High-Rate Codes That Are Linear in Space and Time

Babak Hassibi and Bertrand M. Hochwald

***Abstract*—Multiple-antenna systems that operate at high rates require simple yet effective space–time transmission schemes to handle the large traffic volume in real time. At rates of tens of bits per second per hertz, Vertical Bell Labs Layered Space–Time (V-BLAST), where every antenna transmits its own independent substream of data, has been shown to have good performance and simple encoding and decoding. Yet V-BLAST suffers from its inability to work with fewer receive antennas than transmit antennas—this deficiency is especially important for modern cellular systems, where a base station typically has more antennas than the mobile handsets. Furthermore, because V-BLAST transmits independent data streams on its antennas there is no built-in spatial coding to guard against deep fades from any given transmit antenna. On the other hand, there are many previously proposed space–time codes that have good fading resistance and simple decoding, but these codes generally have poor performance at high data rates or with many antennas. We propose a high-rate coding scheme that can handle any configuration of transmit and receive antennas and that subsumes both V-BLAST and many proposed space–time block codes as special cases. The scheme transmits substreams of data in linear combinations over space and time. The codes are designed to optimize the mutual information between the transmitted and received signals. Because of their linear structure, the codes retain the decoding simplicity of V-BLAST, and because of their information-theoretic optimality, they possess many coding advantages. We give examples of the codes and show that their performance is generally superior to earlier proposed methods over a wide range of rates and signal-to-noise ratios (SNRs).**

***Index Terms*—Bell Labs Layered Space–Time (BLAST), fading channels, multiple antennas, receive diversity, space–time codes, transmit diversity, wireless communications.**

I. Introduction and Model

It is widely acknowledged that reliable fixed and mobile wireless transmission of video, data, and speech at high rates will be an important part of future telecommunications systems. One way to get high rates on a scattering-rich wireless channel is to use multiple transmit and/or receive antennas. In [1], [2], theoretical and experimental evidence demonstrates that channel capacity grows linearly as the number of transmit and receive antennas grow simultaneously.

Early uses of multiple transmit antennas in a scattering environment achieve reliability through "diversity," where redundant information is sent or received on two or more antennas in the hope that at least one path from the transmitter reaches the receiver [3]–[6]. To keep the transmitter and receiver complexity low, linear processing is often preferred [3]. To achieve the high data rates promised in [2], however, new approaches for space–time transmission are needed. One such approach is presented in [7], [8] where a practical scheme, called V-BLAST (Vertical Bell Labs Layered Space–Time), encodes and decodes rates of tens of bits per second per hertz (b/s/Hz) with 8 transmit and 12 receive antennas. The V-BLAST architecture breaks the original data stream into substreams that are transmitted on the individual antennas. The receiver decodes the substreams using a sequence of nulling and canceling steps.

Since then there has been considerable work on a variety of space–time transmission schemes and performance measures [9] such as the space–time trellis codes of [10] and the space–time block codes of [11], [12] for the known channel and [13]–[17] for the unknown channel.

At very high rates and with a large number of antennas, many of these space–time codes suffer from complexity or performance difficulties. The number of states in the trellis codes of [10] grows exponentially with either the rate or the number of transmit antennas. The block codes of [11], [12] suffer from rate and performance loss as the number of antennas grow, and the codes of [14]–[16] suffer from decoding complexity if the rate is too high. Although V-BLAST can handle high data rates with reasonable complexity, the decoding scheme presented in [7] does not work with fewer receive than transmit antennas.

We present a space–time transmission scheme that has many of the coding and diversity advantages of previously designed codes, but also has the decoding simplicity of V-BLAST at high data rates. The codes work with arbitrary numbers of transmit and receive antennas.

The codes break the data stream into substreams that are dispersed in linear combinations over space and time. We refer to them simply as linear dispersion codes (LD codes). The LD codes

1) subsume, as special cases, both V-BLAST [7] and the block codes of [12];
2) generally outperform both;
3) can be used for any number of transmit and receive antennas;
4) are very simple to encode;
5) can be decoded in a variety of ways including simple linear-algebraic techniques such as
    a) successive nulling and canceling (V-BLAST [7], square-root V-BLAST [18]),
    b) sphere decoding [19], [20];
6) are designed with the numbers of both the transmit *and* receive antennas in mind;

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7) satisfy the following information-theoretic optimality criterion:
   — the codes are designed to maximize the mutual information between the transmit and receive signals.

