Practical Coding for QAM Transmission of HDTV

Chris Heegard, *Member, IEEE*, Scott A. Lery, *Member, IEEE*, and Woo H. Paik, *Member, IEEE*

***Abstract*—This paper describes a practical approach to digital transmission of compressed HDTV. We demonstrate how modulation schemes based on QPSK modulation can be directly incorporated into QAM-based modulation systems. We shall argue that this leads directly to an easily implementable structure that is both efficient in bandwidth and data reliability.**

The use of a concatenated code is known to provide an effective and practical approach to achieving low BER, high data rate, and modest implementation complexity. It is our contention that the correct solution to the concatenated coding problem for HDTV transmission is to simply extend the modulation codes developed for QPSK - to - QAM modulation.

In nonconcatenated situations, a trellis code based on a binary code at rate 2/3 is usually best; this fact follows from the study of the asymptotic coding gain of a trellis code. However, this is not the case for higher error rates at the output of the trellis decoder (e.g., when a symbol error correcting decoder follows as in a concatenated code). The reason for this follows from an analysis of the effect of the number of "nearest neighbors" on the error rate.

A four-way partition of QAM is a natural extension of QPSK modulation; it is a simple matter to incorporate any good QPSK code into a trellis coding scheme for QAM modulation. We propose a concatenated coding scheme based on QPSK trellis codes and symbol error correcting coding. A specific example is presented which shows the advantages of this approach.

I. Introduction

THIS paper describes a practical approach to the digital transmission of compressed high definition television (HDTV). The transmission system for this application has the following requirements.

- Data rate: 15–30 Mb/s
- Bandwidth occupancy: 6 MHz
- Data reliability: one error event per minute
- Receiver complexity: low cost in volume production

The data rate requirement arises from the need to provide a high-quality compressed television picture. The bandwidth constraint is a consequence of the U.S. Federal Communications Commission (FCC) requirement that HDTV signals occupy existing television channels; they must coexist with the current broadcast National Television System Committee (NTSC) signals. This combination of data rate and bandwidth occupancy requires a modulation system that has high bandwidth efficiency; the number of transmitted bits per second per unit of bandwidth (i.e., the ratio of data rate to bandwidth) must be on the order of 3 to 5. This means that modulation systems such as quadrature phase shift keying (QPSK), a common scheme for satellite transmission systems (which are usually "power limited"), is unsuitable because its bandwidth efficiency without coding is 2. A more bandwidth-efficient modulation (for "bandlimited" transmission-like terrestrial and cable video systems), most notably quadrature amplitude modulation (QAM), is required.

Manuscript received June 1992; revised August 1992.

C. Heegard is with the School of Electrical Engineering, Cornell University, Ithaca, NY.

S. Lery and W. H. Paik are with the VideoCipher Division, General Instrument Corporation, San Diego, CA 92121.

IEEE Log Number 9204409.

On the other hand, since QPSK systems are so well established, coded modulation schemes for such systems are well understood and routinely implemented. Typically, a binary convolutional code at rate 1/2 (or the same code "punctured" to some higher rate [11], [12]) is incorporated as the modulation code. As a consequence, integrated circuits that realize trellis-coded QPSK modulation are readily available and easily obtained. In this paper, we demonstrate how modulation schemes based on QPSK modulation can be directly incorporated in QAM-based modulation systems. We shall argue that this leads directly to an implementable structure that is both efficient in bandwidth and data reliability.

The need for very high data reliability follows from the fact that highly compressed source material (i.e., compressed video) is intolerant of channel errors; the natural redundancy of the signal has been removed in order to obtain a concise description of the intrinsic value of the source data.

Low error rate requirements are met in practice via the use of a concatenated coding approach (divide and conquer), as depicted in Fig. 1. In such a coding framework, two codes are employed: an "inner" modulation code and an "outer" symbol error-correcting code. The inner code is usually a "coded modulation" that can be effectively decoded using "soft decisions" (i.e., finely quantized channel data). The inner code "cleans up" the channel and exploits the soft decision nature of the received signal. The output of the inner code delivers a small (but unacceptably high) symbol error rate to the outer decoder. This second decoder then removes the vast majority of symbol errors that have eluded the inner decoder in such a way that the final output error rate is extremely small.

The standard concatenated coding approach is to use a convolutional or trellis code [1], [2], [4], [7]–[9] as the inner code with some form of the Viterbi algorithm [3] as the trellis decoder. The outer code is most often a "t error correcting" Reed–Solomon code [2], [4] over a finite field with 2^q symbols (q is usually on the order of 5–10). Such Reed–Solomon coding systems, that operate in the required data rate range, are widely available and have been implemented in the integrated circuits of several vendors.

The optimization of the modulation code for concatenated and nonconcatenated coding systems can lead to different solutions. In a concatenated coding system, one needs to consider the required error rate of the modulation (inner)

