Performance of H.263 Video Transmission over Wireless Channels Using Hybrid ARQ

Hang Liu and Magda El Zarki, *Member, IEEE*

Abstract— **This paper proposes a hybrid ARQ error control scheme based on the concatenation of a Reed–Solomon (RS) code and a rate-compatible punctured convolutional (RCPC) code for low-bit-rate video transmission over wireless channels. The concatenated hybrid ARQ scheme we propose combines the advantages of both type-I and type-II hybrid ARQ schemes. Certain error correction capability is provided in each (re)transmitted packet, and the information can be recovered from each transmission or retransmission alone if the errors are within the error correction capability (similar to type-I hybrid ARQ). The retransmitted packet contains redundancy bits which, when combined with the previous transmission, result in a more powerful RS/convolutional concatenated code to recover information if error correction fails for the individual transmissions (similar to type-II hybrid ARQ).**

Bit-error rate (BER) or signal-to-noise ratio (SNR) of a radio channel changes over time due to mobile movement and fading. The channel quality at any instant depends on the previous channel conditions. For the accurate analysis of the performance of the hybrid ARQ scheme, we use a multistate Markov chain (MSMC) to model the radio channel at the data packet level. We propose a method to partition the range of the received SNR into a set of states for constructing the model so that the difference between the error rate of the real radio channel and that of the MSMC model is minimized. Based on the model, we analyze the performance of the concatenated hybrid ARQ scheme. The results give valuable insight into the effects of the error protection capability in each packet, the mobile speed, and the number of retransmissions. Finally, the transmission of H.263 coded video over a wireless channel with error protection provided by the concatenated hybrid ARQ scheme is studied by means of simulations.

Index Terms— **H.263 video, hybrid ARQ, multistate Markov model, wireless channels.**

I. INTRODUCTION

THE advances in low bit-rate video coding technology have led to the possibility of delivering video services to users through band-limited wireless networks. Both ITU-T/SG15 and ISO-MPEG4 are working to set standards for very low bit-rate video coding. Recently, ITU-T/SG15 finished the first draft recommendation of H.263 which targets the transmission of video telephony through the public switched telephone network (PSTN) at data rates less than 64 kbit/s [1]. The expert's group is starting to adapt H.263 for wireless applications because the low bit rate makes it well suited for current band-limited wireless networks.

Real-time video services require high reliability with a low bounded time delay and a reasonably high transmission rate. Wireless channels, on the other hand, are error prone, time varying, and band limited. Proper error control is necessary to obtain acceptable quality video transmission. Traditionally, forward error correction (FEC) codes are used for real-time services because they maintain a constant throughput and a bounded delay. Wireless channels are time varying. FEC codes can be chosen to guarantee certain quality of service (QOS) requirements for the worst channel conditions. However, this causes unnecessary overhead and reduces throughput while the channel is in a good state. A possible approach to solve this problem is to use unequal error protection techniques [21]. For these schemes, to improve the channel utilization, different error protection capabilities are provided to the data in the coded video bit stream based on the importance of the data to the video quality and its resilience to errors. The other approach is to use adaptive error protection schemes. The advantages of the adaptive error protection schemes are that these schemes do not require extra overhead for the priority layering, and a change in the syntax of the standard single-layer coding schemes such as H.263. Recently, it has been shown that truncated hybrid automatic repeat request (ARQ) schemes can significantly improve video transmission quality because of their adaptability to channel conditions [2]. In this paper, we investigate this further.

There are two conventional classes of hybrid ARQ schemes: type-I and type-II. Type-I hybrid ARQ schemes include parity bits for both error detection and error correction in every transmitted packet. If the number of erroneous bits in a received packet is within the error correction capability of the code, the errors are corrected and the decoded message is accepted by the receiver. If an uncorrectable error pattern is detected, the packet is rejected and a retransmission is requested. The transmitter sends the original packet again. In type-I hybrid ARQ schemes, information can be recovered from each transmitted packet. Appropriately designed error correction overhead can correct the errors in the transmitted packets, thereby reducing the number of retransmissions compared to an error-detection-only pure ARQ scheme and enhancing the system throughput. Since wireless channels are time varying, some algorithms have been developed to adapt the code rate in each (re)transmission to the channel conditions [7]–[9]. A disadvantage of type-I hybrid ARQ schemes is that the uncorrectable packets are discarded by the decoder even if they might contain some useful information.