We briefly summarize the general structure of the LD codes. Suppose that there are M transmit antennas, N receive antennas, and an interval of T symbols available to us during which the propagation channel is constant and known to the receiver. The transmitted signal can then be written as a $T \times M$ matrix S that governs the transmission over the M antennas during the interval. We assume that the data sequence has been broken into Q substreams (for the moment we do not specify Q) and that $s_1, \ldots, s_Q$ are the complex symbols chosen from an arbitrary, say r-PSK or r-QAM, constellation. We call a rate $R = (Q/T)\log_2 r$ linear dispersion code one for which S obeys

$$S = \sum_{q=1}^{Q} (\alpha_q A_q + j\beta_q B_q) \tag{1}$$

where the real scalars $\{\alpha_q, \beta_q\}$ are determined by

$$s_q = \alpha_q + j\beta_q, \quad q = 1, \ldots Q.$$

The design of LD codes depends crucially on the choices of the parameters T, Q and the dispersion matrices $\{A_q, B_q\}$. To choose the $\{A_q, B_q\}$ we propose to optimize a nonlinear information-theoretic criterion: namely, the mutual information between the transmitted signals $\{\alpha_q, \beta_q\}$ and the received signal. We argue that this criterion is very important for achieving high spectral efficiency with multiple antennas. We also show how the information-theoretic optimization has implications for performance measures such as pairwise error probability. Section IV has several examples of LD codes and performance comparisons with existing schemes.

We now present the multiple-antenna model considered in this paper.

A. The Multiple-Antenna Model

In a narrow-band, flat-fading, multiple-antenna communication system with M transmit and N receive antennas, the transmitted and received signals are related by

$$x = \sqrt{\frac{\rho}{M}}\, Hs + v \tag{2}$$

where $x \in \mathcal{C}^N$ denotes the vector of complex received signals during any given channel use, $s \in \mathcal{C}^M$ denotes the vector of complex transmitted signals, $H \in \mathcal{C}^{N\times M}$ denotes the channel matrix, and the additive noise $v \in \mathcal{C}^N$ is $\mathcal{CN}(0, 1)$ (zero-mean, unit-variance, complex-Gaussian) distributed that is spatially and temporally white. The channel matrix H and transmitted vector s are assumed to have unit variance entries, implying that

$$\mathrm{E}\,\mathrm{tr} HH^* = MN \quad \text{and} \quad \mathrm{E}s^*s = M.$$

Since the random quantities H, s, and v are independent, the normalization $\sqrt{\frac{\rho}{M}}$ in (2) ensures that ρ is the signal-to-noise ratio (SNR) at each receive antenna, independently of M. We often (but not always) assume that the channel matrix H also has independent $\mathcal{CN}(0, 1)$ entries.

The entries of the channel matrix are assumed to be known to the receiver but not to the transmitter. This assumption is reasonable if training or pilot signals are sent to learn the channel, which is then constant for some coherence interval. The coherence interval of the channel should be large compared to M [21]. When the channel is known at the receiver, the resulting channel capacity (often referred to as the *perfect-knowledge* capacity) is [2], [1]

$$C(\rho, M, N) = \max_{R_s \geq 0, \mathrm{tr} R_s = M} \mathrm{E} \log\det\left(I_N + \frac{\rho}{M} HR_sH^*\right) \tag{3}$$

where the expectation is taken over the distribution of the random matrix H.[1] The capacity-achieving s is a zero-mean complex Gaussian vector with covariance matrix $\mathrm{E}ss^* = R_{s,\mathrm{opt}}$, where $R_{s,\mathrm{opt}}$ is the maximizing covariance matrix in (3). When the distribution on H is rotationally invariant, i.e., when $p(H) = p(\Theta H) = p(H\Phi)$ for any unitary matrices Θ and Φ (as is the case when H has independent $\mathcal{CN}(0, 1)$ entries), the optimizing covariance is $R_{s,\mathrm{opt}} = I_M$, and (3) becomes

$$C(\rho, M, N) = \mathrm{E} \log\det\left(I_N + \frac{\rho}{M} HH^*\right). \tag{4}$$

This expectation can sometimes be computed in closed form (see, for example, [22]).