0733–8716/93$03.00

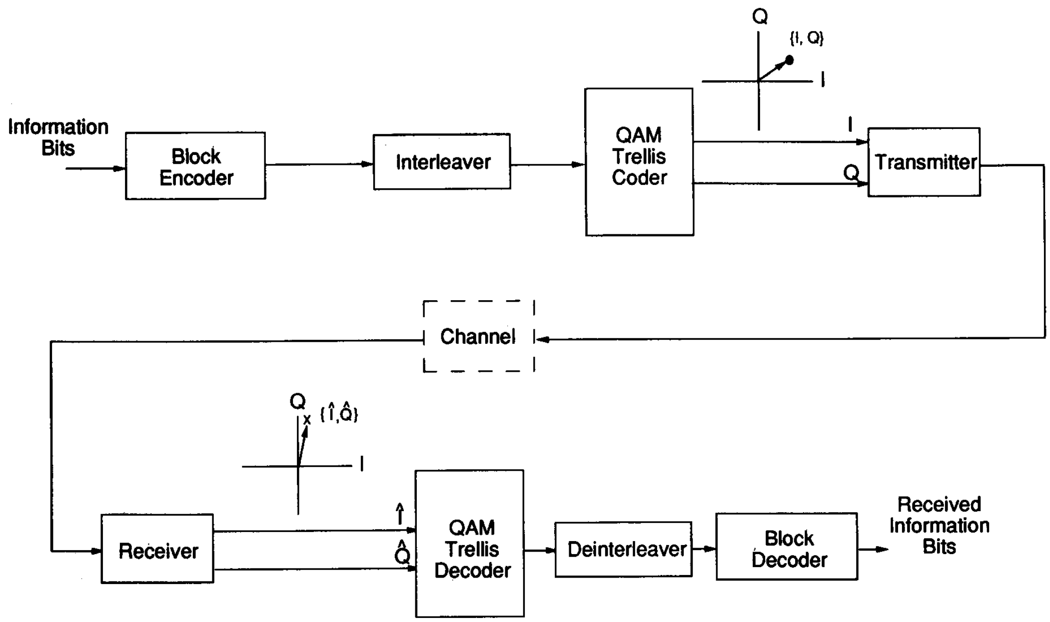


Fig. 1. QAM transmission system using concatenated coding.

code to achieve a specified bit or block (codeword) error rate from the outer code. In a nonconcatenated coding system, the required error rate of the modulation code is usually much lower than in a concatenated coding system. For example, modulation code A may perform better at "low" signal-to-noise ratio (SNR) where the error rate is "large" than modulation code B, which performs better at "high" SNR where the error rate is "small." Code A may be the better choice for a concatenated coding system, and code B may be the better choice for a nonconcatenated coding system.

In light of this, it is our contention that simple extensions of modulation codes developed for QPSK-to-QAM modulation provide the correct solution to the concatenated coding problem. This is true even though these extensions are known to be weaker than other known modulation codes used in nonconcatenated systems (i.e., when used in the domain where the output error rate of the modulation code is "small").

The organization of this paper is as follows. In Section II, a brief description of trellis/modulation coding is given. In Section III, the optimization of coding gain for trellis codes used in concatenated and nonconcatenated systems is discussed. In Section IV, QPSK-based trellis codes are described in detail, along with a trellis decoder implementation. In Section V, performance comparisons between QPSK-based trellis codes and Ungerboeck codes are shown with and without outer coding. A short summary in Section VI concludes the paper.

II. Trellis/ Modulation Coding

In uncoded QAM transmission, n bits per symbol are transmitted by mapping n data bits onto the 2^N points of a QAM constellation. Thus, in the uncoded case, the number of data bits is equal to the number of input bits of the QAM modulator $N = n$. In a trellis code [1], [7]–[9], the constellation is expanded by one bit (i.e., the constellation size is doubled), and the number of data bits per symbol is one less than the number of input bits of the QAM modulator $N = n + 1$. This expansion of the signal constellation is what allows for a coding gain to be achieved (i.e., it allows for redundancy to be introduced in the transmitted signal).[1]

A two-dimensional trellis code for the transmission of n bits per QAM symbol is obtained by the combination of an n-bit input, an $n+1$ bit output, a 2^{ν}-state finite state machine (FSM) (i.e., encoder), and an $N = n + 1$ bit QAM mapper (i.e., a 2^N point QAM mapper/modulator), as depicted in Fig. 2. As discovered by Ungerboeck [7], the most economical approach to this problem involves two components. The signal constellation is partitioned into 2^{N-m} subsets, each of size 2^m, in such a way that the distance between points within each subset is maximized.[2] The n input bits are split into k "coded" bits and $m = n - k$ "uncoded" bits. The k-bit coded data is then encoded by a FSM with $k + 1$ output bits (i.e., redundancy in time is introduced) and used to select the subset, while the m-bit uncoded data is used to select the point within the subset selected by the FSM. One way of labeling the QAM constellation points, corresponding to a mapping of the $n + 1$ bits to a QAM constellation point, is described in [7]–[9].

[1] In the "pragmatic" approach described in [10], QAM modulation is obtained from the one-dimensional PAM model. This approach leads to a quadrupling of the QAM signal constellation $N = n + 2$. In many applications, this extra expansion is undesirable.

[2] Caulderbank and Sloane [1] have shown that Ungerboeck's method of set partitioning is best described in terms of lattices and their cosets.

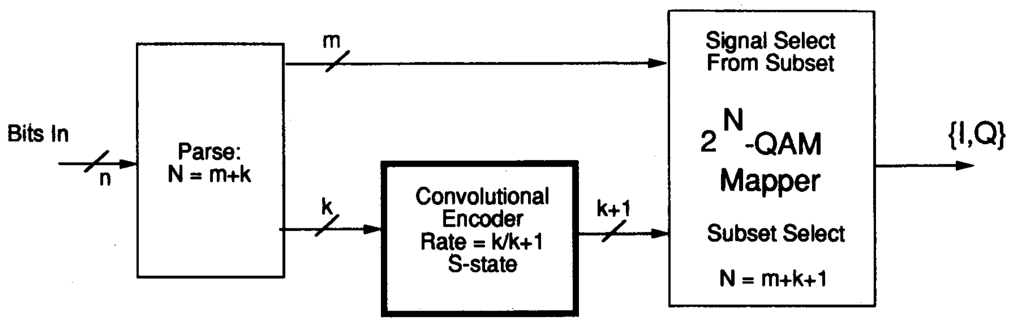


Fig. 2. Trellis encoder.

