Progressive Source Transmissions using Joint Source-Channel Coding and Hierarchical Modulation in Packetized Networks

Suayb S. Arslan, Pamela C. Cosman and Laurence B. Milstein
Department of Electrical and Computer Engineering
University of California, San Diego, La Jolla, CA 92093-0407
Email: {sarslan, pcosman, lmilstein}@ucsd.edu

***Abstract*—With Unequal Error Protection (UEP), more important symbols are given greater protection against channel errors than are less important symbols. The protection can be accomplished by various methods, including joint source-channel coding (JSCC) and hierarchical modulation. In this paper, a robust progressive source transmission system using forward error correction (FEC) and a bits-to-symbol assignment methodology that provides UEP is proposed. UEP is not only provided by hierarchical modulation but also by the packetization methodology combined with channel coding. It is demonstrated by simulation that our system improves performance compared to an Equal Error Protection (EEP) technique and to a baseline JSCC-only mechanism.**

I. INTRODUCTION

In progressive bitstreams, the importance of bits decreases with each successive bit. Because of their finite state nature, progressive encoders are often very vulnerable to noisy channel effects. A single bit error may cause loss of synchronization. This necessitates robustness that can be provided by techniques such as channel coding, or by modifying the source encoder. The problem is usually characterized as a JSCC problem in which the goal is to allocate the source and channel coding rates optimally. However, the choice of modulation scheme also becomes critical in the rate allocation problem, especially in recent high rate multimedia applications. Hierarchical modulation has become a popular strategy to provide UEP and is included in various standards [1], [2].

Hierarchical modulation is utilized to give unequal transmission reliability to High Priority (HP) and Low Priority (LP) bits. Fig. 1 illustrates a Hierarchical 4PAM (H-4PAM(α)) constellation, where α is the hierarchical parameter used to adjust the distances of the symbol points in the constellation. As shown in Fig. 1, α is the ratio of the distances of the two symbols to the origin in the right hand side of the constellation. In going from one constellation to another, we vary α so that HP bits and LP bits will have different error probabilities. When $\alpha = 1$, we have BPSK and where $\alpha = 3$, we have a conventional 4PAM constellation. Even in conventional 4PAM, it can be shown that we have different Bit Error Rates (BERs) for HP and LP bits. We constrain the average bit energy, E_b to be the same for any α.

The idea of hierarchical modulation combined with progressively compressed signals is not new. In [2], the authors consider hierarchical modulation and different layers (streams of bits that have different effects in reconstruction of the source) of an image encoded using an adaptive discrete cosine transform to improve received image quality.

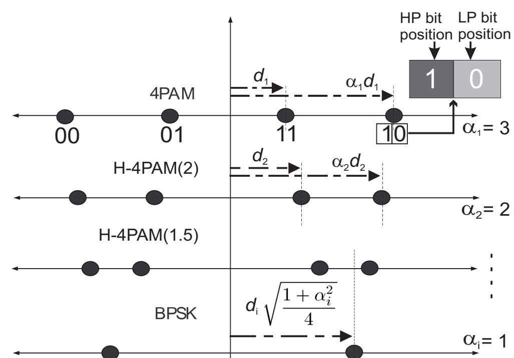

Fig. 1: Hierarchical modulation: 4PAM ($\alpha_1 = 3$, top), H-4PAM ($\alpha_2 = 2$, $\alpha = 1.5$, middle two) and BPSK to which it collapses when $\alpha_i = 1$ (bottom). E_b is kept constant across all constellations.

In [3] and [4], UEP is provided with non-uniform channel codes for different parts of the encoded image. Adaptive modulation is used in [5], however the approach is not hierarchical. Adaptivity is achieved by choosing optimum quantization and channel coding rates. Also, adaptive modulation is considered in [6] within the framework of JSCC; the system adapts to channel conditions but not to the source. Our system adapts to both channel conditions and the source.

