static Rayleigh fading channel.

II. The Compound Linear Gaussian Channel

The mutual information between the input vector $\mathbf{x}$ and output vector $\mathbf{y}$ in Eq. (1) is

$$MI(\mathbf{H}, E_s) = \log_2 \det\left(I_{N_r} + \frac{E_s}{N_o}\mathbf{H}\mathbf{H}^\dagger\right) \tag{2}$$

$$= \sum_{i=1}^{n} \log_2\left(1 + \frac{E_s}{N_o}\lambda_i\right). \tag{3}$$

where E_s is the energy per symbol per transmit antenna, $\mathbf{H}^\dagger$ is the Hermitian matrix of $\mathbf{H}$ and $\lambda_1, \lambda_2, \cdots, \lambda_n$ are the eigenvalues of $\mathbf{H}\mathbf{H}^\dagger$. From this point on, we will assume that $N_t = N_r = N$.

For $N_t \neq N_r$, the same analysis applies with $N = \min(N_r, N_t)$ since the smallest $|N_r - N_t|$ eigenvalues of $\mathbf{H}\mathbf{H}^\dagger$ will be zero.

The search of universal SCTCM in this paper focuses on the 2×2 matrix channels. Assuming that $\lambda_1 \geq \lambda_2$ and defining the eigenvalue skew $\kappa = \frac{\lambda_2}{\lambda_1}$, then the MI is given by

$$MI\left(\mathbf{H}, E_s\right) = \log_2 \left(1 + \frac{E_s}{N_o}\lambda_1\right)\left(1 + \frac{E_s}{N_o}\kappa\lambda_1\right). \quad (4)$$

According to the compound channel coding theorem, if $MI \geq R$, the error probability of a universal code can decrease to zero in the limit as its blocklength goes to infinity. In practice, we design a finite-blocklength code transmitting at R-bits per channel use that achieves its target BER, say 10^{-5}, at $SNR^* = \frac{E_s^*}{N_o}$ for the channel $\mathbf{H}$. The asterisks (*) indicate that this is the SNR at which the target BER is achieved. As in [7], we define the excess mutual information (EMI) of this code for the channel $\mathbf{H}$ as the difference between $MI(\mathbf{H}, E_s^*)$ and R.

$$EMI(\mathbf{H}) = MI(\mathbf{H}, E_s^*) - R. \quad (5)$$

EMI($\mathbf{H}$) measures how closely the code operates within the theoretical limit. A universal code should deliver similar EMI($\mathbf{H}$) for all the possible 2×2 channel realizations. The 2×2 channel is parameterized as follows. First, by singular value decomposition (SVD), $\mathbf{H}$ can be written as

$$\mathbf{H} = U \begin{bmatrix} \sqrt{\lambda_1} & 0 \\ 0 & \sqrt{\lambda_2} \end{bmatrix} V^\dagger. \quad (6)$$

Note that any arbitrary unitary matrix, $V^\dagger$, can be written as

$$V^\dagger = e^{j\mu}\begin{bmatrix} 1 & 0 \\ 0 & e^{j\omega} \end{bmatrix}\begin{bmatrix} \cos\phi & \sin\phi \\ -\sin\phi & \cos\phi \end{bmatrix}\begin{bmatrix} 1 & 0 \\ 0 & e^{j\theta} \end{bmatrix}. \quad (7)$$

In the receiver, we can always multiply by a unitary matrix

$$\mathbf{P} = e^{-j\mu}\begin{bmatrix} 1 & 0 \\ 0 & e^{-j\omega} \end{bmatrix} U^\dagger \quad (8)$$

on the left of $\mathbf{H}$ such that the equivalent channel becomes

$$\begin{aligned}\mathbf{H}_{eq} &= \mathbf{PH} \\ &= \begin{bmatrix} \sqrt{\frac{1}{1+\kappa}} & 0 \\ 0 & \sqrt{\frac{\kappa}{1+\kappa}} \end{bmatrix}\begin{bmatrix} \cos\phi & \sin\phi \cdot e^{j\theta} \\ -\sin\phi & \cos\phi \cdot e^{j\theta} \end{bmatrix}. (9)\end{aligned}$$

Note that $\lambda_1 + \lambda_2$ is normalized to 1. We will use Eq. (9) as the channel model from this point on and evaluate the performance of our SCTCM on the matrix channel space by sampling over κ, ϕ and θ.

