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On Bandwidth Efficient Modulation for High-Data-Rate Wireless LAN Systems

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We address the problem of high-data-rate orthogonal frequency division multiplexed (OFDM) systems under restrictive bandwidth constraints. Based on recent theoretic results, multiple-input multiple-output (MIMO) configurations are best suited for this problem. In this paper, we examine several MIMO configurations suitable for high rate transmission. In all scenarios considered, perfect channel state information (CSI) is assumed at the receiver. In constrast, availability of CSI at the transmitter is addressed separately. We show that powerful space-time codes can be developed by combining some simple well-known techniques. In fact, we show that for certain configurations, these space-time MIMO configurations are near optimum in terms of outage capacity as compared to previously published codes. Performance evaluation of these techniques is demonstrated within the IEEE 802.11a framework via Monte Carlo simulations.

Keywords and phrases: OFDM, WLAN, MIMO, antenna diversity, space-time block codes, TCM.

1. INTRODUCTION

Currently, the IEEE 802.11a standard offers data rates ranging from 6 Mbit/s to 54 Mbit/s. However, there is a growing interest for a 100 Mbit/s mode of operation for the IEEE standard. Unfortunately, to achieve such a rate within the IEEE 802.11a framework requires the system to operate at a spectral efficiency better than 6 bit/s/Hz. The spectral efficiency problem is further complicated by the fact that 20% of the available bandwidth is used for the cyclic prefix to mitigate the effects of frequency selective fading. From a system design perspective, the complexity associated with this mode of operation should not be much greater than that for the 54 Mbit/s mode of the standard. These stringent requirements constitute a very interesting research problem.

Recent information theoretic results [1] suggest that there is a tremendous capacity potential for wireless communication systems using *antenna diversity*. Foschini and Gans [1] and others [2, 3] noted that orthogonal frequency division multiplexed (OFDM) systems are particularly well suited for antenna diversity techniques. Hence, it is expected that multiple element antenna arrays will play an increasingly important role in emerging wireless LAN networks. Indeed, when used in conjunction with appropriately designed signal processing algorithms, these arrays can dramatically enhance performance.

In systems where channel state information (CSI) is not known to the transmitter, space-time coding (STC) is a bandwidth and power efficient solution for communication over wireless Rayleigh or Rician fading channels. STC guarantees transmit diversity and, optionally, receive diversity. Furthermore, the code construction is done such that the diversity advantage is achieved without any sacrifice in the transmission rate. In [4], Tarokh constructed space-time trellis codes (STTC) using design criteria derived for the Rayleigh fading channels, where ideal channel state information is available at the receiver. It was shown that for a quasi-static Rayleigh or Rician channels, performance is determined by the diversity advantage quantified by the rank of certain matrices and by the coding advantage, that is, quantified by the determinants of these matrices.

In contrast, when the temporal and spatial channel gain is available in transmitter, there is no need to use space-time codes. Because of the availability of channel gains in both transmitter and receiver, spatial processing can be performed at both transmitter and receiver to change the statistic of multiple-input multiple-output (MIMO) channel at the receiver into a parallel bank of single-input single-output (SISO) channels [2]. When the channel has severe nulls in its frequency response or there is a powerful narrowband noise, the signal frequency components are completely canceled out in those frequencies. It is obvious that transmitting energy in these frequencies is a waste of power. Therefore, to approach the capacity of these channels, a kind of spectral shaping should be applied to the transmitted signal. Correspondingly, new coding schemes, which are adapted to the frequency response of the channel, are required. In other words, the power and rate should be optimally distributed over frequency components of the transmitted signal by an algorithm called *water-filling*. It has long been known that multicarrier modulation [5] could in principle be used to achieve the power and rate allocations prescribed by water-filling. However, practical realizations of multicarrier modulation in combination with powerful codes have been achieved only in recent years. Goldsmith in [6, 7] was one of the first researchers to develop powerful codes for the wireless channel.

The remainder of this paper is organized as follows. Section 2 describes the basic channel model used for MIMO systems, which is the basis for later development in the paper. Section 3 is dedicated to the design criteria for STC. It briefly summarizes the popular work by Tarokh in [4], but also includes the work of Ionescu [8], who developed a new criterion for the coding gain for STC. Section 4 leverages heavily on the work presented in Section 3 to derive simple construction techniques for near-optimum STC. Section 5 reviews the previous work in adaptive modulation and introduces novel extensions to this work. Finally, we summarize the work presented and salient points for future research.

2. CHANNEL MODEL FOR MIMO SYSTEMS

We provide sufficient details to generally characterize the channel for any multiple-input multiple-output (MIMO) systems employing an OFDM modulation. The final form for the channel model is denoted in matrix/vector for notational convenience. Now, we consider a communication link comprising N transmitter antennas and M receiver antennas that operates in an OFDM MIMO channel. Each receiver antenna responds to each transmitter antenna through a statistically independent fading path. The received signals are corrupted by additive noise, that is, statistically independent among the M receiver antennas and the transmission periods. For ease of presentation, discrete baseband notation is used, that is, at sample time index l, the complex symbols, $s_i(l)$, sent by the N transmit antennas and, subsequently, detected by the kth receive antenna, is denoted by $y_k(l)$. Then $y_k(l)$ can be expressed as

$$y_k(l) = \sqrt{\frac{P}{N}} \sum_{i=1}^{N} h_{ki}(l) \star s_i(l) + v_k(l), \quad 1 \le k \le M, \tag{1}$$

where $h_{ik}(l)$ is the complex scalar associated with path between the ith transmitter antenna and the kth receive antenna at time index l. The noise samples at time index i, $v_k(l)$, are complex zero-mean spatially and temporally white Gaussian random variable with variance $N_0/2$ per dimension. It is further assumed that the transmitted signals $s_i(l)$ are normalized such that the average energy for the signal constellation is unity. Recall that, for OFDM signal, the channel is made circulant by prepending a cyclic prefix (CP) to the data sequence prior to transmission. This simple mechanism allows us to replace the linear convolution in (1) with a circular one. Hence, the frequency selective channel given in (1) is transformed into a parallel bank of flat fading channels via an L-point FFT, that is,

$$Y_k(m) = \sqrt{\frac{P}{N}} \sum_{i=1}^{N} H_{ki}(m) S_i(m) + V_k(m), \quad \text{for } 1 \le m \le L, \tag{2}$$

where $Y_k(m), H_{ki}(m), S_i(m)$, and $V_k(m)$ denote the frequency domain representations of the mth subcarrier of the received signal for the kth antenna, complex channel gains between the ith transmitter antenna and the kth receive antenna, transmitted signal for the ith antenna, and noise signal for the kth receive antenna, respectively. If the channel gains are slowly fading, then it can be assumed that, during a period of T time indexes, the gains are constant and we can approximate the channel as a block fading channel. Thus, the received signal vector for the mth subcarrier using matrix notation is given by

$$\mathbf{Y}(m) = \sqrt{\frac{P}{N}} \mathbf{H}(m) \mathbf{S}(m) + \mathbf{V}(m), \tag{3}$$

where

$$\mathbf{Y}(m) = \left[Y_1(m), Y_2(m), \ldots, Y_M(m)\right]^T,$$
$$\mathbf{H}(m) = \begin{bmatrix} \mathbf{H}_1(m) \\ \mathbf{H}_2(m) \\ \vdots \\ \mathbf{H}_M(m) \end{bmatrix}, \tag{4}$$
$$\mathbf{S}(m) = \left[S_1(m), S_2(m), \ldots, S_N(m)\right]^T.$$

For ease of presentation, the MIMO analysis herein will be developed for the mth subcarrier, knowing that analysis applies equally to the remaining subcarriers. Hence, the index for the subcarriers will be dropped here for notational simplicity and re-introduced when channel coding across subcarriers are discussed later in this paper.