III. Coding Gain

A. Asymptotic Coding Gain

In nonconcatenated situations, a trellis code based on a binary code at rate 2/3 (not a punctured rate 1/2 code) is usually the best solution.[3] This fact follows from the study of the asymptotic coding gain (ACG) of a trellis code.

The ACG, γ, of a QAM-based trellis code is given by [8]

$$\gamma = \frac{d^2_{\text{free}}(\text{coded})/Ec(\text{coded})}{\Delta_o^2/Ec(\text{uncoded})} \tag{1}$$

where Ec(coded) and Ec(uncoded) denote the average constellation energies of the coded and uncoded schemes, respectively, and Δ_o is the minimum spacing of the QAM constellation points. The free Euclidean distance of the trellis code is given by [8]

$$d^2_{\text{free}}(\text{coded}) = \Delta_o^2 \min\{d^2_{\text{free}}[\text{FSM}(k,\nu)], 2^{k+1}\} \tag{2}$$

where the Euclidean distance of the FSM (i.e., the convolutional encoder) is $d^2_{\text{free}}[\text{FSM}(k,\nu)]$. The free distance of the FSM depends on the structure of the encoder.

For a given number of encoder states parameterized by ν and the number of inputs given by k, the encoders that maximize the FSM free distance have been tabulated for small values of ν [1], [8]. The results show why $k = 2$ (i.e., an encoder FSM at rate 2/3) is the most practical for maximizing the ACG. For a given value of k and small ν, the ACG is determined by the FSM

$$\gamma = \frac{Ec(\text{uncoded})}{Ec(\text{coded})} d^2_{\text{free}}[\text{FSM}(k,\nu)]. \tag{3}$$

However, because the ACG involves a minimum as the complexity of the encoder (as measured by ν) is allowed to grow, it becomes

$$\gamma = \frac{Ec(\text{uncoded})}{Ec(\text{coded})} 2^{k+1}. \tag{4}$$

Thus, for a given value of k, there is a natural value of trellis complexity $\nu^*(k)$ such that the ACG is not improved by making $\nu > \nu^*(k)$. We note that $\nu^*(k)$ is monotonically increasing in k, since the rate of the encoder $k/(k+1)$ increases with increasing k. Now from [9] it is seen that, for QAM trellis coding, $\nu^*(1) = 2$ and $\nu^*(2) = 7$. Thus, for a four-way partition ($k = 1$, rate 1/2 encoding), the maximum ACG (equal to 3.01 dB) is achieved with four states while, for an eight-way partition ($k = 2$, rate 2/3 encoding), the maximum ACG (equal to 6.02 dB) is achieved with 128 states. Note that since the complexity of the decoder (with Viterbi decoding [2], [4], [7]) depends on the number of states of the encoder, a 16-way partition ($k = 3$, rate 3/4 encoding) may not be practical since the number of states required to achieve a large ACG might be prohibitive.

[3]This assumes a two-dimensional trellis code [8].

B. Optimization of Coding Gain

The conclusion from the above discussion is that if the ACG is to be maximized, then the obvious practical choice for trellis coding is an eight-way partition of the QAM constellation with a rate 2/3 encoder. Furthermore, in nonconcatenated systems, where the output error rate of the trellis code is to be small, maximizing the ACG is the appropriate thing to do. However, this is not the case for higher error rates at the output of the trellis decoder (e.g., when a symbol error correcting decoder follows). The reason for this follows from an analysis of the effect on the error rate of the number of "nearest neighbors."

The probability of error, P_{sym}, at the output of the trellis decoder can be predicted by the behavior of the formula [6], [8]

$$P_{\text{sym}} \approx M_{nn} Q\left[\frac{d_{\text{free}}(\text{coded})}{\sqrt{2N_o}}\right] \tag{5}$$

or, using (1) and (2),

$$P_{\text{sym}} \approx M_{nn} Q\left[\Delta_o \frac{\sqrt{\gamma\rho}}{\sqrt{2Ec(\text{uncoded})}}\right] \tag{6}$$

where the Q-function is

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty \exp(-y^2/2)\, dy.$$

M_{nn} is the average number of nearest neighbors, N_o is the one-sided spectral density of the additive Gaussian noise, and ρ is the signal-to-noise ratio (i.e., the energy per transmitted symbol divided N_o).

For "high" signal-to-noise ratio ρ, the error probability is more seriously effected by the ACG, γ, than the number of nearest neighbors M_{nn}. However, in the domain of "low" ρ, this is simply not the case; the number of the nearest neighbors has a significant effect. Thus, a code with a smaller ACG and smaller M_{nn} may be more reliable than a code optimized for the ACG.

For example, for the rate 2/3 32-QAM Ungerboeck code with 16 states, the ACG is 4.77 dB [9]. However, the number of nearest neighbors is estimated to be 56 [9]. On the other hand, if a strong QPSK-based code modulation code is used with a four-way partition of 32-QAM, the ACG is only 3.01 dB yet the number of nearest neighbors is 2.5. (The free distance of this code is determined by the "uncoded" bits. A four-way partition of 32-QAM yields four subsets, each with eight points. Among these eight points, five have two nearest neighbors, two have three nearest neighbors, and one has four nearest neighbors. So the average number of nearest neighbors is 2.5.) For low signal-to-noise ratios, the latter code has a smaller probability of error.