For type-II hybrid ARQ schemes, the erroneous packet is kept for future use rather than discarded. Redundancy bits

Manuscript received September 1, 1996; revised March 1, 1997.

The authors are with the Video Processing and Telecommunications Laboratory, Department of Electrical Engineering, University of Pennsylvania, Philadelphia, PA 19104 USA.

Publisher Item Identifier S 0733-8716(97)07698-1.

are transmitted only when they are needed, which provides the ability to adapt to changing channel conditions. In the conventional type-II hybrid ARQ schemes based on rate-compatible punctured convolutional (RCPC) codes[1] [3], [4], the initial transmission of a data packet only includes the bits required for a very high-rate code. If the high-rate code is not powerful enough to combat channel errors, then supplemental bits, which were previously deleted by puncturing, are transmitted in the retransmission. The receiver, upon receiving the retransmitted redundancy bits, tries to recover the information by combining the redundancy bits with the previously transmitted data to produce a lower rate error correction code. If the combined decoding scheme fails, then this process continues as needed, decreasing the coding rate down to that of the mother code. If the starting code rate is high and the incremental step is small, it could give good throughput. A special case of type-II hybrid ARQ schemes, first introduced by Lin *et al.* [5], [6], employs a rate 1/2 invertible block or convolutional code. The original data packet and the parity packet are alternatively transmitted until either an error-free packet is received or error correction is possible with the rate 1/2 code obtained by combining the original data packet and the parity packet. However, each individual packet (information packet or parity packet) contains no error protection capability.

For type-II hybrid ARQ schemes, the redundancy bits are added through retransmissions, that is, the type-II hybrid ARQ schemes trade off the number of retransmissions for a higher throughput. For real-time services, delay limits the number of retransmissions that can be used. It is impractical to increase the redundancy step by step starting from a very high-rate code. We need to compromise the throughput to increase the error correction capability for faster convergence and to reduce the delay. Moreover, the conventional type-II hybrid ARQ scheme based on RCPC codes cannot recover information from the retransmitted packet alone [3], [4]. Both random and bursty errors exist on wireless channels. When a packet is in a long deep fade, almost all of the data in the packet are corrupted, and additional parity information for that packet will be useless for the decoder even if the parity retransmission is received with a high signal-to-noise ratio (SNR) value. In addition, the initially transmitted packet may be lost due to header errors so that the parity bits are completely useless. Although the type-II hybrid ARQ scheme based on rate 1/2 block or convolutional codes can recover information from each transmission alone [5], [6], like a pure ARQ scheme, each transmitted packet itself contains no error correction capability. It is desirable that the information can be recovered from each transmission alone, and the transmitted packet itself has some error correction power dependent upon the channel conditions.

In this paper, we design a hybrid ARQ scheme based on the concatenation of a Reed–Solomon (RS) code and an RCPC code (called concatenated hybrid ARQ, CH-ARQ). The CH-ARQ scheme combines the advantages of both type-I and type-II hybrid ARQ schemes, i.e., certain error correction capability is provided in each (re)transmitted packet, and the information can be recovered from each transmission or retransmission alone if the errors are within the error correction capability (similar to type-I hybrid ARQ), and the retransmitted packet consists of redundancy bits which, when combined with the previously transmitted packet, result in a more powerful RS/convolutional concatenated code to recover information if error correction has failed for every single packet (similar to type-II hybrid ARQ). Future networks will support multimedia services. A mobile terminal may set up sessions for voice, data, as well as video, and modify them during a call. The proposed CH-ARQ scheme can flexibly accommodate the individual QOS requirements of different services. For example, the RCPC code can be used to guarantee the QOS requirements of voice services. The RS code and the ARQ scheme can be assigned to an upper layer of the system to improve the QOS for the requirements of video services.