3. REVIEW OF STC DESIGN CRITERIA

We briefly review STC design criteria for quasi-static channels. In general, a wireless communication system is comprised of N transmit antennas and M receive antennas. The information data is encoded using a channel code to protect it against imperfections caused by the channel. The encoded data stream is split into N parallel streams each of which is modulated and then transmitted using separate antennas. Each path for the separate antennas is assumed *i.i.d.* and quasi-static, that is, the complex gains of the paths are constant over each data frame but change from frame to frame.

To develop codes that perform well over fading channels, we choose to minimize the pairwise error probability. That is, the probability that the code word $\mathbf{c}$ is transmitted over the channel and a maximum-likelihood (ML) receiver decides in favor of a different code word $\mathbf{e}$. The ML receiver makes decoding decisions based on a performance metric $m(\mathbf{Y}, \mathbf{c}, \mathbf{H})$ provided that estimates of the fading amplitudes, $\mathbf{H}$, are available at the receiver. Formally, the maximum likelihood criterion for the optimum decoder requires the conditional probability of receiving $\mathbf{Y}$ given that the code word $\mathbf{c}$ was transmitted to be greater than the probability of receiving $\mathbf{Y}$ assuming any other code word $\mathbf{e}$ was transmitted, that is,

$$\Pr(\mathbf{c} \longrightarrow \mathbf{e} \mid \mathbf{H}) = \Pr\left[m(\mathbf{Y}, \mathbf{e}; \mathbf{H}) \geq m(\mathbf{Y}, \mathbf{c}; \mathbf{H}) \mid \mathbf{H}\right], \quad (5)$$

where

$$m(Y_k, e_k; H_k) = -\left| Y_k - \mathbf{H}_k \mathbf{e}_k \right|^2; \quad (6)$$

Y_k is again the received signal for the kth antenna; $\mathbf{H}_k$ and $\mathbf{e}_k$ are the complex path gains and transmitted symbols, respectively, from the N transmit antennas to the kth received antenna, and $|\cdot|^2$ represents the squared Euclidean norm. The pairwise error probability is found by taking the statistical expectation of (5). Rather than solving for the exact pairwise error probability, which can only be evaluated numerically, an upper bound can be found for (5) using the Chernoff bound techniques. Evaluation of the Chernoff bound for Rician and Rayleigh channels leads to the following two design criteria.

First, we define the quantity referred to the *code word difference matrix* $\mathbf{D}(\mathbf{c}, \mathbf{e})$ defined as

$$\mathbf{D}(\mathbf{c}, \mathbf{e}) = \begin{bmatrix} c_1(0) - e_1(0) & \cdots & c_1(l) - e_1(l) \\ \vdots & \ddots & \vdots \\ c_N(0) - e_N(0) & \cdots & c_N(l) - e_N(l) \end{bmatrix}^T, \quad (7)$$

where l represents the length of an error event path. In (7), the columns of $\mathbf{D}(\mathbf{c}, \mathbf{e})$ index the transmit antennas and the rows index the symbol epochs[1] of the codes $\mathbf{c}$ and $\mathbf{e}$. The rank of $\mathbf{D}(\mathbf{c}, \mathbf{e})$ determines the diversity advantage of the code. That is, in order to achieve a diversity of pM in a rapid fading environment for any two code words $\mathbf{c}$ and $\mathbf{e}$ the strings $c_1(l)c_2(l)\cdots c_N(l)$ and $e_1(l)e_2(l)\cdots e_N(l)$ must be different at least for p values of $1 \leq l \leq N$. The coding advantage d_P^2 is determined from the geometric mean of the eigenvalues λ_i of the matrix of $\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger \mathbf{D}(e, c)$, that is,

$$d_P^2(r) = \left(\prod_{i=1}^{r} \lambda_i\right)^{1/r}. \quad (8)$$

Typically, the minimum coding advantage $d_P^2(r)$ amongst all code word pairs is the dominate factor in the performance of STC. Further, the minimum Euclidean distance $d_{\min}$ between any two code words $\mathbf{c}$ and $\mathbf{e}$ determines the minimum coding advantage. Hence, codes with larger $d_{\min}$ have better coding advantages.

3.1. Improvement of design criterion

Ionescu [8] demonstrated that the determinant criterion can be strengthened by requiring the eigenvalues of $\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger \mathbf{D}(\mathbf{e}, \mathbf{c})$ to be as close as possible, for any pair of code words $\mathbf{c}$, $\mathbf{e}$. Formally, the criterion was expressed as follows.

Theorem 1 (new determinant criterion). *The equal eigenvalue criterion: for N-transmit antenna system operating in i.i.d. Rayleigh fading with perfect channel state information (CSI), an upper bound to the pairwise error probability is made as small as possible if and only if, for all pair of code words* $\mathbf{c}$, $\mathbf{e}$, *the squared Euclidean distance* $\mathrm{tr}[\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger \mathbf{D}(\mathbf{e}, \mathbf{c})]$ *is made as large as possible. Further, the nonsquare matrices* $\mathbf{D}(\mathbf{e}, \mathbf{c})$ *are semiunitary—up to a scale factor—, that is,* $\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger \mathbf{D}(\mathbf{e}, \mathbf{c}) = (\mathrm{tr}[\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger \mathbf{D}(\mathbf{e}, \mathbf{c})]/N) * I_N$. *Essentially, maximizing* $\min_{\mathbf{c},\mathbf{e}} \det[\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger \mathbf{D}(\mathbf{e}, \mathbf{c})]$, *as specified by the determinant criterion, requires maximizing the minimum eigenvalue product over all* $\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger \mathbf{D}(\mathbf{e}, \mathbf{c})$.

Proof. By Hadamard's theorem, the eigenvalue product for a square, positive definite matrix $\mathbf{A}$, with elements $a_{i,j}$ assumes its maximum value $\Pi_i a_{i,i}$, *if and only if* $\mathbf{A}$ is diagonal. Once $\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger \mathbf{D}(\mathbf{e}, \mathbf{c})$ is diagonalized, the product of its diagonal elements is maximized *if and only if* they are rendered equal and their sum, $\mathrm{tr}[\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger \mathbf{D}(\mathbf{e}, \mathbf{c})]$, is maximized. Consider the arithmetic-mean geometric mean inequality, that is,

$$\sqrt[n]{\prod_{i=1}^{N} \lambda_i} \leq \left(\frac{1}{N} \sum_{i=1}^{N} \lambda_i\right). \quad (9)$$

Hence, the product distance $d_P^2(r)$ given in (8) is upper bounded by the arithmetic mean of eigenvalues of $\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger \mathbf{D}(\mathbf{e}, \mathbf{c})$ with equality achieved when all eigenvalues are equal. □

Ionescu recognized that it might be difficult to enforce this condition for all pairs of code words $\mathbf{c}$, $\mathbf{e}$ thus proposed

[1] For this case, the symbol epochs corresponds to subcarriers of the OFDM symbol.

a suboptimal solution to enforce the condition of the code words corresponding to the shortest error event paths in the code trellis. In the next section, we demonstrate a simple method for constructing near optimal space-time trellis code (STTC) using the equal eigenvalue criterion.