It is this simple realization that leads to the conclusion that, in fact, a four-way partition with $\nu \gg \nu^*(1) = 2$ is a very efficient method of trellis coding in a concatenated coding system. Furthermore, a four-way partition of QAM is a natural extension of QPSK modulation. It is a simple matter to incorporate any good QPSK code into a trellis coding scheme for QAM modulation.

IV. Practical QPSK-Based Trellis Code

Two issues showing how a QPSK code is incorporated into a QAM modulation system are detailed. The first addresses transmission (encoding): how the "codewords" of the QPSK code and the "uncoded" bits are assigned to the QAM constellation. The method described has the following desirable features: 1) it addresses the 90° phase ambiguity of QAM; and 2) the most significant digits control the constellation size. The second issue involves the decoder: how the received signal is prepared for decoding by the soft-decision QPSK decoder, and how the "uncoded" bits are decided.

A. Labeling of QAM Points

For purposes of QAM transmission, the codewords of the QPSK code and the uncoded bits must be assigned to the QAM constellation. This is accomplished by labeling the QAM constellation points by a modulation function $\mathrm{MOD}(\boldsymbol{m}) \in R^2$,

$$(\mathrm{mod}:)\{0,1\}^N \rightarrow R^2.$$

The method described has the following desirable features: 1) the consequences of the 90° phase ambiguity of QAM is imposed on the QPSK codewords, while the uncoded bits are invariant to the ambiguity (i.e., the 90° phase ambiguity can be dealt with in the same manner as the QPSK system); and 2) the most significant digits control the constellation size (i.e., a nested scheme for 16/32/64-QAM).

Consider the labeling (modulation function $\mathrm{MOD}(\boldsymbol{m})$) given in Fig. 3, and depicted in Fig. 4. The outputs of the QPSK encoder form the least significant bits (LSB's), m_1, m_0, of the constellation label; the LSB's select the column of the matrix. The most significant bits (MSB's) determine the constellation size. With no uncoded bits, QPSK is generated; with two uncoded bits, 16-QAM is generated; with three uncoded bits, 32-QAM is generated; and with four uncoded bits, 64-QAM is generated. Furthermore, the effect of rotating the QAM constellation by 90° is to rotate the columns of the matrix

$$00 \rightarrow 01 \rightarrow 11 \rightarrow 10 \rightarrow 00$$

which leaves the rows invariant. Thus, the label of the uncoded bits is unaffected by 90° rotations. The handling of the 90° phase ambiguity at the receiver (decoder) is left solely to the QPSK encoder. The same method used to resolve the ambiguity with a QPSK receiver can be incorporated into the QAM system using this labeling. For example, differential encoding could be used if the QPSK code is itself rotationally invariant.

As a final note, the assignment of the two coded bits, m_1, m_0, to the four constellation subsets is such that the intersubset Hamming distance is proportional to the intersubset Euclidean distance squared (the proportionality factor is Δ_o^2, the square of the minimum spacing of the constellation) as is normally done in coded QPSK systems. (See Fig. 4 for the coded bit assignment.)

B. Pruning and Decoding

Consider the process of signal detection when a soft decision QPSK decoder is incorporated into a system employing the previously described QAM modulator. First, in hard decision detection of QPSK or QAM signals, the received signal $y_k = x_k + w_k$ is quantized, where the signal x_k belongs to the QPSK or QAM constellation (i.e., in the range of $\mathrm{MOD}[\boldsymbol{m}]$), and w_k is the noise. The quantization function provides an estimate of both the signal x'_k and the data $\boldsymbol{m}'$ according to the relation $x'_k = \mathrm{MOD}(\boldsymbol{m}')$. For maximum likelihood (ML) detection, the log-likelihood function $-\log(p(y_k|\mathrm{MOD}(\boldsymbol{m})))$ is minimized over the possible messages $\boldsymbol{m} \in \{0,1\}^N$, where $p(y_k|x_k)$ is the conditional probability of receiving y_k given that x_k was transmitted. For random messages, ML detection minimizes the probability of error. The most common method of quantization is nearest (Euclidean) neighbor detection, which satisfies

$$\|y_k - x'_k\|^2 = \min \|y_k - \mathrm{MOD}(\boldsymbol{m})\|^2$$

where the minimum is taken over $\boldsymbol{m} \in \{0,1\}^N$, and $\| \bullet \|^2$ is the Euclidean distance squared (i.e., the sum of squares). In the case of additive Gaussian noise, nearest neighbor detection is ML.

In coded QPSK and QAM systems, soft decision information should be provided to the decoder for more effective decoding of the codeword. This soft decision information is often described as a symbol metric, which indicates the quality of deciding a particular symbol was sent when y_k is received. For nearest neighbor decoding, the metric of choice is

$$\mathrm{metric}(y_k; \boldsymbol{m}) = \|y_k - \mathrm{MOD}(\boldsymbol{m})\|^2.$$

(In practice, the metric itself is quantized for purposes of implementation.) In QPSK, for example, for each possible

For QPSK with m5=m4=m3=m2=0, and 16-QAM:

$$\begin{array}{c|c}
 & m1m0 \\
m5\,m4\,m3\,m2 & \begin{array}{cccc} 00 & 01 & 11 & 10 \end{array} \\
\begin{array}{c} 0000 \\ 0001 \\ 0011 \\ 0010 \end{array} & \begin{bmatrix} +1,+1 & -1,+1 & -1,-1 & +1,-1 \\ +1,-3 & +3,+1 & -1,+3 & -3,-1 \\ -3,-3 & +3,-3 & +3,+3 & -3,+3 \\ -3,+1 & -1,-3 & +3,-1 & +1,+3 \end{bmatrix}
\end{array}$$

