Optimal Code Rates for Concatenated Codes on a PR4-Equalized Magnetic Recording Channel

William E. Ryan

***Abstract*—At some nominal recording density, the read signal in digital magnetic recording resembles a Class IV partial response (PR4) signal and, hence, may be equalized to the PR4 shape with relatively little noise enhancement. When coding is added, for a fixed user density, the recording density must increase as a result of coding overhead, and the read signal will resemble PR4 to a lesser extent. Equalization to PR4 in this case will produce excessive noise enhancement. Thus, coding overhead (or rate) must be selected for optimum tradeoff between code strength and noise enhancement. Toward this end, we provide results for high-rate concatenated codes, assuming a Lorentzian recording channel model. In addition to examining optimal code rates, we compare parallel and serial concatenated code performance on the PR4 channel.**

***Index Terms*—Concatenated codes, magnetic recording channels, partial-response channels.**

I. Introduction

RECENT papers have examined the applicability of concatenated codes with iterative (turbo) decoders to digital magnetic recording and have demonstrated very promising results for both parallel and serial concatenations [1]–[5]. The focus in those papers is on Class IV partial response (PR4) signaling and high rate codes. PR4 signaling is considered because it is a commonly employed signaling scheme in magnetic recording. This is because the read signal at medium recording densities already resembles PR4 (which admits a simple sequence detector) so that equalizing to PR4 suffers relatively little noise enhancement. High rate codes are considered because, for a fixed user density, excessive coding overhead would result in a read signal that is more difficult to equalize to the PR4 shape.

In this paper, we examine the tradeoff between code strength and increased noise enhancement as a result of equalization at the higher densities required for coded signaling. This tradeoff is examined to some extent in [1], where rate 4/5, 8/9, and 16/17 parallel concatenated codes are considered, but an improper SNR definition is used.[1] In this paper, we consider rate 4/5, 8/9, 16/17, 32/33, and 64/65 parallel and serial concatenated codes using an improved SNR definition. The optimal code rate for a given user density can be determined from the approach demonstrated here. The results presented also allow a comparison between the two classes of concatenated codes. Comparisons between the two classes for the white noise case for selected code rates may be found in [3]–[5].

II. Channel Model

We consider the record head, the magnetic medium, and the read head to be a single linear system. A well-accepted model for the transition response for this system is the Lorentzian pulse, which can be written as

$$h(t) = \sqrt{\frac{4E_i}{\pi pw_{50}}}\, \frac{1}{1 + (2t/pw_{50})^2} \tag{1}$$

where pw_{50} is the width of the pulse measured at half its height and E_i is the energy in this "isolated pulse." Thus, when an NRZ signal of the form $\sum_k a_k\, p(t - kT_c)$ is to be recorded ($a_k \in \{\pm 1\}$, $p(t) = 1$ for $t \in [0, T_c]$, and 0 elsewhere), the response of the channel is

$$r(t) = \sum_k (a_k/2)[h(t - kT_c) - h(t - (k+1)T_c)] + w(t). \tag{2}$$

In (2), T_c is the recorded bit duration and is related to the user bit duration T_u via the code rate r as $T_c = rT_u$; $w(t)$ is assumed to be white Gaussian noise with power spectral density $N_0/2$ for simplicity, although a more accurate model would include correlated media noise.

We assume the optimal receiver front end, which is a filter matched to the "dibit" $s(t) = h(t) - h(t - T_c)$ followed by a symbol-rate ($1/T_c$) sampler. This leads to a discrete-time equivalent channel response $f(D)$, where $f(D)f(D^{-1})$ is a factorization of the sampled autocorrelation function $R_s(D)$ of $s(t)$. We use the (noncausal) factorization of Bergmans [8] for which $f(D) = (1 - D)g(D)$ with the coefficients of $g(D)$ given by

$$g_k = \left[\frac{E_i S_c}{2\pi} \tanh(\pi S_c/2)\right]^{1/2} \frac{k + S_c/2}{k^2 + (S_c/2)^2}, \qquad k \in \mathbb{Z} \tag{3}$$

where $S_c \triangleq pw_{50}/T_c$ is defined to be the channel density and is related to the user density $S_u \triangleq pw_{50}/T_u$ via $S_c = S_u/r$.

