

Multiresolution Broadcast for Digital HDTV Using Joint Source/Channel Coding

Kannan Ramchandran, Antonio Ortega, K. Metin Uz, and Martin Vetterli, *Senior Member, IEEE*

Abstract—The use of multiresolution (MR) joint source-channel coding in the context of digital terrestrial broadcasting of high definition television (HDTV) is shown to be an efficient alternative to traditional single-resolution techniques. While the single-resolution schemes suffer from a sharp threshold effect in the fringes of the broadcast area, we show how a matched multiresolution approach to both source and channel coding can provide a stepwise graceful degradation. Furthermore, this multiresolution approach improves the behavior, in terms of coverage and robustness of the transmission scheme, over systems that are not specifically designed for broadcast situations. This paper examines the alternatives available for multiresolution transmission through embedded modulation and error correction codes. We also show how multiresolution trellis-coded modulation (TCM) can be used to increase coverage range. We present coding results and simulations of noisy transmission. From a systems point of view, we also discuss the tradeoffs involved in the choice of the low- and high-resolution coverage areas as well as the comparative costs and complexities of the different multiresolution transmission alternatives.

I. INTRODUCTION

A. General Discussion of the Problem

RECENT advances in video compression techniques have spurred interest in the idea of digital HDTV. Even the most demanding delivery mechanism, namely terrestrial broadcast, might turn digital. Digital broadcast differs from digital point-to-point transmission in that different receivers have channel capacities which decrease with distance from the emitter. Furthermore, in a digital environment the transition from reliable to unreliable reception is very abrupt, creating the so-called threshold effect. Hence, if digital broadcast is tackled as a single-resolution (SR) problem, one would in effect be designing for the fringes of the coverage area, thus reducing the spectral efficiency in areas close to the emitter, as pointed out in [1]. In light of the current interest in digital terrestrial broadcast of HDTV in the U.S., the concern for *spectral efficiency* becomes even more pressing, especially given the conditions set by the FCC in terms of bandwidth allocation.

Manuscript received June 1992; revised August 1992. This was supported in part by the New York State Science and Technology Foundation's CAT, the Fulbright Commission, the Ministry of Education of Spain, and the National Science Foundation under Grants ECD-88-11 111 and MIP-90-14 189.

K. Ramchandran, A. Ortega, and M. Vetterli are with the Department of Electrical Engineering and Center for Telecommunications Research, Columbia University, New York, NY 10027-6699.

K. M. Uz is with the David Sarnoff Research Center, Princeton, NJ 08543. IEEE Log Number 9204422.

The approach of designing for the fringes is known from information theory to be suboptimal. When dealing with different channels, one can do better than to transmit only for the worst one or to perform "naive" time or frequency multiplexing between the different channels! Cover [2] showed that one could trade off channel capacity from the poor channels to the better ones, and that this tradeoff can in theory be worthwhile. These ideas point out the efficiency of using a multiresolution approach to digital broadcast. However, to the best of the authors' knowledge, no real end-to-end system has been designed using these results.

We approach this problem as one of joint source and channel coding in a *multiresolution (MR) framework*, extending our work of [3]. In the two-resolution case (see Fig. 1), the source is split into "base" information, the coarse channel, and "refinement" information, the fine channel.¹ As in Fig. 1, the idea is to match the different resolution levels to different channel capacities, creating an *MR channel coding scheme* in which the receiver closer to the emitter can decode the full quality signal while the distant receiver has access to the lower-resolution quality, thus providing a stepwise graceful degradation. Furthermore, we show that the use of error concealment in the source decoder of an MR system (see Fig. 1) improves the robustness of the full-resolution signal, thus increasing the coverage of "indistinguishable quality" delivery over SR schemes.

We explore the available alternatives to an embedded transmission design and show how MR modulation schemes, combined with trellis coded modulation (TCM) techniques, can be used for this purpose while pointing out the relative difficulty of designing efficient error correction codes (ECC's) to solve the same problem. We consider, in our experiments, a specific high-quality MR HDTV coder [4] whose coarse-to-refinement channel bit rates are in the ratio of 1 : 2. We assume a spectral efficiency of 6 bits/symbol for our specific example; though, depending on the available broadcast bandwidth, other scenarios may use 3–4 bits/symbol. We evaluate the performance of the system in terms of both coverage area and subjective quality.

B. Past and Current Work

Most proposals to the FCC for digital terrestrial broadcast in the U.S. initially approached the problem as one of point-to-point transmission. The idea of graceful degradation,

¹Note that, throughout this paper, we use "coarse" synonymously with the lower-resolution channel and "detail" or "fine" with the refinement or augmentation channel of the two-resolution hierarchy.

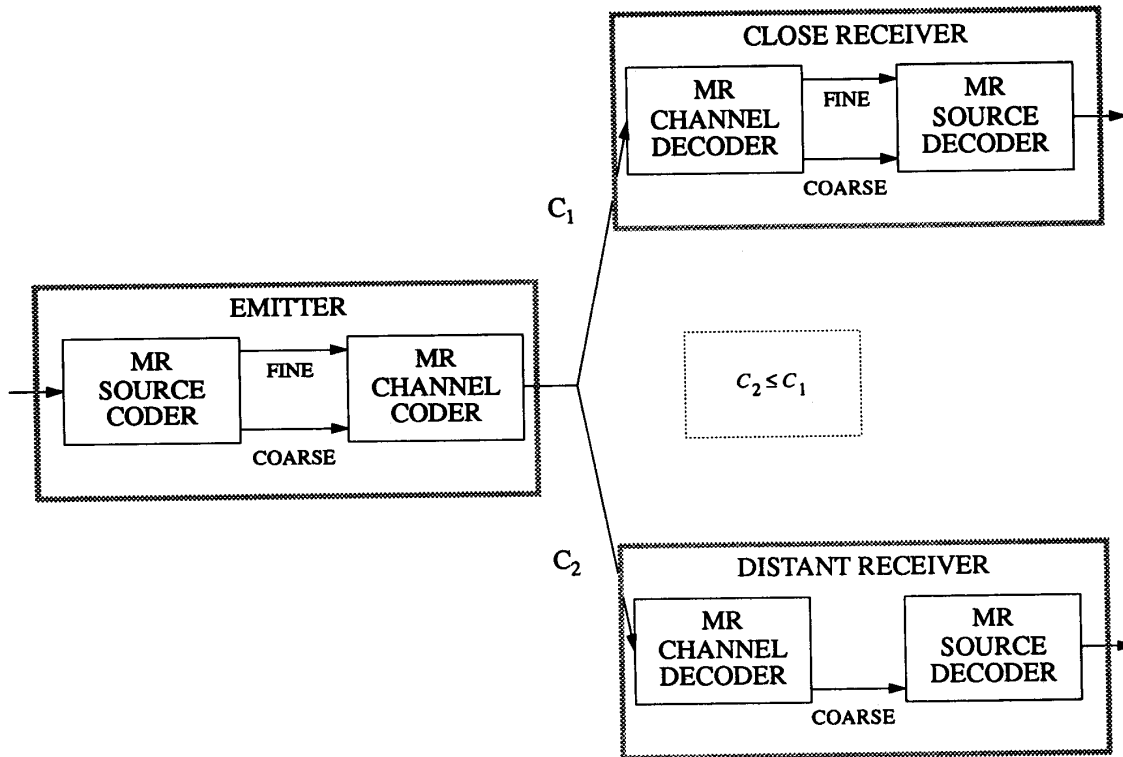


Fig. 1. Block diagram of a multiresolution digital broadcasting scheme shown for two receivers with channel capacities C_1 and C_2 with $C_2 \leq C_1$.

previously proposed as a natural advantage of multiresolution systems [5], has been recently included in the AT&T/Zenith proposal [6], a change from their single-resolution scheme advocated earlier [7]. The Sarnoff/NBC/Philips/Thomson [8] proposal includes prioritization in its coding scheme, but does not possess the “embedded” MR transmission to be described in this paper. The idea of efficient multiplexing of the different resolutions of a MR transmission scheme has been studied, using multidimensional constellations, in [9] although a joint source and channel coding design is not addressed. Schreiber has pointed out [1], [10] the problem of spectral efficiency for broadcast, and has proposed a hybrid analog-over-digital scheme which, though multiresolution in nature, does not fully exploit recent advances in digital compression technology. Note that though several works [11]–[13] have considered, in different contexts, the problem of joint source channel coding of images, none has tackled the problem in a broadcast scenario.

C. Outline of Paper

The outline of the paper is as follows. Section II presents the digital broadcast problem and proposes a multiresolution formulation. Section III reviews MR video coding [14] and summarizes the specific scheme [4] used in this paper for HDTV source coding. Section IV discusses the idea of MR transmission for broadcast channels. It reviews the classic idea of embedding [2] and shows how it can be applied

to digital broadcast. We introduce the concept of *embedded constellations* and show, through a series of examples, how these, possibly combined with TCM and ECC’s, can provide an efficient solution. Section V discusses the alternatives and proposes a recipe for the broadcast problem as posed in Section II. Finally, Section VI verifies the benefits of using an embedded multiresolution design and illustrates the robustness achievable by using efficient *error concealment* techniques in an MR coding environment.

II. THE DIGITAL BROADCAST PROBLEM: A MULTIREOLUTION FORMULATION

While Shannon [15] established the theoretical optimality of the separation of source coding or removal of redundancy from a source, from channel coding, or insertion of redundancy to combat a noisy channel, his results hold only in the limit of infinitely complex and long codes and, more important, for a single-channel or point-to-point communication system. For the broadcast or multichannel environment where a source communicates with a multitude of receivers of varying strengths, as will be explained in detail in Section IV, Cover [2] established that optimal broadcast scenarios are multiresolution or embedded in character. This justifies the choice of a multiresolution (MR) source coding scheme to represent a source compactly in a hierarchy of resolutions, to which a “matched” MR transmission can be designed in order to produce an efficient end-to-end design.

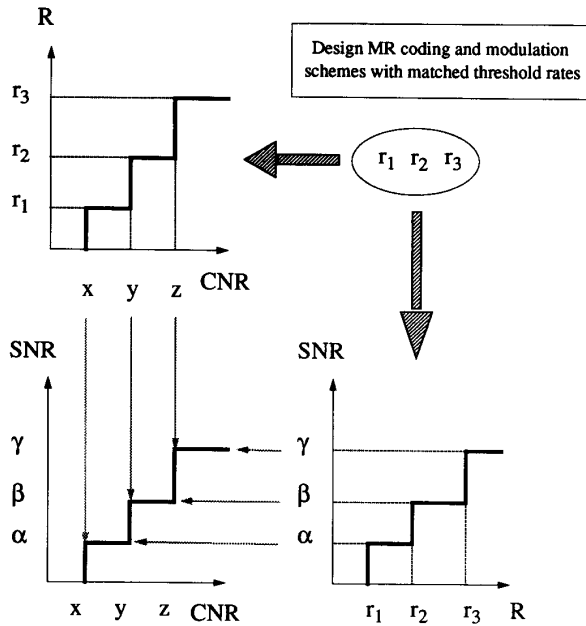


Fig. 2. Matching of MR source and channel coders for desired broadcast characteristics. (a) MR channel coder characteristics (rate versus CNR). (b) Matching of threshold rates of channel and source coders to achieve desired broadcast characteristics. (c) Achieved broadcast characteristics. (d) MR source coder characteristics (SNR versus rate).

A. Matched MR Source Channel Coding

While the problem of joint source and channel coding has been addressed previously in various coding contexts, as stated in Section I, in this paper we propose the idea of designing an end-to-end joint MR system, i.e., one which includes an MR channel coding scheme (an analog MR constellation, possibly using an MR TCM, and/or a digital MR ECC) that is matched to the MR source coding scheme outlined in Section III).

Fig. 2 outlines the importance of employing a joint design. For the different receiver carrier-to-noise ratios (CNR's) throughout the broadcast area, the MR digital transmission system [see Fig. 2(a)] can reliably deliver different user bit rates.

The idea is to design the MR source and channel coders so that their delivered rates are efficiently matched. The channel rates correspond to the MR modulation scheme while the source rates refer to the different resolutions of the source coder, whose characteristics are shown in Fig. 2(d),² resulting in the broadcast characteristics of Fig. 2(c).