When the channel is constant for at least T channel uses we may write

$$x_\tau = \sqrt{\frac{\rho}{M}}\, Hs_\tau + v_\tau, \qquad \tau = 1, \ldots, T$$

so that defining

$$X = [x_1 \quad x_2 \quad \cdots \quad x_T]^t$$
$$S = [s_1 \quad s_2 \quad \cdots \quad s_T]^t$$

and

$$V = [v_1 \quad v_2 \quad \cdots \quad v_T]^t$$

(where the superscript t denotes "transpose"), we obtain

$$X^t = \sqrt{\frac{\rho}{M}}\, HS^t + V^t.$$

It is generally more convenient to write this equation in its transposed form

$$X = \sqrt{\frac{\rho}{M}}\, SH + V \tag{5}$$

where we have omitted the transpose notation from H and simply redefined this matrix to have dimension $M \times N$. The matrix $X \in \mathcal{C}^{T\times N}$ is the received signal, $S \in \mathcal{C}^{T\times M}$ is the transmitted signal, and $V \in \mathcal{C}^{T\times N}$ is the additive $\mathcal{CN}(0, 1)$ noise. In X, S, and V, time runs vertically and space runs horizontally. We are concerned with designing the signal matrix S obeying the power constraint $\mathrm{E}\,\mathrm{tr} SS^* = TM$.

[1]Equation (3) actually slightly generalizes [2], [1], which assume that H has independent $\mathcal{CN}(0, 1)$ entries.

We note that, in general, the number of $T \times M$ matrices S needed in a codebook can be quite large. If the rate in bits per channel use is denoted R, then the number of matrices is 2^{RT}. For example, with $M = 4$ transmit and $N = 2$ receive antennas the channel capacity at $\rho = 20$ dB (with $\mathcal{CN}(0, 1)$ distributed H) is more than 12 bits per channel use. Even with a relatively small block size of $T = 4$, we need $2^{48} \approx 10^{14}$ matrices at rate $R = 12$. The huge size of the constellations generally rules out the possibility of decoding at the receiver using exhaustive search.

The LD codes that we present easily generate the very large constellations that are needed. Moreover, because of their structure, they also allow efficient real-time decoding. In the next section, we briefly describe and analyze some existing space–time codes so that we may motivate the LD codes and explain how they are different.

II. Information-Theoretic Analysis of Some Space–Time Codes

Since the capacity of the multiple-antenna channel can easily be calculated, we may ask how well a space–time code performs by comparing the maximum mutual information that it can support to the actual channel capacity. The distribution for the $T \times M$ matrix S that achieves (4) is independent $\mathcal{CN}(0, 1)$ entries, but we cannot easily use this by itself as a guideline for constructing and decoding a (random) constellation of $T \times M$ matrices because of the sheer number of matrices involved. Therefore, a constellation of matrices that has sufficient structure for efficient encoding and decoding is generally needed. One such structure is that of an *orthogonal design*, originally proposed in [11] and later generalized in [12].

A. Mutual Information Attainable With Orthogonal Designs

An orthogonal design is introduced by Alamouti in [11] for $T = M = 2$ and has the structure

$$S = \begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{bmatrix}. \tag{6}$$

The complex scalars s_1 and s_2 are drawn from a particular (r-PSK or r-QAM) constellation, but we may simply assume that they are random variables such that $\mathrm{E}(|s_1|^2 + |s_2|^2) = 2$. We show that this particular structure can be used to achieve capacity when there is one receive antenna but *not* when there is more than one. Portions of our argument may also be found in [23], [24].