JSCC-only mechanisms are considered extensively in the literature for the transmission of progressively compressed signals. In [7], Rate Compatible Punctured Convolutional (RCPC) channel codes are applied to source packets to obtain improved performance for memoryless channels. RCPC codes are also considered in [8] by applying different channel code rates for different kinds of bits (e.g., sign bits, significance bits, etc) of Set-Partitioning-In-Hierarchical-Trees (SPIHT) [9] encoded data. Turbo codes are considered in [10] and LDPC codes in [11]. However none of those studies consider alphabet

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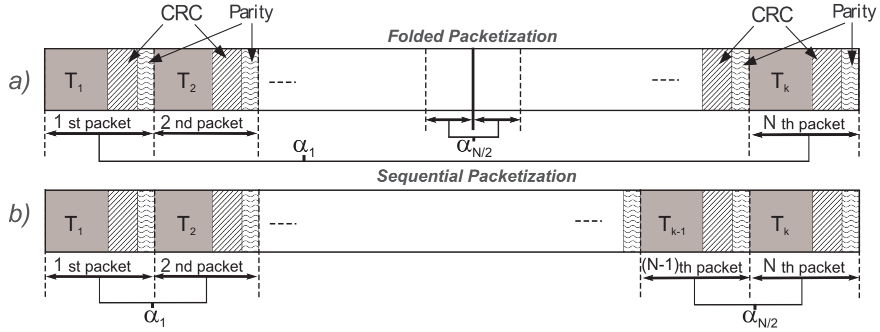

Fig. 2: Two different assignments of packetized bits to symbols in a progressively encoded source. T_i : ith chunk of information.

sizes greater than two.

To our knowledge, there is no study of progressive JSCC that considered higher order multi-resolution constellations with packetization. Thus, we aim to consider hierarchical modulation, source coding and channel coding together for memoryless channel scenarios. The major contribution of this work is a new UEP scheme for progressive sources using packetization in conjunction with the joint optimization of two known UEP schemes: hierarchical modulation and JSCC. We develop a systematic optimization framework to find the best source-channel code rates and hierarchical parameters for the given bandwidth and transmission time constraints. A distortion-minimal approach is adopted throughout the paper.

This paper is organized as follows: Two different bits-to-symbol assignment strategies are discussed in Section II. Section III describes the system model, and Section IV discusses the performance analysis and the optimization of the system. Finally, simulation results are given in Section V and conclusions in Section VI.

II. Bits-to-Symbol Assignment

We consider a packetized bits-to-symbol assignment methodology for the hierarchical modulation, as shown in Fig. 2. We use hierarchical parameters $\{\alpha_i \in \mathbb{R} : \alpha_i \in [l_i, u_i]\}$ with $i = 1, 2, \ldots, N/2$, where N is the total number of packets, assumed to be even, and l_i, u_i are lower and upper bounds, respectively, for the hierarchical modulation parameter α_i. We consider combining in two different ways. They are named *Folded Packetization (FP)* and *Sequential Packetization (SP)*. In Fig. 2a, FP is shown, where α_i is used to combine packets i and $N-i+1$. This scheme is proposed in [12] and shown in [12] to be optimal under high SNR conditions for an uncoded system, and we are evaluating its performance here for a coded system under moderate SNR values. Thus, the zth bit of packet i and the zth bit of packet $N-i+1$ are encoded together using an H-4PAM(α_i) constellation, for all bits in the packets, i.e. for any z. For SP, as shown in Fig 2b, $\alpha_{\frac{k+1}{2}}$ is used to combine packets k and $k+1$, where $k = 1, 3, 5, \ldots, N-1$. Two bytes of Cyclic Redundancy Check (CRC) are appended to each sequence of $b_s^{(l)}$ bits of information for packet l, where $l = 1, 2, \ldots, N$, along with m additional bits to flush the memory and terminate the decoding trellis in the all-zero state. A total of $b_s^{(l)} + CRC + m$ bits are then encoded using the code rate for packet l, r_l .

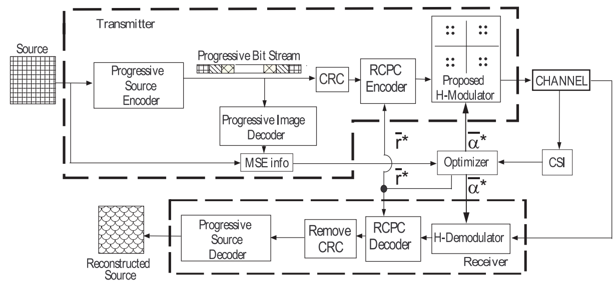

Fig. 3: Baseline system model with related block diagrams.

Once an error is detected at the receiver, the encoded stream is truncated and the decoded packets are used to reconstruct the source. If we fix the packet size, then the number of information bits (b_s^l) within each packet varies based on the chosen channel code rate. The parameters $\{\alpha_i\}$ determine the relative reliability among the packets. Thus, we need to determine the set of hierarchical parameters, $\overline{\boldsymbol{\alpha}} = [\alpha_1, \alpha_2, \ldots, \alpha_{N/2}]$ that yields the best performance.