III. SCTCM SCHEME

The proposed space-time SCTCM scheme consists of a rate-R^o outer convolutional encoder, C^o, an interleaver, a rate-R^i inner convolutional encoder, C^i, and a two-dimensional 2^n-point constellation mapper. The symbols are de-multiplexed into the N symbol streams. Then the TVLT multiplies each N-symbol vector by a time-varying $N \times N$ unitary matrix. The signal symbols are then transmitted on the N transmit antennas. Therefore, the throughput of the overall scheme is nR^oR^iN bits per transmission. The decoder has perfect channel side information and the decoder also knows exactly the TVLT unitary matrices. Therefore, the log-likelihood (LLR) of the symbols can be calculated and passed to the turbo decoder. Note that the received symbols form an $N\times1$ vector and each received symbol is a superposition of N transmitted symbols. So, the inner Soft-Input Soft-Output (SISO) module is based on the *collapsed trellis* which combines N trellis stages together into a super-trellis.

IV. CODE DESIGN CRITERIA

In [13], [14], universal SCTCMs were designed for the period-2 fading channels with fading pattern $[1\ q]$ where q is a real number and $0 \leq q \leq 1$. Note that a 1-dimensional SCTCM over the scalar $[1\ q]$ period-2 fading channels is equivalent to the proposed de-multiplexed space-time SCTCM system over the 2×2 matrix channels in (9) when $\kappa = q^2$ and $\phi = \theta = 0$, i.e. it is diagonal. The performance of a diagonal channel is usually the best one among the set of channels with the same eigenvalue skew. The universal SCTCMs over the $[1\ q]$ channel are good starting point since they deliver universal performances over κ, but there is still the issue of non-diagonal channels.

Benedetto et al. [16] proposed design criteria for SCTCMs for AWGN channels using maximum-likelihood (ML) decoding which maximizes the free Hamming distance of the outer code and the effective free distance of the inner code. The actual performance of SCTCMs under iterative decoding depends on convergence properties not captured in the ML analysis. Codes with early convergence and an error floor below the target BER are desired. According to [13], [14], the following are the design criteria for universal SCTCMs over the $[1\ q]$ channel:

- Constituent code complexity: Start from low complexity convolutional codes to have earlier convergence. Increase the complexity if the error floor is too high.
- Outer code: Since the outer code does not interface with the channel, it has a smaller effect on universal performance. So the maximum free Hamming distance codes are used.
- Inner code: Must be systematic and recursive. A trellis collapse check narrows down the number of candidate codes. If the code has defects under certain channels, high error floors may occur. If trellis collapse induces a complexity reduction, earlier convergence results.
- Constellation labeling: Use a symmetric-ultracomposite Gray labeling [17] and map the systematic bits of the inner code bits in the constellation that are most protected.
- Interleaver: Use extended spread interleaver [18] with a further constraint that depends on the outer code and the periodic erasure patterns.

The best SCTCMs found for periodic fading at 0.5, 1.0 and 1.5 bits per symbol with BER error floors lower than 10^{-7} are listed in Table I. The same codes will be used in the corresponding 1, 2 and 3 bits per transmission space-time SCTCMs.

V. TIME-VARYING LINEAR TRANSFORMATION (TVLT)

For a fixed eigenvalue skew, the channel is parameterized using the channel angles ϕ and θ as shown in Eq. (9). The angle ϕ basically determines the amount of interference between the two antennas. When $|\cos(\phi)|$=1 or 0, the channel is almost diagonal which means there is no interference. On the other hand, the interference is the largest when $|\cos\phi| = |\sin\phi|$. The other parameter, θ, represents the phase difference between the constellations of the two antennas. Our experiments showed that

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TABLE I
SCTCMs PROPOSED IN [8] WITH 0.5, 1.0 AND 1.5 BITS PER SYMBOL.