1) One Receive Antenna ($N = 1$): With $N = 1$, (5) becomes

$$\begin{bmatrix} x_1 \\ x_2 \end{bmatrix} = \sqrt{\frac{\rho}{2}} \begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{bmatrix} \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} + \begin{bmatrix} v_1 \\ v_2 \end{bmatrix}.$$

This can be rewritten as

$$\underbrace{\begin{bmatrix} x_1 \\ x_2^* \end{bmatrix}}_{\triangleq x} = \sqrt{\frac{\rho}{2}} \underbrace{\begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix}}_{\triangleq \mathcal{H}} \underbrace{\begin{bmatrix} s_1 \\ s_2 \end{bmatrix}}_{\triangleq s} + \begin{bmatrix} v_1 \\ v_2^* \end{bmatrix}. \tag{7}$$

It readily follows that

$$\mathcal{H}\mathcal{H}^* = \mathcal{H}^*\mathcal{H} = \left(|h_1|^2 + |h_2|^2\right) I_2. \tag{8}$$

We effectively have an equivalent matrix channel $\mathcal{H}$ in (7) that is a scaled unitary matrix. Maximum-likelihood decoding of s_1 and s_2 is, therefore, decoupled [11].

We may ask how much mutual information the orthogonal design structure (6) can attain? To answer this question we need to compute the mutual information between the transmitted and received vectors s and x in the equivalent channel model (7) and compare it with the capacity of an $M = 2$, $N = 1$ multiple-antenna system.

Since s has the power constraint $\mathrm{E}s^*s = 2$, the maximum mutual information in (7) is simply the channel capacity that is obtained with the structured channel matrix $\mathcal{H}$. If we denote this maximum mutual information by $C_{\mathrm{orth}}(\rho)$, using (3) we obtain

$$\begin{aligned} C_{\mathrm{orth}}(\rho) &= \max_{R_s \geq 0, \mathrm{tr} R_s = 2} \frac{1}{2} \mathrm{E} \log\det\left(I_2 + \frac{\rho}{2} \mathcal{H} R_s \mathcal{H}^*\right) \\ &= \max_{R_s \geq 0, \mathrm{tr} R_s = 2} \frac{1}{2} \mathrm{E} \log\det\left(I_2 + \frac{\rho}{2} \mathcal{H}^* \mathcal{H} R_s\right) \\ &= \max_{R_s \geq 0, \mathrm{tr} R_s = 2} \frac{1}{2} \mathrm{E} \log\det\left(I_2 + \frac{\rho}{2} \left(|h_1|^2 + |h_2|^2\right) R_s\right) \end{aligned}$$

where the factor $\frac{1}{2}$ in front of the expectation normalizes for the two channel uses spanned by the orthogonal design. Since, subject to a trace constraint, the determinant of any positive-definite matrix is maximized when its eigenvalues are all equal, it is easy to see that the maximizing covariance matrix is $R_{s,\,\mathrm{opt}} = I_2$, so that we obtain

$$\begin{aligned} C_{\mathrm{orth}}(\rho) &= \mathrm{E} \log\left(1 + \frac{\rho}{2} \left(|h_1|^2 + |h_2|^2\right)\right) \\ &= C(\rho, M = 2, N = 1). \end{aligned} \tag{9}$$

The expression on the right symbolically denotes the capacity attained by a system with $M = 2$ transmit antennas and $N = 1$ receive antennas at SNR ρ. This equation implies that the orthogonal design (6) can achieve the full channel capacity of the $M = 2$, $N = 1$ system, and there is no loss incurred by assuming the structure (6) as opposed to a general transmit matrix S.

2) Two or More Receive Antennas ($N \geq 2$): With $N = 2$ receive antennas, (5) becomes

$$\begin{bmatrix} x_{11} & x_{12} \\ x_{21} & x_{22} \end{bmatrix} = \sqrt{\frac{\rho}{2}} \begin{bmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{bmatrix} \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} + \begin{bmatrix} v_{11} & v_{12} \\ v_{21} & v_{22} \end{bmatrix}$$

which can be reorganized as

$$\underbrace{\begin{bmatrix} x_{11} \\ x_{21}^* \\ x_{12} \\ x_{22}^* \end{bmatrix}}_{\triangleq x} = \sqrt{\frac{\rho}{2}} \underbrace{\begin{bmatrix} h_{11} & h_{21} \\ h_{21}^* & -h_{11}^* \\ h_{12} & h_{22} \\ h_{22}^* & -h_{12}^* \end{bmatrix}}_{\triangleq \mathcal{H}} \underbrace{\begin{bmatrix} s_1 \\ s_2 \end{bmatrix}}_{\triangleq s} + \begin{bmatrix} v_{11} \\ v_{21}^* \\ v_{12} \\ v_{22}^* \end{bmatrix}. \tag{10}$$