4. SIMPLE STTC CONSTRUCTION FOR EQUAL EIGENVALUE DESIGNS

Here, we postulate that concatenation of an orthogonal space-time block code (STBC) with bandwidth-efficient codes designed for additive white Gaussian noise (AWGN) channels yield near optimum codes in terms of coding gain. It is well known that the STBC [9, 10] are spectrally less efficient than STTC. To compensate for the poor spectral efficiency of STBC, the input symbols $s_1, s_2, \dots, s_k$ of the STBC can be generated from a spectrally efficient modulation such as trellis-coded modulation (TCM) or block-coded modulation (BCM). Note that the use of an orthogonal STBC guarantees satisfaction of the equal eigenvalue criterion by Ionescu [8]. Furthermore, the symbols $s_1, s_2, \dots, s_k$ should be chosen to maximize the Euclidean distance for all pair of code words $\mathbf{c}, \mathbf{e}$. Since STTC assumes flat fading channels in its development, consider space-time block coding of an OFDM system using the Radon-Hurwitz (R-H) unitary transform, popularized by Alamouti [11], defined over OFDM symbols as

$$\text{R-H}\begin{pmatrix}\mathbf{X}_o\\ \mathbf{X}_e\end{pmatrix} = \begin{bmatrix}\mathbf{X}_o & \mathbf{X}_e\\ -\mathbf{X}_e^* & \mathbf{X}_o^*\end{bmatrix}, \tag{10}$$

where the rows in (10) index transmit antennas and the columns index symbol epochs. If two consecutive OFDM symbols are referred to as $\mathbf{X}_o$ and $\mathbf{X}_e$, then at the first antenna, $\mathbf{X}_e$ is transmitted in the first time epoch followed by $\mathbf{X}_o$ in the second time epoch while, at the second transmitter, $\mathbf{X}_o^*$ is transmitted in the first time epoch followed by $-\mathbf{X}_e^*$ in the second time epoch. The appeal of the R-H transform and other orthogonal transmit diversity schemes is because they allow the individual symbols at the receiver to be separated. Denote the diagonal matrices containing the channel frequency response vectors, $\mathbf{H}_1$ and $\mathbf{H}_2$, for a two transmitter and one receiver configuration by $\mathbf{\Lambda}_1$ and $\mathbf{\Lambda}_2$, respectively. Assuming the channel is constant over the two consecutive symbol epochs, the baseband received signals $(\mathbf{Y}_1, \mathbf{Y}_2)$ in the respective symbol periods are given by

$$\begin{aligned}\mathbf{Y}_1 &= \mathbf{\Lambda}_1\mathbf{X}_o + \mathbf{\Lambda}_2\mathbf{X}_e + \mathbf{V}_1,\\ \mathbf{Y}_2 &= -\mathbf{\Lambda}_1\mathbf{X}_e^* + \mathbf{\Lambda}_2\mathbf{X}_o^* + \mathbf{V}_2.\end{aligned} \tag{11}$$

Using simple substitution methods, the noise corrupted estimates of the consecutive OFDM symbols are

$$\begin{aligned}\hat{\mathbf{X}}_o &= \mathbf{\Lambda}_1^*\mathbf{Y}_1 + \mathbf{\Lambda}_2\mathbf{Y}_2^*,\\ \hat{\mathbf{X}}_e &= -\mathbf{\Lambda}_1\mathbf{Y}_2^* + \mathbf{\Lambda}_2^*\mathbf{Y}_1.\end{aligned} \tag{12}$$

Substituting (11) into (12) yields

$$\begin{aligned}\hat{\mathbf{X}}_o &= \left(|\mathbf{\Lambda}_1|^2 + |\mathbf{\Lambda}_2|^2\right)\mathbf{X}_o,\\ \hat{\mathbf{X}}_e &= \left(|\mathbf{\Lambda}_1|^2 + |\mathbf{\Lambda}_2|^2\right)\mathbf{X}_e.\end{aligned} \tag{13}$$

The question that remains is what values of $\mathbf{X}_o$ and $\mathbf{X}_e$ maximize the coding advantage for the system. Recall that, it was noted that the performances of STC are dominated by minimum distance for the code. Also, recall that for any coset code, the minimum distance for the code [12, page 113] is determined by

$$d_{\min} = \min\left(d_{\text{free}}, d_{\text{coset}}\right), \tag{14}$$

where d_{free} is the free distance of the code used to select a coset and d_{coset} is the minimum distance within a coset. Now consider the code difference matrix $\mathbf{D}(\mathbf{e}, \mathbf{c})$ of our concatenated code for the mth subcarrier of an OFDM system with two transmit antennas,

$$\mathbf{D}(\mathbf{e}, \mathbf{c}) = \begin{bmatrix} -\left[c_1(0) - e_1(0)\right]^* & \left[c_2(0) - e_2(0)\right]^*\\ c_2(0) - e_2(0) & c_1(0) - e_1(0)\\ \vdots & \vdots\\ -\left[c_1(l) - e_1(l)\right]^* & \left[c_2(l) - e_2(l)\right]^*\\ c_2(l) - e_2(l) & c_1(l) - e_1(l)\end{bmatrix} \tag{15}$$

which leads to

$$\mathbf{D}(\mathbf{e}, \mathbf{c})^\dagger\mathbf{D}(\mathbf{e}, \mathbf{c}) = \begin{bmatrix} d_{\min}^2 & 0\\ 0 & d_{\min}^2\end{bmatrix}. \tag{16}$$

Clearly, (16) satisfies the equal eigenvalue criterion and thus its coding gain is guaranteed to achieve its upper bound. Hence near-optimum performance is achievable provided that the eigenvalues are made as large as possible. Other researchers [13, 14] have arrived to this same conclusion based on different arguments. The main restriction on the code words $\mathbf{c}, \mathbf{e}$ is that the channel must be static over the two symbol epochs defining the Radon-Hurwitz transform. Hence, we can construct a space-time block coset code (STBCC) by simply selecting a well-known coset code with the desired Euclidean distance, d_{free}, and complexity. One limitation to these codes is that there is no full rate complex orthogonal STBC for more than two antennas [9, 10].

We construct a few simple STBCCs using Ungerboeck codes and compare them to some previously published STTCs [4] of similar complexity.

Example 1. Consider the fully connected 2-STC 4-PSK, 4 states, 2 bit/s/Hz in Figure 1 from [4]. Since the code is fully connected, we can reach any state from any other state. Therefore, its outer product of the code word difference

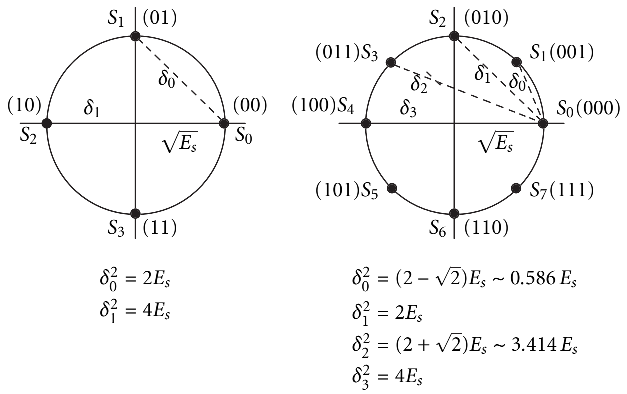

Figure 1: Distance between constellation points for 4-PSK and 8-PSK.