SC	C^o	C^i	Constellation	BPS
2	$[1+D^2 \;\; 1+D+D^2]$	$[1 \;\; \frac{D^2}{1+D+D^2} \;\; 1+D]$	8PSK	0.5
5	$[1+D^2 \;\; 1+D+D^2]$	$\begin{bmatrix} 1 & 0 & \frac{D}{1+D^2} \\ 0 & 1 & \frac{1+D}{1+D^2} \end{bmatrix}$	8PSK	1.0
11	$[1+D^2 \;\; 1+D+D^2]$	$\begin{bmatrix} 1 & 0 & 0 & \frac{D}{1+D^2} \\ 0 & 1 & 0 & \frac{D}{1+D^2} \\ 0 & 0 & 1 & \frac{1+D}{1+D^2} \end{bmatrix}$	16QAM	1.5

although the MI is only a function of κ, different ϕ and θ can result in very different performances for SCTCM especially when κ is close to zero (the singular channel case).

The conventional technique to analyze the iterative decoding convergence of SCTCM is to use Extrinsic Information Transfer (EXIT) charts [19]. In EXIT chart analysis, the input and output LLRs of the SISOs are well approximated as symmetric Gaussian distribution if the channel is Gaussian. Then the characteristics of the inner and the outer SISO can be plotted conveniently and separately. However, for channels other than AWGN, the LLRs do not follow the Gaussian distribution. We can still plot the EXIT curves by tracking the true LLR densities passing between inner and outer SISO but then the computational complexity becomes the same as the numerical simulations.

According to previous experiments of SCTCM on the periodic fading channel [14], the higher the initial extrinsic information (IEI) of the inner SISO is, the more likely the whole inner-SISO EXIT curve stays above the EXIT curve of the outer SISO and thus lower convergence threshold. Note that the IEI can be easily calculated by numerically computing the parallel independent decoding (PID) capacity of the systematic bit of the inner code.

Figure 1 shows the EMI requirement and the corresponding IEI of SC-11 as a function of ϕ where $\theta = \kappa = 0$. It is clear the EMI requirement and the IEI are closely related. Based on the above observations, achieving a uniform IEI over ϕ and θ is way to enhance consistent EMI behavior and thereby approach universal performance. We tested three very different approaches to achieve a uniform IEI over ϕ and θ.

Our first idea was to design a combination of constellation and labeling that delivers a uniform IEI over ϕ and θ than the conventional schemes using PSK/QAM and Gray labeling. The second idea, motivated by [20], was to make the constellations for the two antennas correlated. Unfortunately, both methods which use a fixed constellation and labeling failed to yield a uniform IEI over ϕ and θ.

The third idea, which fortunately does work, uses a time-varying linear transformation (TVLT) to rotate the channel to different angles, hoping that the dependence on ϕ and θ can be "averaged out". This idea was first proposed in [15]. In the current paper, this concept is generalized so that a time-varying unitary matrix, Q_t, which has three parameters α, β and γ, is multiplied to the signal vector, $\mathbf{x}$, before transmitted and thus created a time-varying equivalent channel $\tilde{H} = HQ_t$. Let us use the following notation:

$$M(\psi) = \begin{bmatrix} \cos\psi & \sin\psi \\ -\sin\psi & \cos\psi \end{bmatrix}, D(\psi) = \begin{bmatrix} 1 & 0 \\ 0 & e^{-j\psi} \end{bmatrix}, \quad (10)$$

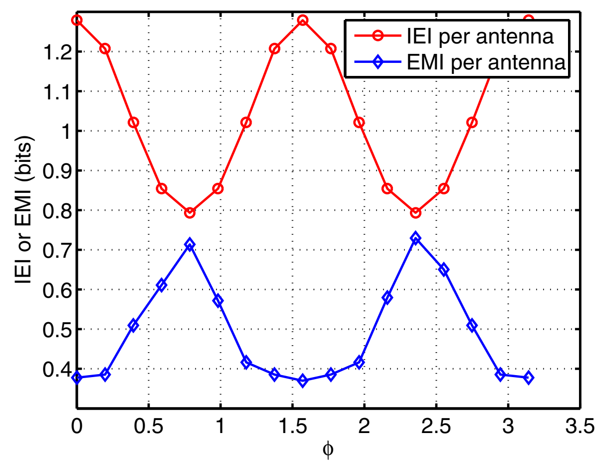

Fig. 1. The IEI and corresponding EMI per antenna. It is observed that IEI is highly correlated with EMI and thus can be used as a tool to facilitate the code search and to understand and approximate the TVLT performance.