For 32-QAM add:

$$\begin{array}{c|c}
 & m1m0 \\
m5\,m4\,m3\,m2 & \begin{array}{cccc} 00 & 01 & 11 & 10 \end{array} \\
\begin{array}{c} 0100 \\ 0101 \\ 0111 \\ 0110 \end{array} & \begin{bmatrix} +5,-3 & +3,+5 & -5,+3 & -3,-5 \\ +1,+5 & -5,+1 & -1,-5 & +5,-1 \\ +5,+1 & -1,+5 & -5,-1 & +1,-5 \\ -3,+5 & -5,-3 & +3,-5 & +5,+3 \end{bmatrix}
\end{array}$$

For 64-QAM add:

$$\begin{array}{c|c}
 & m1m0 \\
m5\,m4\,m3\,m2 & \begin{array}{cccc} 00 & 01 & 11 & 10 \end{array} \\
\begin{array}{c} 1100 \\ 1101 \\ 1111 \\ 1110 \\ 1000 \\ 1001 \\ 1011 \\ 1010 \end{array} & \begin{bmatrix} +5,+5 & -5,+5 & -5,-5 & +5,-5 \\ +5,-7 & +7,+5 & -5,+7 & -7,-5 \\ -7,-7 & +7,-7 & +7,+7 & -7,+7 \\ -7,+5 & -5,-7 & +7,-5 & +5,+7 \\ -3,-7 & +7,-3 & +3,+7 & -7,+3 \\ -7,+1 & -1,-7 & +7,-1 & +1,+7 \\ +1,-7 & +7,+1 & -1,+7 & -7,-1 \\ -7,-3 & +3,-7 & +7,+3 & -3,+7 \end{bmatrix}
\end{array}$$

Fig. 3. Modulation function for 16/32/64-QAM.

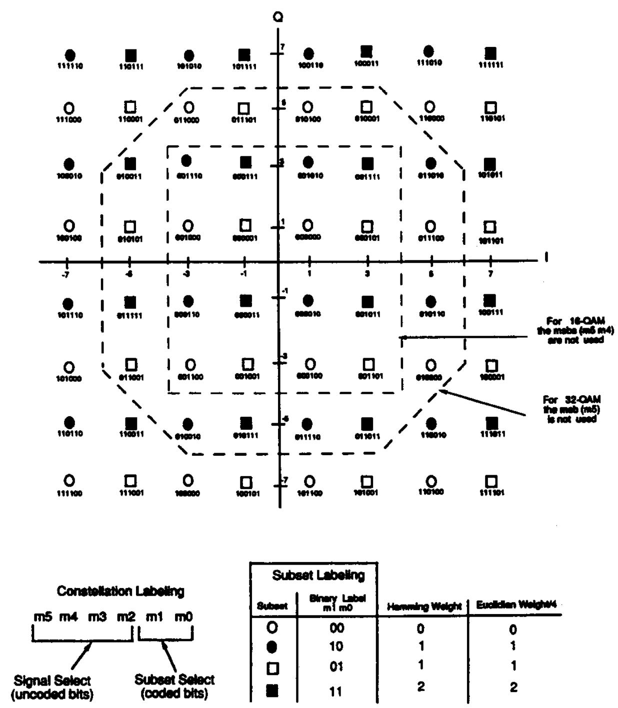


Fig. 4. 16/32/64-QAM mapping.

message $m_1, m_0 \in \{0,1\}^2$, the nearest neighbor metric $\|y_k - \text{MOD}(m_1, m_0\|^2)$ is the ML metric for additive Gaussian noise.

In coded QAM modulation based on a soft-decision decodable QPSK code, four symbol metrics must be supplied to the decoder as well as four conditional hard decisions ("uncoded" bits). For nearest neighbor detection, for each choice of $m_1, m_0 \in \{0,1\}^2$

$$\begin{aligned}\text{metric}\,(y_k; m_1, m_0) \\ = \min \|y_k - \text{MOD}\,(m_{N-1}, \cdots, m_2, m_1, m_0)\|^2\end{aligned}$$

where the minimum is taken over $m_{N-1}, \cdots, m_2 \in \{0,1\}^{N-2}$. The conditional hard decisions correspond to the choice of $m_{N-1}, \cdots, m_2$ that obtain the minimum. The process of determining the symbol metrics and conditional hard decisions is known as pruning. In trellis-coded QAM, the uncoded bits appear as "parallel" branches of the trellis; the computation of the symbol metrics and conditional hard decisions act to prune all but the single best branch from the set of parallel branches.

Once the pruning operation has been completed, the soft decision information is presented to the decoder of the QPSK code. During this time, the conditional hard decisions are stored (delayed) until QPSK decisions become available. The