Previous JSCC UEP techniques protect the progressive stream using a discrete set of channel code rates. If the optimal protection can be provided with a code that falls in between two available code rates in the set, the closest one is chosen to protect the packet [4]. In our system, UEP is provided using both a discrete code set and continuous-valued hierarchical parameters. Thus, a more flexible scheme is proposed by introducing continuous-valued hierarchical parameters.

III. Baseline System Model

As shown in Fig. 3, the transmitter uses a progressive source encoder to produce the bit stream. After reconstruction of the image at the encoder, the distortion (d_l) that results upon receiving packets up to and including packet l is determined. We use Mean Square Error (MSE) as our distortion metric throughout the paper. We have

$$d_l = \frac{1}{L_x L_y} \sum_{x=1}^{L_x} \sum_{y=1}^{L_y} |I(x, y) - R_l(x, y)|^2 \tag{1}$$

where L_x and L_y are the horizontal and vertical dimensions of the image in pixels, $I(x, y)$ is the original image intensity of pixel (x, y), and $R_l(x, y)$ represents the reconstructed image using only the first l packets of the encoded packet stream.

We define the vectors $\overline{\mathbf{d}} = [d_0 \ d_1 \ldots d_l \ldots d_N]$ and $\overline{\mathbf{r}} = [r_1, r_2]$, where r_1 and r_2 are the code rates of the first half

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and the second half of the packets, respectively. The reason for having two code rates for the stream of packets will be discussed in Section V.

For a given $\bar{\mathbf{r}}$, $\bar{\mathbf{d}}$ is determined and used in our optimization algorithm along with the Channel State Information (CSI). After optimization, the optimal α vector ($\overline{\boldsymbol{\alpha}}^* : [\alpha_1^*, \alpha_2^*, \ldots, \alpha_{N/2}^*]$) is constructed. Then we optimize over all possible pairs of (r_1, r_2) to find $\bar{\mathbf{r}}^* = [r_1^*, r_2^*]$. For packet i, the transmitter uses FP along with r_1^* and α_i^* or r_2^* and α_{N-i+1}^* depending on whether packet i is in the first half or the second half of the packet stream.

The optimizer block in the diagram provides the optimal code rates, the set of hierarchical parameters, the puncturing period and the mother code parameters to the channel encoder and decoder blocks.

IV. Performance Analysis

We have a discrete set of RCPC codes [13] with rates $C_r = \{c_1, c_2, ..., c_n\}$. Consider a packet size of ν bits used in conjunction with code rates $r_1 = \frac{a_1}{b_1}$ and $r_2 = \frac{a_2}{b_2}$ with g.c.d$\{a_1, b_1\}$ = g.c.d$\{a_2, b_2\} = 1$. Note that ν is determined by the channel code rates to ensure we have an integer number of information bits. First, a nominal value, ν_n is chosen. If ν_n is divisible by l.c.m.$\{b_1, b_2\}$, then the packet size $\nu = \nu_n$ is used. Otherwise, we use $\nu = \lfloor \frac{\nu_n}{\text{l.c.m.}\{b_1,b_2\}} \rfloor$l.c.m.$\{b_1, b_2\}$.

For a given transmission rate r_{tr} in bits per pixel (bpp), the number of packets N can be found as follows:

$$N = \left\lfloor \frac{r_{tr} \times L_x \times L_y}{\nu} \right\rfloor_{even} \tag{2}$$

where $\lfloor . \rfloor_{even}$ rounds down to the largest even integer not greater than the argument. The source rate r_s in bpp is given by $r_s = \frac{\sum_l b_s^{(l)}}{L_x \times L_y}$, where $b_s^{(l)} = \nu r_l - CRC - m$. We assume ideal coherent detection, perfect CRC error detection and hard decisions at the decoder.