$$\Lambda = \begin{bmatrix} \sqrt{\lambda_1} & 0 \\ 0 & \sqrt{\lambda_2} \end{bmatrix}. \quad (11)$$

Using (9) and decomposing Q_t as in (7), the resulting time-varying channel, $\tilde{\mathbf{H}}$, becomes

$$\begin{aligned} \tilde{\mathbf{H}} &= [\mathbf{H}][Q_t] \\ &= [\Lambda M(\phi)D(\theta)]\,[D(\beta)M(\alpha)D(\gamma)] \quad (12) \\ &= \Lambda\,[M(\phi)D(\theta+\beta)M(\alpha)]\,D(\gamma). \end{aligned}$$

Because $M(\phi)D(\theta+\beta)M(\alpha)$ is also unitary, it can be decomposed into $D(\tilde{\theta}_L)M(\tilde{\phi})D(\tilde{\theta}_R)$. Note that $D(\tilde{\theta}_L)$ can be canceled out in the receiver. Hence the equivalent channel

$$\tilde{\mathbf{H}}_{eq} = \Lambda M(\tilde{\phi})D(\tilde{\theta}), \quad (13)$$

where

$$\tilde{\phi} = \frac{1}{2}\cos^{-1}\left[\cos(2\phi)\cos(2\alpha) + \sin(2\phi)\sin(2\alpha)\cos(\theta+\beta)\right]. \quad (14)$$

$$\tilde{\theta} = \tilde{\theta}_R + \gamma. \quad (15)$$

The ideal TVLT would vary α, β, and γ such that $\tilde{\phi}$ and $\tilde{\theta}$ both sweep uniformly over $[0, 2\pi)$. In that case, the SCTCM sees the same set of channels regardless of what the actual channel angles ϕ and θ are. However, without any information of ϕ and θ in the transmitter, the best strategy is to sample uniformly over the 3-dimensional (α, β, γ) space.

For fixed α and β values, if γ sweeps uniformly over $[0, 2\pi)$, $\tilde{\theta}$ also varies uniformly. However, due to the nonlinearity in Eq. (14), $\tilde{\phi}$ is not swept uniformly and still depends on ϕ and θ. Therefore, the performance of the SCTCM cannot be completely universal using TVLT.

Here we provide an analysis of the approximate performance of an SCTCM with TVLT. From simulation results, we propose a model for EMI as a function of ϕ and θ as

$$EMI(\phi, \theta) \approx EMI_{min} + K[1 - \cos(4\phi)][3 + \cos(2\theta)] \quad (16)$$

To obtain the approximation of EMI with TVLT, we will assume that the average EMI is the average of the instantaneous EMIs. Further, we assume a TVLT in which α, β, and γ are uniformly distributed, the average EMI with such a TVLT is given by

$$\begin{aligned} & EMI^*(\phi, \theta) \\ \approx\ & EMI_{min} + K \cdot \frac{1}{N_\alpha N_\beta N_\gamma} \sum_\alpha \sum_\beta \sum_\gamma EMI(\tilde{\phi}, \tilde{\theta}) \\ =\ & EMI_{min} \\ +\ & K \cdot \frac{1}{N_\alpha N_\beta N_\gamma} \sum_\alpha \sum_\beta [1 - \cos(4\phi)] \sum_\gamma [3 + \cos(2\theta)] \\ =\ & EMI_{min} + \frac{3}{4} K \cdot [5 - \cos(4\phi)]. \qquad (17) \end{aligned}$$

The necessary conditions for (17) to hold are

$$\sum_\alpha e^{j4\alpha} = 0, \ \sum_\beta \cos(2\theta + 2\beta) = 0, \ \sum_\gamma \cos(2\gamma) = 0. \quad (18)$$

Note that the dependence on θ is eliminated but the EMI still depends on ϕ, although with a smaller variance. We propose two TVLT schemes

(1) Fine-sampled TVLT (F-TVLT): Sweep α, β and γ with a step size of $\frac{2\pi}{\sqrt[3]{N_s}}$, where N_s denotes the number of symbols per block.
(2) Simplified TVLT (S-TVLT): $\alpha = 0$ or $\pi/4$, $\beta = 0$ or $\pi/2$, and $\gamma = 0$ or $\pi/2$ yields a total of 8 different unitary matrices compared with N_s matrices in F-TVLT.