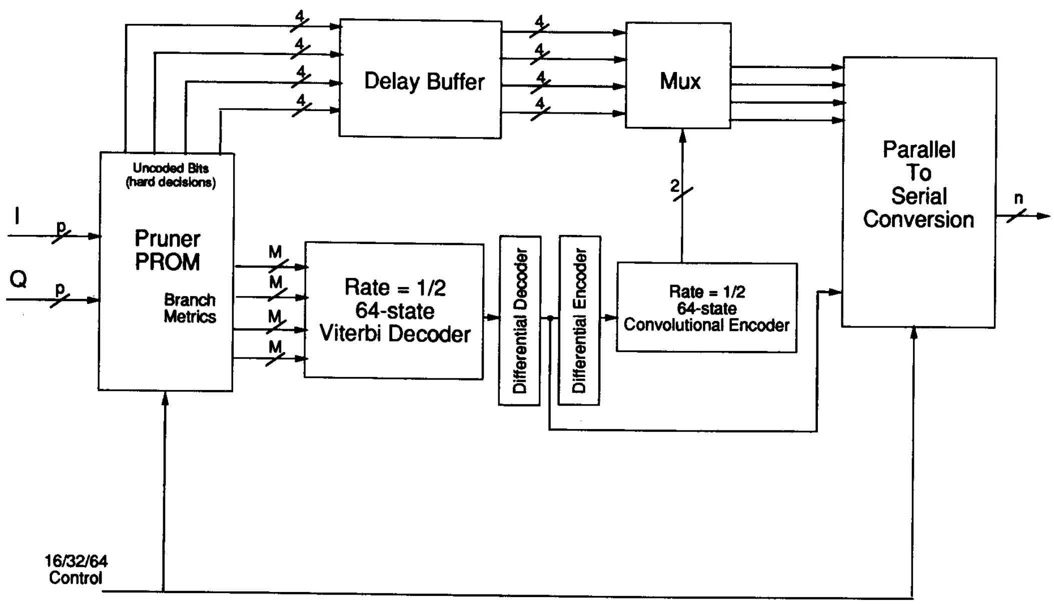


Fig. 5. Trellis decoder for 16/32/64-QAM.

QPSK decoder, using the soft decision information, decodes the QPSK information (i.e., m_1, m_0). The remaining information (i.e., $m_{N-1}, \cdots, m_2$) is then decided using the decoded QPSK information and the previously stored conditional hard decisions.

Note that if the QPSK decoder is ML (for QPSK modulation), then the pruning/QPSK decoding method is also ML. For example, if the QPSK code is a binary convolutional code with nearest neighbor (Viterbi) decoding, then the aforementioned QAM decoding algorithm is also nearest neighbor (i.e., finds the closest codeword to the received sequence) [7]–[9].

Fig. 5 shows a decoder implementation for 16/32/64-QAM. Notice that QPSK-based trellis codes allow a very practical trellis decoder design. The Viterbi decoder can be the "standard" off-the-shelf variety instead of a custom part (as would be the case if, say, a rate 2/3 convolutional code were used).

V. Performance Results

Using (1), (2), and (6) with $\Delta_o = 2$ (i.e., constellations based on the odd integer lattice), the performance of QAM-based trellis coded modulation (TCM) using two different convolutional codes is plotted in Fig. 6. The rate 1/2 code is the "standard" 64-state code with octal generator vector [171 133] found by Odenwalder [5]. The rate 2/3 code is a 16-state code with octal generator matrix rows [5 1 2] and [2 7 0] found by Ungerboeck [9]. In this paper, the rate 1/2 code is referred to as the "practical" code, and the rate 2/3 code is referred to as the "Ungerboeck" code. The Ungerboeck code was chosen because it requires a Viterbi decoder whose complexity is about the same as the decoder for the practical code.

For the concatenated system, the error rate of interest is the Reed–Solomon (RS) code block error rate. The reason the block error rate was chosen as the error rate of interest, as opposed to bit error rate, is because it is natural to block code

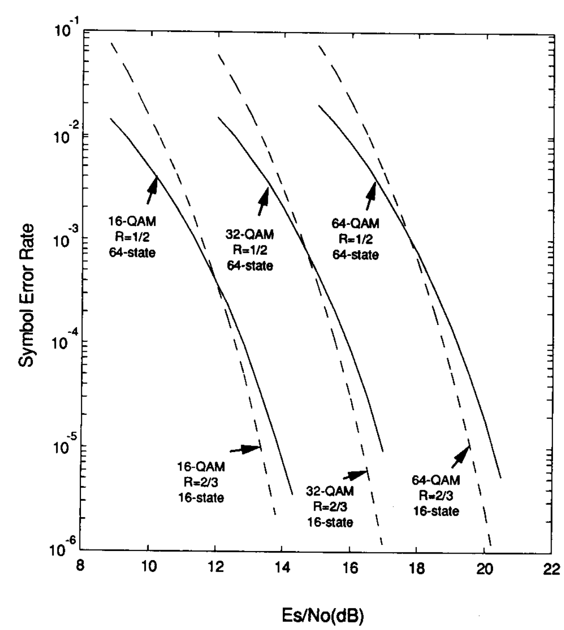


Fig. 6. Theoretical TCM performance using two different convolutional codes.

HDTV lines of data, and when an uncorrectable RS symbol error occurs in the line, some action is taken regardless of the number of bit errors that occurred. The block error rate can be approximated by

$$P_{\text{block}} \approx \sum_{i=t+1}^{L} \binom{L}{i} P_{\text{RSsym}}^{i} (1 - P_{\text{RSsym}})^{L-i} \quad (7)$$

where L is the RS block length (number of m-bit symbols per block), and t is the number of RS symbol errors that can be

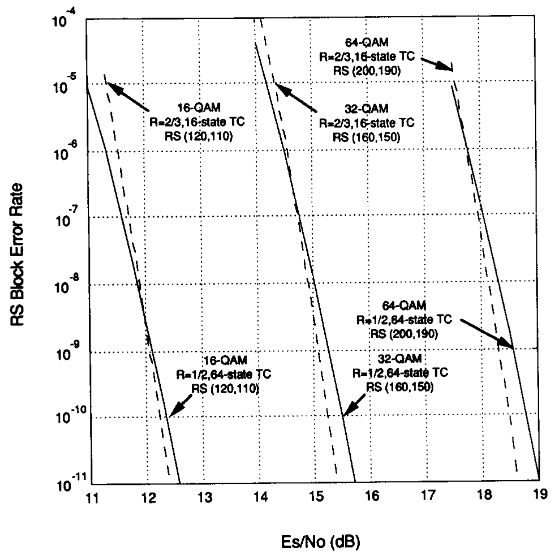


Fig. 7. Theoretical RS+ TCM performance using two different convolutional codes.

corrected per block. P_{RSsym} is the probability of an m-bit RS symbol being in error, and is approximated by

$$P_{\mathrm{RSsym}} \approx 1 - (1 - P_n)^{m/n} \tag{8}$$

where P_n is the n-bit symbol error rate out of the trellis decoder. (Note that the above approximations assume that the channel is memoryless.)