This paper suggests an efficient way to do this matching. We explain how the MR channel coder curve, which we attempt to match to the MR source coder, can be designed using the concept of embedded transmission. Note that while embedded transmission for broadcast is efficient even for a

²Note that, while we use SNR as a source quality measure in this discussion, we do so with the usual disclaimer that while perceptual measures are more meaningful, they are difficult to quantify. Besides, any meaningful measure can be used in place of SNR without changing the nature of the joint source channel coding philosophy we outline here. Also, note that SNR is a source quality measure while CNR is a channel quality measure.

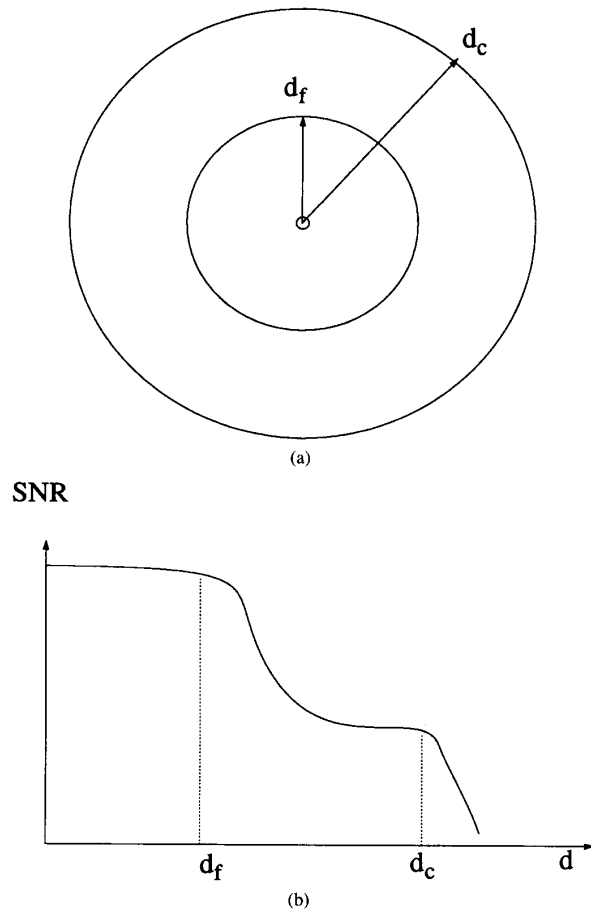


Fig. 3. (a) Definition of the full- and low-resolution coverage areas. (b) Delivered quality as a function of distance from transmitter.

single-resolution source, it is even more natural to invoke when the source coder is hierarchical in nature (as is our case, to be described in Section III).

To solve the matching problem of Fig. 2(b), i.e., design the source and channel coders with matched rates, one needs a broadcast performance criterion over which to optimize the parameters. We now address this problem and suggest a tractable formulation.

B. The Problem of Choosing a Cost Function

The main difficulty in assessing the performance of a digital broadcast system is that of defining a cost function. In other words, one would like to have some way of measuring the performance of a system in terms of, say, the coverage area and the delivered quality for a given set of resources to be used, such as bandwidth and power. When studying a digital broadcast problem, this measure is not simple: the threshold effect mentioned earlier, simplistically stated, boils down to a tradeoff between coverage area and quality of reception in the case of a single-resolution scheme. A multiresolution scheme will face the same tradeoff but in a more complex way. For example, in the two-resolution case, one can trade off

high-quality (full-resolution) coverage area for a lower quality (lower resolution) coverage area as well as the quality of each resolution for a larger coverage area without affecting the area of the other resolution. Would it be better to cover a wide area with relatively low quality or a small area with high quality? The answer is not obvious and points to the lack of a clear cost function for this problem. However, making some assumptions about both the system and the requirements helps us set the system parameters without resorting to a cost analysis.

C. Setting the Objectives for the System

Assume a two-resolution system. It is reasonable to expect the system to provide the two possible grades of service (full resolution closer to the emitter and a reduced but still acceptable quality further away) for the respective areas defined by distances of d_c and d_f from the emitter ($d_c \geq d_f$). The crucial point is to define what those distances represent in terms of quality. Since different systems will deliver different qualities, it is convenient to define those distances as the maximum distances at which each channel is received reliably (see Fig. 3). We can, for instance, equate reliability with the delivered error rate being below a desired threshold. In summary, the system requirements can be set up in terms of providing full-resolution and lower-resolution quality at certain specified distances from the emitter. Now, source and channel coding have to be chosen so as to guarantee that the required areas are covered while maximizing the received quality.

In order to address the stated problem (see Section V), we first analyze the system components, namely the source and channel coders.

III. MULTIREOLUTION SOURCE CODING

Many popular and efficient source coding schemes are either directly or indirectly MR in nature. Methods like subband and wavelet coding have a natural multiresolution interpretation while others, like discrete cosine transform (DCT)-based techniques which represent a common theme in all the digital HDTV proposals to the FCC, have an "acquired" MR interpretation. For a comprehensive review of multiresolution digital coding techniques, the reader is referred to [14].

Multiresolution source coding schemes can be seen as successive approximation methods. While they can be slightly suboptimal in terms of compression over a single-resolution (SR) scheme that achieves the same full-resolution quality for point-to-point communications, they can be superior in a broadcast situation, which is a multiuser communications problem. Even for point-to-point communications, it has been shown [16], [17] that MR source coding, using a successive refinement approach, can be theoretically optimal under certain conditions and, recently, an efficient practical MR source coder has been suggested [18] that compares very favorably with single-resolution approaches that achieve the same full-resolution quality. The advantage of MR-based schemes over SR schemes in a broadcast environment comes from the presence, in the former, of a coarse channel (which comes as a "byproduct") that, combined with error concealment techniques used at the source decoder, can be used to increase

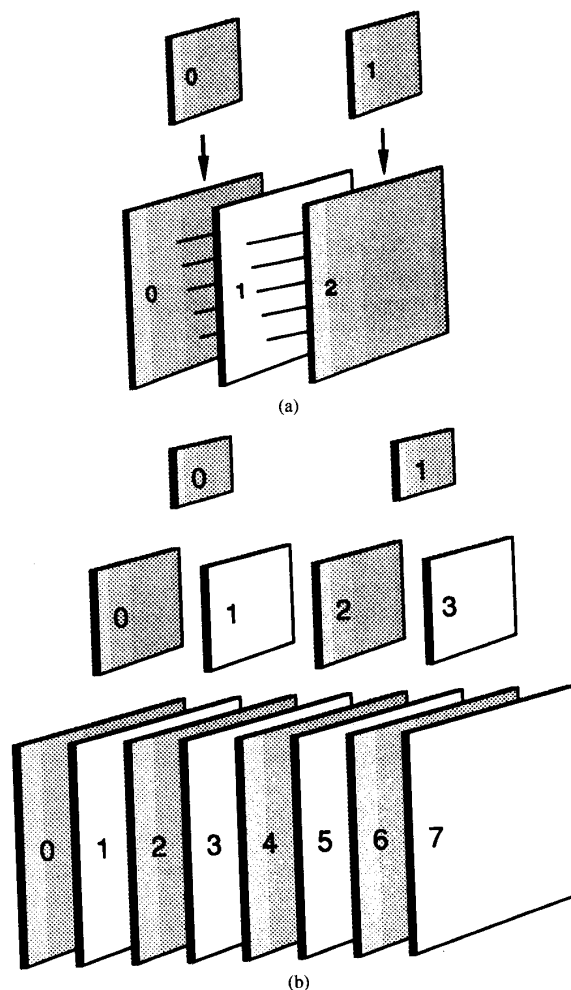


Fig. 4. Reconstruction of the spatiotemporal pyramid. (a) One step of coarse-to-fine scale change. (b) The reconstructed pyramid. Note that approximately one-half of the frames in the structure (shown as shaded) are spatially coded/interpolated.

robustness. A more detailed analysis of this robustness issue will be made in Section VI. A note of interest, especially in the wake of the ongoing standards and compatibility debates, is that MR decomposition affords a hierarchy of resolutions that are both natural and useful for compatibility and broadcast problems.

A. A Specific MR Source Coding Scheme for HDTV

Our MR video coder [4] is a three-dimensional pyramidal decomposition, based on spatiotemporal interpolation, forming a hierarchy of video signals at increasing temporal and spatial resolutions [see Fig. 4(b)]. The structure is formed in a bottom-up manner, starting from the finest resolution, and obtaining a hierarchy of lower-resolution versions. Spatially, images are subsampled after antialiasing filtering. Temporally, the reduction is achieved by simple frame skipping, since temporal filtering would be inadequate when motion is present (it would produce double images).

The encoding is done in a stepwise fashion, starting at the top layer and working down the pyramid in a series of successive refinement steps. The coarse-to-fine scale change step is illustrated in Fig. 4(a). At each step, the spatial resolution is increased by linear interpolation and then the temporal motion-based interpolation is done based on these new frames at the finer scale. We describe the interpolation procedure only briefly, and refer the reader to [4] for details. See Fig. 20 for the different resolutions of a three-layer pyramid for the "mit" sequence.

The unshaded frames shown in Fig. 4(b) are interpolated in time. For these frames, the encoder computes a set of motion vectors that are transmitted along with the residual, i.e., the difference between the original and the interpolated frame. The motion vectors are computed in an MR fashion, using a hierarchical blockmatching algorithm [4]. For each block in the interpolated frame, three different motion vector candidates for the following interpolation modes are considered: *backward interpolation*, the motion vector that yields the best replacement from the previous frame; *forward interpolation*, the motion vector that yields the best replacement from the next frame; *motion averaged interpolation*, the motion vector \mathbf{d} that yields the best replacement by averaging the block displaced by \mathbf{d} in the previous frame and displaced by $-\mathbf{d}$ in the next frame. The mode that results in the best interpolated block (in the mean squared error or MSE sense) is selected, and the mode selection information is also encoded and transmitted to the receiver.

A DCT-based coder is used to encode the top layer and the subsequent bandpass difference images. Quantizer steps and, consequently, bit allocation at different levels in the hierarchy are determined to obtain good perceptual quality. Another major consideration in the bit allocation scheme is in "matching" the subsequent channel coding, to be described later in the paper.

It is important to note that for the MR source coder we consider in our system, if one resorts to a two-resolution hierarchy where the two coarsest layers of the spatiotemporal pyramid comprise the coarse resolution source channel and the difference layer comprises the detail channel, then the bit ratio of coarse-to-detail information is roughly 1 : 2 at high perceptual quality for typical sequences. This ratio is more accurate if one includes the "vital" overhead information associated with motion vectors and synchronization in the lower-resolution channel. The coarse and fine channel bit streams could be interpreted as entering virtual independent buffers with throughputs in the ratio of 1 : 2, with instantaneous temporal mismatches in the input channel rates being absorbed by the buffers and (if necessary) to prevent overflow or underflow, resolved by exchange of data between the buffers, resulting in minimal degradation for slight mismatches. This 1 : 2 ratio is a key parameter in the development of our joint MR source channel coding system.

IV. MULTIREOLUTION TRANSMISSION: EMBEDDING

The problem of efficient communication of digital information from a single source to multiple receivers with various

carrier-to-noise ratios is key to digital broadcast of HDTV. While the theory of digital broadcast has received attention in early information-theoretical literature [2], [16], [19], there is no evidence of the application of the theoretical maxims proffered in [2] to the design of practical digital broadcast channels. An efficient end-to-end broadcast system should have its *transmission constellation matched to its source coding scheme*, and this is the crux of our work, which we undertake in a *multiresolution environment*.

A. Efficiency of Using Embedding for Digital Broadcast

Fig. 5(a) depicts a typical broadcast environment, with a source wishing to convey information $\{r, s_1\}$ to a stronger receiver and $\{r, s_2\}$ to a weaker one. Note that r represents the common message to be conveyed to both receivers. In [2], Cover establishes the efficiency of superimposing information, i.e., broadcasting in a multiresolution embedded fashion, where the *detailed information meant for the stronger receiver necessarily includes the coarse information meant for the noisier receiver*. The efficiency of embedded broadcast, in terms of theoretically deliverable bit rate, compared to independent sharing of the broadcast channel resources in time or frequency among the receivers is depicted in Fig. 5(b), where the superior curve is obtained by superimposing the detail information within the coarse information. That is, the superior receiver 1, in an optimal scenario, necessarily has access to the information $\{r, s_2\}$ meant for the weaker receiver 2. Note that the plot portrays the potentially deliverable bit rates which are upper bounded by the Shannon capacities of the channels, and has the same drawback of providing no more than existential knowledge, as in Shannon's classical results on channel coding [15]. In this work, we show a practical way of realizing this embedding gain.

