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Time-Division Versus Superposition Coded Modulation Schemes for Unequal Error Protection

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***Abstract*— This paper is concerned with the design problem of coded modulation schemes which provide unequal error protection (UEP) against Gaussian noise. Starting with the case of two bit streams, we consider two classes of UEP schemes: time-division coded modulation (TDCM) and superposition coded modulation (SCM). The early result of Bergmanns and Cover [1] ensures the asymptotic superiority of SCM. However, accounting for the fact that practical channel codes do not achieve capacity, we show that the validity of this result depends on the prescribed degree of inequality in protection of the streams. To complement the straightforward design and use of TDCM schemes, we propose methods to decode and design SCM schemes. The design examples substantiate the conclusions obtained from performance analysis. We extend the results to the case of UEP of multiple (more than two) bit streams, where a hybrid of SCM and TDCM must also be considered. We show that, for a large range of unequal protection levels, the hybrid scheme outperforms both SCM and TDCM.**

***Index Terms*—Coded modulation, quadrature amplitude modulation, time-varying channels, unequal error protection.**

I. Introduction

IN THIS PAPER, we study the performance and design of unequal error protection (UEP) coded modulation schemes. The term UEP implies that the resources available to provide protection to the various bit streams are not equally distributed, but instead, each bit stream may be protected so that it withstands a different level of channel noise. A natural motivation for UEP is due to the fact that bit errors in compressed data may cause vastly different levels of damage depending on their precise location. In such applications, it is advantageous to provide varying levels of protection to the different bits. Moreover, the type of unequal protection coding schemes that we consider in this paper are mainly motivated by applications where the coded signal is transmitted over a channel whose precise characteristics are not known to the transmitter. Such communication channels include: 1) broadcast channels and 2) mobile communication channels without a feedback path from the receiver to the transmitter.

In the case of broadcast applications, a single transmitter transmits an encoded signal to many receivers. Depending on its location, each receiver experiences a different channel condition in terms of the received signal strength and the level of interference (and noise) power. An appropriate source-channel coding strategy [2] for this case consists of employing a multi-resolution source coder followed by unequal protection of its different bit streams. The coarse level bit stream is given the heaviest protection while the finer resolution bit streams are given lighter protection. Each receiver estimates its channel condition and decodes as many bit streams as can be reliably decoded. Thus, the coarse level bit stream can be decoded even when the channel conditions are relatively poor. The receivers that experience better channel conditions will also decode the finer resolution bit streams and hence will obtain a reproduced signal of improved quality.

In a mobile communication scenario, the information about the time-varying channel characteristics is often available at the receiver and can be used during the decoding process. However, if a feedback path is not available (or is not feasible), the transmitter has no access to this information. Hence, the encoding operation should take into account the range of possible channel conditions that the receiver may experience. Clearly, this situation is quite similar to the broadcast case. Here, too, it is advantageous to adopt a source-channel coding scheme which consists of a multiresolution source coder with unequal protection of the different resolution bit streams.

UEP can be achieved by employing either time-division coded modulation (TDCM) or superposition coded modulation (SCM). TDCM is a form of resource sharing in which bit streams of differing importance are transmitted on disjoint modulation intervals. In SCM, the different bit streams are transmitted on the same modulation intervals. It has been shown by Bergmanns and Cover [1] that, ideally (i.e., asymptotically), SCM always outperforms TDCM. Motivated by

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that result, much of the earlier work on the design of UEP schemes focused on SCM schemes [2], [4]. An important exception to this is the work of Wei [3], where both TDCM and SCM schemes are considered. In fact, Wei observes that there exist examples where TDCM outperforms SCM. He also speculates on the possible reasons for this discrepancy between performance of ideal and practical UEP channel coding schemes. The aim of this paper is to conduct a systematic investigation into performance and design of UEP channel codes while taking into account the limitations of practical channel codes. The primary contribution of this paper is to clearly identify *why* and *when* TDCM or a hybrid of TDCM and SCM is expected to significantly outperform SCM.

This paper is organized as follows. We first formulate the design problem where two bit streams need to be given unequal protection. We then describe the two main categories of UEP schemes, namely, TDCM and SCM. Results of ideal performance analysis are summarized, followed by investigation of practical (nonideal) performance analysis. The analysis leads to characterization of the conditions under which TDCM outperforms SCM. We subsequently describe methods to decode and design SCM schemes. Simulation results on the design examples are presented to validate the conclusions of the preceding performance analysis. Finally, we extend the results to the case of UEP coding for multiple (three or more) bit streams and show that a hybrid TDCM-SCM scheme will normally yield superior performance.

II. Problem Statement

To introduce the basic problem of UEP design, we initially consider the case of two bit streams. Let B_1 and B_2 denote, in decreasing order of importance, the two bit streams to be unequally protected against channel noise, and let r_1 and r_2 denote their respective rates. The rates are given in terms of bit rate normalized by the total number of modulation intervals[1] available for transmission of these bit streams. To specify the unequal protection requirements, let N_1 and N_2 denote the prescribed variances of the Gaussian noise that the respective bit streams need to withstand, where it is naturally assumed that, $N_1 > N_2$. That is, as long as the variance of the channel noise is less than N_i, the bit stream B_i can be decoded with a bit error rate below some prescribed value. For the sake of numerical calculations and simulations in this paper, we will assume that a bit error rate of about 10^{-3} is "acceptable." The problem can then be stated as: adjust the coding system parameters to minimize the average transmission energy, while providing the required level of protection.

There are two main categories of UEP coded modulation schemes, which we describe next.

A. Time-Division Coded Modulation

Time-division coded modulation (TDCM) is a scheme in which the bit streams B_1 and B_2 are transmitted on distinct modulation intervals. Bit stream B_i is transmitted over the fraction α_i of the available modulation intervals (where $\alpha_1 + \alpha_2 = 1$) using channel code Γ_i and transmission energy e_i. The rate of the code Γ_i is, therefore, r_i/α_i. The design problem is that of selecting the TDCM parameters (α_1, α_2) and (e_1, e_2) such that the prescribed levels of UEP are achieved while minimizing the average transmission energy per modulation interval: $\bar{e}_T = \alpha_1 e_1 + \alpha_2 e_2$.

[1] A modulation interval is the interval of time corresponding to the duration of one modulation symbol. Throughout this paper, we will assume transmission of two-dimensional modulation symbols on the inphase and quadrature phase components of a carrier.