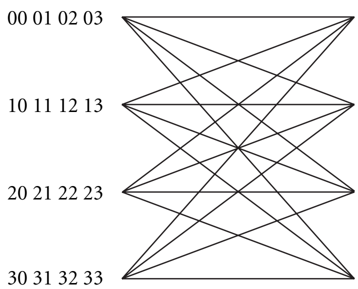

Figure 2: Tarokh et al. [4]: a fully connected 4-state STC using quadrature phase shift keying (QPSK).

matrix is given by

$$\mathbf{D}(\mathbf{e},\mathbf{c})^{\dagger}\mathbf{D}(\mathbf{e},\mathbf{c}) = \begin{bmatrix} \delta_0^2 & 0 \\ 0 & \delta_0^2 \end{bmatrix}, \tag{17}$$

where $\delta_0^2 = 2$ for 4-PSK as shown in Figure 2. Now consider a 2 bit/s/Hz Ungerboeck code with 4 state, defined by the octal generator polynomials $g_1 = 2$ and $g_2 = 5$, in [12, page 120], with $\delta_3^2 = 4$ for 8-PSK. Hence, the coding advantage using this simple construction is 3 dB better than the 4-state STC of the same rate.

Example 2. Now consider the 2-space-time code, 4-PSK, 8 states, 2 bit/s/Hz shown in Figure 3. The code word different matrix is equal to

$$\mathbf{D}(\mathbf{e},\mathbf{c}) = \begin{bmatrix} \delta_1^2 & 0 \\ 0 & \delta_1^2 \end{bmatrix}, \tag{18}$$

where $\delta_1^2 = 4$ for 4-PSK. Now consider a 2 bit/s/Hz with 8 states, with octal generator polynomials $g_1 = 04$, $g_2 = 04$, and $g_3 = 11$. The minimum free distance for this code is $d_{\text{free}}^2 = 4.586$, which is 1.15 dB better than the equivalent STC code by Tarokh [4].

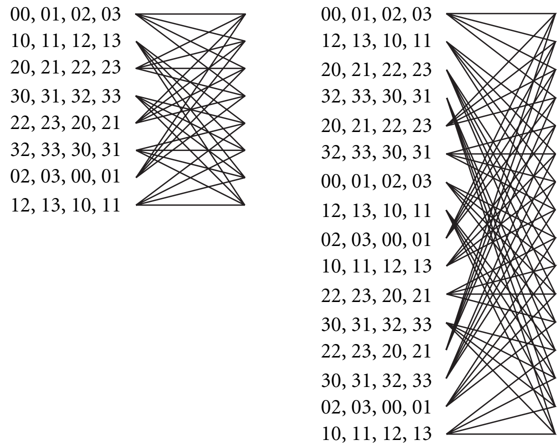

Figure 3: Tarokh et al. [4]: trellises for 8 states and 16 states with diversity order 2.

Example 3. The final example of this section considers a 2-space-time code, with 16 states, 2 bit/s/Hz. The trellis for this code is given in Figure 3. The code word difference matrix for this code is

$$\mathbf{D}(\mathbf{e},\mathbf{c}) = \begin{bmatrix} 6 & 2\sqrt{2} \\ 2\sqrt{2} & 6 \end{bmatrix} \tag{19}$$

with eigenvalues of $\mathbf{D}(\mathbf{e},\mathbf{c})^{\dagger}\mathbf{D}(\mathbf{e},\mathbf{c})$ equal to 3.17 and 8.83. The geometric mean of the eigenvalues equals 5.29. An Ungerboeck code—$g_1 = 16$, $g_2 = 04$, $g_3 = 23$ octal generator polynomials—with 16 states and rate of 2 bit/s/Hz has a minimum distance of 5.17, which is slightly less than the STTC. However, the performance of this code should be better than the STTC since half of the data is transmitted over a channel whose eigenvalue is 3.17 and it should dominant the performance for the overall code while, for our construction, each path to the receiver has eigenvalues of 5.17 associated with it.

Perhaps, the most attractive property of these STBCCs is their simpler decoder complexity compared to their STTC counterparts. When multiple receive antennas are used, maximal ratio combining of each estimate of the transmitted symbols in (12) can be performed for all the receive antennas.

4.1. Improving diversity gain

In Section 3, the design criterion is optimum in terms of Euclidean distance. However, for fast fading channels, the Hamming distance rather than the Euclidean distance is the appropriate design metric. Recall (see [15]) that the design of a TCM scheme that achieves diversity of order p requires all pairs of possible encoded sequences to differ in at least p symbols. A simple method of improving the Hamming distance of a coset scheme would be to use a signal constellation whose symbols differ in every coordinate yet maintain the desired minimum Euclidean distance d_{free} amongst all signal points. The authors in [16] show that rotation of well-known constellations, rather than designing such a constellation, can achieve the same purpose. Once each symbol has distinct coordinates over the whole signal constellation or within a subconstellation, then the bound on the pairwise error probability is computed over coordinates, that is,

$$P(\mathbf{s} \longrightarrow \hat{\mathbf{s}}) \leq \frac{1}{(\text{SNR}_c/4)^{p_c}} \Pi_{k=1}^{p_c} \frac{1}{|s_k - \hat{s}_k|^2}, \quad (20)$$

where p_c is the number of distinct coordinates between pairs of code sequences, and $\text{SNR}_c = \text{SNR}/2$ for 2D constellations (SNR stands for signal-to-noise ratio). In other words, if each coordinate experiences independent fading, then we anticipate a *two-fold* improvement in diversity level at moderate-to-high SNR. To ensure independent fading, we interleave over coordinates rather than symbols.

We revisit Example 1 with respect to Hamming distance. For Example 1, the symbol Hamming distance for the STC and STBCC are both two. However, if this coordinate interleaving technique was employed, then the coordinate Hamming distance for the STBCC would increase to 4. Similar arguments can be made for Examples 2 and 3. The remaining sections in this paper consider cases when CSI is available at the transmitter.

5. POWER OPTIMIZATION AND BIT ALLOCATION ALGORITHM

In this section, power optimization and bit allocation algorithm for MIMO systems are considered. Information theory determines the theoretical maximum informational data rate through parallel channels via a water-filling power allocation scheme. This theoretic solution relies on a channel-coding scheme that achieves Shannon's channel capacity for each subcarrier. For the past fifty years, designing, coding, and modulation schemes with practical implementation complexity that achieve this theoretic upper bound are still an open research topic. In Section 6, we outline the problem.

Let the singular value decomposition of $\mathbf{H}$ be $\mathbf{H} = \mathbf{U}\Sigma\mathbf{V}^H$, where $\mathbf{U}$ and $\mathbf{V}$ are unitary matrices and $\mathbf{\Sigma}$ is a diagonal matrix with positive real values on diagonal elements standing for the singular values of the channel. If the transmitted vector is premultiplied by $\mathbf{V}$ in transmitter, and received vector be postmultiplied by $\mathbf{U}^H$ in receiver, then (3) could be rewritten as

$$\tilde{\mathbf{y}} = \mathbf{\Sigma}\mathbf{s} + \tilde{\boldsymbol{\eta}}, \quad (21)$$

where $\tilde{\mathbf{y}}$ and $\tilde{\boldsymbol{\eta}}$ denote the received vector and noise vector after postmultiplication by $\mathbf{U}$. Note that because $\mathbf{U}$ is a unitary matrix, there is no noise amplification and the noise vectors remain spatially white. Thus, the error rate of maximum likelihood decoder remains the same as it was for decoding (3), but its complexity is completely reduced because the entries of $\mathbf{s}$ are decoded separately. If r denotes the rank of $\mathbf{\Sigma}$, then this MIMO channel is a set of parallel SISO channels,

$$\tilde{y}_k = \sigma_k s_k + \tilde{\eta}_k, \quad k = 1, \ldots, r, \quad (22)$$

where σ_k is the gain of kth channel and $\tilde{\eta}_k$ is white Gaussian noise. For an OFDM framework, (22) becomes

$$\tilde{Y}_k(m) = \sigma_k S_k(m) + \tilde{V}_k(m), \quad k = 1, \ldots, r; \; m = 1, \ldots, L. \quad (23)$$

The total capacity of the MIMO channel is equal to the aggregate capacity of all these SISO channels. If spatial water-filling is applied for this MIMO channel, then the capacity of this MIMO channel is the sum of the individual capacities per subcarrier.