B. Embedding in the Modulation Domain

Cover's concept of embedding the coarse information within the detailed information is generic in scope, and places no restrictions on the domain in which the embedding should be performed. To describe the effect of an analog domain-embedded modulation, we refer to Fig. 6. The basic idea is that each constellation consists of "clouds" of miniconstellations or "satellites," where the detail information is represented in the satellites, while the coarse information is carried in the clouds. Thus, the loss of coarse information is associated with the receiver's inability to decipher correctly which cloud was transmitted, while the loss of refinement information occurs from the receiver's confusing one intracloud signal point for another. The receiver first decodes the likeliest cloud (coarse information), "subtracts" the decoded cloud value from the received point, and then decodes the likeliest satellite within the cloud (detail information). Thus, the MR 16 QAM (quadrature amplitude modulation) constellation of Fig. 6(a) has four bits/symbol, of which two bits are coarse (four clouds) and two bits are detail (four satellites/cloud). Similarly, the MR 16 PSK (phase shift keying) scheme has two coarse bits/symbol and two detail bits/symbol, while the 4 PAM (pulse amplitude modulation) constellation has one bit/symbol

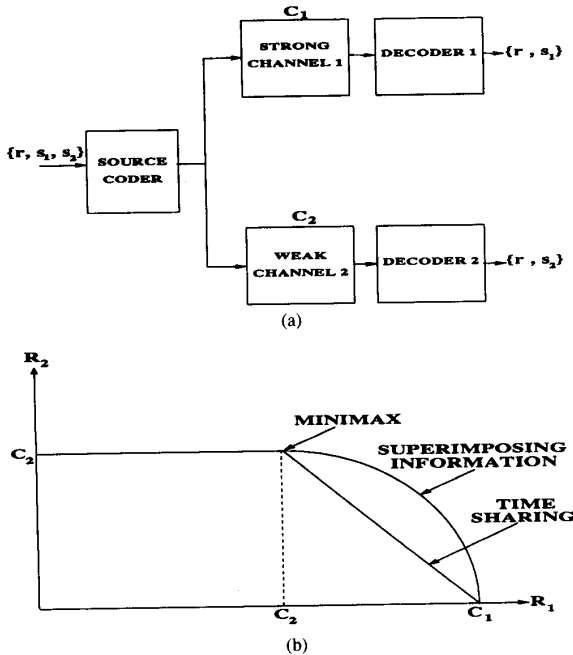


Fig. 5. Typical broadcast environment. (a) Single-source broadcasting to receivers of channel capacities C_1 and C_2 . (b) Set of achievable broadcast bit rates for receivers 1, 2.

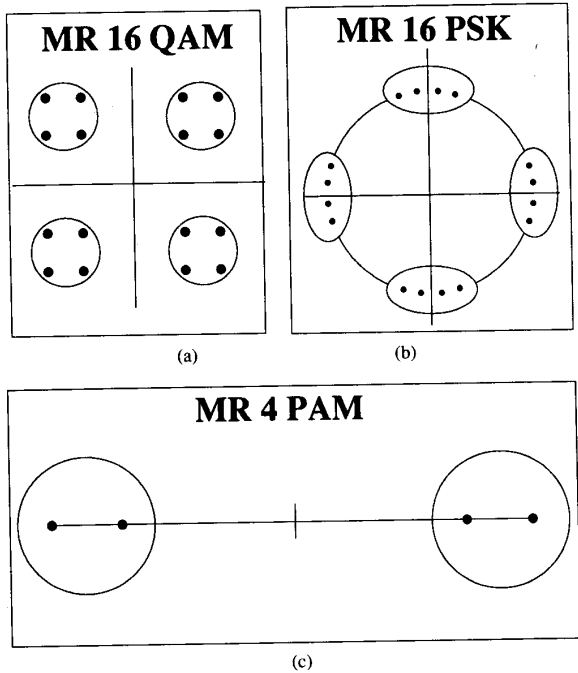


Fig. 6. Some multiresolution constellations. (a) MR 16 QAM. (b) MR 16 PSK. (c) MR 4 PAM.

of each. For our specific source coder, we consider the MR 64 QAM constellation of Fig. 7.

While we present a two-resolution hierarchy, the principles hold for any number of hierarchical levels desired and would

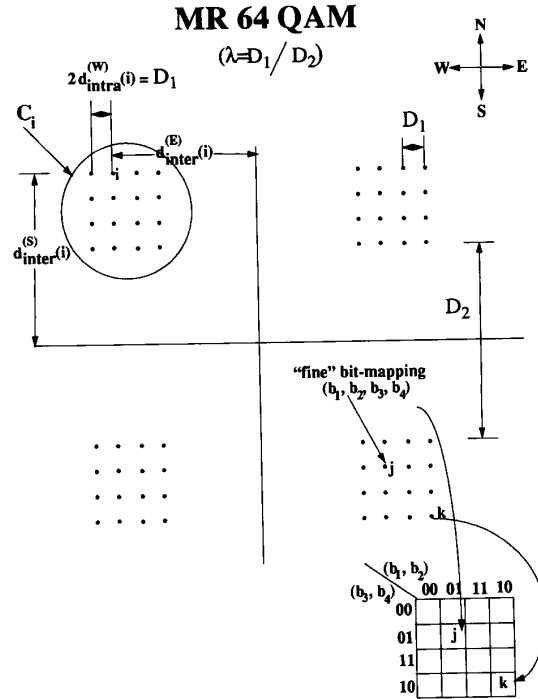


Fig. 7. Example A: MR 64 QAM system constellation with definitions of λ , C_i , $d_{intra}^k(i)$, $d_{inter}^k(i)$, and the "fine" bit mapping of the constellation signal points according to the well-known Karnaugh map partitioning. Note that $\lambda = 0$ represents uniform 4 QAM, while $\lambda = 1$ corresponds to uniform 64 QAM.

result in a "fractal" modulation constellation [20], although at increased complexity and decreased practicality. We point out later how one can combine an embedded ECC scheme with an embedded modulation scheme to increase the number of broadcast resolutions in a practical manner without sacrificing efficiency in the information-theoretical sense.

1) *MR 64 QAM*: Consider the constellation of Fig. 7 as Example A. For every six composite bits per channel symbol emitted by the 1 : 2 source (see Section III), two coarse bits select one of the four clouds while the remaining four detail bits select one of the 16 satellites within the selected cloud. By "matching" the relative distances between intracloud constellation points (D_1) and intercloud points (D_2), whose ratio is a design parameter λ , to the relative degrees of desired protection for the two bit streams, one obtains an efficiently designed joint MR source/MR transmission system. An optimal "broadcast λ " could be obtained if a meaningful cost function over the broadcast area (which would probably include factors like population density) were available. On the other hand, due to the difficulty of a cost function analysis (see Section II), one could operate at the maximum value of λ that meets the full-resolution coverage range requirement (see Section V).

The Appendix contains the mathematical analysis of the coarse and detail channel performance of the MR 64 QAM of Fig. 7, on which the curves shown in Fig. 9 are based. Note that those curves reflect packet error rates for the two

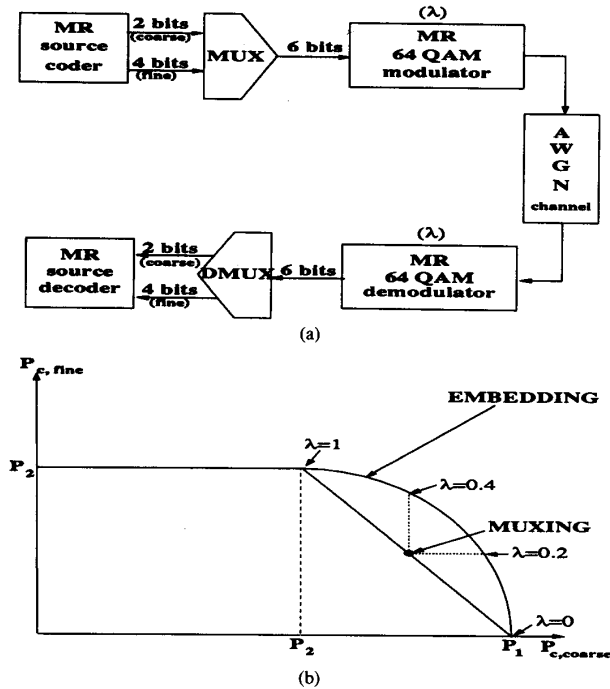


Fig. 8. (a) MR QAM system block diagram (Example A). Note that the modulator and demodulator are operated at transmission parameter λ (see Fig. 3). (b) Achievable performance (probability that packet is correct) for practical system of Fig. 8(a). See analogy with theoretical curve of Fig. 5(b).

channels, where a composite packet of length 1080 bits (with 1/3 coarse and 2/3 detail information embedded in it) is used to prevent error propagation.

While the details are provided in the Appendix, it is important to mention a few salient features. Note that $d_{\text{intra}}^k(i)$ and $d_{\text{inter}}^k(i)$ represent half the Euclidean distances between signal point i and its nearest coarse and detail neighbors, respectively, in the k -direction. Also, it must be emphasized that the topology of the equivalent constellation at the broadcast receiver is a function of the CNR and λ . Qualitatively, the CNR affects the "radius" of the constellation as seen as the receiver for a fixed noise variance while λ affects the relative distances between intercloud and intracloud points. As λ goes from 0 to 1, the intracloud and intercloud thresholds decrease and increase, respectively, for a fixed power budget, indicating the quantitative tradeoffs involved in coarse and detail channel robustness as shown in Fig. 9. Also, note that as we can always form a Gray code fine-channel digital bit mapping of the constellation points, exactly as in Karnaugh maps used in digital logic design [21] (see Fig. 7), we can guarantee that every point in the constellation is at Hamming distance one away from each of its intracloud nearest neighbors. Thus, assuming that single-bit errors dominate when symbol errors occur, we can equate symbol errors with bit errors. This leads to an efficient mapping and aids in the mathematical analysis.

Due to favored protection of the coarse stream via the parameter λ , it is possible for the fine packet component to be corrupted while the coarse packet component is received reliably for the same composite packet. The dotted curves in

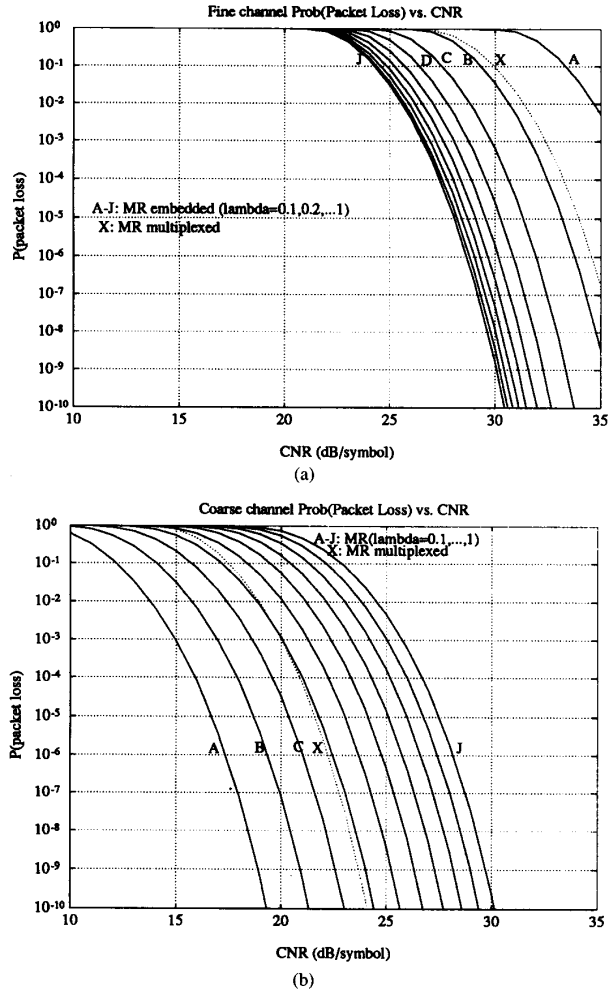


Fig. 9. Example A. Probability of packet error versus receiver CNR over the entire range of transmission parameter λ for the embedded MR 64 QAM case. Composite packet length is 1080 bits. (a) Fine channel packet loss. (b) Coarse channel packet loss.