Based on our experience with video compression systems, a block error rate of 10^{-6} defines an acceptable viewing threshold. If blocks are transmitted at the NTSC TV horizontal line rate of 15.734 kHz, a 10^{-6} block error rate corresponds to about one block error per minute. If in addition the baud rate is restricted to 5 MHz (sufficient for transmission over a 6 MHz NTSC channel), then it is required that each of the three TCM's corresponding to three, four, and five RS coded bits per symbol (16, 32, and 64-QAM, respectively) be concatenated with RS codes with block lengths of 120, 160, and 200 RS symbols, respectively. Commercial RS chips that can correct five, eight-bit symbol errors are readily available. Therefore, the RS codes (over GF[256]) chosen for concatenation with the 16, 32, and 64-QAM TCM's are RS(120, 110), RS(160, 150), and RS(200, 190), respectively.

Fig. 7 shows the theoretical performance of concatenating the aforementioned RS codes with practical and Ungerboeck trellis codes. Notice that all practical codes are better at a block error rate of 10^{-6} or more. However, simulations reveal even better results for practical codes.

Fig. 8 compares the simulated performance of the practical and Ungerboeck codes. Notice here that the curves cross at 10^{-4} compared to the theoretical curves in Fig. 6, which cross near 10^{-3}. This result is reflected in the concatenated case, shown in Fig. 9, where the practical codes are shown to be better than the Ungerboeck codes at a block error rate of 10^{-10} or more. This clearly shows the performance advantage of the practical codes.

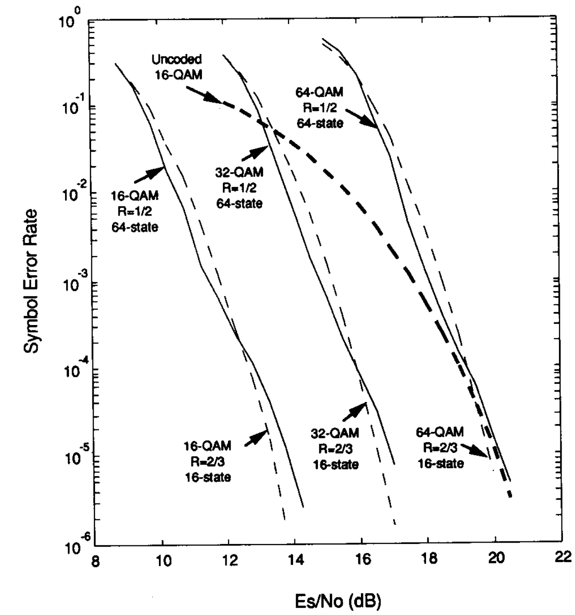


Fig. 8. Simulated TCM performance using two different convolutional codes.

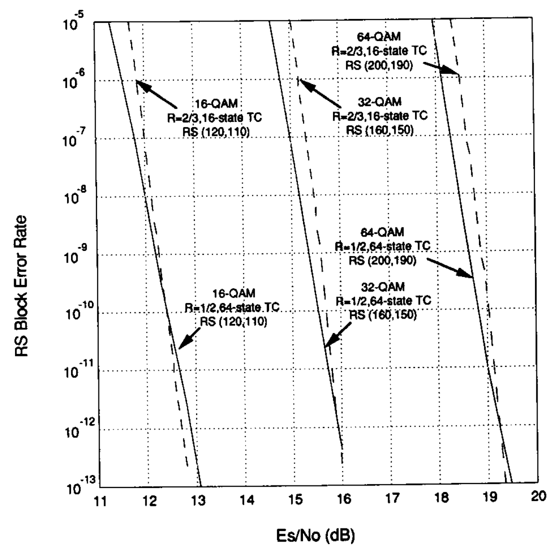


Fig. 9. Simulated TCM+ calculated RS performance using two different convolutional codes.

VI. SUMMARY

In this paper, it was argued that for concatenated coding systems employing QAM-based trellis coded modulation, optimization of coding gain is achieved by analysis of the number of nearest neighbors, where as for nonconcatenated systems the ACG is the parameter of interest. A practical concatenated coding scheme based on a QPSK trellis code employing a "standard" rate 1/2 64-state code was shown to perform better than a rate 2/3 16-state code. Furthermore, the practical scheme has a distinct implementation advantage over other trellis coding schemes due to the fact that a standard off-the-shelf Viterbi decoder can be used in the trellis decoder rather than a custom part.

REFERENCES

[1] A. R. Caulderbank and N. J. A. Sloane, "New trellis codes based on lattices and cosets," *IEEE Trans. Inform. Theory*, vol. IT-33, pp. 177–195, Mar. 1987.

[2] G. C. Clark and J. B. Cain, *Error-Correction Coding for Digital Communications*. New York: Plenum, 1981.

[3] G. D. Forney, Jr., "The Viterbi algorithm," *Proc. IEEE*, vol. 61, Mar. 1973.

[4] S. Lin and D. J. Costello, *Error Control Coding: Fundamentals and Applications*. Englewood Cliffs, NJ: Prentice-Hall, 1983.