The discrete-time magnetic recording channel (MRC) model then consists of a filter with response $f(D)$ and a white noise process $w(D)$ with spectral density $N_0/2$ added to this filter's output. The front end of the discrete-time receiver is the matched filter $f(D^{-1})$ [although some authors include $f(D^{-1})$ with the equalizer].

Manuscript received February 4, 2000; revised August 16, 2000. This work was supported in part by NASA and NSIC. This paper was presented in part at the 1998 Conference on Information Sciences and Systems.

The author is with the Department of Electrical and Computer Engineering, University of Arizona, Tucson, AZ 85721 USA.

Publisher Item Identifier S 0018-9464(00)09951-9.

[1]We also point out that in addition to the improper SNR definition, the coded performance curves in [1, Fig. 4] are in error and should be shifted to the right by 3 dB.

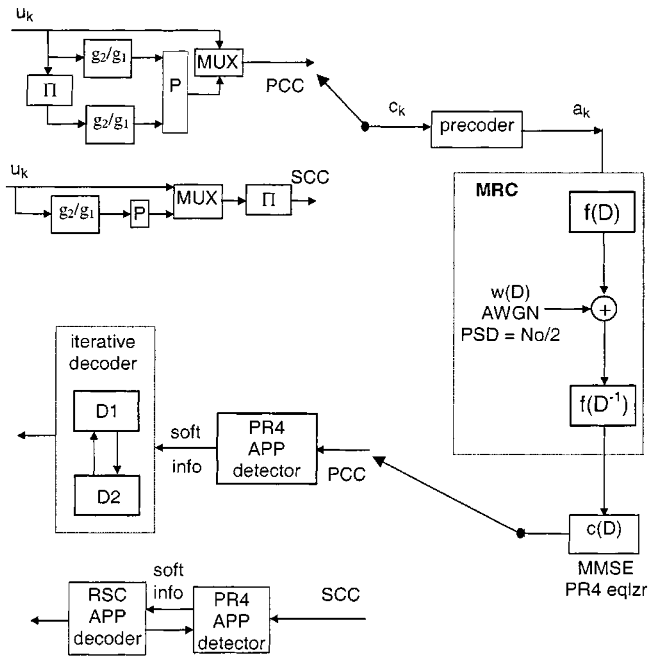

Fig. 1. Model for PCC and SCC encoding and interative decoding on the PR4-equalized MRC. [Π indicates a permuter (interleaver), and P indicates a puncturer.]

III. Concatenated Codes Under Study

The parallel concatenated codes (PCCs) we consider employ the polynomials $(g_1, g_2) = (23, 31)$ for each constituent recursive systematic convolutional (RSC) code, with each RSC code separated by a 4096-bit pseudorandom interleaver. For $K_0 = 4, 8, 16, 32,$ and 64, rate $K_0/(K_0+1)$ PCCs are achieved by saving the second bit in every $2K_0$-bit parity block of each constituent RSC encoder output (and "puncturing" the rest) [9] Only the top RSC encoder (see Fig. 1) is terminated (i.e., is forced to the zero state at the end of the data block). All PCCs operate within 1 dB of the capacity limit at $P_b = 10^{-6}$ on an AWGN channel with antipodal signaling [9].

To use these codes over the PR4 channel (modeled by the polynomial $1-D^2$), we first precode with the $1/(1 \oplus D^2)$ precoder so that a **1** is mapped to ± 2 and a **0** is mapped to 0 over the precoded PR4 channel (after the precoder, a **1** is converted to a $+1$ and a **0** is converted to $a-1$). This has the effect of converting code words with a large minimum Hamming distance to channel sequences with a large minimum Euclidean distance [10]. The PCC encoder and iterative decoder combined with the channel model of the previous section are depicted in Fig. 1, with the switches in position "PCC." Included in the channel model is the matched filter $f(D^{-1})$ and an MMSE PR4 equalizer $c(D)$. The PR4 *a posteriori probability* (APP) detector employs the BCJR algorithm [11] to compute the log-likelihood ratio $L(c_k)$ for each code bit c_k. The PCC decoder treats these inputs $L(c_k)$ as extrinsic information, but otherwise iterates in the usual fashion. An alternative setup includes the PR4 APP detector in the iterative loop with the two PCC APP decoders [2]–[5], increasing the coding gains for all code rates by about 0.5 dB, but we have not done that here.