Fig. 9 refer to a "naive" multiplexing of the broadcast channel between the coarse and detail information streams under conditions of equal power, bandwidth, and average spectral efficiency, as will be explained in Section V. The curves clearly show the superiority of embedding over multiplexing. For example, for values of λ from 0.2 to about 0.4, both coarse and detail channel performances are better than those of the multiplexed case. The particular multiplexing point shown in the figure is obtained when the power in the coarse and detail constellations are made equal, though similar performance improvement can be obtained by embedding over any other multiplexing point also, corresponding to different values of λ . This is a verification of the information theoretical result that embedding outperforms multiplexing. See Fig. 8(b).

C. Embedded TCM Constellation

In order to increase reliability of reception over the de-

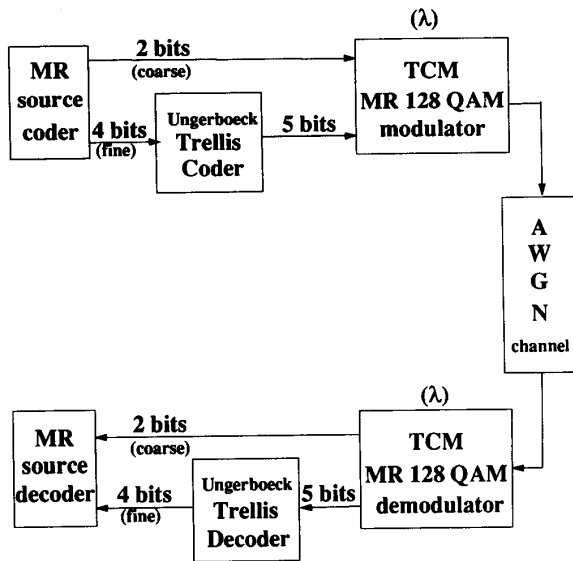


Fig. 10. Example B. Block diagram of an embedded MR TCM system using a 128 QAM constellation. Note that it consists of four clouds of trellis code modulated 32 QAM constellations.

manding broadcast channel and to increase coverage area, it may be necessary to add more redundancy to protect the broadcast information. As is well known, convolutional codes using a Euclidean distance metric can achieve better performance for the same complexity than the more commonly used block ECC's, which use a "hard-limiting" Hamming distance metric. Convolutional (trellis) codes achieve coding gain by using soft decoding with the Viterbi algorithm [22]. Conventional convolutional coding, like block coding, would require an increase in bit rate to accommodate the redundant bits, which must come at the expense of lowered source coding quality, for a fixed total throughput. However, it is possible to achieve almost all the coding gain theoretically possible, i.e., to approach the Shannon limit, by expanding the 2-D modulation constellation by a factor of 2 and employing a redundant constellation via trellis coded modulation, as established by Ungerboeck [23]. While multidimensional TCM [24], [25] can provide the same gain for a smaller expansion factor than the 2-D Ungerboeck constellations, we restrict ourselves to the latter in the interest of simplicity of design and analysis. *The novelty here is that we combine the concept of multiresolution with the power of TCM to propose an embedded TCM modulation for efficient broadcast of a MR source* [see Fig. 11(a)].

An Ungerboeck TCM scheme requires an expansion factor of 2 in the constellation size. Thus, our original MR 64 QAM constellation would be expanded to 128 QAM, using the same power as the former. Of course, this large constellation size is for our specific example (Example B: see Fig. 10): a more practical example for HDTV broadcast might be expansion of a MR 16 QAM scheme (with a 1 : 1 coarse-to-detail bit rate ratio, as in [26], using 2 bits/symbol for each resolution) into an embedded TCM 32 QAM, which is certainly practical in size. The principle of operation is what is important.

Embedded TCM 128-QAM

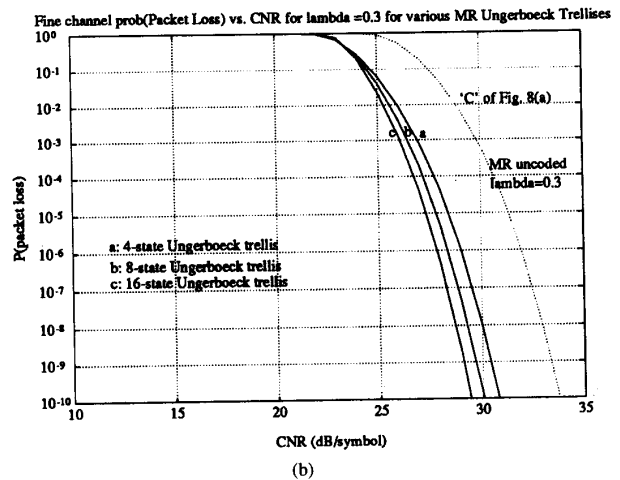
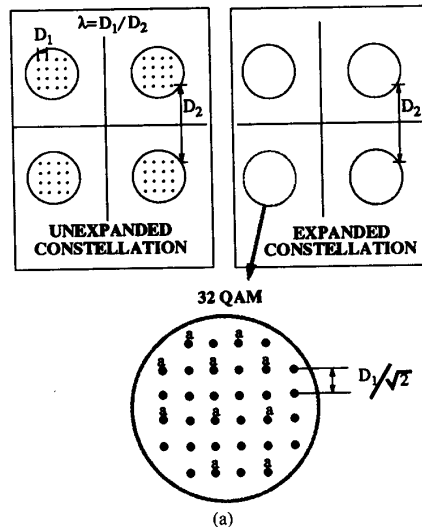


Fig. 11. Example B. (a) Expansion of MR 64 QAM into MR-TCM 128 QAM with an expansion in constellation points of each cloud from 16 to 32. Note that the coarse channel is unaffected. (b) Coding gain over MR 64 QAM for the detail channel using MR-TCM 128 QAM for $\lambda = 0.3$.

The idea for the TCM 128 QAM scheme [see Fig. 11(a)] is that the coarse information retains preferential protection through λ , while the detail information gets expanded from 16 points to 32 points per cloud via a TCM coding scheme. Fig. 11(a) shows the first-level set partitioning for each 32-point cloud into the subset marked "a" and its complement (unmarked), each subset enjoying a 3 dB gain in squared Euclidean minimum distance over that of its parent, as needed for an Ungerboeck code. Note that here we maintain the same intercloud distance after expansion, while the intracloud distance is halved (3 dB fine channel gain). Alternatively, one could maintain the intracloud distance (6 dB fine channel gain) while slightly degrading the coarse channel performance, a tradeoff that is worthwhile for small values of λ [33].

Fig. 11(b) shows the coding gain for the fine channel (the

coarse channel remains unchanged) for $\lambda = 0.3$ for trellises with 4, 8, and 16 states. The coding gain over the unexpanded MR-64 QAM constellation is seen to be consistent with that tabulated in [23]. Thus, the simple four-state trellis is seen to provide a coding gain of 3 dB/symbol in CNR. Identical gains in detail channel protection will occur for any desired value of λ . Thus, for an efficient end-to-end MR design, one may ensure the coarse channel robustness through λ while using a TCM code of acceptable complexity to achieve the desired full-resolution coverage area. An important feature of our MR system is that due to inclusion of error concealment techniques at the decoder (see Section VI-B), it is possible to obtain indistinguishable full-resolution quality even at a fine-channel packet loss rate exceeding 10^{-1} . As seen from Fig. 11(b), at this high loss rate one gets marginal return from using trellises over four states, thus making our TCM design nearly optimal with only a simple four-state trellis! It is important to note that this scheme permits operation with *no decrease in source coding bit rate* over that of an uncoded system.

D. Embedding in the ECC Domain: UEP Codes

While unequal degrees of robustness can be achieved for the coarse and refinement channels using analog modulation, similar results can be obtained using digital ECC's with unequal error protection (UEP) [13], [27], [28]. Although the TCM constellation mentioned earlier is indeed efficient, there may be practical limitations to expanding the constellation size. Moreover, one may need ECC's to "bridge" any mismatches in rate between the source and channel modulators (see Fig. 2).

It can be seen that embedding in the modulation and ECC domains are essentially equivalent. In the ECC domain, codewords of length n in $(GF(2))^n$ are clustered into "clouds" whose members ("satellites") are closer in Hamming distance, with respect to one another, than to members of other clouds. Codes having this behavior are referred to as UEP codes.

A two-level UEP code can be described as an (n, k_1, k_2, t_1, t_2) code (where t_i represents the number of channel errors the code can withstand for the information k_i). It has to be noted that using a UEP code is by no means the only way to provide unequal error protection. As a first approach, one could use two different codes for each category of information but it is essential to note that embedding the codes can yield better (in terms of the rate k/n) codes than using two separate codes. In other words, combining two (n_1, k_1, t_1) and (n_2, k_2, t_2) codes to obtain an $(n_1 + n_2, k_1, k_2, t_1, t_2)$ code can potentially be outperformed by an (n, k_1, k_2, t_1, t_2) embedded code. As an example, consider a $(63, 12, 24, 5, 3)$ binary cyclic UEP code listed in [29]. Alternatively, one can consider two smaller BCH codes with characteristics $(31, 11, 5)$ and $(31, 12, 3)$. The BCH codes can provide the same protection but clearly their rates are worse than those obtained with the embedded code. To further the analogy with the modulation domain, the use of different codes for the different classes of information (as in [11]) can be likened to the "naive" multiplexing for transmission for the two-user broadcast channel.

It follows that resorting to UEP codes is more efficient than using separate ECC's to provide unequal protection. However,

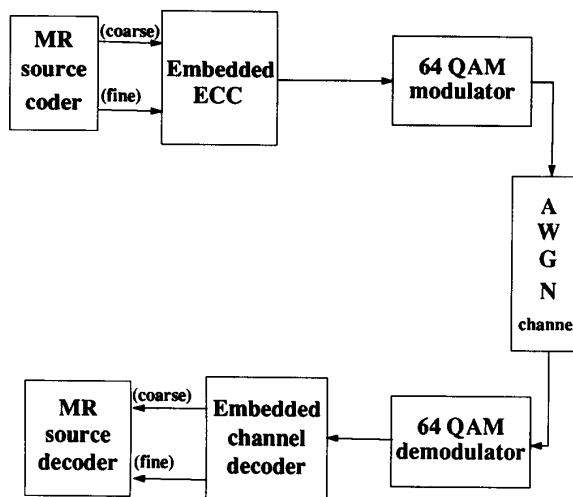


Fig. 12. Example C. Block diagram of an MR system with embedded ECC's for the coarse and detail channels.

UEP codes are hard to find and no structured method (that does not require brute-force computer search) has been described to design them. See Lin *et al.* [29] for a tabulation of all possible embedded ECC's of odd lengths up to 65. The codes listed in [29] are not appropriate for the application considered in that, of those codes with ratio of coarse-to-detail information (k_2/k_1) close to 2, few are efficient (i.e., with rates $(k_1 + k_2)/n$ close enough to 1). Fig. 12 (Example C) presents the block diagram of a scheme that uses embedded UEP codes, and Fig. 13 shows the results, in terms of packet loss, for different CNR's when several of the codes tabulated in [29] are used.

Thus, while UEP codes can be used instead of MR modulation schemes to perform the MR transmission, the issue of designing good UEP codes is largely open and involves a high degree of complexity. Following the above considerations, for our application we consider unequally error protected ECC's designed *independently* for the coarse and detail information channels. Using the same coarse packet size of 360 user bits (k) and various levels of redundancy ($n - k$), we simulated the performance of various (n, k, t) ECC's (Example D). This example consists of protection of only the coarse channel to varying degrees of robustness, while leaving the detail channel uncoded [see Fig. 14(a)]. Fig. 14(b) shows how using ECC's lowers the probability of coarse packet loss over the range of CNR's of interest.