B. Superposition Coded Modulation

Superposition coded modulation (SCM) consists of transmitting both bit streams on all the available modulation intervals using superposition of channel codes in the modulation space as described next. We choose constellations S_1 and S_2 with average energies e_1 and e_2, respectively. Bit stream B_i is encoded with channel code Γ_i and transmitted using constellation S_i. More specifically, the code Γ_i generates the codeword $\boldsymbol{x}_i$, which is a sequence of signal points $\{x_{ij}\}$, where $x_{ij} \in S_i$. The codewords $\boldsymbol{x}_1$ and $\boldsymbol{x}_2$ are superimposed and transmitted on the channel as $\boldsymbol{x} = \boldsymbol{x}_1 + \boldsymbol{x}_2$. We refer to Γ_1 and S_1 as the *outer code* and *outer constellation* respectively, and, Γ_2 and S_2 as *inner code* and *inner constellation*, respectively. At the decoding end, the received sequence is $\boldsymbol{y} = \boldsymbol{x} + \boldsymbol{n}$, where $\boldsymbol{n} = \{n_j\}$ is a sequence of iid Gaussian noise samples of variance N. If, $N_2 < N < N_1$, only the bit stream B_1 is reliably decoded. If $N < N_2$, then both bit streams are reliably decoded. The SCM design method aims at achieving the prescribed levels of protection for the two bit streams while minimizing the average transmission energy.

III. Performance Analysis of UEP Codes

A. Ideal Channel Coding

We first summarize the performance of TDCM and SCM assuming ideal channel coding. These results are due to the work of Bergmanns and Cover [1]. We have chosen, however, to present these results from a different angle such that they are directly extendible to the case where nonideal (practical) channel coding is employed. By ideal channel coding we mean that Shannon's channel coding bound is achieved. Hence, given Gaussian channel noise of variance N and transmission energy per modulation interval e, we can reliably transmit r bits per modulation interval where e, N, and r are related through

$$\frac{e}{N} = 2^r - 1. \tag{1}$$

1) TDCM: The bit stream B_i is transmitted over a fraction α_i of the modulation intervals using transmission energy e_i. The corresponding channel code has rate r_i/α_i. From (1), we obtain

$$e_i = N_i(2^{r_i/\alpha_i} - 1). \tag{2}$$

The resulting average transmission energy is therefore

$$\bar{e}_T = \sum_{i=1}^{2} \alpha_i N_i(2^{r_i/\alpha_i} - 1). \tag{3}$$

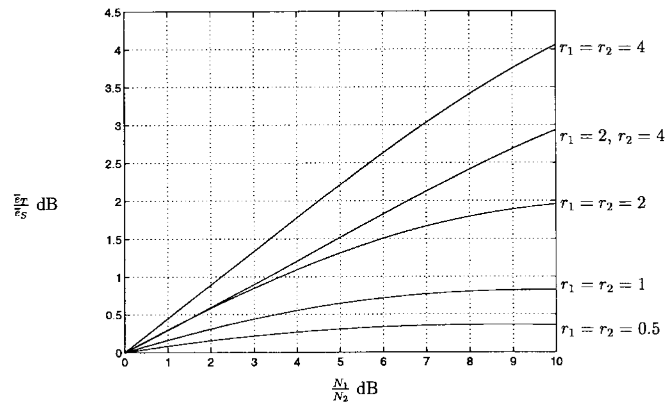

Fig. 1. The computed superposition gain $\bar{e}_T/\bar{e}_S$ versus the UEP ratio N_1/N_2. The plot depicts the ideal case with perfect channel coding (which achieves the Shannon bound).

Minimizing $\bar{e}_T$ over all possible choices of $\{\alpha_i\}$ determines the optimal design parameters and provides the best performance achievable by TDCM.

2) SCM: Here we transmit the bit streams on all available modulation intervals. The r_2 bits of B_2 are transmitted employing transmission energy e_2. Applying (1) to this case

$$e_2 = N_2(2^{r_2} - 1). \tag{4}$$

Bit stream B_1 needs to withstand, in addition to the channel noise of variance N_1, the effective "noise" introduced by the presence of the inner code. This effective noise is modeled as Gaussian with variance e_2 (see [1]). Plugging the overall noise into (1), we get for B_1

$$e_1 = (N_1 + e_2)(2^{r_1} - 1). \tag{5}$$

Finally, the average transmission energy of SCM is given by

$$\bar{e}_S = \sum_{i=1}^{2} N_i(2^{r_i} - 1) + N_2(2^{r_2} - 1)(2^{r_1} - 1). \tag{6}$$

3) Comparison: To compare the relative performance of TDCM and SCM, we consider the ratio $\bar{e}_T/\bar{e}_S$—the superposition gain. (Note that a higher value for this ratio implies that SCM requires less transmission energy relative to TDCM). It has been established in [1] that $(\bar{e}_T/\bar{e}_S) > 1$, that is, SCM always outperforms TDCM. It can be verified by inspection that for prescribed values of r_1 and r_2 the superposition gain only depends on the UEP ratio N_1/N_2. In Fig. 1, we plot the computed superposition gain against the UEP ratio for various values of (r_1, r_2). We observe that the performance gain of SCM over TDCM is pronounced at high spectral efficiencies[2] and is small at low spectral efficiencies.

An intuitive explanation of the above observations, which is particularly useful for the sequel, is the following. Let us first neglect the effective noise due to the inner code of SCM. That is, we neglect the last term in (6). Then the contribution of bit stream B_i to the average transmission energy is always larger for TDCM than for SCM. This follows from the easily verifiable inequality $\alpha_i(2^{r_i/\alpha_i} - 1) > (2^{r_i} - 1)$. We can therefore consider the ratio $(2^{r_i} - 1)/[\alpha_i(2^{r_i/\alpha_i} - 1)]$ as a quantitative measure of this advantage of SCM over TDCM for transmission of B_i. We refer to it as the *bandwidth expansion gain*. Next, we must take into account the effective noise due to the inner code. As a consequence of this effective noise, the average transmission energy required by SCM is increased by $(2^{r_1} - 1)(2^{r_2} - 1)N_2$, which we simply call the *performance loss due to effective noise*. Under the assumption of ideal channel coding, the bandwidth expansion gains outweigh the performance loss due to the effective noise, and SCM outperforms TDCM. In fact, this is strictly true when there is unequal protection; for equal protection, the gain and loss exactly cancel, and the two schemes provide identical performance. We note, finally, that the difference between the bandwidth expansion gain and the performance loss due to effective noise is more pronounced at higher spectral efficiencies, and this explains the increase in overall superposition gain at higher spectral efficiencies (that is, for larger values of r_1 and r_2).

In Section III-B, we will show that practical SCM does *not* always outperform TDCM. Moreover, wherever SCM is preferable, its performance advantage *decreases* with spectral efficiency.

B. Practical Channel Coding

We now re-analyze the performance of the TDCM and SCM schemes while taking into account the limitations of practical channel codes.

1) Modeling the Performance of Practical Channel Codes: We first describe a heuristic framework to model the performance of a practical channel code. This framework will be subsequently used to analyze the performance of practical UEP schemes.