The serial concatenated codes (SCCs) we consider comprise a single RSC as the outer code and the precoded PR4 channel as the inner code, with the inner and outer codes separated by an L-bit interleaver, $L = 4096(K_0+1)/K_0$. The outer code is terminated, whereas the inner code is not. Note the interleaver sizes are selected for a data word size of $N = 4096$ bits for all code rates, consistent with the PCC case. The RSC code employs the polynomials $(g_1, g_2) = (23, 31)$ and rate $K_0/(K_0+1)$ SCCs are achieved by saving a single bit in every K_0-bit parity block of the RSC encoder output. We have not performed an exhaustive search among all puncturers, but we have found that (for our specific interleaver) saving bits 4, 5, 15, 2, and 15 for $K_0 = 4, 8, 16, 32,$ and 64, respectively, yields effective codes. Nor do we consider optimal precoders here, but the $1/(1 \oplus D^2)$ precoder appears to be effective and will suffice for our study.[2]

As indicated in Fig. 1 (with the switches in the "SCC" position), the decoder consists of the PR4 APP detector cooperating iteratively with the RSC APP decoder, with each APP processor passing only extrinsic information to the other.

We remark that the noise seen by the various APP processors is certain to be colored. Although it is possible to design the APP processors so that they use a colored noise metric, we shall not do this here and instead use only the Euclidean distance metric, which is optimal only for white Gaussian noise.

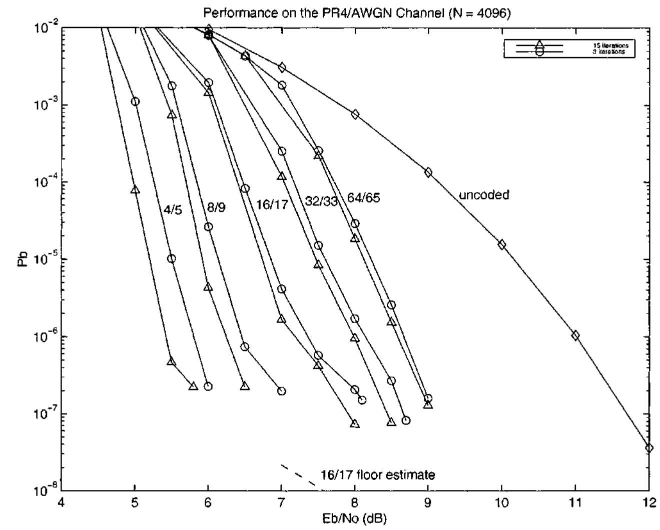

Fig. 2. PR4/AWGN performance for the various PCCs.

IV. Code Performance Comparisons

A. White Noise Baseline Performance

As a baseline, we plot in Fig. 2 the bit error probability P_b of the PCCs versus E_b/N_0 for PR4 signaling and AWGN, where E_b is the user bit energy. This is achieved in Fig. 1 by setting $f(D) = 1 - D^2$ and removing the $f(D^{-1})$ and $c(D)$ blocks. Note that this results in a mathematical model useful for setting a baseline, as is commonly done in the literature. It does not correlate well with an actual magnetic storage channel, although it correlates well with certain communication channels. The channel signal levels for this model are $\{\pm 1, 0\}$, resulting in an average channel energy of $E_c = 1$. The received SNR is E_c/N_0 [actually, $E_c/(N_0/2),$ but the factor of 2 is generally

[2]The effect of removing the $1/(1 \oplus D)$ precoder on the $1 - D$ channel is studied in [3] and [4]. A study on optimal precoder design may be found in [6] and [7].

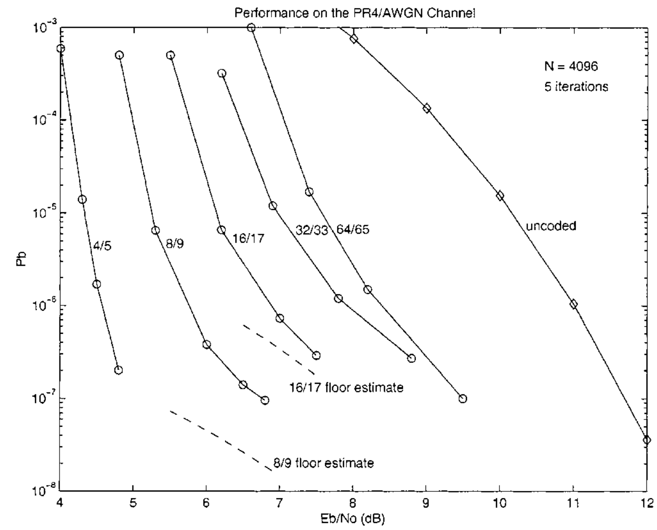

Fig. 3. PR4/AWGN performance for the various SCCs.

dropped]. In terms of the user bit energy E_b, the received SNR is rE_b/N_0, where r is the code rate. Thus, when plotting error rate curves versus the uncoded SNR parameter E_b/N_0, as is customary, there is a "code rate loss" of r [or 10 $\log_{10}(r)$ dB] because the received SNR is scaled by r for a given E_b/N_0.