E. Hybrid Embedded Modulation/ECC Scheme

ECC's and MR embedded modulation may have to be used jointly in an efficient end-to-end system. For example, one could use a nonuniform QAM scheme rather than the uniform QAM constellation in Fig. 14(a) (Example D). Thus, the ECC scheme could be used as a "bridge" to achieve a match between the bit rates (coarse and fine) required by the MR constellation and the bit rates (source bits plus ECC bits) sent through each of the channels (see Fig. 2). Also, embedding in both the ECC domain and the MR modulation domain

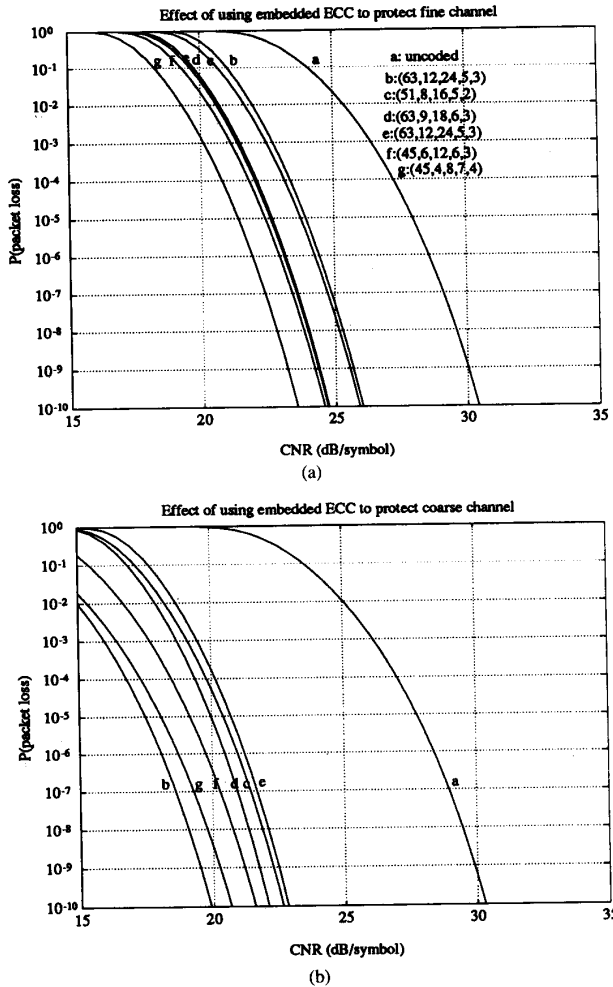


Fig. 13. Example C. Probability of packet error versus receiver CNR for some known embedded ECC's. Note that the five-tuple (n, k_1, k_2, t_1, t_2) listed refers to the embedded code length, the coarse bits per block, the detail bits per block, the error correction capability for the coarse bits per block, and the error correction capability for the detail bits per block, respectively. (a) Fine channel packet loss. (b) Coarse channel packet loss.

would lead to an efficiently designed MR joint source channel system with more than two resolutions without resorting to a complex "fractal" modulation constellation. This could be accomplished, for example, for a three-resolution design by having the two coarsest resolutions being embedded in the ECC domain, and the resultant composite coarse bit stream being embedded in the third (detail) channel bit stream in the modulation domain as a two-layer embedding (see Fig. 15).

V. AN EFFICIENT END-TO-END SYSTEM DESIGN

In the previous section, we have illustrated, through Examples A–E, the different tools that can be used to design an efficient broadcast system. Using these tools, we now examine the tradeoffs involved in the various schemes to provide a solution to the problem stated in Section II. Table I gives the coordinates of Examples A–D.

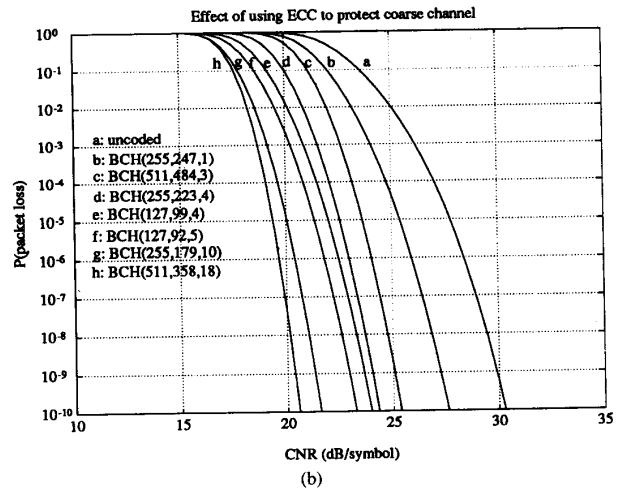
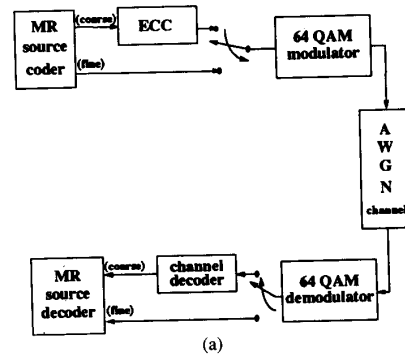


Fig. 14. Example D: Multiplexed (nonembedded) ECC using a family of BCH codes. Note that error correction is applied to the coarse channel only, with the detail channel sent unprotected. Note that the three-tuple (n, k, t) listed refers to the code length, the coarse bits per block, and the error correction capability for the coarse bits per block. The packet loss rate refers to a coarse packet length of 360 bits. (a) Block diagram. (b) Simulation of coarse channel performance.

A. Comparison of λ Modulation, TCM, and ECC Schemes

The λ modulation scheme of Example A might be used to provide a desired coverage range for the coarse-resolution signal, and a "basic" coverage for the fine channel. The MR TCM scheme of Example B increases the full-resolution coverage area at the cost of an expanded modulation constellation. The scheme of Example C (embedded ECC's), while efficient in an information theoretical sense, is unlikely to meet the bit rate ratios of the different resolutions required of most practical HDTV schemes and is, hence, omitted from our discussion. The ECC scheme of Example D can be used to make the coarse channel more robust but at the cost of reduced quality, for a fixed total bit rate budget for source and channel coding.

Table II gives a comparison of Examples A, B, and D for a typical problem. We fix the coarse channel quality and coverage requirement (delivered coarse packet error rate (PER) of less than 10^{-3} at receiver CNR's above 19 dB/symbol) and compare the full-resolution quality and coverage range for the different schemes. As seen, all schemes perform well with respect to an uncoded system. Note that, with MR schemes,

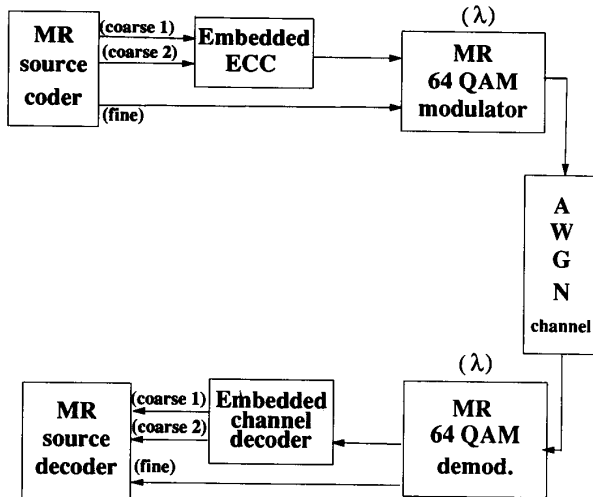
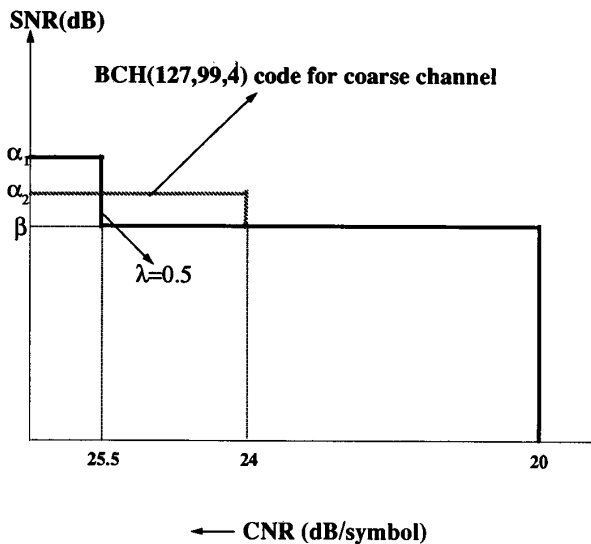


Fig. 15. Example E: Block diagram of a MR system with three levels of resolution using both embedded ECC's and embedded modulation to make overall design efficient and practical.



Thresholding assumed at $\Pr(\text{coarse packet error})=10^{-3}$
and $\Pr(\text{fine packet error})=10^{-1}$

Fig. 16. Tradeoff between using modulation domain protection via transmission parameter $\lambda = 0.5$, and ECC domain protection using a BCH (127, 99, 4) code applied to the coarse channel.

a fine channel packet loss of 10^{-1} is reasonable to get good full-resolution quality if the coarse channel is near-perfect and error concealment is invoked at the decoder (see Section VI-B and Fig. 21). While Example A is a good low-complexity solution, Example B, operating at $\lambda = 0.3$ with an embedded four-state trellis, is the best choice if constellation expansion is tolerable. As seen from Fig. 11(b), for a probability of fine channel packet loss of 10^{-1} , one needs only a simple four-state Ungerboeck trellis to get most of the coding gain.

TABLE I
SUMMARY OF PRESENTED ALTERNATIVES

Example	A	B	C	D
Description	MR QAM	MR TCM	Embedded ECC	Multiplexed ECC
Section	4.2	4.3	4.4	4.4
Block Diagram	Fig. 8	Fig. 10	Fig. 12	Fig. 14
Simulation	Fig. 9	Fig. 11	Fig. 13	Fig. 14

TABLE II
COMPARISON OF SCHEMES A, B, AND D. WE REQUIRE PACKET ERROR RATE (PER) LESS THAN 10^{-3} FOR THE COARSE CHANNEL AT 19 dB/SYMBOL CNR. WE COMPARE PERFORMANCE FOR FINE CHANNEL PER LESS THAN 10^{-1}

Example	Uncoded	A	B	D
PER	$\leq 10^{-3}$	$\leq 10^{-3}$	$\leq 10^{-3}$	$\leq 10^{-3}$
CNR(dB/symbol) Range	≥ 26	≥ 19	≥ 19	≥ 19
Low-Resolution Quality	—	same as uncoded	same as uncoded	same as uncoded
PER	$\leq 10^{-3} (*)$	$\leq 10^{-1}$	$\leq 10^{-1}$	$\leq 10^{-1}$
CNR (dB/symbol) Range	≥ 27	≥ 27	≥ 24	≥ 24
High-Resolution Quality	—	same as uncoded	same as uncoded	fine channel bit rate 15% less than uncoded (**)
Design Parameter	—	$\lambda = 0.3$	$\lambda = 0.3$ four-state trellis	BCH (255, 179, 10)
Complexity	—	same order as uncoded	higher than uncoded	higher than uncoded
Increase in Coverage over Uncoded	—	coarse: +7dB full: +0 dB	coarse: +7dB full: +3 dB	coarse: +7dB full: +3 dB

(*)For the uncoded system, the fine channel error rate cannot be 10^{-1} as error concealment requires "perfect" coarse channel performance at that fine channel error rate. See Section VI-A.

(**)The reduction in bit rate available for source coding is due to the use of an ECC.

Example D gives the same coverage as Example B, but it requires a complicated ECC which also results in 15% reduced fine channel bit rate and, therefore, a degraded full-resolution quality.

Fig. 16 shows the different performances obtained with Example A for $\lambda = 0.5$ and Example D using a BCH (127, 99, 4) code to protect the coarse channel. For the same coarse channel performance, with CNR's below 20 dB/symbol receiving no signal, the crossover from coarse resolution quality ($\text{SNR} = \beta$) to full resolution occurs for Example A and Example D at 25.5 dB/symbol and 24 dB/symbol, respectively. Note, however, that the parity bits needed by the ECC-protected coarse channel must necessarily come at the expense of a lower fine-channel bit rate, resulting in degraded full resolution quality as noted in Table II. Thus, if a comparison is to be made on the basis of equal bandwidth, the ECC scheme would necessarily have lower full-resolution

quality ($\text{SNR} = \alpha_1$) than the MR modulation scheme ($\text{SNR} = \alpha_2$) for all receiver CNR's better than 25.5 dB/symbol, but the full-resolution gain in CNR is 1.5 dB for the ECC scheme (24 dB versus 25.5 dB). The assessment of the tradeoff depends on the values of α_1, α_2 , and β , which in turn depend on the source coding used.

The following points of comparison between the two schemes of Examples A (MR modulation) and D (ECC scheme) are worthy of note.