Consider a code Γ that transmits r bits per modulation interval using transmission energy e. Let N be the maximum variance of the Gaussian noise per modulation interval that the code can withstand. If Γ were an ideal code—a code that achieves capacity—then the ratio e/N and rate r would satisfy (1), which we repeat here for convenience

$$\left(\frac{e}{N}\right)_{\text{ideal}} = 2^r - 1.$$

However, a practical channel code of rate r requires a value of e/N that is larger than $(e/N)_{\text{ideal}}$ by some factor which we denote by λ. Thus, we have

$$\frac{e}{N} = \lambda(2^r - 1). \tag{7}$$

We call λ the signal-to-noise ratio (SNR) loss factor of the code. The value of the SNR loss factor is related to the target bit error rate and the complexity of the code. In particular: 1) Given a fixed code, decrease in target bit error rate implies increase in λ; 2) given a fixed target bit error rate, codes with larger blocklength (or constraint length) typically have lower value of λ; and 3) for given code and target bit error rate (BER), λ is a constant that determines the relation between e and N.

[2] Spectral efficiency is commonly measured by the number of bits transmitted per modulation interval. Thus, r_1 and r_2 are a measure of spectral efficiency.

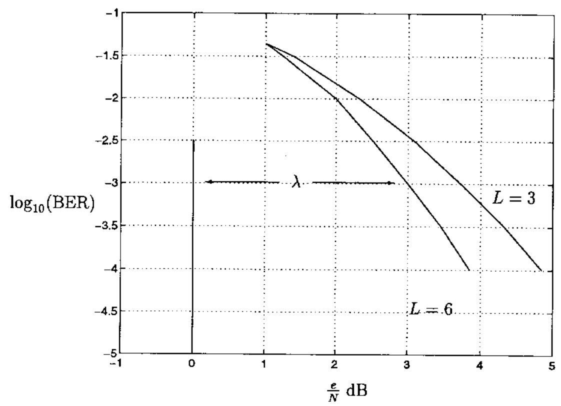

Fig. 2. The BER versus e/N for coding schemes with overall rate $r = 1$. The schemes consist of 4-QAM coded by rate $1/2$ convolutional codes of constraint lengths $L = 3$ and $L = 6$. The value of $(e/N) = 0$ dB required by an ideal channel code is marked for reference: the distance between this line and the curves gives the SNR loss factor λ (in decibels) at the prescribed BER.

Consider the example of coded modulation schemes with rate $r = 1$. In Fig. 2, we plot the BER versus the ratio e/N for two rate $1/2$ convolutional codes (of differing constraint length) in conjunction with 4-QAM. The value of $(e/N)_{\text{ideal}}$ is also shown in the figure for reference.[3] The (horizontal) difference in the values of e/N and $(e/N)_{\text{ideal}}$ gives in decibels the SNR loss factor (λ) of the code. Observe that the SNR loss factor increases significantly as we lower the target BER. Consider next the example of BER of 10^{-3}. The low complexity code (with constraint length 3) achieves a value of $\lambda = 2.4$ (3.8 dB), while the higher complexity code (of constraint length 6), results in a value of λ about 2 (3 dB). If one is willing to invest large complexity in the coding (and shaping) scheme, one can achieve a value of λ close to 1.

2) Performance of TDCM: Let λ_1 and λ_2 be the SNR loss factors of the codes Γ_1 and Γ_2, respectively. We obtain the transmission energy values e_i by rewriting (2) for practical channel codes as

$$e_i = \lambda_i N_i (2^{r_i/\alpha_i} - 1), \quad \text{for } i = 1, 2.$$

Given the prescribed values of (r_1, r_2) and (N_1, N_2), the optimal solution is obtained by choosing the pair (α_1, α_2) which minimizes

$$\overline{e}_T = \alpha_1 e_1 + \alpha_2 e_2 = \sum_{i=1}^{2} \alpha_i \lambda_i N_i (2^{r_i/\alpha_i} - 1). \tag{8}$$

[3] Strictly speaking, the value of $(e/N)_{\text{ideal}}$ cannot be used as a reference. This due to the fact that $(e/N)_{\text{ideal}}$ refers to probability of bit error equal to zero while each point on the performance curves corresponds to a specific nonzero value of BER (which is depicted on the vertical axis). To find the minimum (e/N) required to achieve transmission at rate r with probability of bit error ϵ, we proceed as follows. Using the results of distortion rate theory, we can show that the rate required to to code an equiprobable binary source of rate r, with probability of bit error ϵ, is $r' = r(1 - \mathrm{H_b}(\epsilon))$, where $\mathrm{H_b}(\epsilon)$ is the binary entropy for bit error probability of ϵ, e.g., [7]. The minimum value of (e/N) can now be found as $2^{r'} - 1$. However, for small values of ϵ, the difference between r' and r is negligible. Consequently, the difference between $2^{r'} - 1$ and $2^r - 1$ too can be ignored. In fact, for the range and scale shown, $(e/N)_{\text{ideal}}$ is practically indistinguishable from the correct curve.

Comparing (8) with the expression obtained from the ideal analysis (3), we see that λ_1 and λ_2 simply scale the respective terms in the average energy expression.

3) Performance of SCM: Consider employing codes Γ_1 and Γ_2 with SNR loss factors λ_1 and λ_2, as the outer and inner codes, respectively.

For the inner code, we obtain the value of the transmission energy e_2 by modifying (4) to

$$e_2 = \lambda_2 N_2 (2^{r_2} - 1). \tag{9}$$

The outer code Γ_1 needs to withstand the combined effect of channel noise of variance N_1 and the effective noise resulting from the presence of the inner constellation. In the ideal performance analysis of Section III-A, the effective noise due to the inner code was modeled as Gaussian with variance e_2. Here we propose to model the effective noise due to inner constellation as Gaussian, with variance N_{eff}. If shaping is employed for the inner code, the sequence of signal points transmitted by the inner code closely approximates Gaussian noise of variance e_2—the average transmission energy used by the code Γ_2. For this case, the value of variance N_{eff} can be expected to be e_2. Our experiments indicate that when we do not use shaping for the inner code the variance of the effective noise can be estimated to be a scaled version the value of e_2, that is, $N_{\text{eff}} = f e_2$, where f is a scaling factor whose value depends on the constellations S_1, S_2 and the UEP ratio N_1/N_2. That is, $f = f((N_1/N_2), S_1, S_2)$. Moreover, the value of f has to be determined empirically. It should be mentioned that f is almost always smaller than 1 and approaches the value of 1 for large UEP ratios.