We now readily see

$$\mathcal{H}^*\mathcal{H} = \left(|h_{11}|^2 + |h_{21}|^2 + |h_{12}|^2 + |h_{22}|^2\right) I_2. \tag{11}$$

As with $N = 1$, maximum-likelihood estimation of s_1 and s_2 is decoupled. However, unlike with $N = 1$, the orthogonal design structure prohibits us from achieving channel capacity.

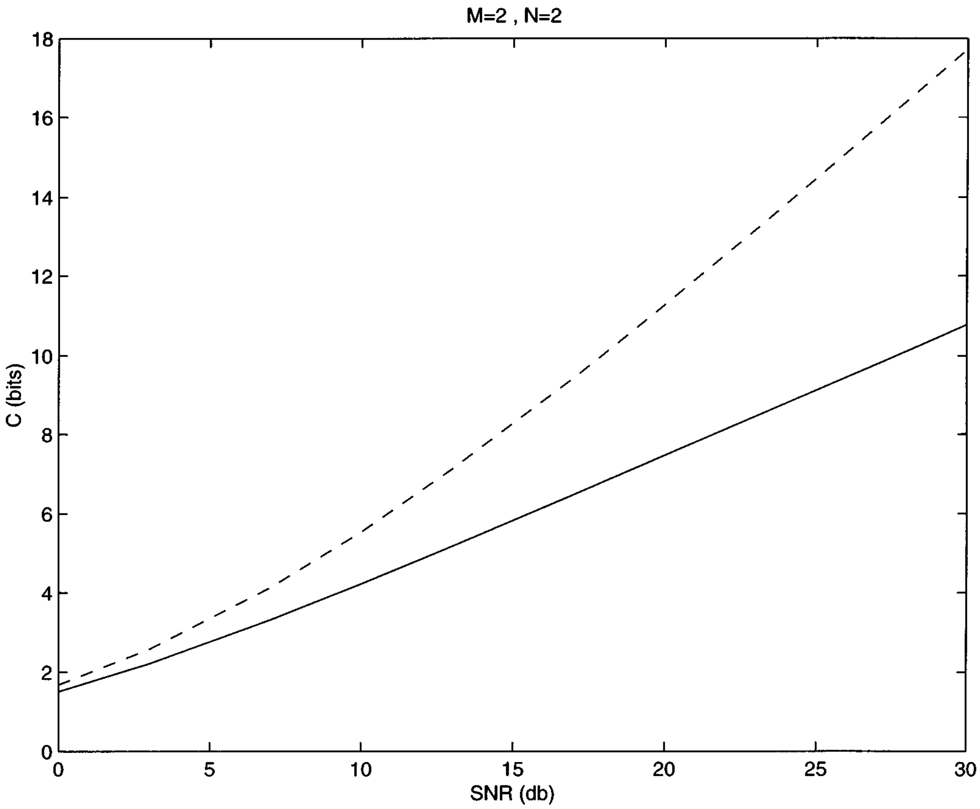

Fig. 1. Maximum mutual information achieved by 2×2 orthogonal design (6) compared to actual channel capacity for the $M = 2$, $N = 2$ system. Solid line: maximum mutual information for 2×2 orthogonal design. Dashed line: capacity of the $M = 2$, $N = 2$ system.

To see this, we compare the maximum mutual information between s and x in (10) with $C(\rho, M = 2, N = 2)$, the actual channel capacity for the system.