- The coverage tradeoff is between the modulation scheme's degradation of quality by $(\alpha_2 - \beta)$ dB for receivers between 24 dB/symbol and 25.5 dB/symbol CNR versus the ECC scheme's degradation of full-resolution quality by $(\alpha_1 - \alpha_2)$ dB for all receivers with CNR better than 25.5 dB/symbol.

- Note the complexity disparity in the two schemes, with the ECC scheme resorting to a complicated BCH code, while the MR embedded modulation QAM scheme comes at relatively little excess cost over that of a uniform QAM scheme which must be used for transmission anyway.

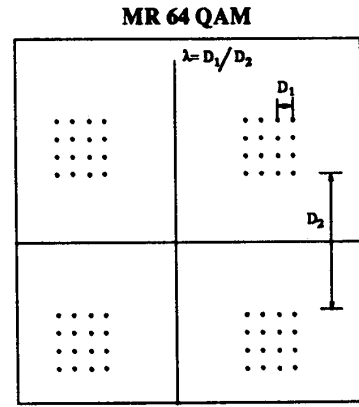
- The MR modulation parameter λ , being a continuous variable, also affords any desired operating point over the range of CNR's of interest while the ECC scheme, being discrete in nature, may not afford a solution at any desired operating point.

- In an information-theoretic sense, an embedded MR coding scheme outperforms a nonembedded one, and embedding is accomplished much more easily in the modulation domain. As the ECC scheme uses a Hamming distance metric compared to a softer Euclidean distance criterion for the modulation scheme, the latter is more efficient.

B. An Efficient Choice of System Parameters

We now have the tools to propose an efficient end-to-end systems solution to the broadcast problem stated in Section II. Assume we are constrained by the modulation constellation to operate at coarse and fine channel bit rates of R_c and R_f (for our MR 64 QAM example, $R_c:R_f = 2:4$). Note that R_c and R_f represent the combined bit budgets to be allocated between source coding and error protection for each channel. Refer to Section II and Fig. 3. The coarse channel is the "anchor" for the MR system as it represents the fallback mode and is vital for error concealment (see Section VI-B). Therefore, the desired coarse channel coverage requirement d_c of Fig. 3 should be satisfied at the maximum low-resolution quality possible. To this end, a sensible strategy for the coarse channel would be to allocate its bit budget R_c *completely for source coding*, while using the embedded λ modulation scheme of Example A to provide the desired robustness.

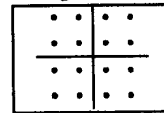
Assume that the R_f bits of the fine channel have to be allocated between error protection ($R_{f,c}$) and source coding ($R_{f,s}$), where $R_f = R_{f,s} + R_{f,c}$. Once the coverage distances (i.e., d_c and d_f) have been fixed and the modulation scheme has been selected (i.e., R_c and R_f are known), the joint source channel coding problem can be converted into a simpler one, allowing us to determine the remaining free variables of the system as follows.



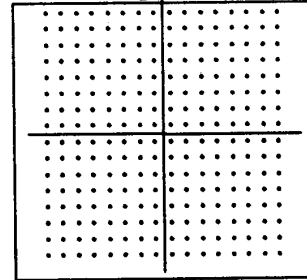
MR "EMBEDDED" CONSTELLATION

(a)

16 QAM (coarse)



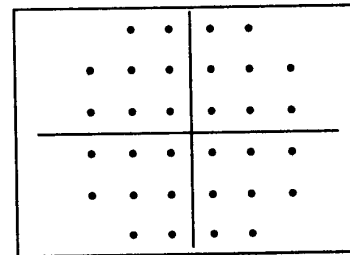
256 QAM (detail)



INDEPENDENTLY MULTIPLEXED CONSTELLATIONS

(b)

32 QAM



SINGLE-RESOLUTION CONSTELLATION

(c)

Fig. 17. (a) MR 64 QAM constellation of parameter λ . (b) Independent modulation constellations (16/256 QAM) for coarse and fine channels. (c) Single-resolution 32 QAM constellation. All constellations use equal power.

Step 1: Use the budget allocated to the coarse channel R_c to maximize the quality of the source coder for that channel (e.g., see the bit allocation strategy for our coder described in [30].)

Step 2: Use a MR modulation scheme (Example A) and set λ to the maximum value for which, at distance d_c from the emitter, the error rate for the coarse channel is below the desired threshold. If this results in an impractically low value

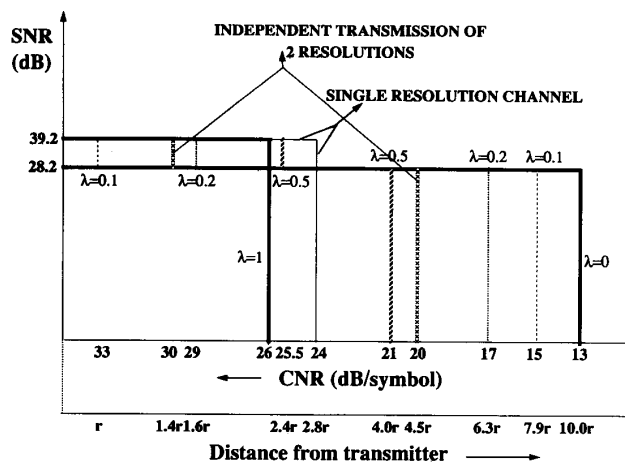


Fig. 18. Typical broadcast environment SNR versus receiver CNR.

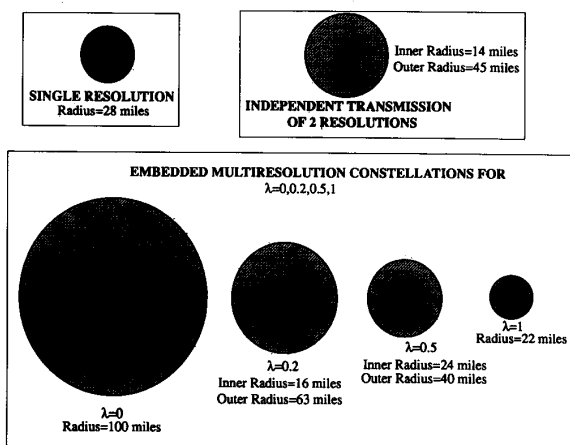


Fig. 19. Broadcast ranges for the different constellations.

of λ , one could resort to a hybrid scheme using ECC's to "boost" the coarse channel coverage albeit at the expense of a lower coarse-resolution quality, since part of the total budget must now be diverted from source coding to channel coding.

Step 3: Use a MR TCM scheme (Example B) to protect the fine channel if the constellation size is reasonable. An embedded multidimensional TCM scheme may be used [24] to reduce the expansion factor if complexity permits. If an additional increase in coverage (coding gain) is desired beyond that affordable by TCM, find an efficient error correction code (convolutional or block) that satisfies the desired fine channel error probability at distance d_f . This code will use $R_{f,c}$ bits and, therefore, the remaining $R_{f,s} = R_f - R_{f,c}$ will be used for source coding. However, due to the effect of error concealment techniques feasible with an MR design, it is unlikely that the fine channel would require additional protection (see Section VI-B). Note that if the MR TCM scheme will suffice to meet the requirements, no extra channel bits would be needed and $R_{f,s} = R_f$.

Step 4: Finally, use the remaining $R_{f,s}$ bits for the fine channel source allocation in an efficient manner, as in Step 1.

VI. ADVANTAGES OF MR BROADCAST

We now describe the advantages of using MR broadcast, i.e., joint MR coding and embedded transmission, over single-resolution as well as MR-independently transmitted schemes. We address the transmission comparison in Section VI-A and show the advantages of MR coding through error concealment in Section VI-B. While a terrestrial broadcast channel incurs various degradations like multipath distortion and fading, we consider only the more analytically tractable additive white Gaussian noise (AWGN) channel. The AWGN assumption is reasonable since actual systems would resort to adaptive channel equalization or multicarrier (e.g., orthogonal frequency division multiplexing or OFDM) techniques to combat multipath distortion.

A. Comparison of MR Embedded, MR Independent, and SR Constellations

Simulations were carried out for the AWGN channel for the MR-embedded constellation, the MR nonembedded constellation (i.e., independently transmitted constellations for the two resolutions), and the single-resolution constellation, as shown in Fig. 17. The independent case refers to separate transmission of the coarse and fine channels using "naive multiplexing" of the frequency spectrum. To ensure fairness of comparison, all three cases were tailored to operate under conditions of equal average power (it can be shown that the comparison under equal peak power constraint would be similar) and equal spectral efficiency (i.e., throughput/bandwidth).

To compare the MR versus independent constellations, a MR 64 QAM and a 16/256 QAM (coarse/fine) independent constellation pair were picked. The independent channels have a spectral efficiency of four and eight bits/symbol, or an average spectral efficiency (six bits/symbol) identical to that of the MR 64 QAM. Fig. 9 shows the performance of these two schemes for several values of λ . As was mentioned in Section IV-B, for the range of values of λ from about 0.2 to 0.4, *the embedded MR scheme outperforms the multiplexed MR scheme for both coarse and fine channels*. In order to get a comprehensive picture, a plot of received quality (SNR) versus receiver CNR is shown in Fig. 18 using perceptually consistent thresholding of the curves of Fig. 9 at coarse and fine packet loss probabilities of 10^{-3} and 10^{-1} , respectively, as justified earlier. As can be seen from Fig. 18 [and Fig. 8(b)], the MR constellation outperforms the independent one over all ranges of CNR's for an entire range of λ values (e.g., $\lambda = 0.2$).

In comparing MR with a single-resolution scheme, we assume that the SR source coder is 16% more efficient than the MR coder. We take this as a worst-case figure, as based on empirical comparisons using a nonMR-friendly framework like JPEG [31]. Under these conditions, the SR channel could afford a 32 QAM modulation scheme for the same transmitter power as the MR 64 QAM scheme, due to a source compression advantage of roughly 5/6. For fairness of comparison, the SR scheme received the same thresholding (10^{-3}) as the coarse resolution packet stream, as they both achieve transitions from the region of no signal to the region of discernible signal.

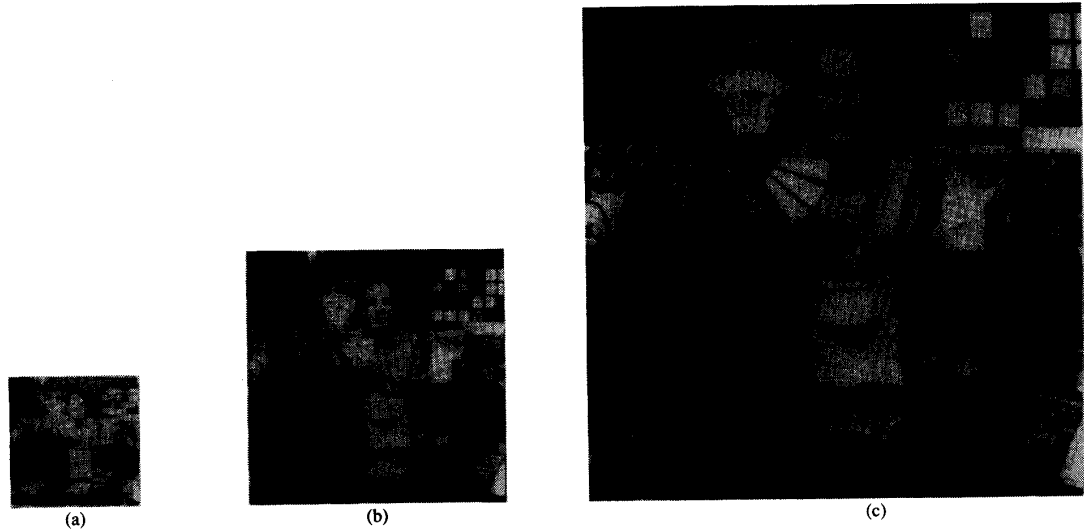


Fig. 20. Resolutions of the pyramid. (a) Coarsest layer. (b) Intermediate layer. (c) Full-resolution layer.

The results shown in Fig. 19 indicate the tradeoffs involved. As can be seen by comparing the SR scheme with, say, the MR-embedded scheme with $\lambda = 0.5$, the broadcast coverage area is much greater for the MR scheme at the price of some midregion suboptimality.

A point to note in favor of the MR scheme is the increase in full-resolution quality coverage area made possible by performing error concealment techniques to be described next. The SR scheme loses this advantage, as it has no coarse resolution channel to fall back upon.