For practical implementation, designers must choose a baseband modulation scheme such as MPSK or QAM for bit assignments. Thus, the symbol error rate (SER) is bounded strictly away from zero for finite transmission power in the presence of random noise. This still leaves us the question on how to maximize the total data rate R,

$$\max[R] = \sum_{m=1}^{L} b_m, \quad (24)$$

subject to a total power constraint,

$$\sum_{m=1}^{L} \sum_{k=1}^{r} E\left(|S_{km}|^2\right) \leq P_{\text{total}}, \quad (25)$$

as well as to an upper bound on the symbol error rate per subcarrier ρ_m

$$\rho_m \leq \varsigma, \quad m = 1, \ldots, L, \quad (26)$$

where b_m is the number of bits per subcarrier, ς is the upper bound on the SER across subcarrier, and P_{total} is the total power allocated for the system. Our proposed power optimization and bit-loading algorithms for MIMO configurations are derived in Section 6.

Throughout this section, we have only considered M-ary QAM constellations for subcarrier bit mappings with two or more bits, $b_m \geq 2$, and BPSK signaling for single bit subcarrier allocation. In addition, the analysis presented herein is for the kth path and the dependence on k is suppressed. Thus, we are able to upper bound the SER per subcarrier [17] as

$$\rho_m \leq 4Q\left(\sqrt{\frac{3|H_m|P_m}{(2^{b_m}-1)N_0}}\right), \quad m = 1, \ldots, L, \quad (27)$$

where N_0 is the noise power density and Q is complementary error function. This bound is tight for high SNRs. Using this upper bound as an equality on SER and letting the bit values be continuous, we make the aforementioned problems mathematically tractable in order to suggest a solution for each. Due to the "looseness" of this upper bound for low

SNRs, any solution derived from the equality approximation will be suboptimal, but from existing literature, the loss in performance is small ($\sim$ 0.1 dB) and still one of the better techniques.

To maximize the total rate with a constraint on total power, we rewrite (24) using the Lagrange multiplier method, that is,

$$J_R = \sum_{m=1}^{L} b_m + \lambda \sum_{m=1}^{L} P_m \tag{28}$$

which can be solved relatively straightforward for a power allocation and bit allocation as

$$\begin{aligned} P_m &= \frac{P_{\text{total}}}{L} - \frac{Q^{-1}(\varsigma/4)^2}{3\text{CNR}_m} + \frac{1}{L}\sum_{jj=1}^{L} \frac{Q^{-1}(\varsigma/4)^2}{3\text{CNR}_{jj}}, \\ b_m &= \log_2\left(1 + \frac{3\text{CNR}_m P_m}{Q^{-1}(\varsigma/4)^2}\right), \quad \text{for } m = 1, \dots, L, \end{aligned} \tag{29}$$

where the channel-to-noise-ratio (CNR) is

$$\text{CNR}_m = \frac{|H_m|^2}{N_0}. \tag{30}$$

During implementation, we must take the necessary care that we also satisfy positive power constraints and finite granularity (i.e., integer number of bits per subcarrier). Obviously, for unused subcarriers determined by the bit allocation expression above, the transmitter cannot apply power to the subcarriers with no bits assigned, thus in practice we set

$$P_m = 0 \quad \forall m, \text{ where } b_m = 0. \tag{31}$$

Furthermore, we do not use the subcarriers with poor gain characteristics which are identified by negative solutions due to the equality constraint in our Lagrange multiplier formulation. Thus, we remove these subcarriers by not allocating power to those subcarriers,

$$P_m = \max[0, P_m], \tag{32}$$

and due to logarithm of one (in any base) in our bit expression, we do not allocate any rate to the poorly operating subcarrier. Although through creative means fractional rates are achievable for QAM mapping, we consider here only an integer number of bits per subcarrier, so we floor the result from the above bit allocation equation,

$$b_m = \max[0, \lfloor b_m \rfloor]. \tag{33}$$

By flooring the bit assignment value, we ensure that the SER for each subcarrier meets the upper bound requirement in the problem formulation. Similar expressions can be derived for minimization of the average SER and transmission power under rate constraints. In Section 6, bit allocation concatenated with coding and modulation, referred to as *adaptive modulation*, is considered.

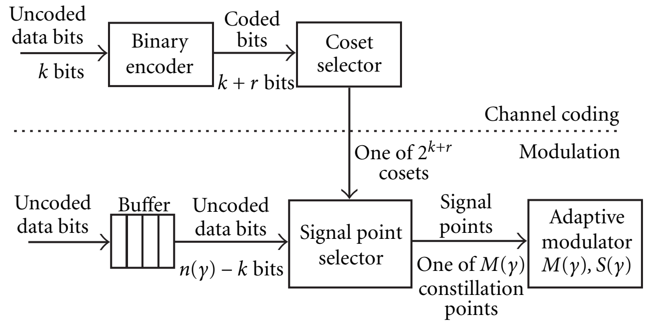

Figure 4: Goldsmith [6] proposed adaptive TCM scheme.

6. REVIEW OF ADAPTIVE MODULATION

Recall that, as mentioned in previous sections, when channel state information (CSI) is not available in transmitter, space-time codes should be used to get diversity and coding gain. Likewise for the SISO systems, when CSI is available in transmitter, we can increase the efficiency of the system optimal power and rate adaptation in transmitter. Provided that the fading channel gains are available in transmitter, codes can be constructed such that it adapts its output power and data rate to channel variation. Such codes can potentially achieve maximum channel capacity and are termed adaptive modulation. With adaptive modulation, rather than transmitting the same information rate for good channel and degraded channel, power optimization and bit allocation [6, 18, 19] are adapted to transmit more information when channel is good, and less information when channel is degraded. Thus, without sacrificing bit error rate (BER), these schemes provide high average spectral efficiency by transmitting at high speeds under favorable channel conditions, and reducing the throughput as the channel degrades. Adaptive coded modulation does not require interleaving since error bursts are eliminated by adjusting the power, size, and duration of the transmitted signal constellation relative to the channel fading. However, adaptive modulation does require accurate channel estimates at the receiver which are fed back to the transmitter with minimal latency.

The general structure of an adaptive coded modulation, which was introduced by Goldsmith and Chua [6] is shown in Figure 4. Specifically, a binary encoder operates on k uncoded data bits to produce $(k + r)$ coded bits, and the coset selector uses these coded bits to select one of the 2^{k+r} cosets from a partition of the signal constellation. For the nonadaptive modulation, $(n - k)$ additional bits are used to select one of the 2^{n-k} signal points in the selected coset, while in the adaptive modulator $(n(\gamma) - k)$ additional bits are used to select one of the $2^{n(\gamma)-k}$ preselected signal points out of the 2^{n-k} available signal points in the selected coset. The preselection is done in such a way that maximizes the minimum distance between signals.

In a fading channel the instantaneous SNR varies with time, which will cause the distance between signals varies. The basic premise for using adaptive modulation with coset codes is to keep these distances constant by varying the size of the signal constellation relative to γ, subject to an average transmit power constraint P. Therefore, maintaining a constant minimum distance $d_{\min}$, the adaptive coded modulation exhibits the same coding gain as coded modulation designed for an AWGN channel with minimum coded distance $d_{\min}$.