As before, the maximum mutual information in (10) is simply the channel capacity for the structured channel matrix $\mathcal{H}$. Denoting this maximum mutual information by $C_{\rm orth}(\rho)$, we obtain

$$\begin{aligned} C_{\rm orth}(\rho) &= \max_{R_s \geq 0, {\rm tr} R_s = 2} \frac{1}{2} \mathrm{E} \log\det\left(I_2 + \frac{\rho}{2}\mathcal{H}^*\mathcal{H}R_s\right) \\ &= \max_{R_s \geq 0, {\rm tr} R_s = 2} \frac{1}{2} \mathrm{E} \log\det\left(I_2 + \frac{\rho}{2}\left(|h_{11}|^2 + |h_{21}|^2 \right.\right. \\ &\qquad\qquad \left.\left. + |h_{12}|^2 + |h_{22}|^2\right) R_s\right) \\ &= \frac{1}{2}\mathrm{E}\log\det\left(I_2 + \frac{\rho}{2}(|h_{11}|^2 + |h_{21}|^2 + |h_{12}|^2 + |h_{22}|^2) I_2\right) \\ &= \mathrm{E}\log\left(1 + \frac{2\rho}{4}\left(|h_{11}|^2 + |h_{21}|^2 + |h_{12}|^2 + |h_{22}|^2\right)\right) \\ &= C(2\rho, M = 4, N = 1) \\ &< C(\rho, M = 2, N = 2). \end{aligned} \tag{12}$$

The last equation implies that the orthogonal design (6) is restrictive and *does not* allow us to achieve the full channel capacity of the $M = 2$, $N = 2$ system, but rather the capacity of an $M = 4$, $N = 1$ system at twice the SNR. Thus, when $N = 2$ we take a loss with the structure (6). The amount of this loss is substantial at high SNR and is depicted in Fig. 1 which shows the actual channel capacity compared to the maximum mutual information obtained by the orthogonal design (6).

For $N > 2$ receive antennas, the analysis is similar and is omitted. We simply state that for $M = 2$ transmit antennas and N receive antennas the 2×2 orthogonal design allows us to attain only $C(N\rho, M = 2N, N = 1)$, rather than the full $C(\rho, M = 2, N)$.

3) Other Orthogonal Designs: The preceding subsection focuses on the $M = 2$ orthogonal design but there are also orthogonal designs for $M > 2$. The complex orthogonal designs for $M > 2$ are no longer "full-rate" (see [12]) and therefore generally perform poorly in the maximum mutual information they can achieve, even when $N = 1$. We give an example of these nonsquare orthogonal designs [12], [25].

For $M = 3$, we have, for example, the rate $\frac{3}{4}$ orthogonal design

$$S = \sqrt{\frac{4}{3}} \begin{bmatrix} s_1 & s_2 & s_3 \\ -s_2^* & s_1^* & 0 \\ -s_3^* & 0 & s_1^* \\ 0 & -s_3^* & s_2^* \end{bmatrix}. \tag{13}$$

The factor $\sqrt{4/3}$ ensures that $\mathrm{E}\,\mathrm{tr}\, SS^* = T \cdot M = 12$. It can be shown that maximum-likelihood estimation of the variables s_1, s_2, s_3 is decoupled. Again using an argument similar to