We can now estimate the value of the transmission energy e_1, as follows: Since the outer code needs to withstand the combined effect of channel noise N_1 and the effective noise of variance $f e_2$, we have

$$e_1 = \lambda_1 [N_1 + f e_2](2^{r_1} - 1). \tag{10}$$

The resulting total transmission energy is given by

$$\overline{e}_S = \sum_{i=1}^{2} \lambda_i N_i (2^{r_i} - 1) + \lambda_1 \lambda_2 f N_2 (2^{r_1} - 1)(2^{r_2} - 1). \tag{11}$$

Comparing (11) and (6), we see that the first two terms in the expression for $\overline{e}_S$ are simply scaled by the SNR loss factors λ_1 and λ_2. However, the third term that corresponds to the performance loss due to effective noise is scaled by $f\lambda_1\lambda_2$. Note that, for the case of TDCM, each term in the average energy expression was scaled by a single SNR loss factor. However, for the case of SCM, the term corresponding to performance loss due to the effective noise, is scaled by two SNR loss factors (and f). It is therefore possible that while in the ideal coding case SCM's performance advantage over TDCM only vanished at the limit of equal protection, in the practical coding case, depending on the values of f, the

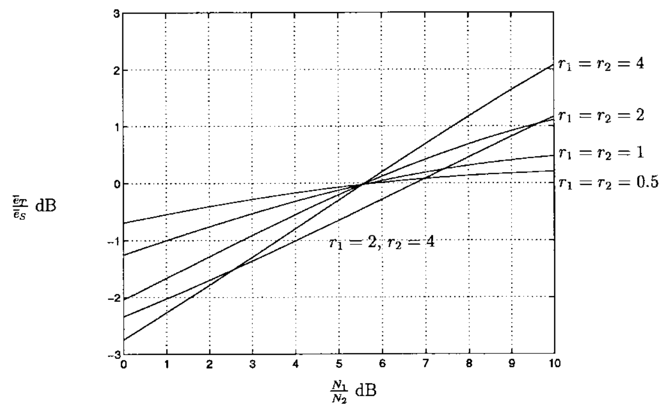

Fig. 3. The computer superposition gain $\overline{e}_T/\overline{e}_S$ versus the UEP ratio N_1/N_2 for the case of nonideal channel coding. The computation assumes SNR loss factors of 2.0.

"crossover point" may not be at the limit, but rather at some distinct level of unequal protection.

It is clear that, to perform a meaningful comparison of the performance of TDCM and SCM, we require estimates on the value of the scaling factor f. However, some insights into the relative performance of TDCM and SCM can still be obtained by using the representative value of $f = 1$. The corresponding plots of $\overline{e}_T$ and $\overline{e}_S$ are shown in Fig. 3. Compared to the ideal results depicted in Fig. 1, we see that for moderate values of the UEP ratio TDCM outperforms the SCM scheme. Only at high values of the UEP ratiois the bandwidth expansion gain large enough to overcome the amplified performance loss due to effective noise. At these high UEP ratios, SCM indeed outperforms TDCM. It can also be seen from the plots that, at low spectral efficiencies, the performance advantages of superposition, which emerge at high levels of unequal protection, are quite small and may be outweighed by the appealing simplicity of TDCM.

We have seen above that if we compute the average energy required by SCM and TDCM using $f = 1$, there exists a range of values for the UEP ratio for which TDCM outperforms SCM. The location of the true crossover point is, however, affected by the values taken by the scaling factor f. In fact, since f is almost always smaller than 1, it shifts the crossover point location to the left. One might wonder if, in practice, the values taken by f can be sufficiently small to push the location of the crossover point to $(N_1/N_2) = 0$ dB. Design examples demonstrate that this is not the case.

In summary, we have demonstrated that, in practice, TDCM may be superior to SCM. In particular, this is likely to be the case at low to moderate values of the UEP ratio. Moreover, at low spectral efficiencies, TDCM may always be preferable over SCM, owing to its implementation simplicity and marginal performance loss. These are important realizations because the known asymptotic superiority of SCM motivated much emphasis on SCM in recent research on UEP.

Remark: It is tempting to extend the analysis framework described here to study the relation between the location of the crossover point and the values of r_1 and r_2. Such a study will, however, require estimates on the value of the scaling factor f. Since such estimates have to be obtained empirically, we feel that the actual location of the crossover point should be determined via the actual design of UEP codes.

IV. Decoding Operation and Overall Design of SCM

A. Decoding the Outer Code Alone

As described in Section II-B, the SCM encoder selects codewords $\boldsymbol{x}_1 = \{x_{1j}\}$ from code Γ_1 and $\boldsymbol{x}_2 = \{x_{2j}\}$ from Γ_2. Their sum $\boldsymbol{x} = \boldsymbol{x}_1 + \boldsymbol{x}_2$ is transmitted on the channel. We receive $\boldsymbol{y} = \boldsymbol{x} + \boldsymbol{n}$, where $\boldsymbol{n} = \{n_j\}$ is a sequence of iid Gaussian noise of variance N. (Recall that the value of channel noise variance N, is known to the decoder.) In this subsection, we consider SCM decoding when the level of the channel noise is such that only the outer code can be decoded reliably, that is, we have $N_2 < N < N_1$.

The optimal maximum likelihood decoding of the outer code searches for the codeword $\boldsymbol{x}_1 \in \Gamma_1$ that maximizes the likelihood $p(\boldsymbol{y}|\boldsymbol{x}_1)$. The likelihood can be evaluated as

$$p(\boldsymbol{y}|\boldsymbol{x}_1) = \sum_{\boldsymbol{x}_2 \in \Gamma_2} p(\boldsymbol{x}_2) p(\boldsymbol{y}|\boldsymbol{x}_1 + \boldsymbol{x}_2). \tag{12}$$

The complexity involved in incorporating the precise knowledge of the inner code in terms of the summation over all the codewords $\boldsymbol{x}_2 \in \Gamma_2$ in (12) is often prohibitive. To make the complexity manageable, we resort to the following suboptimal decoding procedure. As in multistage decoding [6], we assume that all sequences of signals from S_2 are equally likely. Thus, we can write

$$p(\boldsymbol{y}|\boldsymbol{x}_1) = \theta \prod_j \sum_{s_2 \in S_2} \exp\left(-\frac{\|y_j - x_{1j} - s_2\|^2}{2N}\right) \tag{13}$$

where θ is a normalizing constant and $\|\cdot\|^2$ is the squared Euclidean norm. Consequently, the decoding rule becomes

$$\hat{\boldsymbol{x}}_1 = \arg\max_{\boldsymbol{x}_1} \sum_j \log \sum_{s_2 \in S_2} \exp\left(-\frac{\|y_j - x_{1j} - s_2\|^2}{2N}\right) \tag{14}$$

where $\hat{\boldsymbol{x}}_1$ denotes the estimate of $\boldsymbol{x}_1$.

Remark: If more than two, say M, codes are superimposed, we can generalize (14) as follows. We transmit on the channel the sum of M codewords $\boldsymbol{x} = \boldsymbol{x}_1 + \cdot + \boldsymbol{x}_i + \cdot + \boldsymbol{x}_M$, where $\boldsymbol{x}_i$ is a codeword in the ith code Γ_i and consists of a sequence of signal points from the constellation S_i. That is, $\boldsymbol{x}_i = \{x_{ij}\}$, where $x_{ij} \in S_i$. On receiving $\boldsymbol{y} = \boldsymbol{x} + \boldsymbol{n}$, we decode $\boldsymbol{x}_1$ as

$$\hat{\boldsymbol{x}}_1 = \arg\max_{\boldsymbol{x}_1} \sum_j \log \sum_{s_2 \in S_2, \cdots, s_M \in S_M} d(y_j, x_{1j}, s_2, \cdots, s_M) \tag{15}$$

where

$$d(y_j, x_{1j}, s_2, \cdots, s_M) = \exp\left(-\frac{\|y_j - x_{1j} - s_2 - \cdots - s_M\|^2}{2N}\right).$$

B. Decoding Both Outer and Inner Codes

Let the level of channel noise N be small enough to allow reliable decoding of both the inner and outer codes, i.e., $N < N_2$. For this case, we consider two variants of the multistage decoding technique, e.g., [6].