[5] J. P. Odenwalder, "Optimal decoding of convolutional codes," Ph.D. thesis, Dept. of Elec. Eng., Univ. of Calif. at Los Angeles, Jan. 1970.

[6] E. Biglieri, D. Divsalar, P. McLane, and M. Simon, *Introduction of Trellis-Coded Modulation With Applications*. New York: Macmillan, 1991.

[7] G. Ungerboeck, "Channel coding with multilevel/phase signals," *IEEE Trans. Inform. Theory*, vol. IT-28, pp. 55–67, Jan. 1982.

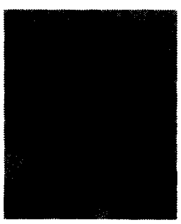
[8] G. Ungerboeck, "Trellis-coded modulation with redundant signal sets—Part I: Introduction," *IEEE Commun.*, vol. 25, pp. 5–11, Feb. 1987.

[9] G. Ungerboeck, "Trellis-coded modulation with redundant signal sets—Part II: State of the Art," *IEEE Commun.*, vol. 25, pp. 12–21, Feb. 1987.

[10] A. Viterbi, J. K. Wolf, E. Zehavi, and R. Padovani, "A pragmatic approach to trellis-coded modulation," *IEEE Commun. Mag.*, vol. 27, pp. 11–19, July 1989.

[11] Y. Yasuda, Y. Hirata, K. Nakamura, and S. Otani, "High rate punctured convolutional codes for soft decision Viterbi decoding," *IEEE Trans. Commun.*, vol. COM-32, pp. 315–319, Mar. 1984.

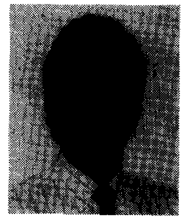
[12] Y. Yasuda, Y. Hirata, K. Nakamura, and S. Otani, "Development of a variable-rate Viterbi decoder and its performance characteristics," in *Proc. 6th Ann. Conf. Satel. Commun.*, Phoenix, AZ, Sept. 1983.



Chris Heegard (S'75–M'76–S'79–S'80–81) was born in Pasadena, CA, on Oct. 4, 1953. He received the B.S. and M.S. degrees in electrical and computer engineering from the University of Massachusetts, Amherst, in 1975 and 1976, respectively, and the Ph.D. degree in electrical engineering from Stanford University, Stanford, CA, in 1981.

From 1976 to 1978, he was an R&D Engineer at Linkabit Corp., San Diego, CA, where he worked on the development of a packet-switched satellite modem and several sequential decoders for the decoding of convolutional codes. In 1981, he joined the faculty of the School of Electrical Engineering, Cornell University, Ithaca, NY, as an Assistant Professor; he was appointed to Associate Professor, with tenure, in 1987. At Cornell, he teaches courses in digital communications, error control codes, information theory, detection and estimation theory, digital systems, and audio engineering. His current research interests include information, coding, and communication theory; algorithms for digital communications; coding for computer memory systems; signal processing and error control in optical and magnetic recording systems; audio signal compression and analysis/synthesis using multiresolution and wavelet filtering; algebraic coding theory; and symbolic and numerical computer methods. He is an active member of the consulting community. He has worked on problems of digital HDTV and cable TV transmission, DSP and hardware-based trellis-coded modems, modulation and error control for optical LAN's, and modulation and coding for recording systems. In 1984, he received the Presidential Young Investigator Award from the National Science Foundation and the IBM Faculty Development Award. He has ongoing research support from the NSF as well as AT&T and IBM. He has been involved in the organization of several IEEE workshops and symposia. In 1986, he was elected to the Board of Governors of the Information Theory Society of the IEEE and reelected in 1989; he is currently Second Vice President of the group.

Dr. Heegard is a member of IEEE, AES, and Eta Kappa Nu.



Scott A. Lery (M'85) received the B.S. degree in electrical engineering and computer science from the University of California, Berkeley, in 1980, and the M.S. degree in communications theory and systems from the University of California, San Diego, in 1984.

From 1980 to 1982, he was employed by the NCR Corp., San Diego, CA, as an Engineer working in the area of large-systems CPU hardware design. From 1982 to 1990, he was employed by the Linkabit Corp., San Diego, CA, as a Senior Engineer working on the performance analysis and design of military modems and satellite systems. In 1990, he joined the Advanced Development Group at the VideoCipher Division of General Instrument Corp., San Diego, CA, where he has been the Principal Staff Engineer on the HDTV transmission project, currently being tested by the FCC in competition for the American HDTV standard. He has six patents pending related to the HDTV project. His main interests are coding, adaptive filtering, and communications system simulation.

Woo H. Paik received the M.S. and B.S. degrees in electrical engineering from Seoul National University, Seoul, Korea, and the Ph.D. degree in electrical engineering from the Massachusetts Institute of Technology, Cambridge.

He is Vice President of the Advanced Development Group of the VideoCipher Division of General Instrument Corporation, San Diego, CA. He is currently leading the research and development effort in the Advanced Development Group, where he is responsible for developing the digital HDTV/NTSC compression algorithm and computer simulation as well as for designing real-time demonstration hardware of the DigiCipher^{TM} system. He is one of the coinventors of the VideoCipher satellite television encryption system, which became the de-facto standard of the C-band satellite video transmission used by most cable television programmers. He joined M/A-Com Linkabit, Inc., in June 1978. Over the past twelve years, he has led numerous development programs in the area of military and commercial satellite communications and digital signal processing. He has held a variety of positions with M/A-Com and General Instrument, including Senior Scientist, Director of Engineering, Assistant Vice President of Engineering, Assistant Vice President of Advanced Development, and his current position as Vice President of Advanced Development.