The curves in Fig. 2 are presented for three and 15 decoder iterations. Note the gains over the uncoded case[3] at $P_b = 10^{-6}$ are substantial and far greater than any trellis code: 5-plus dB for the rate 4/5 code and 2-plus dB for the rate 64/65 code. Observe also that the well-known "error-rate floor" for PCCs occurs below 10^{-7} for all of the codes, although the floor has been reached for the rate 16/17 code. The figure includes an estimate of the level of the floor for this code, the computation of which will be discussed shortly. Finally, we point out that little is lost in opting for three iterations over 15 iterations. Thus, our PCC results presented below will be for three iterations.

The baseline PR4/AWGN curves for the SCCs are presented in Fig. 3 for five decoder iterations. We may discern from the SCC curves of Fig. 3 the fact that they are superior to their PCC counterparts by about 0.7 dB in the $P_b = 10^{-5}$ region, although including the PR4 APP detector in the PCC decoding iteration would close this gap. We notice also that they are inferior in the floor region, which starts below $P_b = 10^{-5}$ for all code rates. (These characteristics are also seen in [3] and [4] for a rate 8/9 code.) We attribute the higher floor to the fact that the "inner code" is not of the recursive type required for large interleaver gain in SCCs [13], although it does have some of the desirable properties required for interleaver gain [5].[4] (See also [6] and [7] for design of the optimal precoders.)

[3]Here, and in all of the following figures, the uncoded curves correspond to a PR4 Viterbi detector with a Euclidean distance metric. That is, although colored noise is present in some cases, the Viterbi detector is not redesigned to account for this.

[4]Although the precoder $1/(1 \oplus D)^2$ combined with the $1 - D^2$ channel is ostensibly a recursive inner code, strictly speaking, it is not. A recursive inner code has the property that an input error sequence of 1000 ... will produce output sequences that are an infinite distance from each other. For the combination above, this error sequence will produce output sequences that are at a Euclidean distance of only 2 from each other.

As is evident from the figures, it is difficult to measure the level of the floor via simulation. We can estimate the level of the floor for a maximum-likelihood (ML) decoder (which the iterative APP decoder approximates) as follows. Let $c(D)$ and $c'(D)$ be any two code words. After passing through the precoder and the $1 - D^2$ response, the corresponding channel sequences become $c_0(D) = (c(D)/(1 \oplus D^2))(1 - D^2)$ and $c_0'(D) = (c'(D)/(1 \oplus D^2))(1 - D^2)$, whose coefficients are taken from the set $\{\pm 1, 0\}$. Now define E to be the set of channel error sequences $e(D)$, given by $E = \{c_0(D) - c_0'(D) : c_0(D) \neq c_0'(D)\}$. Then, P_b for the ML decoder may be upper bounded (and approximated) [12], [3] as

$$P_b \leq \sum_{e(D) \in E} \frac{N_e w_e}{N} Q\left(\frac{\|e\|}{\sqrt{2N_0}}\right) \tag{4}$$

where the *multiplicity* N_e is the average (averaged over c_0) number of code words c_0' satisfying $c_0(D) - c_0'(D) = e(D)$, w_e is the average (averaged over c_0) number of bit errors resulting from choosing c_0' over c_0, and $\|e\| = \sqrt{\sum_k e_k^2}$. This expression, which is applicable only to the floor region (which is of current interest), may be further approximated as

$$P_b \approx \frac{N_{e^*} w_{e^*}}{N} Q\left(\frac{\|e^*\|}{\sqrt{2N_0}}\right) \tag{5}$$

where $e^*(D)$is the $e(D) \in E$ that yields the dominant term in (4).