6.1. Improved adaptive modulation

We outline methods for improving the adaptive TCM modulation proposed by Goldsmith [6, 18], and others [14] when used in an OFDM framework. Goldsmith [6, 18] argues that burst error events are eliminated when power optimization is used. In constrast, we found that adaptive interleaving based on CSI can considerably improve the channel statistics such that greater throughputs are achievable. Goldsmith [6, 18] implemented adaptive modulation in a time-varying channel environment. OFDM systems, on the other hand, are typically deployed in very slowly varying fading environments.

In these scenarios, it is possible to benefit from interleaving. Typically, interleavers are designed to convert slowly varying channels into rapidly changing channel to exploit time diversity. For OFDM systems, the interleavers should exploit the frequency selective of the channel to benefit from frequency diversity. Normal block or convolutional interleaver can be used for this purpose. However, the fullest benefit from interleaving is achieved when the CSI is used to design it. Such interleaving is referred to as *adaptive interleaving*.

6.1.1 Adaptive interleaving

We outline a simple interleaver based on CSI at the transmitter, which improves the performance of an adaptive TCM scheme such that more uncoded bits per subcarrier can be supported than without interleaving. Our adaptive coordinate interleaver algorithm is outlined as follows:

(i) given an estimate of $\mathbf{H}(m)$;
(ii) define $\mathbf{Z}(m) = |\mathbf{H}(m)|$;
(iii) $[ff, kk] = \text{Sort}\{\mathbf{Z}(m)\}$, where kk are indexes associated with the weakest to strongest amplitudes;
(iv) let $jj = [1, \ldots, L]$;
(v) map $\text{real}[\mathbf{S}(jj)] \longrightarrow \mathbf{H}(kk)$;
(vi) map $\text{imag}[\mathbf{S}(jj)] \longrightarrow \mathbf{H}(\text{reverse}(kk))$, where $\text{reverse}([1\ \ 2\ \ 3\ \ 4]) = [4\ \ 3\ \ 2\ \ 1]$;
(vii) combine coordinators as order transmit over channel.

The resulting channel statistics after de-interleaving at the receiver has fewer deep null. A comparison of the channel statistic with and without our new interleaver is shown in Figure 5. Based on this figure, the number of supportable uncoded bits per subcarrier improves considerably. An alternative interleaving scheme based on CSI at the transmitter takes advantage of the pad bits used to generate integer number of OFDM symbol within a packet. The frame format for an OFDM packet for the IEEE 802.11a standard is shown in Figure 6. On average, the number of pad bits required is half an OFDM symbol, which is quite significant. Considering that the pad bits are discarded at the receiver, a better

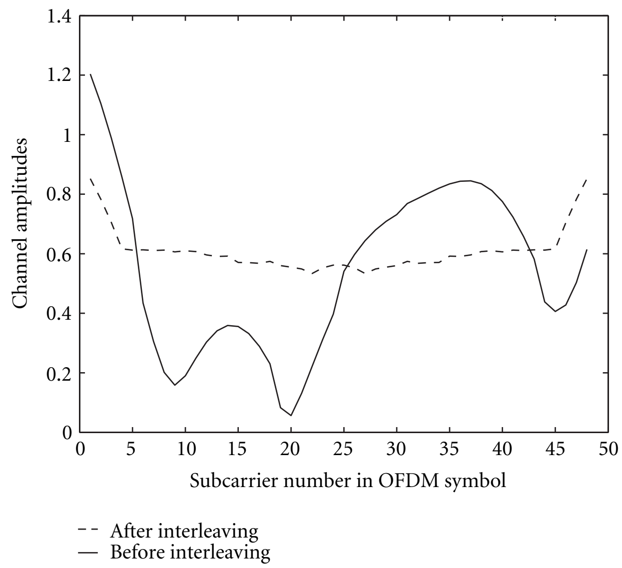

FIGURE 5: Virtual channel statistics at the decoder after de-interleaving.

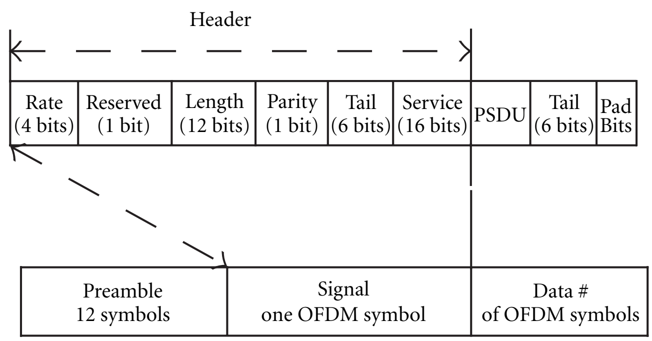

FIGURE 6: Physical layer protocol data unit (PPDU) in 802.11a, including the preamble, header, and PSDU.

utilization of the power and channel allocation for those pad bits would be to interchange data symbols on weak subcarriers for pad bits on strong subcarriers. To achieve the best performance, the adaptive interleaving should be performed in conjunction with a power optimization and bit allocation algorithm. Our proposed algorithm is outlined as follows:

(i) sort the channel amplitudes for all the OFDM data subcarriers in the packet in order of weakest to strongest;
(ii) reverse the order of the bits in data portion of the packet, for example, pad bits, tail bits, and physical layer service data unit (PSDU);
(iii) map the modulation symbols onto the subcarriers by filling the set of weakest subcarriers for each OFDM symbol first;
(iv) continue mapping the modulation symbols onto the group of the next weakest subcarriers for each OFDM symbol;

(v) continue in this fashion until the packet is filled;
(vi) using a power optimization algorithm, water-fill the packet.

This algorithm guarantees that the data symbols are mapped onto the most reliable subcarriers; the pad bits, the least reliable subcarriers. The subcarriers, which are allocated no power, are assigned to the pad bits first. Depending on channel conditions, some of the data subcarriers may still not be allocated any power; however, there are fewer of those data subcarriers than if the adaptive interleaver was not used. A better choice for adaptive modulation for this proposed adaptive interleaving is to use a block-coded modulation scheme. In particular, a Reed-Solomon (RS) constituent encoder provides great flexibility in error protection, code word size, and code rate, which is well suited for this application. The adaptive TCM scheme discussed thus far in this paper has been designed to minimize error probability and maximize average data rate. It is desirable to maintain a fixed number of bits per symbol to facilitate the need for multiple demodulators at the receiver. In this scenario, the data rate is guaranteed for each packet, and the error rate is driven as small as possible for any instance of the channel. That is, these codes minimize the outage probability for a given rate.

6.1.2 Adaptive error correction coding

A code that minimizes the outage capacity for a given rate should be able to easily adjust its coding gain to the channel statistics. The adaptive TCM scheme outlined in this paper assumes a certain number of coded bits per subcarrier, and the number of uncoded bits per subcarrier is adapted. We allow both numbers of uncoded bits per subcarrier and coded bits per subcarriers to varying, under the restriction that the number of bits per symbol per subcarrier is fixed. This is a very interesting problem that stems from practical considerations. For an OFDM system perspective, it is desirable to set the rate for the system prior to transmission. Furthermore, the packet error rate, not the bit error rate, must be minimized. Hence, each packet must meet the rate requirement first and the error requirement second. Since, there are a fixed number of parity symbols available determined by the rate, it is wise to use the adaptive interleaver scheme described in this section to specify erasures for the code. Consider a shortened RS linear block code. Adding erasures, up to the length of the original code, does not reduce the parity symbols available for random symbol errors. The caveat to this statement is that there need to be available sufficient pad bits for the specified erasures.