As shown in Figure 2, TVLT makes the EMI requirements more concentrated and the overall performance is close to universal. The S-TVLT performance, which is not shown in the plot, is almost identical to that of the F-TVLT.

VI. Simulation Results of space-time SCTCM

Figure 2 shows the EMI for SC-2, SC-5 and SC-11 with and without F-TVLT over the eigenvalue skew. For the same eigenvalue skew, the maximum and minimum EMIs are marked and the shaded region represents the region of operation. All the simulations are done with an input blocklength of 10,000 bits, 12 iterations at the decoder, and BER of 10^{-5}. The TVLT technique improves the universality of the SCTCMs. As a result, SC-2 provides a uniform EMI requirement of no more than 0.15 bits per antenna. SC-5 provides a uniform EMI requirement of no more than 0.26 bits per antenna. SC-11 provides a uniform EMI of no more than 0.41 bits per antenna except for the sudden increase to 0.53 bits for the worst-case singular channel.

Next, we will compare the proposed universal SCTCMs with other coding schemes designed specifically for Rayleigh fading environment in terms of both average performance and the universality (channel to channel performance). Among them, the best Rayleigh fading performance comes from Stefanov's [21] parallel concatenated convolutional code with bit-interleaved coded modulation (PCCC-BICM) which is 1.8 dB from the outage probability at FER=10^{-2}. Turbo-TCMs [3], [15], [22] are reported to operate at 2.0 to 2.2 dB from the outage probability. The universal SC-5 with F-TVLT is only 1.5 dB from the outage probability. For the Rayleigh fading performance, SC-5 was simulated with input blocklength of 260 bits, which is the same as that in [21], [3]. However, [15] and [22] use blocklength of 8,196 and 1,024 bits respectively. Moreover, the number of iterations in these papers are not exactly the same. So a fair conclusion would be that these codes all perform similarly on the Rayleigh fading channel.

As for the channel-to-channel performance, Figure 4 shows the EMI requirement region for SC-5, Benedetto's SCTCM [16] and the two PCCC-BICMs. SC-5 clearly has a much narrower EMI

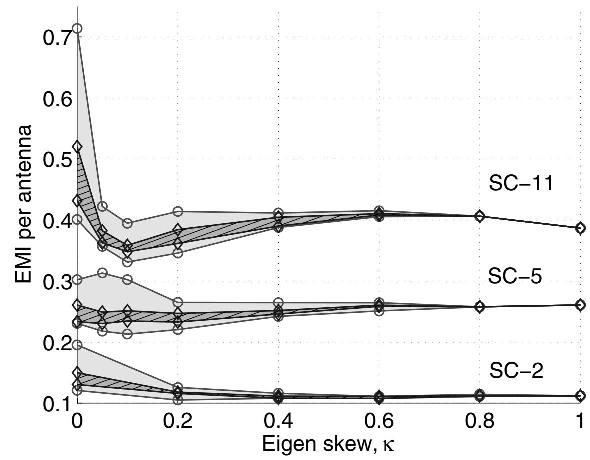

Fig. 2. EMI requirements of SC-2, SC-5 and SC-11 over eigenvalue skew and eigenvectors. The gray area is the EMI region without TVLT while the shaded area is the EMI region with F-TVLT. The proposed SCTCMs perform consistently close to channel capacity for any of the 2×2 channels.

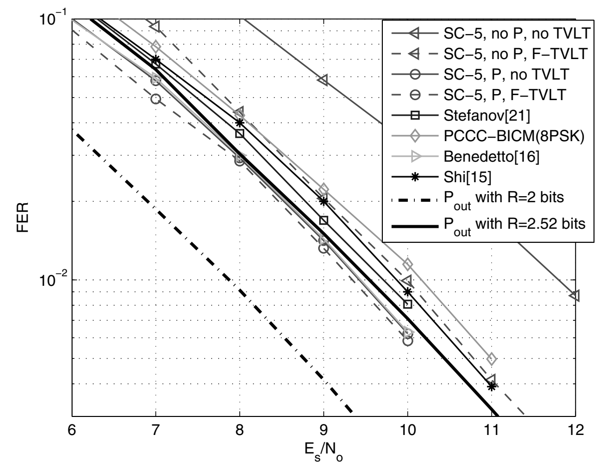

Fig. 3. FER comparison of 2 bits/s/Hz codes including the universal SCTCM (SC-5), an SCTCM (by Benedetto [16]), a PCTCM (by Shi [15]) and two PCCC-BICMs (by Stefanov [21]) over quasi-static Rayleigh fading channels.