Applying (5) to the rate 16/17 code of Fig. 2, we have found that $e^*(D)$ is such that $\|e^*\|^2 = 2$ and corresponds to certain code word pairs, c_0 and c_0', whose inputs differ by $D^k(1+D^{15})$. Clearly, $w_{e^*} = 2$, and we have estimated N_{e^*} to be 0.057 (via a Monte Carlo technique) so that for this code

$$P_b(\text{floor}) \approx 2.8 \times 10^{-5} Q\left(\sqrt{\frac{1}{N_0}}\right) \tag{6}$$

from which a large interleaver gain is evident. In fact, all of the coding gain is interleaver gain because there is no distance gain over the uncoded case for which $\|e^*\|^2 = 2$ is also true and for which $P_b = 4Q(\sqrt{1/N_0})$. This expression in (6) yields the dashed line in Fig. 2. For comparison, a limited search has indicated that $\|e^*\|^2$ is much larger than 2 for the rate 8/9 PCC code.

We have also applied (5) to the rate 8/9 and 16/17 codes of Fig. 3. For the rate 8/9 code, we found that $\|e^*\|^2 = 2$, which corresponds to certain code word pairs, c_0 and c_0', whose inputs differ by $D^k(1 + D^{30})$. We have estimated N_{e^*} to be 0.04 so that for this code, with $w_{e^*} = 2$

$$P_b(\text{floor}) \approx 2.0 \times 10^{-5} Q\left(\sqrt{\frac{1}{N_0}}\right)$$

where again a large interleaver gain is observed together with a lack of distance gain. This expression yields the dashed line in Fig. 3 under the rate 8/9 simulation curve. We remark that, in addition to two-bit error patterns of the form $D^k(1 + D^{30})$, we observed several four-bit error patterns in our simulations, to which we can attribute the large separation between the truncated bound and the simulated performance.

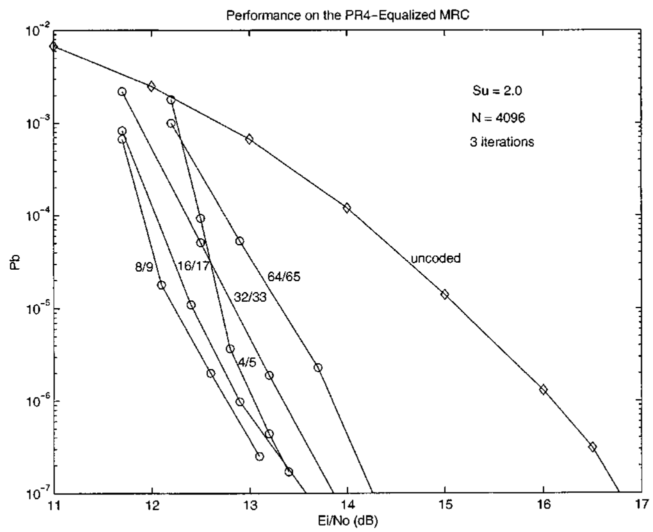

Fig. 4. Performance of the various PCCs on the PR4-equalized MRC with $S_u = 2.0$.

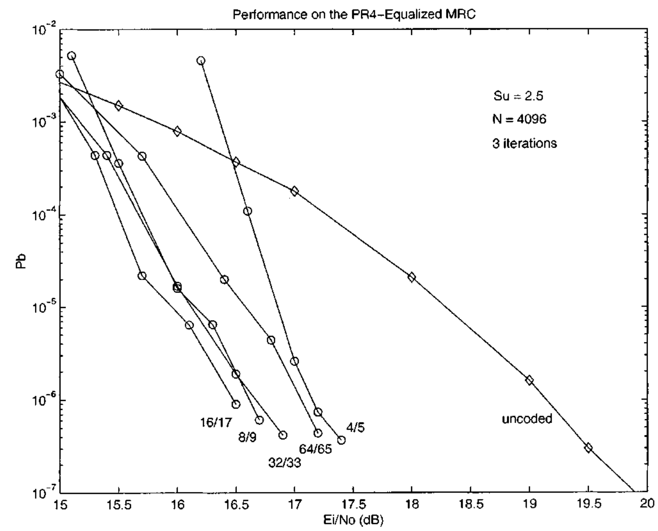

Fig. 5. Performance of the various PCCs on the PR4-equalized MRC with $S_u = 2.5$.