Now, we focus on applying bit and power allocation ideas to improve error correction codes. If hard erasure decoding is implemented, RS codes can correct up to $(NN - KK)$ erasures or $\lfloor (NN - KK)/2 \rfloor$ random errors, where KK is the number of information symbols and NN is the length of the RS code. Thus for an AWGN channel (spectrally flat channel), the word error rate (WER) is upper bounded by the expression

$$\text{WER} \leq \sum_{k=(NN-KK+1)}^{NN} \binom{NN}{k} \rho^k (1-\rho)^{NN-k}, \tag{34}$$

which assumes that the erasure locations are known or can be predicted. Note that for a spectrally shaped system such as an OFDM system, the SER ρ_m varies from subcarrier to subcarrier. Hence, the average SER over the $NN - \chi$ best subcarriers $\bar{\rho}_\chi$ is defined by

$$\bar{\rho}_\chi = \frac{1}{NN - \chi} \sum_{m=\chi+1}^{NN} \rho_m, \tag{35}$$

and (34) becomes

$$\min_{\chi \in [0,\ldots,(NN-KK+1)]} (\text{WER}), \tag{36}$$

where

$$\text{WER} \leq \sum_{k=(NN-\chi-KK+1)}^{NN-\chi} \binom{NN-\chi}{k} \bar{\rho}_\chi^k \left(1-\bar{\rho}_\chi\right)^{NN-\chi-k}. \tag{37}$$

Although the erasures are not actually transmitted across the channel, there need to be enough "empty" or "unmodulated" subcarriers to account for those symbols. Power saved by not transmitting the erasures should be reallocated amongst the data bearing subcarriers in a manner that minimizes (37). An outline for the algorithm is given as follows:

(1) given an estimate of $\mathbf{H}(m)$, sort the channel amplitudes for all the OFDM data subcarriers in the packet in order of weakest to strongest;
(2) determine the zero power locations from power and bit allocation algorithms and specify them as erasures for the shortened RS code;
(3) determine WER_i from (37);
(4) specify the next weakest subcarrier as an erasure and reallocate power;
(5) recalculate WER_{i+1};
(6) if $\text{WER}_i > \text{WER}_{i+1}$, repeat steps 4 and 5 until this is no longer true;
(7) using targeted SER bound ς, compute the SER per subcarrier ρ_m using (27), determine the location and number of symbols, which can be unprotected;
(8) finally, determine the number of bits needed to meet the total data rate R and allocate bits using the adaptive coded modulation structure introduced by Goldsmith and Chua [6], under the restriction of a fixed constellation size.

Note that the coset selector for this problem is the shortened RS code and not a convolutional code. An example of allocation of variable coded bits and uncoded bits for a fixed constellation size is depicted in Figure 7. In Figure 7, the number of uncoded bits supportable per subcarriers varies

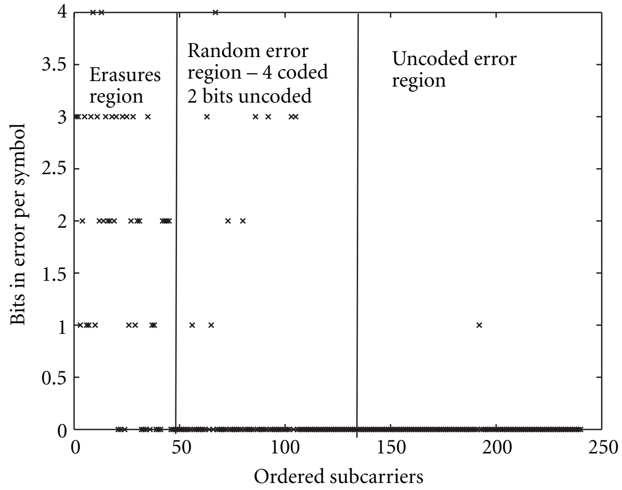

FIGURE 7: An adaptive bit allocation per subchannel for BCM scheme with 6 bits per symbol.

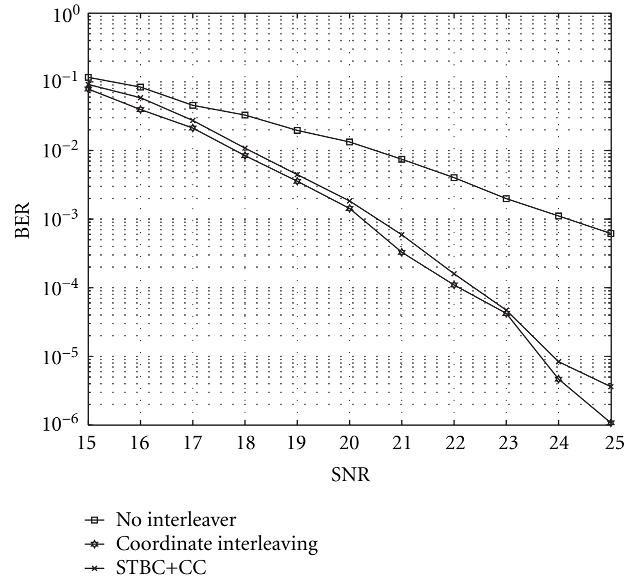

FIGURE 8: Diversity improvement for STBCC using coordinate interleaving in a Rayleigh fading channel.

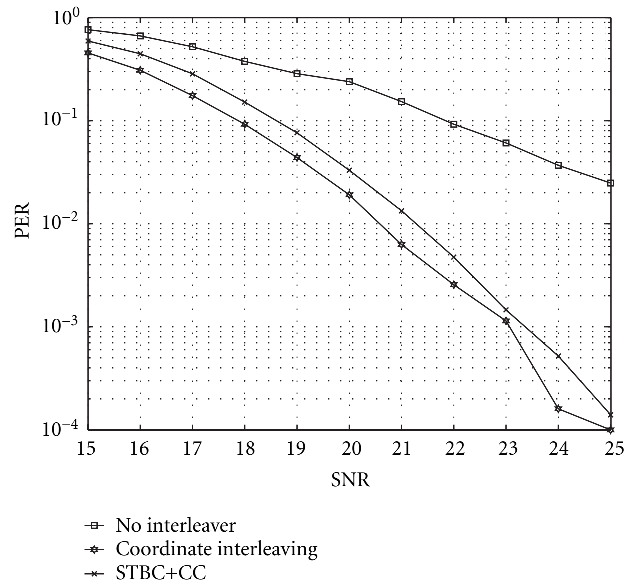

FIGURE 9: PER performance improved with coordinate interleaving for TCM 48 Mbit/s.

over the packet. The first 48 subcarriers of the 240 subcarriers in the packet are specified as erasures. The overall rate for the system is 2/3 not including the erasures specified by the stuff bits. When the power is re-distributed, the single error in the uncoded region is corrected. A summary of steps for performing adaptive error.

In the next section, we validate some of the algorithms presented in this document within the IEEE 802.11a framework. Note, that the IEEE 802.11a framework assumes quasi-static channels. That is, the channel is constant over a packet but changes from packet to packet. Hence, this is akin or similar to the environment used for outage probability.

7. SIMULATION RESULTS

We present a performance evaluation of several MIMO configurations within the IEEE 802.11a framework. The constituent TCM scheme uses a 128-state, rate 2/3 convolutional encoder for all the data rates considered. The IEEE 802.11a baseline system parameters are

(i) 64 pt. FFT: 48 data carrier and 4 pilot carriers,
(ii) 20 MHz sampling frequency,
(iii) 3.2 microsecond FFT period and 0.8 microsecond cyclic prefix,
(iv) symbol rate of 12 Msymbols/s.