Multistage Decoding Method 1: This is the well-known technique employed in decoding of multilevel codes (e.g., [6]). The objective is to find a pair of codewords $(\boldsymbol{x}_1, \boldsymbol{x}_2)$ such that the probability $p(\boldsymbol{y}|\boldsymbol{x}_1, \boldsymbol{x}_2)$ is maximized. This would, in principle, be achieved by minimizing the square distance

$$(\hat{\boldsymbol{x}}_1, \hat{\boldsymbol{x}}_2) = \arg \min_{\boldsymbol{x}_1, \boldsymbol{x}_2} \sum_j ||y_j - x_{1j} - x_{2j}||^2. \quad (16)$$

Since this operation is too complex, we approximate it in two steps:

$$\hat{\boldsymbol{x}}_1 = \arg \min_{\boldsymbol{x}_1} \sum_j \min_{s_2 \in S_2} ||y_j - x_{1j} - s_2||^2 \quad (17)$$

and

$$\hat{\boldsymbol{x}}_2 = \arg \min_{\boldsymbol{x}_2} \sum_j ||y_j - \hat{x}_{1j} - x_{2j}||^2. \quad (18)$$

Multistage Decoding Method 2: Here, instead of estimating $\hat{\boldsymbol{x}}_1$ using (17), we employ the distance measure (14). Subsequently, we use the estimate of $\hat{\boldsymbol{x}}_1$ in (18) to decode the codeword $\boldsymbol{x}_2$.

In all our experiments, whenever the noise level N was small enough to allow decoding of both the outer and the inner codes, the decoding errors produced in estimating $\hat{\boldsymbol{x}}_1$ by both methods were negligible. (Of course, this observation does depend on the assumed level of "acceptable" BER, but note that the value of 10^{-3} is relatively high). We chose to use method 2.

Remarks:

1) When the level of channel noise N is such that only the outer code can be reliably decoded, we found that the estimate of $\hat{\boldsymbol{x}}_1$ obtained using (14) is somewhat better than the one obtained using (17).
2) Using the generalization of the decoding procedure for the outer code described in (15), method 2 can be extended in a direct manner to decode multiple superimposed codes.

C. Optimized Superposition of Constellations

An important question regarding the design of SCM schemes is this: how can we superimpose constellations S_1 and S_2 so as to minimize the effective noise? The distance properties of the inner code are unchanged if the constellation S_2 is rotated about its center prior to superposition. However, this rotation of S_2 can influence the value of the effective noise due to the inner code.

One approach is based on maximizing the effective minimum distance (EMD) of the outer code, which is defined as follows. Let $(\boldsymbol{x}_1, \boldsymbol{x}_1')$ be any two distinct code sequences of the code Γ_1. We define the EMD of the outer code d_{out}^2 as

$$d_{\text{out}}^2 = \min_{\boldsymbol{x}_1 \neq \boldsymbol{x}_1'} \sum_j \min_{s_2, s_2' \in S_2} ||(x_{1j} + s_2) - (x_{1j}' + s_2')||^2. \quad (19)$$

The optimization consists of evaluating the outer code EMD for various rotations of S_2 and selecting the rotation that maximizes it.

A major drawback of this EMD based approach is its focus on the distance of the nearest codeword while neglecting the effect of other neighboring codewords. In the context of equal error protection channel coding, it is known that minimum distance is an important characteristic of channel codes, yet it does not always provide sufficient information. Moreover, for the case of SCM, and especially the outer code, EMD may be a very poor measure of performance, as we illustrate next. Consider an SCM scheme consisting of a 4-QAM outer constellation, with energy $e_1 = 4$, coded by a constraint length of 6, rate $1/2$ convolutional code, while the inner constellation is 4-QAM with energy $e_2 = 1$, and coded by the same channel code. (The precise channel code employed for the inner code is irrelevant to this discussion and is only mentioned for completeness.) With this scheme, the outer code can withstand channel noise of variance $N_1 = 1.0$. The EMD of the outer code for this example is $d_{\text{out}}^2 = 28$. Now it is possible to simultaneously reduce the value of e_1 and e_2 in a way that keeps d_{out}^2 constant. However, as a result of this reduction in value of e_2, more and more codewords are at a distance equal to or close to 28. Consequently, the level of channel noise that the outer code can truly withstand is decreased. At the limiting case of $e_1 = 1$ and $e_2 \approx 0$, although $d_{\text{out}}^2 = 28$, the noise that the outer code can actually withstand is reduced to $N_1 = 0.5$—a reduction by a factor of 2 compared to the initial configuration. This, admittedly contrived, example clearly illustrates that restricting attention only to the EMD of the outer code is not the correct strategy. We need to account for the distance as well as the number of neighboring codewords. We achieve this by employing the channel capacity measure, as described next.

The capacity of the channel available for transmission of bit stream B_1 using constellations S_1 in the presence of channel noise N and constellation S_2 is given by [5]

$$C_{\text{out}}(N) = h[s_1 + s_2 + N] - h[s_2 + N] \quad (20)$$

where $h(\cdot)$ denotes the differential entropy and s_1 and s_2 are random variables that are uniformly distributed over the constellations S_1 and S_2. Since we are interested in protecting bit stream B_1 against channel noise N_1, we attempt to maximize $C_{\text{out}}(N_1)$. In a straightforward manner we evaluate $C_{\text{out}}(N_1)$ for different rotations of the constellation S_2 and choose the one which maximizes the capacity.

Choosing an appropriate rotation of the inner constellation can contribute to improving the design. We demonstrate this with the aid of the following examples. First consider the case where S_1 is a scaled version of the 8-point signal constellation shown in Fig. 8(a), while S_2 is a scaled version of the 32-point constellation shown in Fig. 8(b). In particular, we consider the constellations with average transmission values $e_1 = 84.9$ and $e_2 = 19.9$. Evaluation of the capacity at noise value of $N_1 = 1.0$ shows that rotation of constellation S_2 by 45 degrees (which is the near optimal rotation) achieves a capacity increase that translates to increasing the value of e_1 by about 0.4 dB. Next consider the case when both S_1 and S_2 are scaled versions of the 8-point constellation of Fig. 8(a), with energy values $e_1 = 25.6$ and $e_2 = 4.1$. In this case, evaluation of the capacity at a noise value of $N_1 = 1.0$ shows that rotation of constellation S_2 by 45 degrees (which is again the near optimal rotation) achieves a capacity increase that translates to raising the value of e_1 by about 0.1 dB. That is, for this case, choosing the correct rotation may not be very important, while for the former example it yields significant overall performance improvements.

To summarize, in view of the significant drawback of EMD, we deviate from the approach taken in most earlier work on UEP coding (e.g., [4], [3]), which measures the performance of the outer code by its EMD. Instead, we have chosen to quantify the performance in terms of the level of noise that the code can withstand while achieving a BER below the prescribed value.