Let $\rho_l(\gamma)$ be the average bit error probability for the lth packet as a function of $\gamma = \frac{E_b}{N_0}$, and assume the all-zero codeword is transmitted. For a given code rate $\beta \in C_r$, let $d^{(\beta)}$ represent the distance to the all-zero codeword of the path being compared with the all-zero path at some node in the trellis. For a BSC with crossover probability $\rho_l(\gamma)$, the probability of selecting the incorrect path is given in [15] by

$$P^l_{d^{(\beta)}} = \begin{cases} \sum_{j=\frac{d^{(\beta)}+1}{2}}^{d^{(\beta)}} \binom{d^{(\beta)}}{j} (1-\rho_l(\gamma))^{d^{(\beta)}-j} \rho_l^j(\gamma) \\ \qquad\qquad \text{if } d^{(\beta)}\text{:odd} \\ \sum_{j=\frac{d^{(\beta)}}{2}+1}^{d^{(\beta)}} \binom{d^{(\beta)}}{j} (1-\rho_l(\gamma))^{d^{(\beta)}-j} \rho_l^j(\gamma) \ldots \\ \quad + \frac{1}{2} \binom{d^{(\beta)}}{d^{(\beta)}/2} (1-\rho_l(\gamma))^{d^{(\beta)}/2} \rho_l^{d^{(\beta)}/2}(\gamma) \\ \qquad\qquad \text{if } d^{(\beta)}\text{:even} \end{cases} \tag{3}$$

The number of different values of $P^l_{d^{(\beta)}}$ is equal to the number of different BERs that can be provided by the hierarchical M-ary constellation. In 4-PAM, we need to compute (3) for two different BERs; HP and LP BERs $(\rho^{HP}(\alpha, \gamma), \rho^{LP}(\alpha, \gamma))$, which are given at the bottom.

Based on these results, one can evaluate the union bound for bit error probability (P_b^l) [15],

$$P_b^l \leqslant \frac{1}{p} \sum_{d^{(\beta)}=d_{free}}^{\infty} c_{d^{(\beta)}} P^l_{d^{(\beta)}} \tag{4}$$

where p is the puncturing period, d_{free} is the free distance of the code and $c_{d^{(\beta)}}$ is the information error weight distribution for a given code $\beta \in C_r$ [13]. Using the formulation in [16], the packet error rate for packet l (PER_l) can be upper bounded by

$$PER_l \leq 1 - \left(1 - \frac{1}{p} \sum_{d^{(\beta)}=d_{free}}^{\infty} c_{d^{(\beta)}} P^l_{d^{(\beta)}}\right)^{b_s^l}. \tag{5}$$

In this formulation, errors at the output of the decoder are assumed to be uniform, which is not the case because of the channel coding operation and the Viterbi algorithm. Thus, we will use a nonlinear least squares (NLS) curve fitting technique to approximate PER_l in (5). It is approximated using an exponential function and denoted $\widehat{PER_l}$ for packet l:

$$\widehat{PER_l} = 1 - Ae^{B \times \overline{P}_b^l} \tag{6}$$

where A, B are design parameters chosen according to the following criterion:

$$\min_{A,B \in \mathbb{R}} \left\{ \sum_{j=1}^{s} \left| \underbrace{\overline{PER}_l^{(\gamma_j)}}_{Simulation} - \left(1 - Ae^{B \times \overline{P}_b^{l\,(\gamma_j)}}\right) \right|^2 \right\} \tag{7}$$

where s is the number of discrete SNR values and $\overline{PER}_l^{(\gamma_j)}$ is the random variable that is the outcome of Monte Carlo simulation at each SNR γ_j. $\overline{P}_b^{l(\gamma_j)}$ is the bound in (4) evaluated at each SNR γ_j. For example, the results of a simulation for code rate $\beta = 1/2$ are shown in Fig. 4 with the NLS curve

$$\rho^{HP}(\alpha, \gamma) = \frac{1}{2}\left[Q\left(\sqrt{\frac{8\gamma}{(1+\alpha^2)}}\right) + Q\left(\sqrt{\frac{8\gamma\alpha^2}{(1+\alpha^2)}}\right)\right]$$

$$\rho^{LP}(\alpha, \gamma) = \frac{1}{2} + \frac{1}{2}\sum_{s=0}^{1}\sum_{m=0}^{1}(-1)^{s+m} Q\left(\left[\frac{(-1)^s(1+\alpha)}{2} + \alpha^m\right]\sqrt{\frac{8\gamma}{1+\alpha^2}}\right), \text{ where } Q(x) = \frac{1}{\sqrt{2\pi}}\int_x^{\infty} e^{-\frac{z^2}{2}} dz.$$

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fitting. As illustrated, the upper bounds are not tight for the range of SNRs of interest, and the NLS gives a better match to simulation results.