For the rate 16/17 SCC code, we found that $||e^*||^2 = 2$, which corresponds to certain code word pairs, c_0 and c_0', whose inputs differ by $D^k(1 + D^{15})$, $D^k(1 + D^{30})$, and $D^k(1 + D^{45})$. We have estimated N_{e^*} to be 0.9 so that for this code, with $w_{e^*} = 2$

$$P_b(\text{floor}) \approx 4.4 \times 10^{-4} Q\left(\sqrt{\frac{1}{N_0}}\right).$$

This expression yields the dashed line in Fig. 3 under the rate 16/17 simulation curve. In this case, the bound and simulated performance curves are closer, consistent with the fact that weight-two error patterns dominated in our simulations.

B. Colored Noise Performance

As for the performance of the system of Fig. 1 and PCCs, we first point out that presenting the error rate curves as a function of E_b/N_0 is not as simple as in the previous case. The reason is that the energy per code bit E_c does not simply scale with the energy per information bit E_b as in the above case (i.e., $E_c \neq rE_b$). This can be seen from (2), where the basic signaling pulse is $s(t) = h(t) - h(t - T_c) = h(t) - h(t - rT_u)$. Observe that as r decreases, $h(t)$ and $-h(t - rT_u)$ will occur closer together in time (and space) so that the energy in $s(t)$ will also decrease, but not in a linear fashion. (In fact, it is shown in the Appendix that the "code rate loss" is approximately r^2 when $S_c \geq 3$). A simple way to circumvent this problem is to define the SNR to be E_i/N_0. Code comparisons will be fair in this case because the various codes will be facing the same recording channel characterized by the SNR parameter E_i/N_0.[5]

The error rate P_b is plotted against E_i/N_0 in Fig. 4 for the various PCCs for a user density of $S_u = 2.0$ and three decoder iterations. In the figure, we observe that although the rate 4/5 code is superior in the scenario of Fig. 2, it is inferior to the rate 8/9 and 16/17 codes over a large range of error rates (corresponding to "medium" SNRs). This follows from our argument in the introduction where we stated that the higher recording density required by the rate 4/5 code will lead to a discrete-time response $f(D)$ that is more difficult to equalize to the PR4 target and, hence, will result in greater noise enhancement. We observe that 8/9 is the optimal code rate for error rates down to about 1e-7 at this user density. Lastly, we observe that the rate 64/65 PCC is only about 1 dB away from the rate 8/9 PCC near $P_b = 10^{-7}$ (compared with about 3.5 dB in Fig. 2).

[5]We remark that, following a conference version of the present paper, Moon [14] has adopted and extended this SNR definition to include media noise.

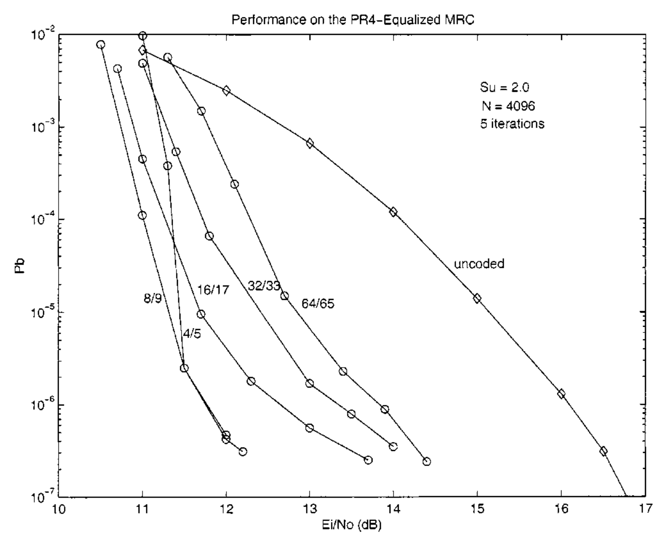

Fig. 6. Performance of the various SCCs on the PR4-equalized MRC with $S_u = 2.0$.

Fig. 5 presents the P_b-curves of the various PCCs versus E_i/N_0 for a user density of $S_u = 2.5$ (and three iterations). As observed in the figure, the rate 4/5 code has the poorest performance, owing to the recording density of $S_c = (5/4)S_u = 3.125$ and the resulting extreme mismatch from PR4 target. The rate 16/17 code is optimal at this higher user density and provides a coding gain of about 2.5 dB over the uncoded case near $P_b = 10^{-6}$.