D. Design Method

The SCM design procedure is summarized as follows.

1) Choose an inner constellation S_2 and a code Γ_2 to achieve a BER below the prescribed value level for bit stream B_2. This can be achieved by choosing a standard trellis code of rate r_2 bits with affordable complexity. The choice of code determines the value of the SNR loss factor λ_2. The level of transmission energy e_2 is determined by λ_2 and r_2 via

$$e_2 = \lambda_2 N_2 (2^{r_2} - 1).$$

2) Choose an outer code Γ_1 of rate r_1, along with the corresponding constellation S_1. Let λ_1 be the SNR loss factor for the outer code. Using the approximation that the effective noise is Gaussian with variance e_2, allocate transmission energy e_1 to the outer code such that it can tolerate a noise of variance $N_1 + e_2$. Thus, the transmission energy e_1 is related to $N_1 + e_2, r_1,$ and λ_1 via

$$e_1 = \lambda_1 (N_1 + e_2)(2^{r_1} - 1).$$

3) Rotate the inner constellation to maximize the capacity $C_{\text{out}}(N_1)$ given by (20).
4) Test the performance of the outer code by simulation. If the BER is below the prescribed value, reduce e_1. If the error rate is above the prescribed value, increase e_1.
5) Repeat steps 3)-5) until the prescribed value for the BER is obtained.

It should be mentioned that step 3) can be ignored in all but the first iteration of SCM design. This is due to the fact that, in subsequent iterations, the value of the transmission energy e_1 changes by small amounts. Consequently, there is little to gain by further optimization of the rotation.

V. Design Results

A. Codes for High Spectral Efficiency

Example A: We designed SCM schemes to provide UEP to bit streams with $r_1 = r_2 = 2$. The codes Γ_1 and Γ_2 used in the design were 16-state nonrecursive trellis codes.

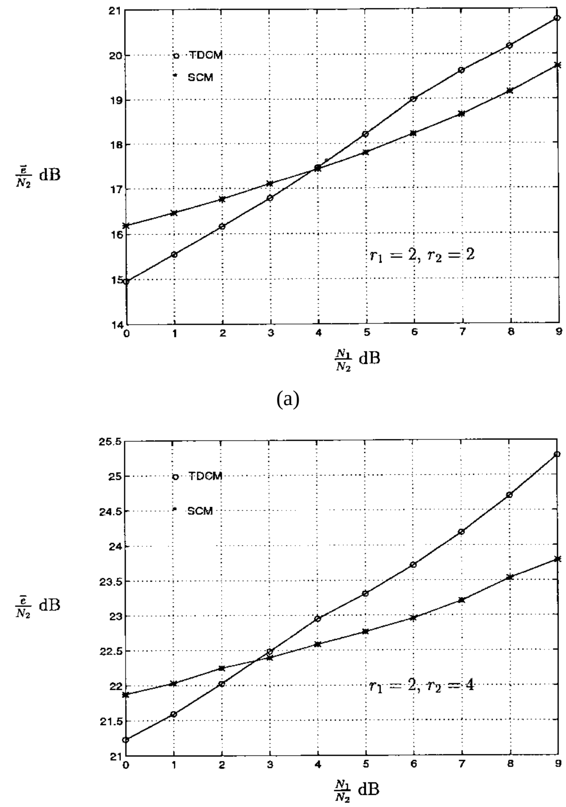

Fig. 4. Average transmission energy ($\bar{e}$) required by the different UEP schemes for high-spectral-efficiency design examples. (a) Rates: $r_1 = 2$, $r_2 = 2$. (b) Rates: $r_1 = 2$, $r_2 = 4$.

The block diagram of a nonrecursive trellis encoder with two coded bits is shown in Fig. 7, and the specific generator vectors used in this example are given by $g_0 = (0,0,1,0,0,0)$, $g_1 = (1,1,1,1,0,1)$, and $g_2 = (0,1,0,0,1,1)$. The signal constellations S_1 and S_2 were scaled versions of the 8-point signal constellations depicted in Fig. 8. The level of average transmission energy $\bar{e}_S$ required by the SCM scheme varies with the UEP ratio (N_1/N_2) as shown in Fig. 4(a). As a reference, we also included a plot of $\bar{e}_T$, the average transmission energy required by the TDCM design. To construct the TDCM scheme, we selected a collection of pairs of codes with rates $((r_1/\alpha_1), (r_2/\alpha_2))$ such that $\alpha_1 + \alpha_2 = 1$. The pair that required the smallest level of $\bar{e}_T$ for the prescribed UEP ratio was chosen.

We can see that TDCM outperforms SCM at low UEP ratios, and the "cross-over" point is at about UEP ratio = 4 dB.

The value of variance of the effective noise N_{eff} was estimated for these designs as follows. The performance of the outer code was evaluated in the absence of the inner code. For this case, let $\tilde{N}$ denote the level of noise that the outer code can withstand. The difference $\tilde{N} - N_1$ can be considered as the estimate of N_{eff}. A plot of the scaling factor $f = (N_{\text{eff}}/e_2)$ is shown in Fig. 6. We can see that the values of f are in the range of 0.7–0.85 and increase monotonically with the UEP ratio.

Example B: Here we consider the design of SCM and TDCM unequal protection schemes for the case of unequal

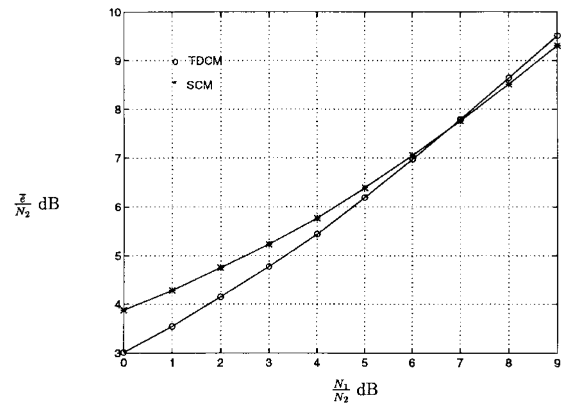

Fig. 5. Average transmission energy ($\bar{e}$) required by the different UEP schemes for a low-spectral-efficiency design example, with $r_1 = r_2 = 0.5$.

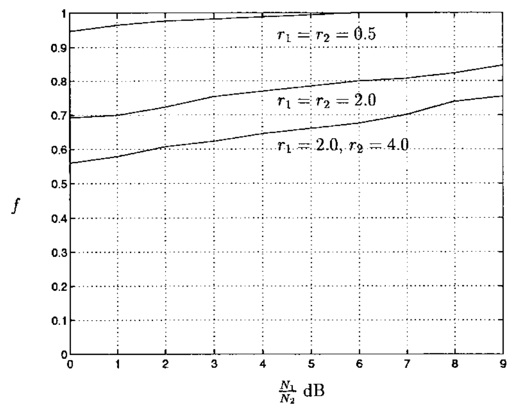

Fig. 6. A plot of how the scaling factor $f = N_{\text{eff}}/e_2$ varies with the UEP ratio N_1/N_2 for different design examples.