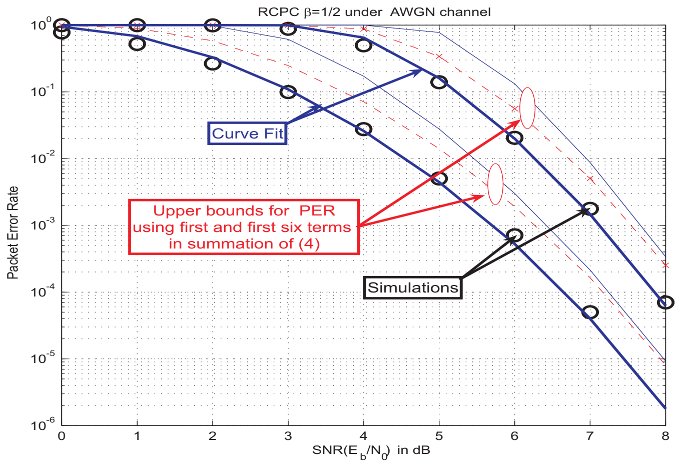

Fig. 4: Proposed Curve Fitting

A. Construction of the optimization problem

In this section, we address the optimization problem and its solution. For a given γ and $\bar{\mathbf{r}}$, we want to minimize the average distortion ($E[\mathbf{D}]$) by finding the vector $\overline{\boldsymbol{\alpha}}^*$, where $E[.]$ is the expectation operation. It can be shown that the expected distortion can be expressed as follows:

$$E[\mathbf{D}(\bar{\mathbf{r}}, \overline{\boldsymbol{\alpha}}, \gamma)] = d_0 + \sum_{l=1}^{N} \prod_{i=1}^{l} (1 - \widehat{PER}_l)(d_l - d_{l-1}) \quad (8)$$

For a given γ, the minimum average distortion is the solution to the following optimization problem:

$$E[\mathbf{D}^*_\gamma] = \min_{\substack{\bar{\mathbf{r}} \\ r_l \in C_r \\ l=1,\ldots,N}} \left\{ \min_{\substack{\alpha_i \\ i=1,\ldots,N/2}} E[\mathbf{D}(\bar{\mathbf{r}}, \overline{\boldsymbol{\alpha}}, \gamma)] \right\} \quad \text{subject to} \quad l_i \le \alpha_i \le u_i \quad (9)$$

Since we have $d_0 \ge 0$, we can rewrite the previous expression using (8) as

$$\min_{\substack{\bar{\mathbf{r}} \\ r_l \in C_r \\ l=1,\ldots,N}} \left\{ \min_{\substack{\alpha_i \\ i=1,\ldots,N/2}} \left\{ \sum_{l=1}^{N} \xi_{\mathbf{r}_l,l} \Delta_l \right\} \right\} \text{ s. t. } l_i \le \alpha_i \le u_i \quad (10)$$

where the distortion reduction associated with the l-th packet is $\Delta_l = d_l - d_{l-1}$, and the weights $\xi_{\mathbf{r}_l,l}$ are given by

$$\xi_{\mathbf{r}_l,l} = \prod_{i=1}^{l} P_c^i = \begin{cases} \prod_{i=1}^{l} P_c^i(r_i, \alpha_i, \gamma) & \text{if } 1 \le l \le \frac{N}{2} \\ \xi_{r_i,\frac{N}{2}} \prod_{i=\frac{N}{2}+1}^{l} P_c^i(r_i, \alpha_{N-i+1}, \gamma) & \text{if } \frac{N}{2}+1 \le l \le N. \end{cases}$$

B. Optimization of hierarchical parameter set: $\overline{\boldsymbol{\alpha}}$

Since the code set is discrete, we first condition on $\bar{\mathbf{r}}$ and optimize the hierarchical parameters. Then we optimize over the channel code rate vector $\bar{\mathbf{r}}$.

Let us consider the following piecewise linear function:

$$\text{For } 1 \le i \le N/2, \quad \bar{\mathbf{g}}(\alpha) = \begin{cases} g_i(\alpha) = \alpha \\ g_{\frac{N}{2}+i}(\alpha) = -\alpha \end{cases}$$

where $\bar{\mathbf{g}}(\alpha) = [g_1(\alpha)\ g_1(\alpha)\ \ldots\ g_N(\alpha)]$. Then our optimization problem can be expressed as

$$\min_{\substack{\bar{\mathbf{r}} \\ r_l \in C_r \\ l=1,\ldots,N}} \left\{ \min_{\substack{\alpha_i \\ i=1,\ldots,N/2}} \left\{ \sum_{l=1}^{N} \xi_{\mathbf{r}_l,l} \Delta_l \right\} \right\} \text{ s. t. } g_l(\alpha) \le x_l \quad (11)$$