The quality of compressed video transmission strongly depends on the residual errors that remain after the error correction process. The performance analysis and characterization of the residual error are important to understand the effects of errors on video transmission quality, and to effectively design the communication systems and video coding algorithms. Although a lot of work on the performance analysis of hybrid ARQ schemes has appeared in the literature [10]–[14], unfortunately, they are either based on a fully interleaved Rayleigh channel, or an AWGN channel, or a two-state Markov channel without the derivation of the Markov parameters from the real radio channel. Radio channels are time varying, and the channel quality at each instant depends on the previous channel conditions. A multistate Markov chain (MSMC) has been shown to be a much more accurate channel model than the two-state Gilbert–Elliott model [15], [16]. However, previously, MSMC was only used as a bit level model for radio channels [16]. In fact, data are transmitted in packets that contain many bits/symbols. Various error correction/detection, interleaving, and retransmission schemes are employed. Only the residual error is relevant to the quality of video transmission and the effective design of higher layer video encoding algorithms. Furthermore, a complete statistical analysis of the bit/symbol error process can be complex, and often, because of coding and interleaving, quite some approximations are needed to obtain results. In this paper, we use an MSMC as a data packet level model of the radio channel, and propose a method to partition the range of the received SNR into a set of states for constructing the model so that the difference between the error rate of the real radio channel and that of the MSMC model is minimized. Based on the model, we analyze the performance of the CH-ARQ scheme. The effects of the error protection capability in each packet and the mobile speed on the throughput and the residual packet error rate (RPER) are determined.

Unlike data, video transmission can tolerate certain errors (less than 10^{-4}–10^{-3} packet error rate), but it has stringent de-

---

[1] A low-rate $1/\Omega$ code can be periodically punctured (i.e., deleting or not transmitting certain code bits) with period p to obtain a family of codes with rate $p/(p+b)$ where b can be varied between 1 and $(\Omega-1)p$. The concept of RCPC codes was developed by Hagenauer [3]. For a family of RCPC codes, all code bits of high-rate codes can be used by the lower rate codes of the family; this allows transmission of incremental redundancy in hybrid ARQ schemes.

lay requirements. This suggest that we can limit the maximum allowed number of retransmissions in a hybrid ARQ scheme to satisfy the QOS requirements of video communications. It is important to understand the impact of the maximum allowed retransmission number on reliability, throughput, and delay performance. In this paper, the impact of the number of retransmissions on the residual packet error rate, the throughput, and the transmission delay are investigated. Finally, the transmission of H.263 coded video over a wireless channel with error protection provided by the CH-ARQ scheme is studied by means of simulations.

The paper is organized as follows. In the next section, the concatenated hybrid ARQ scheme is described. In Section III, the multistate Markov chain channel model is defined and used to analyze the performance of the CH-ARQ protocol. In Section IV, we investigate the transmission of H.263 coded video over a wireless channel with error protection provided by the CHARQ scheme. Section V concludes our work.

II. Description of the Concatenated Hybrid ARQ Scheme

The concatenated hybrid ARQ scheme combines the type-I and type-II hybrid ARQ. The type-II hybrid ARQ is realized with a half-rate code [5], [6]. The CH-ARQ scheme employs three codes, C_0, C_1, and C_2. C_0 is an (N, K) cyclic redundancy check (CRC) code [5], which is used as the error detection code. C_1 is an RCPC code for error correction, and C_2 is a half-rate invertible shortened RS code $(2k, k, t)$ with m bits per symbol for both error detection and correction. For the invertible code, the information block and the parity block have the same length, and the information block can be obtained uniquely from the parity block by a simple inverting algorithm [6].

Information data are first divided into blocks with k symbols (km bits) per block. An information block D is encoded using the half-rate invertible RS code C_2. The k-symbol-long parity block $P(D)$ corresponding to D is formed. $(D, P(D))$ is a codeword in C_2. After RS encoding, λ consecutive information blocks are transmitted in the initial transmission, and λ corresponding parity blocks are saved in the transmitter buffer for possible transmission at a later time. Before transmission, these λ consecutive information blocks are input row by row into a $\lambda \times k$-symbol interleaving buffer, and then read out column by column in symbols. These interleaved information symbols are converted to $K = \lambda \times k \times m$ bits. Based on the (N, K) CRC code C_0, $(N - K)$ CRC bits are attached to K information bits to form an N-bit information macroblock I. The information macroblock I is appended by M tail bits for terminating the convolutional encoder memory, and then encoded by the RCPC encoder to form an information packet J. J is transmitted. The code rate of the RCPC encoder can be selected based on channel conditions. The code rate determines the redundancy added in the information packet J by the RCPC encoder. A timer at the transmitter is set when J is transmitted.