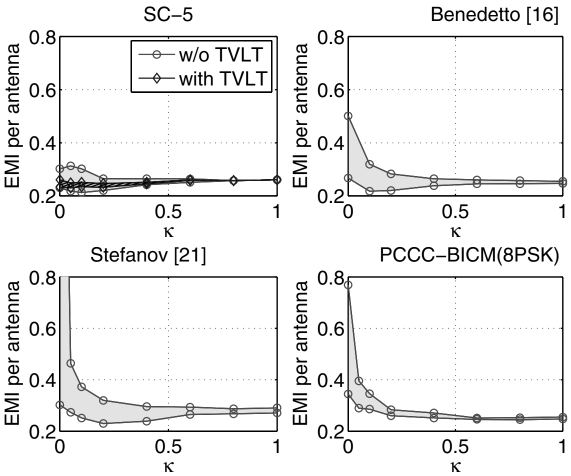

Fig. 4. EMI requirement region of 2 bits/s/Hz codes including the universal SCTCM (SC-5), an SCTCM (by Benedetto [16]), and two PCCC-BICMs (by Stefanov [21]) over the 2×2 matrix channel in Eq. (9).

region and hence is more universal than others, especially after applying TVLT. Note that the PCCC-BICM in [21] was originally rate-1/2 and mapping on QPSK constellation to transmit 2 bits/s/Hz over two antennas. The system, in the worst condition of $\kappa = 0, \phi = 0$, becomes uncoded and thus requires a very large EMI. Therefore, we also designed a rate-1/3 PCCC-BICM mapping an 8-PSK constellation. The worst case EMI is reduced to 0.8 bits per antenna but it is still not as universal as SC-5.

In Sec. IV-C in [15], Shi found that by multiplying $\mathbf{P}$ in Eq. (8), i.e. the $\mathbf{R}$ matrix in Shi's notation, at the receiver, the FER improves significantly. Shi also found that the effect of TVLT is invisible with or without $\mathbf{P}$. We also found the $\mathbf{P}$ matrix crucial for the SCTCM performance on the Rayleigh fading channel. However, unlike [15] who uses only real unitary matrices in their experiments, our generalized TVLT matrices scan the whole complex unitary matrix space and provides a better averaging effect. As a result, for SC-5, TVLT improves the FER by 1.8 dB at FER=10^{-2} in the case of no $\mathbf{P}$ and merely 0.1 dB when $\mathbf{P}$ is applied.

Since SC-5 requires a uniform EMI no more than 0.26 bits per antenna, SC-5 should be able to transmit 2 bits successfully over any channel which provides an MI of more than 2.52 bits. Figure 3 shows that the FER of SC-5 with F-TVLT is close to the outage probability of MI=2.52 bits. The beauty of the universal SCTCM is that SC-5 can perform close to the outage probability of a MI=2.52 bits not only for Rayleigh fading but *any* quasi-static fading channels.

VII. Conclusions

Designs of universal space-time SCTCM schemes are proposed to deliver close-to-capacity performance on any of the quasi-static 2×2 channels. First we design universal SCTCMs over periodic fading which is equivalent to diagonal MIMO channels and show their universality over eigenvalue skew. Then for non-diagonal channels, it is found that the performance of SCTCM may degrade significantly due to the convergence behavior of the iterative decoder. We observed that the relation between EMI and IEI can be used to facilitate the comparison of different constellations and labelings. Experiments with fixed constellations and labelings could not achieve universal IEI. Hence, TVLT is used to rotate the channel matrix over time and is shown to effectively mitigate the dependency of EMI on the eigenvectors. As a result, the proposed SCTCMs of 1, 2 and 3 bits per channel use require a consistent EMI of 0.11-0.15, 0.23-0.26 and 0.35-0.53 bits per antenna and are 1.1, 1.5, and 2.1 dB respectively from the outage probability at FER=10^{-2} on the quasi-static Rayleigh fading channels. However, the main point of designing such codes is that they are not designed specifically for Rayleigh fading, and will work close to the theoretical limit on any quasi-static fading channel.

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