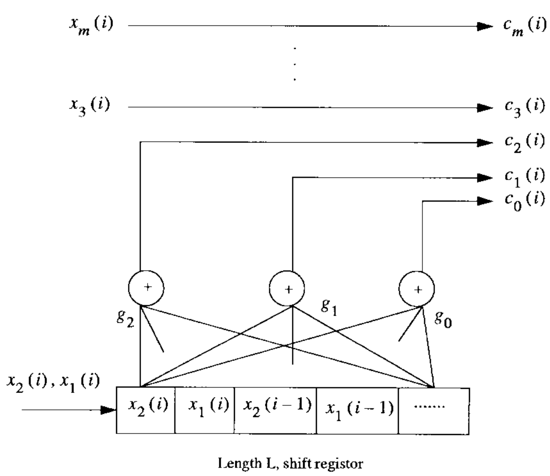

Fig. 7. Encoder of a nonrecursive trellis code with two coded bits. The m input bits $(x_m, \cdots, x_1)$ are mapped to $m+1$ label bits $(c_m, \cdots, c_1, c_0)$. These label bits select a signal point from a signal constellation of size 2^{m+1} labeled via Ungerboeck's set partitioning [8]. g_2, g_1, and g_0 are binary length L generator vectors that specify the particular trellis code.

rates for the two bit streams. Specifically we designed SCM and TDCM schemes for the rates $r_1 = 2$ and $r_2 = 4$. Again, all the codes used in the design were 16-state trellis codes with generator vectors given in example A. The constellations S_1 and S_2 were scaled versions of the 8-point and 32-point

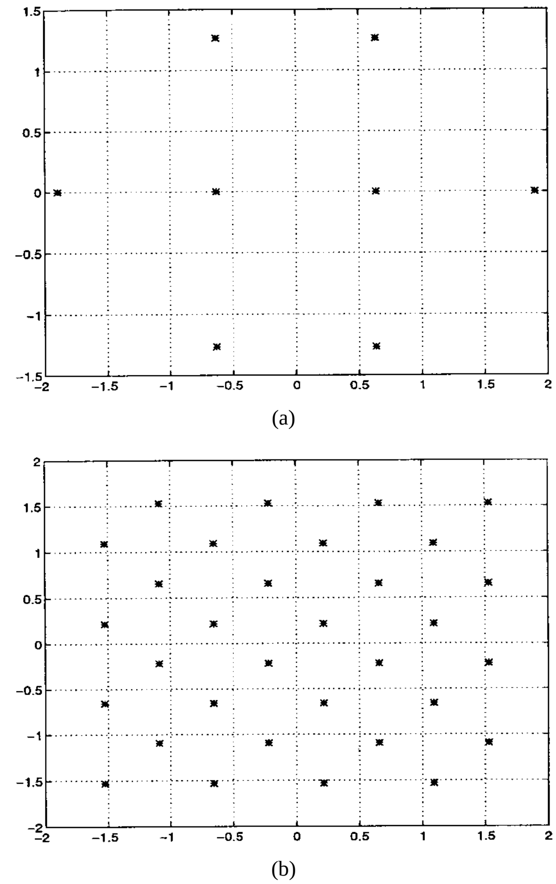

Fig. 8. (a) Eight-point and (b) 32-point signal constellations used in design examples.

signal constellations depicted in Fig. 8. The average energy required by the SCM scheme $\bar{e}_S$ and the TDCM scheme $\bar{e}_T$ is compared in Fig. 4(b). In this case, we see that for a UEP ratio in the approximate range of 0–3 dB, TDCM outperforms SCM. The variation in the value of the scaling factor f with the UEP ratio is also depicted in Fig. 6. Here the values of f lie between 0.57 and 0.75.

B. Unequal Protection at Low Spectral Efficiency

To compare the performance of SCM and TDCM for unequal protection at low spectral efficiencies, consider the design results for the case of $r_1 = r_2 = 1/2$. All the codes used in the design were constraint length-6 convolutional codes coupled with 4-QAM. Fig. 5 shows the average transmission energy required by the SCM and TDCM designs as functions of the UEP ratio. It is evident that at such low spectral efficiencies, SCM has little or no performance advantage over TDCM. For this example, the value of the scaling factor f, is very close to 1, as shown in Fig. 6.

Some complexity aspects should not be ignored. At low spectral efficiencies, TDCM schemes often involve binary modulation (4-QAM) based channel codes. Superposition schemes, however, require the use of multilevel modulation since the superposition of constellations S_1 and S_2 results in a multilevel constellation even if S_1 and S_2 themselves are binary constellations. Demodulating binary constellations is simpler as it does not require accurate estimates of the magnitude of the received signal. Moreover, even when SCM outperforms TDCM, the performance margin is small and may not justify the additional complexity. It is thus reasonable to conclude that under these circumstances it is preferable to use TDCM irrespective of the level of UEP.

VI. UEP Codes for Three or More Bit Streams

We now extend the ideas presented in the earlier sections to the case of unequal protection of more than two bit streams. The basic problem can be summarized as follows: let $B_1, B_2, \cdots, B_M$ be the M bit streams that we desire to protect unequally. Let r_i denote the rate (normalized by the total modulation intervals) of the bit stream B_i and let N_i be the noise level that we wish to protect the bit stream against. Our aim is to achieve the prescribed levels of protection while minimizing the average transmission energy required by the coding technique.

A straightforward extension of our results would be to compare TDCM with SCM schemes for transmission of M bit streams. However, for $M > 2$, there is a third alternative, namely, partial superposition coded modulation (PSCM). This is a hybrid of the two extremes of TDCM and SCM. We will demonstrate here that PSCM is often superior to both TDCM and SCM schemes.

A. Partial Superposition Coded Modulation

Consider the following strategy. We first group the M bit streams into M' sets, where $M' \leq M$. Each set of bit streams is transmitted on exclusive modulation intervals. Specifically, the kth set is transmitted on a fraction β_k of the modulation intervals, where $\Sigma\beta_k = 1$. Within each set, the bit streams are transmitted using superposition. Thus, the overall system is a hybrid between the two extremes of pure SCM and pure TDCM. Note in particular that the special cases of $M' = 1$ and $M' = M$ correspond to "pure" SCM and TDCM, respectively.

To design a PSCM scheme, we need to group the M bit streams into M' sets and assign a fraction β_k of modulation intervals to each set. Subsequently, we have to design appropriate channel codes for the sets of bit streams. We have already described in Section IV-D the method for designing superposition codes for two bit streams. The design techniques of Section IV-D are extendible to more bit streams in a straightforward manner. Using the extended design method, we design PSCM schemes for different groupings of bit streams and allocations of modulation intervals. The design which yields the smallest average transmission energy is selected.