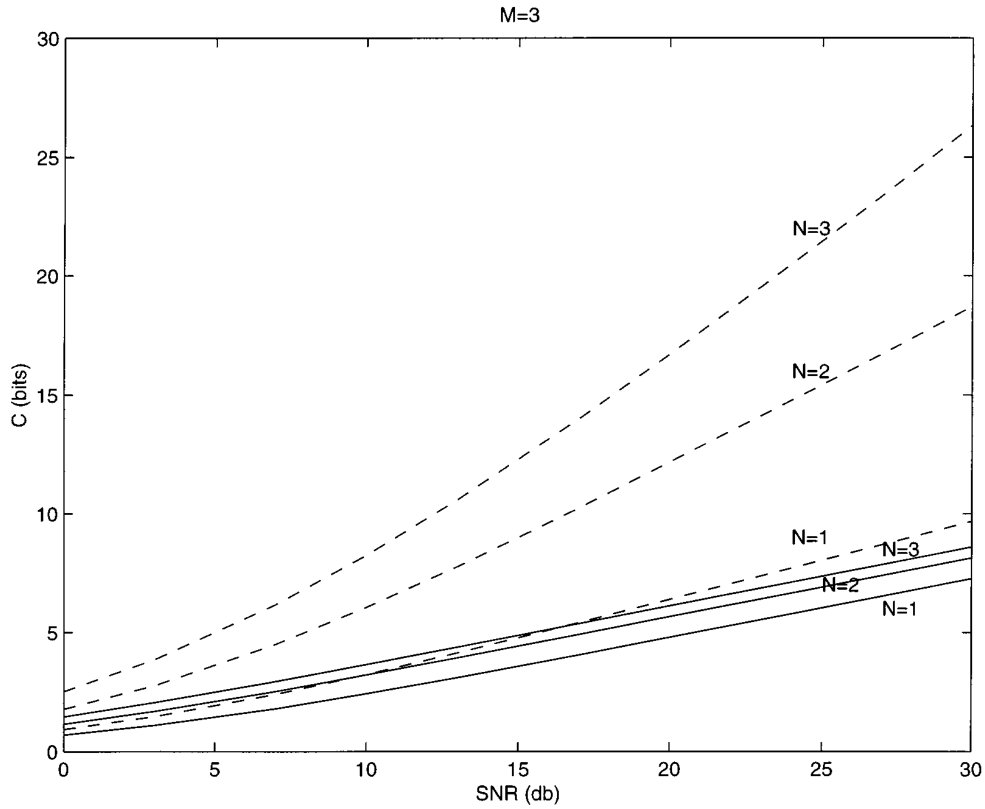

Fig. 2. Maximum mutual information achieved by 4×3 orthogonal design (13) compared to actual channel capacity. Solid lines: maximum mutual information of 4×3 orthogonal design for $N = 1, 2, 3$ receive antennas. Dashed lines: capacity of $M = 3$, $N = 1, 2, 3$ systems.

the one presented for $M = 2$, it is straightforward to show that the maximum mutual information attainable with (13) with N receive antennas is $\frac{3}{4}C(N\rho, M = 3N, N = 1)$ which is (much) less than the true channel capacity $C(\rho, M, N)$. We omit the proof and refer instead to Fig. 2 which shows the actual channel capacity compared to the maximum mutual information obtained by the orthogonal design (13).

B. Mutual Information Attainable With V-BLAST

We showed in Section II-A that, even though orthogonal designs allow efficient maximum-likelihood decoding and yield "full-diversity" (the appearance of the sum of the $|h_{ij}|^2$ in the mutual information formulas attests to this), orthogonal designs generally cannot achieve high spectral efficiencies in a multiple-antenna system, no matter how much coding is added to the transmitted signal constellation. This is especially true when the system has more than one receive antenna. An examination of the model (10) (and its counterparts for other orthogonal designs) reveals that the orthogonal design does not allow enough "degrees of freedom"—there are only two unknowns in (10) but four equations.

We can conclude that orthogonal designs are not suitable for very-high-rate communications. On the other hand, a scheme that is proven to be suitable for high spectral efficiencies is V-BLAST [7]. In V-BLAST each transmit antenna during each channel use sends an independent signal (often referred to as a substream). Thus, over a block of T channel uses, the $T \times M$ transmit matrix takes on the form

$$S = \begin{bmatrix} s_{11} & s_{12} & \cdots & s_{1M} \\ s_{21} & s_{22} & \cdots & s_{2M} \\ \vdots & \vdots & \ddots & \vdots \\ s_{T1} & s_{T2} & \cdots & s_{TM} \end{bmatrix} \tag{14}$$

where each s_{ij} is an independent symbol drawn from a complex constellation. Since the transmitted symbols are not dispersed in time, T is arbitrary. (We could, for example, take $T = 1$.)

When $N \geq M$ (the number of receive antennas is at least as large as the number of transmit antennas), there exist efficient schemes for decoding the V-BLAST matrices. These are based on "successive nulling and canceling" [7], and its more efficient variants [18], as well as more recently on sphere decoding [19]. All these decoding schemes essentially solve a well-conditioned system of linear equations. Successive nulling and canceling provides a fast approximate solution to the maximum-likelihood decoding problem with the benefit of cubic complexity in the number of transmit antennas (M^3). Sphere decoding, on the other hand, finds the exact maximum-likelihood estimate and benefits from avoiding an explicit exhaustive search. Recent work [20] has shown analytically that for a wide range of SNRs, the expected computational complexity of sphere decoding is also roughly cubic in the number of transmit antennas.

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