B. Error Concealment

Due to the nature of the broadcast communication, it is impossible (or perhaps impractical) to achieve error-free transmission. Therefore, any real system has to be able to function in the presence of transmission errors. An error concealment scheme is often required to mask those errors and provide a gracefully degrading picture. Recursive systems (motion-compensated hybrid DCT being the typical example) cannot recover from errors until the next restart of the prediction loop.

The source coder we have used is based on a finite memory structure, and errors will not accumulate but die out within a few time samples. The structure used in conjunction with the MR modulation also allows very successful error concealment. Concealment strategies for single-resolution schemes are typically based on packet loss rates on the order of 10^{-5} (as in [32]), since transmission systems based on prediction loops are extremely fragile. However, in our MR scheme, we perform effective concealment at fine channel packet error rates exceeding 10^{-1} since the coarse channel is almost perfect (error rate $< 10^{-9}$) at the same CNR for typical values of λ (see Fig. 9).

Therefore, most of the errors will occur in the fine detail and a coarse version and motion vectors will be available for concealment. The concealment strategy differs slightly for the frames that are interpolated spatially or temporally, and

assumes that the information transmitted in the coarse channel is intact. The spatially interpolated frames of the finest layer are called anchor frames, as they have no temporal dependence on any other frame.

For the temporally interpolated frames, motion vectors and the selected interpolation mode for each block are available but the actual interpolation error (or residual) is lost. In the packetized transmission we have implemented, the typical region affected by a packet loss is a narrow strip 8 pixels in height and 1000–2000 pixels long. Because the encoder uses a smooth motion vector field (enforced by the hierarchical motion estimation algorithm) and both previous and next frames are available for interpolation, errors tend to be very small. Most artifacts show up as “blockiness” and are almost invisible, even in a still frame. Since these frames are not used to predict any other frames, the errors do not need to be processed further.

The errors are more visible, and potentially last longer for the spatially interpolated (anchor) frames. The artifacts appear as blurred blocks or decreased spatial resolution, and are clearly visible in still frame [see Fig. 22(b)]. Furthermore, since the previous and next frames (which are temporally interpolated) are based on this frame, errors can be annoying in real time. The concealment is based on replacing the region affected by the lost packet from the previous anchor frame. Since motion vectors are not available for this frame, an approximation is computed based on the motion vectors of the previous frame. Then, this is used to interpolate blocks from the previous anchor frame. This works very well in practice, as motion vectors resemble the true motion.

This concealment strategy gives excellent results even in extreme cases of packet loss. Complete loss of a frame can be tolerated, and sustained 15% packet loss rate causes no visible loss in quality. Fig. 21 shows the effect of 15% fine packet loss (obtained for $\lambda = 0.5$, CNR = 25.5 dB/symbol) on the spatial residual of the sequence with Fig. 21(c) showing the

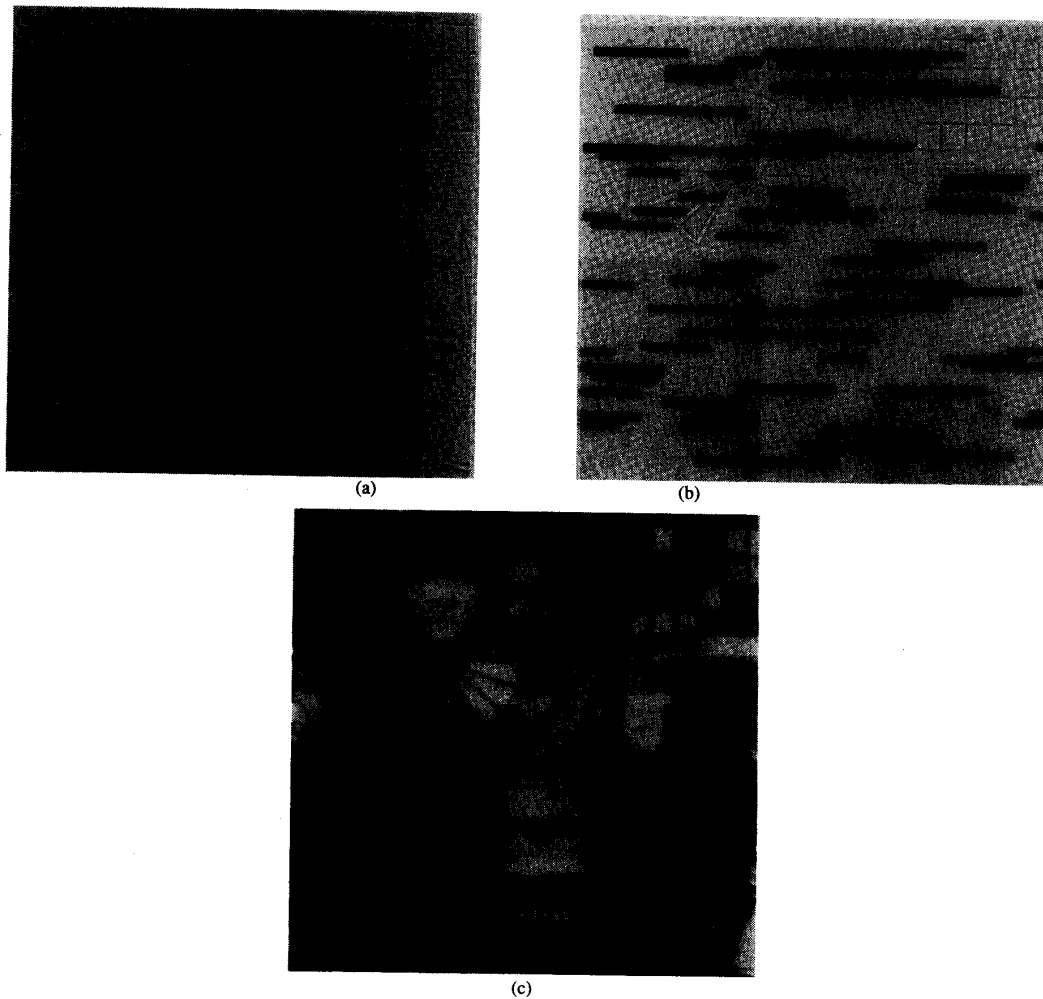


Fig. 21. Effect of channel noise for $\lambda = 0.5$, CNR = 25.5 dB/symbol (15% fine-channel packet loss). (a) Spatial residual frame. (b) Packets corrupted by channel. (c) Full-resolution reconstruction after error concealment. Image is 512×512 .

reconstructed quality, while Fig. 22 illustrates the power of error concealment in a MR environment.

VII. CONCLUSIONS

We have demonstrated a multiresolution (MR) joint source channel coding system, where using a source coder matched to an embedded trellis-coded modulation constellation (with/without error correction coding) has been shown to provide an efficient end-to-end MR system. The threshold effect plaguing single-resolution (SR) systems is softened by a stepwise graceful degradation reminiscent of analog systems without sacrificing the source coding advantage of digital schemes. We show the superiority of an embedded MR transmission scheme over independent transmissions of the MR source resolutions, and point out the tradeoffs in robustness and broadcast area coverage of low- and high-resolutions between embedded MR and SR digital systems for QAM constellations, highlighting the benefits of using joint MR source and channel coding.

APPENDIX

ANALYSIS OF MRQAM OF SECTION IV-B-1)

Let us introduce some definitions associated with Fig. 7.

- $S = \{j | j \in S \text{ and } i, j \text{ are in the same cloud}\}$, i.e., the set of all points which are in the same cloud as signal i .
- $N = |S|$ is the number of signals in the constellation.
- $D = \{\text{set of all "directions" (N, S, E, W) representing the one-sided independent degrees of freedom for the additive Gaussian noise, with unit directional vectors } (\mathbf{u}_N, \mathbf{u}_S, \mathbf{u}_E, \mathbf{u}_W) \text{ respectively.}\}$
- $C_i = \{j | j \in S \text{ and } i, j \text{ are in the same cloud}\}$, i.e., the set of all points which are in the same cloud as signal i .
- $d_{\text{intra}}^k(i) (d_{\text{inter}}^k(i)) = \text{half the Euclidean distance between } i \text{ and its nearest "fine" ("coarse") neighbor in the (positive) } \mathbf{u}_k \text{ direction. Thus, } \{d_{\text{intra}}^k(i)\} (\{d_{\text{inter}}^k(i)\}) \forall k \in D, \forall i \in S \text{ is the minimum instantaneous noise amplitude component in the } \mathbf{u}_k \text{ direction that will cause the receiver to incorrectly decode the intracloud (intercloud) information in that direction. Note also that if a signal point should have no neighbor in the positive}$

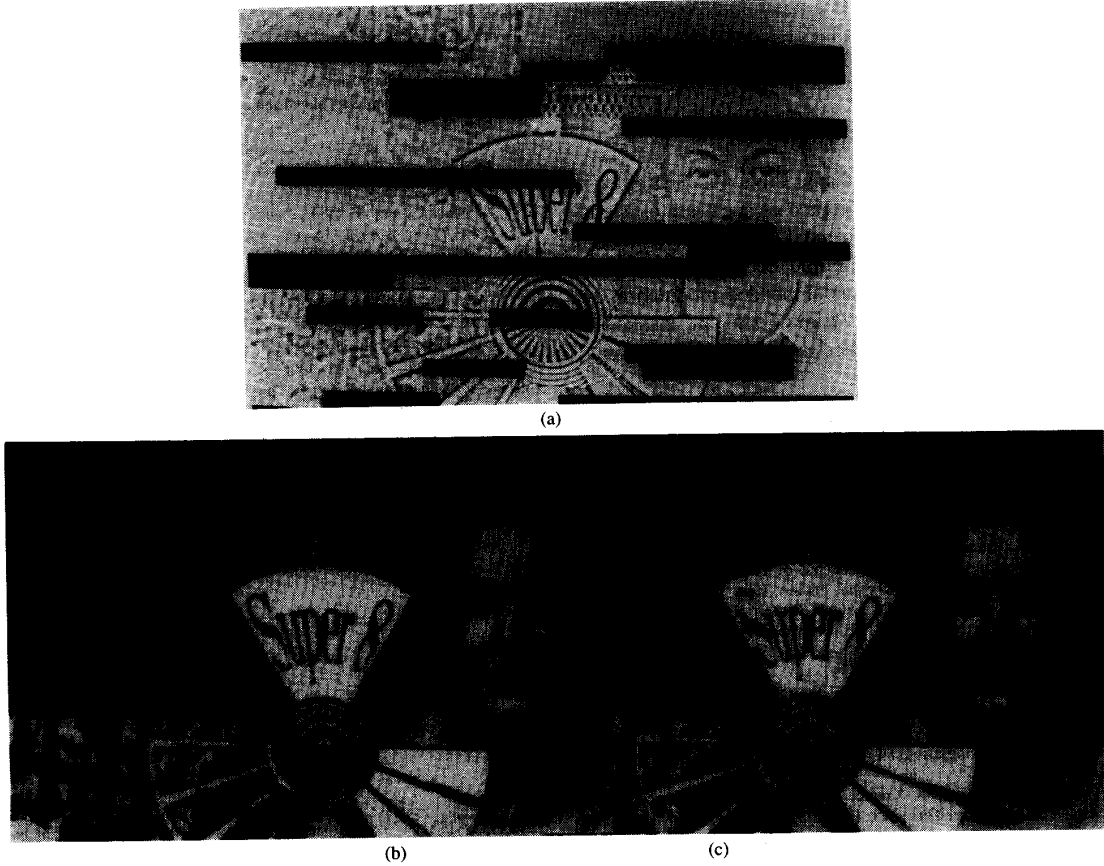


Fig. 22. Effect of error concealment for 15% fine-channel packet loss (blow up of Fig. 5). (a) Corrupted spatial residual frame. (b) Reconstruction without error concealment. (c) Reconstruction with error concealment.

u_k direction, then its corresponding nearest neighbor distance will be ∞ .