Fig. 6 presents P_b versus E_i/N_0 curves for the various SCCs for a user density of $S_u = 2.0$ (and five iterations). As for the

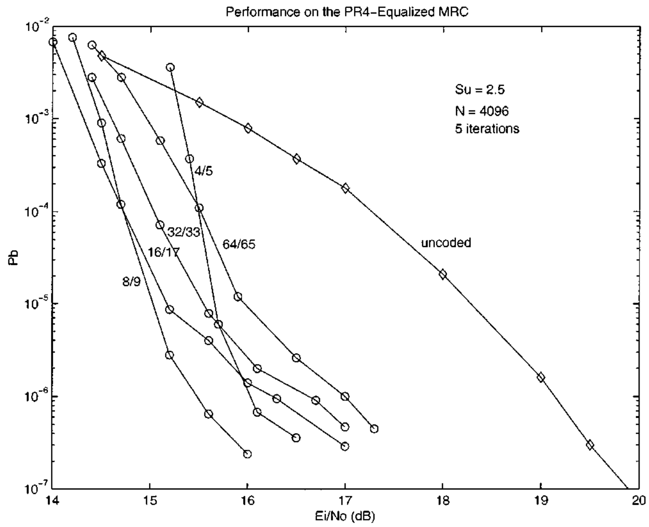

Fig. 7. Performance of the various SCCs on the PR4-equalized MRC with $S_u = 2.5$.

PCC case, the rate 8/9 code is optimal for medium SNRs, although the rate 4/5 code appears to have equal performance beyond $E_i/N_0 = 11.5$ dB. We observe again the flooring effect below $P_b = 10^{-5}$.

Fig. 7 repeats Fig. 6, except for a user density of $S_u = 2.5$. In this case, the rate 16/17 code is (marginally) optimal until about 14.5 dB, beyond which the rate 8/9 code is optimal.

Observe that the coding gains for this colored noise model are on the order of 2 dB less than the corresponding gains for the AWGN model. This is a consequence of the r^2-loss in the colored noise case (mentioned above) versus the r-loss in the AWGN case. It can also be attributed to the fact that the colored noise is more geometrically aligned with PR4 error sequences than is the AWGN noise.

V. Conclusion

We presented a method for determining the optimal code rate for concatenated codes on a PR4-equalized magnetic recording channel modeled by the Lorentzian pulse. Simulation results for rate 4/5, 8/9, 16/17, 32/33, and 64/65 parallel and serial concatenated codes and user densities of $S_u = 2$ and 2.5 were presented. We point out that the optimal code rates may change if the model was modified to include media noise.

We remark that these results are extendible to the general ISI channel, whether or not PR equalization is involved. That is, an optimal code rate exists for any given ISI channel. When PR equalization is not involved, the tradeoff is between code strength and increased ISI due to coding overhead.

Appendix

We examine here the code rate loss by first examining the energy $E_{\rm dibit}$ in the dibit $s(t)$, and then by considering the SNR in the sampled matched filter (MF) output.

The energy in the dibit is simply the zeroth coefficient $R_{s,0}$ in the sampled autocorrelation function $R_s(D)$. This was found by Bergmans [8] (with different notation) to be

$$E_{\rm dibit} = \frac{2E_i}{S_c^2 + 1}.$$

Observe that $E_{\rm dibit} \to 2E_i$ as $S_c \to 0$, as it should. Note also that for $S_c > 3$, say

$$E_{\rm dibit} \simeq \frac{2E_i}{S_c^2} = \frac{2r^2 E_i}{S_u^2} \tag{7}$$

so that $E_{\rm dibit}$ goes as r^2 for large recording densities.

The SNR in the MF output samples may be derived as follows. The response of the MF $s(-t)$ to the kth signal term $(a_k/2)s(t - kT_c)$ in (2) is the convolution

$$(a_k/2)s(t - kT_c) * s(-t) = (a_k/2)R_s(t - kT_c)$$

so that the signal sample at time $t = kT_c$ is

$$(a_k/2)R_{s,0} = (a_k/2)E_{\rm dibit}.$$

Now observe that noise variance at the MF output is

$$\begin{aligned}\sigma^2 &= \frac{N_0}{2}\int_{-\infty}^{\infty} |S(f)|^2\, df \\ &= \frac{N_0}{2}\int_{-\infty}^{\infty} s^2(t)\, dt \\ &= \frac{N_0}{2} E_{\rm dibit}\end{aligned}$$

where $S(f) \triangleq \mathcal{F}\{s(t)\}$. Thus, the SNR in the MF output samples is

$$\text{SNR} = \frac{[(a_k/2)E_{\rm dibit}]^2}{(N_0/2)E_{\rm dibit}} = \frac{E_{\rm dibit}}{2N_0}$$

because $a_k^2 = 1$. Thus, from (7), when $S_c > 3$

$$\text{SNR} \simeq \frac{r^2 E_i}{S_u^2 N_0},$$

and the SNR goes as r^2.