where $\bar{\mathbf{x}} = [x_1\ x_2\ \ldots\ x_N] = [u_1\ u_2\ \ldots\ u_{N/2}\ l_1\ \ldots\ l_{N/2}]$. From Lagrangian theory, the Lagrangian function of (11) can be written as follows:

$$\Lambda_D = \Lambda(\alpha_1, \alpha_2, \ldots, \alpha_{N/2}, \lambda_1, \ldots, \lambda_N) \quad (12)$$
$$= \sum_{l=1}^{N} \xi_{\mathbf{r}_l,l}\Delta_l - \lambda_l\big(g_l(\alpha) - x_l\big) \quad (13)$$

where the parameters $\lambda_1, \lambda_2, \ldots, \lambda_N$ are the Lagrange multipliers. The unconstrained minimization problem is

$$\min_{\bar{\mathbf{r}}} \min_{\overline{\boldsymbol{\alpha}}} \{\Lambda_D\}. \quad (14)$$

The necessary conditions for our optimization problem are given by the Karush-Kuhn-Tucker (KKT) conditions:

$$\nabla\Lambda_D(\overline{\boldsymbol{\alpha}}^*) = [\frac{\partial \Lambda_D}{\partial \alpha_1^*}, .., \frac{\partial \Lambda_D}{\partial \alpha_{N/2}^*}, \frac{\partial \Lambda_D}{\partial \lambda_1^*}, .., \frac{\partial \Lambda_D}{\partial \lambda_N^*}] = 0 \quad (15)$$
$$\lambda_l^*\big(g_l(\alpha^*) - x_l\big) = 0 \quad (16)$$
$$\lambda_l^* \ge 0 \quad (17)$$
$$l_i \le \alpha_i^* \le u_i \quad (18)$$

In (15), the equations are not linear. To our knowledge, there is no analytical way to solve them. Thus, we use numerical approaches to solve (15) for the $\overline{\boldsymbol{\alpha}}^*$ that minimizes the overall $E[\mathbf{D}]$ at a given γ.

V. Numerical Results and Discussions

Our first simulation demonstrates the average BER performances of different packetized bits-to-symbols assignments. We present simulation results for the coded cases and numerical computation of BER expressions for the uncoded case. We only consider 4PAM, but the results can be extended to higher rectangular constellations.

Fig. 5 shows the performance for different packetization schemes using conventional modulation. Note that there is a BER performance gap between uncoded HP bits (bits in the first half of the packet stream) and LP bits (bits in the second half of the packet stream) which is inherent to PAM signalling, and therefore to higher QAM constellations. This gap is naturally widened when we use lower channel code rates because, in this case, parity bits are also protected unequally,

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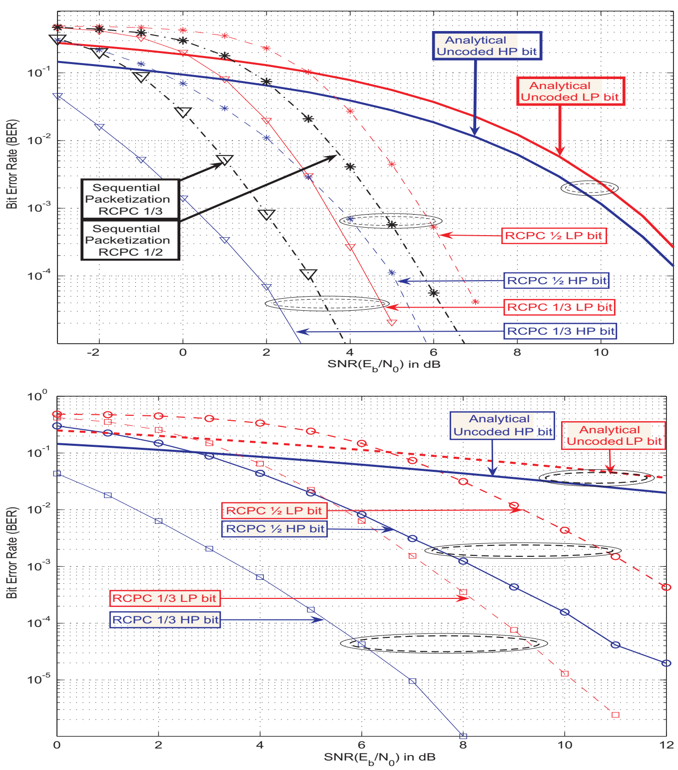

Fig. 5: HP and LP BER performances using uncoded and coded conventional 4PAM under AWGN (top) and flat Rayleigh fading channels (bottom).