Remark: Designing PSCM schemes for several groupings of bit streams can be a computationally demanding procedure. The design burden can, however, be largely reduced as follows. Using the framework developed in Section III-B for performance analysis of practical channel codes, we can estimate the performance achievable by different groupings of bits streams (with the representative value of $f = 1$ during calculation). These estimates allow us to restrict the search to a small number of groupings that are the most promising candidates. Hence, the actual code design can be performed only for these selected bit stream groupings. It is in particular expected that a configuration where bit streams with small UEP ratios are transmitted via superposition will be quickly eliminated in favor of configurations where bit streams with large UEP ratio are transmitted using SCM.

TABLE I
AVERAGE TRANSMISSION ENERGY REQUIRED BY THE UEP SCHEMES FOR THE CASE OF THREE BIT STREAMS AT DIFFERENT UNEQUAL PROTECTION LEVELS

	$r_1 = r_2 = r_3 = 2.0$			$r_1 = r_2 = r_3 = 0.33$		
$\frac{N_1}{N_3}$ dB	8	12	16	8	12	16
$\frac{N_2}{N_3}$ dB	4	6	8	4	6	8
$\frac{\overline{e}_P}{N_3}$ dB	24.6	26.6	29.0	8.1	11.2	14.6
$\frac{\overline{e}_S}{N_3}$ dB	25.7	27.0	28.9	8.3	11.3	14.6
$\frac{\overline{e}_T}{N_3}$ dB	26.4	29.7	33.2	8.2	11.5	15.0

$\overline{e}_P$, $\overline{e}_T$, and $\overline{e}_S$ denote the transmission energies of PSCM, TDCM, and SCM, respectively.

B. Design Examples and Comparisons

We illustrate the relative performance of PSCM, SCM, and TDCM with the following examples.

1) High Spectral Efficiency:

Example A: We designed the UEP schemes for the case of three bit streams with $r_1 = r_2 = r_3 = 2.0$. Again, all the codes used in the design of SCM, TDCM, and PSCM were 16-state trellis codes described in Section V. The corresponding results are summarized in Table I. One must keep in mind that SCM and TDCM are special (extreme) cases of PSCM. From Table I, we observe that, for a large range of unequal protection ratios N_1/N_3 and N_2/N_3, the optimal solution is almost never at these extremes.

The superiority of PSCM over TDCM is not surprising and can easily be explained as follows: consider the TDCM scheme for bit streams r_1, r_2, and r_3. For all the cases under consideration, we know from the results of Section V-A that the ratio N_1/N_3 is high enough to benefit from transmitting r_1 and r_3 via superposition, rather than transmission on distinct modulation intervals. Thus, a PSCM scheme, where we transmit r_1 and r_3 via superposition on all the modulation intervals allocated to them by TDCM, must outperform TDCM. The best PSCM scheme will, of course, perform at least as well as this specific PSCM scheme. We also observe the superiority of PSCM over the other limiting special case, SCM, in most of the tests. This result can be interpreted and explained with the aid of the SNR loss factors as in Section III-B. In the case of two bit streams, the transmission energy required by the outer code was scaled by its loss factor λ_1 and indirectly by the inner code loss factor λ_2. With more bit streams, the loss is exacerbated as more loss factors impact the level of the transmission energy required by outer and intermediate codes. Specifically, the code Γ_1 needs to withstand the effective noise due to the two codes Γ_2 and Γ_3, while the code Γ_2 experiences the effective noise due to Γ_3. Consequently, the transmission energy e_1 is scaled directly by λ_1 and indirectly by λ_2 and λ_3, while the transmission energy e_2 is scaled directly by λ_2 and indirectly by λ_3. It is, therefore, expected that PSCM will outperform SCM in most circumstances, with the exception

TABLE II
AVERAGE TRANSMISSION ENERGY REQUIRED BY THE UEP SCHEMES FOR THE CASE OF FOUR BIT STREAMS WITH $r_1 = r_2 = r_3 = r_4 = 2.0$

$\frac{N_1}{N_4}$ dB	12	18	24
$\frac{N_2}{N_4}$ dB	8	12	16
$\frac{N_3}{N_4}$ dB	4	6	8
$\frac{\bar{e}_P}{N_4}$ dB	32.1	35.2	39.0
$\frac{\bar{e}_S}{N_4}$ dB	33.7	35.5	38.2
$\frac{\bar{e}_T}{N_4}$ dB	35.3	40.2	43.6

$\bar{e}_P$ $\bar{e}_T$, and $\bar{e}_S$ denote the transmission energies of PSCM, TDCM, and SCM, respectively.

of unequal protection ratios N_1/N_3 and N_2/N_3, which are sufficiently high for the bandwidth expansion gain to overcome the performance loss due to the effective noise.

Example B: Here we consider providing UEP to four bit streams with $r_1 = r_2 = r_3 = r_4 = 2.0$. As in the previous example, all the codes used in the design were 16-state trellis codes. The performance of the different UEP schemes is listed in Table II. We again see that, for a wide range of UEP ratios N_1/N_4, N_2/N_4, and N_3/N_4, PSCM outperforms SCM and TDCM by a substantial margin.

2) Design Example 2—Low Spectral Efficiency: Here we compare the performance of the different UEP schemes at low spectral efficiencies. We consider the example of $r_1 = r_2 = r_3 = 0.33$. The overall transmission energy required by different UEP schemes for this case is also tabulated in Table I. All the codes employed in the different UEP schemes are convolutional codes with constraint length of 6 coupled with 4-QAM constellations. As expected, PSCM almost always outperforms SCM and TDCM. However, as noted in the design example of Section V-B, TDCM with binary coded modulation may have significant implementation advantages over the other schemes which employ nonbinary modulation. The ease of implementation of TDCM should be weighed against the performance gains of PSCM, especially when these gains are small.

VII. SUMMARY AND DISCUSSION

In this paper, we considered the design of UEP coded modulation schemes. The ideas were initially developed for the case of two bit streams. UEP modulation schemes can be classified into two categories: 1) superposition coded modulation (SCM) and 2) time-division coded modulation (TDCM). An important known result states that under ideal channel coding the performance of SCM is superior to that of TDCM. The most recent research on UEP was therefore focused on SCM. However, we have shown here that when one accounts for the fact that practical channel codes do not achieve capacity, then this theoretical result does not hold in general, and there exist important cases where TDCM outperforms SCM. In particular, this happens whenever the UEP ratio (which measures the degree of inequality in protection) is below a critical value. To complement the straightforward design of TDCM, we developed methods to design and decode superposition codes. Various simulation results on design examples validate the conclusions obtained from performance analysis. Finally, we extended the approach to UEP coding for multiple (more than two) bit streams. It is not a trivial extension because, beside TDCM and SCM, it involves an entire class of hybrid coding schemes, which we named partial superposition coded modulation (PSCM). This class, in fact, includes pure SCM and pure TDCM as limiting special cases. We demonstrated that, most often, a hybrid PSCM scheme achieves the best performance.

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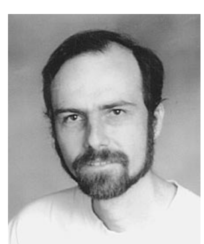

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