First, we examined the packet error rate performance (PER) of STBCC schemes, 128 states, 4 bit/s/Hz over a five-tap Rayleigh channel. The channel taps were spaced 50 *n*sec apart with an exponential channel delay profile, which is less than the cyclic prefix. The first observation concerning Figures 8 and 9 is that the diversity order of the STBCC with no interleaving is two not one. Considering that this code has two parallel transitions per state; this simulation validates that STBCCs are guaranteed to have a diversity order of at least two, which was determined from their code word difference matrices. A second observation is that coordinating interleaving with constellation rotation definitely improves the diversity order of the code. In fact, the diversity order is double at moderate-to-high SNR as was predicted in Section 4.1. Furthermore, the bit error rate shown in Figure 8 is greatly improved as well as the error event probability illustrated by the PER in Figure 9. For comparison, a STBC concatenated with the IEEE 802.11a 48 Mbit/s mode is shown also in the figures. The purpose was to illustrate how well the coordinate interleaving improves the Hamming distance. The IEEE 802.11a uses a bit interleaver prior to the STBC.

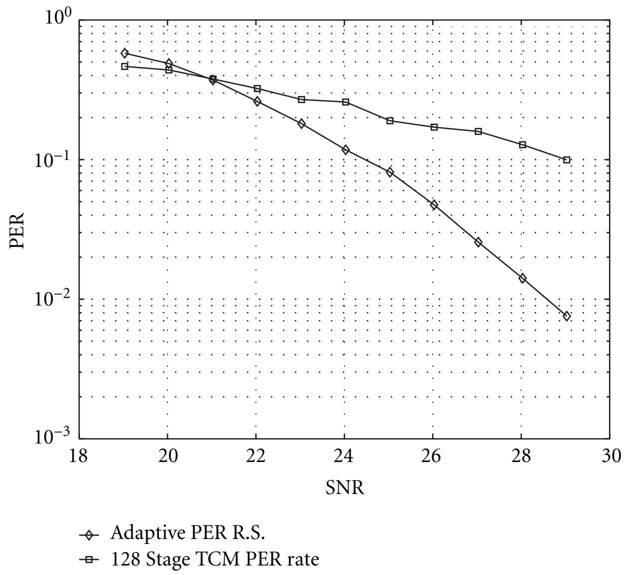

FIGURE 10: Adaptive modulation PER comparison: BCM versus TCM for five-tap Rayleigh fading and 48 Mbit/s mode.

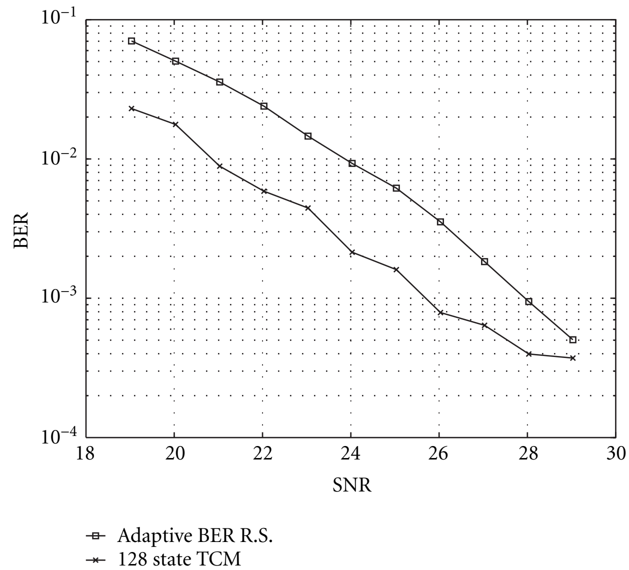

FIGURE 11: Adaptive modulation BER comparison: BCM versus TCM for five-tap Rayleigh fading and 48 Mbit/s mode.

To illustrate our proposed adaptive BCM, we consider the same TCM scheme with two parallel transitions per state. Both schemes are adaptively interleaved according to the algorithms specified in Section 6.1.1. The adaptive BCM scheme uses shortened versions of RS(63, 31) and uncoded symbols across the packet in a similar fashion, as depicted in Figure 7. The power optimization algorithm specified 48 erasures, of which 43 were pad symbols to fill the packet. Also, ninety-four of the subcarriers had four coded bits and two coded bits. The remaining subcarriers were uncoded. The power from the 48 erasures were distributed in a water-filling fashion across the data bearing subcarriers, for example, the uncoded symbol were placed on the strongest subcarriers and distributed a greater percentage of the power. The adaptive BCM performs significantly better in terms of PER as shown in Figure 10, although the adaptive TCM scheme performs better in terms of BER performance as shown in Figure 11. This validate our assertion that an adaptive BCM is better suited for environments where the outage probability is the appropriate capacity measure.

Our final example illustrates STBCC with coordinate interleaving using the TCM scheme described earlier and 6 uncoded bits per symbol. The overall rate is 96 Mbit/s in a 20 MHz band, which has a spectral efficiency of 4.8 bit/s/Hz. For comparison, the 54 Mbit/s mode of the IEEE 802.11a is shown in Figure 12, which has a spectral efficiency of 2.7 bit/s/Hz. At high SNR, the STBCC would eventually outperforms the 54 Mbit/s although it carries, on average, an additional 2.1 bit/s/Hz capacity. If receive diversity is available, power saving of the order of 10 dB is possible with only two-receive antennas. Further, because of the structure of the STBCC, the additional signal processing required for multiple receiver is greatly diminished when compared to the maximum likelihood decoders need for STTC.

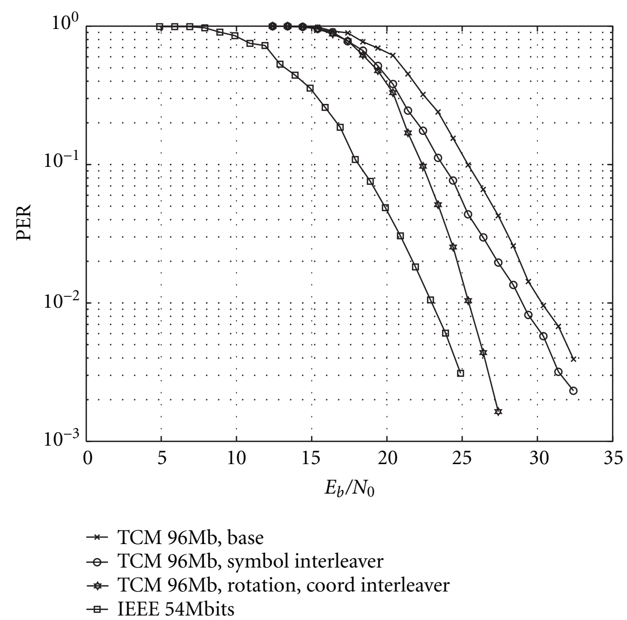

FIGURE 12: Comparison of STBCC with symbol interleaving to a STBCC with coordinate interleaving, and constellation rotations and 54 Mbit/s mode of the IEEE 802.11a in five-tap Rayleigh fading.

8. CONCLUSIONS

In this paper, we demonstrated the benefits of the concatenating coset codes with orthogonal STBC, which we termed STBCC. It was found that STBCC with outer codes designed for AWGN channels produces suitable codes for fading channels. In fact, the coding advantage of these codes, based on the code word difference matrix, is as good or better than the STTC found from computer search. Although STBCCs are guaranteed a minimum diversity level of 2, we showed that 2D constellation rotation and coordinate interleaving can double that diversity order. Performance evaluation over a frequency selective fading channel was demonstrated using Monte Carlo simulations, which showed substantial improvements in PER performances. Our final example showed a STBCC configuration for 96 Mbit/s in 20 MHz bandwidth, which performed comparably to the 54 Mbit/s mode of the IEEE 802.11a in terms of E_b/N_0.

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