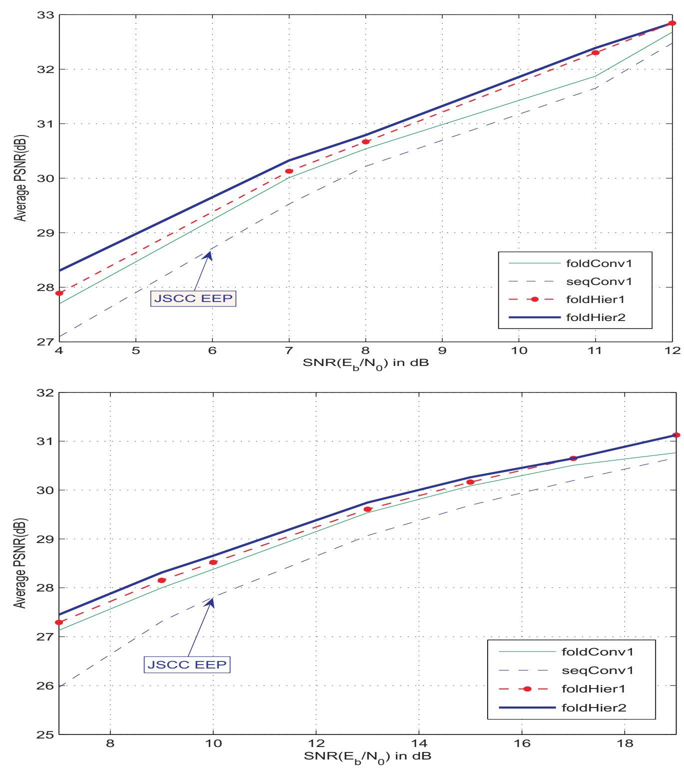

Fig. 6: Performance of different systems under AWGN (top) and flat Rayleigh fading channels (bottom). JSCC EEP is also drawn for comparison.

just like the information bits. There is no gap in SP, because the average HP BER is equal to the average LP BER.

In the next set of results, we consider an image transmission system with $\bar{\mathbf{r}} = [r_1 \ r_2]$, where we apply code rates r_1 and r_2 for the first and second half of the total packet stream, respectively. We use $\nu_n = 450$ bits and the RCPC code with constraint length K=7 from [13]: the code rate set is given by $C_r = \{\frac{8}{9}, \frac{4}{5}, \frac{2}{3}, \frac{4}{7}, \frac{1}{2}, \frac{4}{9}, \frac{2}{5}, \frac{4}{11}, \frac{1}{3}, \frac{4}{13}, \frac{2}{7}, \frac{4}{15}, \frac{1}{4}\}$. A CRC code from [7] is used for error detection. A standard grayscale (8 bpp) image $Lena(512 \times 512)$ is encoded using the SPIHT algorithm without arithmetic coding and sent over the channel.

The transmission rate (r_{tr}) is 0.25bpp. Throughout this section, we determine the MSE values and average them before converting the average MSE to average Peak Signal to Noise Ratio (PSNR).

In the proposed system, hierarchical parameters and channel code rates are found by solving the optimization problem. The complexity of the problem is equivalent to a line search algorithm [17]. We introduce the following systems:

- **seqConv1**: No hierarchical modulation ($\alpha_l = 3$). A fixed optimal code rate chosen from C_r using SP.
- **foldConv1**: No hierarchical modulation ($\alpha_l = 3$). A fixed optimal code rate chosen from C_r using FP.
- **foldHier1**: Hierarchical modulation ($\overline{\boldsymbol{\alpha}}^*$). A fixed optimal code rate chosen from C_r using FP.
- **foldHier2**: Hierarchical modulation ($\overline{\boldsymbol{\alpha}}^*$). Two optimal code rates chosen from C_r using FP.

Note that there are four other possible combinations (*seqConv2*, *seqHier1*, *seqHier2*, *foldConv2*). However, these other combinations are not meaningful. For example, since the difference in importance of adjacent bits is negligibly small, it does not make sense to use hierarchical modulation in conjunction with sequential packetization.