From these definitions, using the simple Gaussian error function we can obtain closed-form solutions to the probabilities of fine and coarse channel bit errors ($P_{e,b}^f$ and $P_{e,b}^c$, respectively). It is easy to show that the probability of fine bit error for a given λ and CNR are given by

$$P_{e,b}^f(\lambda, \text{CNR}) = \sum_{i \in S(\lambda, \text{CNR})} q(i) \sum_{k \in D} [0.5 * \text{erfc}(d_{\text{intra}}^k(i)/\sqrt{2})] \quad (1)$$

where $\text{erfc}(x)$ is the standard complementary Gaussian error function defined as

$$\text{erfc}(x) = (2/\sqrt{\pi}) \int_{-\infty}^x e^{-t^2} dt \quad (2)$$

and $q(i)$ in (1) refers to the symbol probabilities which, if assumed to be equal, would simplify it to

$$P_{e,b}^f(\lambda, \text{CNR}) = \frac{1}{2N} \sum_{i \in S(\lambda, \text{CNR})} \sum_{k \in D} \text{erfc}(d_{\text{intra}}^k(i)/\sqrt{2}). \quad (3)$$

Similarly, for the coarse bit stream, we have, assuming equiprobable symbols,

$$P_{e,b}^c(\lambda, \text{CNR}) = \frac{1}{2N} \sum_{i \in S(\lambda, \text{CNR})} \sum_{k \in D} \text{erfc}(d_{\text{inter}}^k(i)/\sqrt{2}). \quad (4)$$

In order to prevent error propagation, we packetize the streams into a composite length of L bits/packet comprising $L/3$ bits of coarse data and $2L/3$ bits of fine information (as demanded by the 1 : 2 ratio in coarse-to-fine bit rate). In the absence of ECC, we assume that a single bit error anywhere in an entire packet corrupts that packet and causes it to get lost. As was shown, due to the Karnaugh mapping, single bit errors will dominate. Defining the packet error probabilities as $P_{e,p}^c$ and $P_{e,p}^f$, respectively, for the coarse and fine channels, we have

$$P_{e,p}^f(\lambda, \text{CNR}) = 1 - (1 - P_{e,b}^f(\lambda, \text{CNR}))^{L/6} \quad (5)$$

and

$$P_{e,p}^c(\lambda, \text{CNR}) = 1 - (1 - P_{e,b}^c(\lambda, \text{CNR}))^{L/6} \quad (6)$$

See Fig. 9 for a plot of the curves for $L = 1080$ using the MR 64 QAM constellation for both coarse and fine packet

probability of loss performance as a function of the broadcast-area CNR for a multitude of λ values encompassing its region of definition from 0 to 1.

ACKNOWLEDGMENT

The authors would like to thank Prof. W. Schreiber of M.I.T. for pointing out the importance of spectrum efficiency for television broadcast. We also thank Dr. R. Calderbank of AT&T Bell Labs and Prof. G. Pottie of UCLA for fruitful discussions on multiplexed versus embedded modulation and TCM. Profs. W. Schreiber and J. Lim of MIT are thanked for providing the progressive sequence used in the simulations.

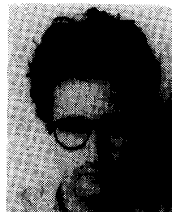
REFERENCES

- [1] W. F. Schreiber, "Considerations in the design of HDTV systems for terrestrial broadcasting," *Electron. Imag.*, Oct. 1990.
- [2] T. Cover, "Broadcast channels," *IEEE Trans. Inform. Theory*, vol. IT-18, pp. 2-14, Jan. 1972.
- [3] K. M. Uz, K. Ramchandran, and M. Vetterli, "Multiresolution source and channel coding for digital broadcast of HDTV," in *Proc. Fourth Int. Workshop on HDTV and Beyond*, Torino, Italy, Sept. 1991, vol. 1.
- [4] K. M. Uz, M. Vetterli, and D. LeGall, "Interpolative multiresolution coding of advanced television with compatible subchannels," *IEEE Trans. CAS for Video Technol.*, Spec. Iss. Signal Process. Advanc. Televis., vol. 1, pp. 86-99, Mar. 1991.
- [5] D. Anastassiou and M. Vetterli, "All digital multiresolution coding of HDTV," in *Proc. Nat. Assoc. Broadcast.*, Las Vegas, NV, Apr. 1991, pp. 210-216.
- [6] "Digital spectrum compatible HDTV system," AT&T/Zenith Tech. Rep., Sept. 1991.
- [7] A. Netravali, E. Petajan, S. Knauer, K. Matthews, R. Safranek, and P. Westerink, "A high performance digital HDTV codec," in *Proc. Nat. Assoc. Broadcast.*, Las Vegas, NV, Apr. 1991.
- [8] "Advanced digital television: System description," Sarnoff/NBC/Philips/Thomson Tech. Rep., Feb. 1991.
- [9] A. R. Calderbank and N. Seshadri, "Multilevel codes for unequal error protection," AT&T tech. memo., 1992.
- [10] W. F. Schreiber, "All-digital HDTV terrestrial broadcasting in the U.S.: Some problems and possible solutions," in *Proc. Workshop on Advanced Televis.*, ENST, Paris, France, May 1991.
- [11] J. Modestino, D. G. Daut, and A. Vickers, "Combined source-channel coding of images using the block cosine transform," *IEEE Trans. Commun.*, vol. COM-29, pp. 1261-1274, Sept. 1981.
- [12] G. Karlsson and M. Vetterli, "Sub-band coding of video for packet networks," *Opt. Eng.*, vol. 27, pp. 574-586, July 1988.
- [13] K. Fazel and J. L'Huilier, "Application of unequal error protection codes on combined source-channel coding," in *Proc. IEEE Int. Conf. Commun.*, Atlanta, GA, Apr. 1990, pp. 320.5.1-320.5.6.
- [14] M. Vetterli and K. M. Uz, "Multiresolution coding techniques for digital video: A review," *Multidimen. Syst. and Signal Proces.: Spec. Iss. on Multidimens. Proces. of Video Signals*, vol. 3, pp. 161-187, Mar. 1992.
- [15] C. E. Shannon, "A mathematical theory of communication," *Bell Syst. Tech. J.*, vol. 27, pp. 379-423, 1948.
- [16] R. M. Gray and A. D. Wyner, "Source coding for a simple network," *Bell Syst. Tech. J.*, vol. 53, pp. 1681-1721, Nov. 1974.
- [17] W. H. Equitz and T. M. Cover, "Successive refinement of information," *IEEE Trans. Inform. Theory*, vol. 37, pp. 269-275, Mar. 1991.
- [18] J. M. Shapiro, "An embedded wavelet hierarchical image coder," in *Proc. ICASSP '92*, Mar. 1992, pp. 657-660.
- [19] A. E. Gamal and T. Cover, "Multiple user information theory," *IEEE Trans. Inform. Theory*, vol. 68, pp. 1466-1483, Dec. 1980.
- [20] G. W. Wornell and A. V. Oppenheim, "Wavelet-based representations for a class of self-similar signals with applications to fractal modulation," *IEEE Trans. Inform. Theory*, vol. IT-38, pp. 785-800, Mar. 1992.
- [21] M. M. Mano, *Digital Logic and Computer Design*. Englewood Cliffs, NJ: Prentice-Hall, 1979.
- [22] J. G. Proakis, *Digital Communications*. New York: McGraw-Hill, 1989.
- [23] G. Ungerboeck, "Channel coding with multilevel/phase signals," *IEEE Trans. Inform. Theory*, vol. IT-28, pp. 55-67, Jan. 1982.
- [24] L.-F. Wei, "Trellis-coded modulation with multidimensional constellations," *IEEE Trans. Inform. Theory*, vol. IT-33, pp. 483-501, July 1987.
- [25] A. R. Calderbank and N. J. A. Sloane, "Four-dimensional modulation with an eight-state trellis code," *AT&T Tech. J.*, vol. 64, pp. 1005-1018, May 1985.
- [26] T.-H. Chiang and D. Anastassiou, "Two-layer coding of interlaced HDTV for graceful degradation," in *Proc. Fifth Int. Workshop on HDTV*, Kawasaki, Japan, Nov. 1992, vol. II, pp. 70.1-70.8.
- [27] B. Masnick and J. Wolf, "On linear unequal error protection codes," *IEEE Trans. Inform. Theory*, vol. IT-13, pp. 600-607, Oct. 1967.
- [28] T. Kasami, S. Lin, V. Wei, and S. Yamamura, "Coding for the binary symmetric broadcast channel with two receivers," *IEEE Trans. Inform. Theory*, vol. IT-31, pp. 616-625, Sept. 1985.
- [29] M.-C. Lin, C.-C. Lin, and S. Lin, "Computer search for binary cyclic UEP codes of odd length up to 65," *IEEE Trans. Inform. Theory*, vol. 36, pp. 924-935, July 1990.
- [30] K. Ramchandran, A. Ortega, and M. Vetterli, "Efficient quantization for a multiresolution HDTV source coder," in *Proc. Fifth Int. Workshop on HDTV*, Kawasaki, Japan, Nov. 1992, vol. II, pp. 87.1-87.8.
- [31] JPEG Tech. Spec., Revision (DRAFT), Joint Photographic Experts Group, ISO/IEC JTC1/SC2/WG8, CCITT SGVIII, Aug. 1990.
- [32] F.-C. Jeng and S. H. Lee, "Concealment of bit error and cell loss in inter-frame coded video transmission," in *Proc. IEEE Int. Conf. Commun.*, pp. 496-500, 1991.
- [33] G. Pottie, private communication.



Kannan Ramchandran was born in Madras, India, in 1961. He received the B.S. degree in electrical engineering from the City College of New York in 1982, and the M.S. and M.Phil. degrees in electrical engineering from Columbia University, New York, NY, in 1984 and 1992, respectively.

From 1982 to 1984, he worked for IBM as a Microprocessor Systems Designer. In 1984, he joined AT&T Bell Labs as a Member of the Technical Staff, where he designed digital line multiplexers, phase locked loops, and custom transmission IC's in optical fiber loop carrier systems and fiber-to-the-home development systems. He is currently on educational leave from AT&T to finish his Ph.D. dissertation at the Center for Telecommunications Research at Columbia. His research interests include multiresolution signal processing, fast algorithms and architectures for image and video coding, digital communications, wavelets, and packet video.



Antonio Ortega was born in Madrid, Spain, in 1965. He received the Telecommunication Engineering degree from the Universidad Politécnica de Madrid (UPM), Madrid, Spain, in 1989, and is currently working toward the Ph.D. degree at Columbia University, New York, NY.

In 1986 and 1987, he worked during the summer for Gibbs & Hill, Inc. in Washington, D.C. In 1988, he was a Summer Intern at the Optoelectronics Lab, ENST, Paris, France. During 1990, he was a Research Assistant at the Image Processing Group at UPM. His Ph.D. work is being supported by a scholarship from the Fulbright Commission and the Spanish Ministry of Education.



K. Metin Uz was born in Istanbul, Turkey, on Oct. 11, 1965. He received the B.S. degrees in electrical engineering and physics from Bogazici University, Istanbul, Turkey, in 1987, and the M.S. and Ph.D. degrees from Columbia University, New York, NY, in 1988 and 1992, respectively.

Between 1987 and 1991, he was employed as a Research Assistant at the Center for Telecommunications Research, Columbia University. He is currently with the David Sarnoff Research Center's Digital HDTV Group. His research interests include multidimensional signal processing, multiresolution systems, and algorithms and architectures for image and video coding.



Martin Vetterli (S'86-M'86-SM'90) was born in Switzerland in 1957. He received the Dipl. El.-Ing. degree from the Eidgenössische Technische Hochschule Zürich, Switzerland, in 1981, the Master of Science degree from Stanford University, Stanford, CA, in 1982, and the Doctorat ès Science degree from the Ecole Polytechnique Fédérale de Lausanne, Switzerland, in 1986.

In 1982, he was a Research Assistant at Stanford University, and from 1983 to 1986 he was a Researcher at the Ecole Polytechnique. He has worked for Siemens and AT&T Bell Laboratories. In 1986, he joined Columbia University, New York, NY, where he is currently Associate Professor of Electrical Engineering, member of the Center for Telecommunications Research, and Co-Director of the Image and Advanced Television Laboratory. He is a member of the MDSP Committee of the IEEE Signal Processing Society and of the Editorial Boards of *Signal Processing*, *Image Communication*, and *Annals of Telecommunications*. He received the Best Paper Award of EURASIP in 1984 for his paper on multidimensional subband coding, the Research Prize of the Brown Boveri Corporation (Switzerland) in 1986 for his thesis, and the IEEE Signal Processing Society's 1991 Senior Award (DSP Technical Area) for a 1989 Transactions paper with D.LeGall on filter banks for subband coding. His research interests include multirate signal processing, wavelets, computational complexity, signal processing for telecommunications, and digital video processing.

Dr. Vetterli is a member of SIAM and ACM.