Acknowledgment

The author acknowledges W. Bliss and R. Behrens of Cirrus Logic for introducing him to the E_i/N_0 definition of SNR, A. Ghrayeb of the University of Arizona for help in producing Fig. 2, and the authors of [3]–[5] for providing copies of these papers. He would also like to thank Intel corporation for donating the 450 MHz Pentium II computers on which these simulations were performed.

References

[1] W. Ryan, “Performance of high rate turbo codes on a PR4-equalized magnetic recording channel,” in *Proc. 1998 Int. Conf. Commun*.

[2] W. Ryan, L. McPheters, and S. McLaughlin, “Combined turbo coding and turbo equalization for PR4-equalized Lorentzian channels,” in *Proc. 1998 Conf. Information Systems and Sciences*.

[3] M. Öberg and P. Siegel, “Performance analysis of turbo-equalized dicode partial-response channels,” in *Proc. 36th Allerton Conf. Commun., Control, and Computing*, Sept. 1998.

[4] T. Souvignier, A. Friedmann, M. Öberg, P. Siegel, R. Swanson, and J. Wolf, "Turbo decoding for PR4: Parallel versus serial concatenation," in *Proc. 1999 Int. Conf. Commun*.
[5] L. McPheters, S. McLaughlin, and K. Narayanan, "Precoded PRML, serial concatenation, and iterative (turbo) decoding for magnetic recording," Presented at the 1999 Int. Conf. Magnetics.
[6] W. Ryan, "Design considerations for concatenating convolutional codes with partial response channels," in *Proc. 37th Allerton Conf. Communication, Control, and Computing*, Sept. 1999.
[7] ——, "Concatenated codes for class IV partial response channels," IEEE Trans. Commun., to be published.
[8] J. Bergmans, "Discrete-time models for digital magnetic recording channel," *Philips J. Res.*, vol. 41, pp. 531–558, 1986.
[9] O. Acikel and W. Ryan, "High rate turbo codes for BPSK/QPSK channels," *IEEE Trans. Commun.*, vol. 47, pp. 1315–1323, Sept. 1999.
[10] J. Wolf and G. Ungerboeck, "Trellis coding for partial response channels," *IEEE Trans. Commun.*, pp. 765–773, Aug. 1986.
[11] L. Bahl, J. Cocke, F. Jelinek, and J. Raviv, "Optimal decoding of linear codes for minimizing symbol error rate," *IEEE Trans. Inform. Theory*, vol. IT-20, pp. 284–287, Mar. 1974.
[12] G. D. Forney, "Maximum-likelihood sequence estimation of digital sequences in the presence of intersymbol interference," *IEEE Trans. Inform. Theory*, vol. IT-18, pp. 363–378, May 1972.
[13] S. Benedetto, D. Divsalar, G. Montorsi, and F. Pollara, "Serial concatenation of interleaved codes: Performance analysis, design, and iterative decoding," *IEEE Trans. Inform. Theory*, vol. 44, pp. 909–926, May 1998.
[14] J. Moon, "SNR definition for magnetic recording channels with transition noise," *IEEE Trans. Magn.*, to be published.

William E. Ryan received the B.S.E.E. degree from Case Western Reserve University, Cleveland, OH, in 1981, and the M.S. and Ph.D. degrees in electrical engineering from the University of Virginia, Charlottesville, in 1984 and 1988, respectively.

He has been with The Analytic Sciences Corporation, Ampex Corporation, and Applied Signal Technology prior to his positions in academia. From 1993–1998, he was with the Electrical and Computer Engineering Department faculty at New Mexico State University, Las Cruces. Since August 1998, he has been with the Electrical and Computer Engineering Department at the University of Arizona, Tucson, where he is an Associate Professor. He is an Associate Editor for the IEEE TRANSACTIONS ON COMMUNICATIONS for Modulation, Coding, and Equalization. His research interests are in coding and signal processing for data transmission and storage.