Fig. 6 shows the performance of various systems, and it is observed that the UEP schemes ($foldConv1$, $foldHier1$, $foldHier2$) always perform better than the EEP scheme ($seqConv1$). The non-concave behavior of these curves is a consequence, at least in part, of the code rate set C_r being discrete. In addition to the constraint of the code set (being discrete), our design constraint is that the α values determine the BER for each packet in the first and second half of the packet stream simultaneously. However, the $foldHier2$ system is introduced to somewhat alleviate both constraints by employing two different code rates, one for the first half and the other for the second half of the packet stream. Some performance improvement is likely if we use more than two channel code rates at the expense of greater complexity, however we expect diminishing returns.

The hierarchical values, $\overline{\boldsymbol{\alpha}}^*$, as a function of packet index are plotted in Fig. 7 for the first half of the stream. The figure shows that $\alpha_1 \leq \alpha_2 \leq \cdots \leq \alpha_{N/2}$, meaning that earlier packets in the stream are more heavily protected by the hierarchical modulation. The discrete nature of the code set is the cause of nonuniform gains going from one UEP scheme to another. As one can observe, we pick up different gains at different SNRs, and this is mainly due to the system being forced to use code rates from a discrete set at a given

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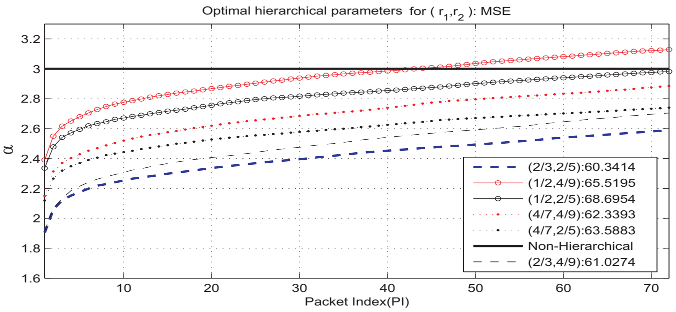

Fig. 7: Optimal hierarchical parameters ($\overline{\alpha}^*$) for different channel code rate pairs at SNR=7dB under AWGN channel. It also shows the corresponding calculated values of MSE using (8). (2/3,2/5) is the optimal pair.

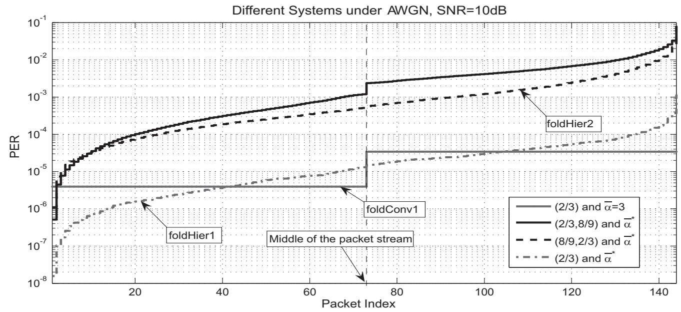

Fig. 8: PER assignment among the packets of proposed hierarchical system.

SNR. At low SNRs, the gap between the curves becomes more pronounced as UEP is more effective when the channel degrades. Another interesting observation is that for system $foldHier2$, channel code rates r_1 and r_2 used for the first and second half of the packet stream, respectively, have the relationship $r_1 \geq r_2$. However, in a JSCC-only UEP scheme, we would expect that $r_1 \leq r_2$, because our RCPC code set satisfies the condition of [14]. In other words, we would expect to protect the first part more heavily than the second part.

This is not the case when JSCC is used with hierarchical modulation. Our conjecture is that this is because the hierarchical parameters adjust themselves to protect the bits of the first half more than the bits of the remaining half. As long as these parameters are able to compensate for the decreased protection due to the FEC, $r_1 \geq r_2$ has the potential to improve the system performance by allocating more information bits in the first half of the packets ensuring their reliable transfer. This leads to better reconstruction quality.

Packet error probability as a function of Packet Index (PI) number at SNR=10dB with both the optimal pair of codes and one with reverse order is plotted in Fig. 8. The case where $r_1 \geq r_2$ is seen to protect almost all the packets better than the $r_2 \geq r_1$ case. Lastly, as seen in Fig. 8, the system can provide as many UEP levels as the number of packets, although the system uses only two code rates.

VI. Conclusion

A reliable and robust progressive source encoding system for memoryless noisy channels based upon the combined use of several UEP methods is considered. Specifically, a packetization methodology that is coupled with both hierarchical modulation and forward error correction coding is considered. We have seen that even though the parameters of the system are subject to design constraints, the different UEP methods can be judiciously combined to provide enhanced reliability for the transmission of the progressive source.

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