Let $\tilde{J}$ be the received version of J at the receiver. $\tilde{J}$ is decoded by the RCPC decoder. The CRC check is then performed on the decoder output, and the CRC bits are removed. The decoded output is denoted by $\tilde{I}$ (after the CRC bits are removed). If no error is detected in the decoded sequence $\tilde{I}, \tilde{I}$ is assumed to be error free, and is accepted by the receiver (after deinterleaving with a $\lambda \times k$-symbol deinterleaver). At the same time, a positive ACK is sent to the transmitter. If the presence of an error pattern is detected in $\tilde{I}, \tilde{I}$ is then deinterleaved and stored in the receiver buffer for possible reprocessing at a later time, and no ACK is sent. At the transmitter, if an ACK is received before the timer expires, the transmitter knows that the initial transmission was successful, and it discards the λ corresponding parity blocks. If not, it assumes that uncorrected errors occurred in the initially transmitted packet. The transmitter then interleaves the λ corresponding parity blocks, adds the CRC bits, and forms a parity macroblock $P(I)$. The parity macroblock $P(I)$ is encoded to form a parity packet Q with the RCPC encoder based on the channel conditions [after the M tail bits are attached to $P(I)$]. The parity packet Q is then sent to the receiver.

Let $\tilde{Q}$ be the received parity packet. After $\tilde{Q}$ is received, the receiver first decodes it with the RCPC decoder. The CRC check is then performed on the decoder output, and the CRC bits are removed. The decoded output is denoted by $\tilde{P}(I)$ (after the CRC bits are removed). If no errors are detected in $\tilde{P}(I)$, after deinterleaving, the receiver inverts all the RS parity blocks in $\tilde{P}(I)$, denoted by $I(\tilde{P})$, and accepts $I(\tilde{P})$ as the original information data (since the RS code is invertible). If the presence of an error pattern is detected in $\tilde{P}(I), \tilde{P}(I)$ is deinterleaved and combined with the erroneous data sequence $\tilde{I}$ in the receiver buffer to form the λ rate 1/2 RS codes. Error correction is then performed on the RS codes. If only one retransmission is allowed due to the delay constraints of real-time services or the errors are correctable by the RS codes, the RS decoded message is accepted.

If multiple retransmissions are allowed and an uncorrectable error pattern is detected by the RS decoder, the erroneous parity data $\tilde{P}(I)$ is saved in the receiver buffer, the old erroneous information sequence $\tilde{I}$ is discarded, and the retransmission of the information packet J is requested. When $\tilde{J}$ is received, it is used to recover the information as described before. If this fails, the new erroneous information data $\tilde{I}$ and the erroneous parity data $\tilde{P}(I)$ (previously stored in the receiver buffer) are combined to form the λ rate 1/2 RS codes for error correction. If the errors are still not correctable, the old $\tilde{P}(I)$ is discarded, and $\tilde{I}$ is stored in the receiver buffer. The next retransmission will be the parity packet Q. This process continues, i.e., alternating transmissions of the information packet J and the parity packet Q, until the allowed maximum retransmission number is reached or the data are successfully accepted.

The length of the output error bursts from Viterbi decoders are widely distributed; we interleave the RS codes so that the error bursts are spread among the RS codes. This increases the probability that errors can be corrected by the RS codes. Otherwise, a long block code should be used. Interleaving is done on a symbol rather than a bit basis.

The proposed hybrid ARQ scheme is very flexible. It can adapt to channel conditions, and provide very powerful error correction capabilities. The RCPC code, whose rate can be selected according to the channel conditions, corrects random errors and short bursty errors. When the channel condition is good, the redundancy added by the RCPC coder can be small, and few retransmissions are required. This results in a high throughput with a low time delay. As a result, the video source rate can be high, and a good video quality can be obtained. When the channel condition becomes bad, the RCPC coder adjusts its rate, and adds more bits for channel coding in order to increase its error correction capability and to reduce the frequency of retransmissions.

If the initial transmission suffers from a long deep fade that swamps the RCPC decoder, the extra parity blocks for the RS codes are sent in the retransmission, i.e., error correction parity symbols for the rate 1/2 RS codes are transmitted only when they are needed. The RS parity blocks can recover the data through inversion alone if they are error free or if the errors in them can be corrected by the accompanied RCPC code. This is very important for time-varying wireless channels, where an error burst or a header error might wipe out most of the initial transmission, yet leave the retransmission relatively error free. In addition, the RS parity blocks can be combined with the previously transmitted information blocks to form a powerful RS-RCPC concatenated code to correct errors when error correction has failed for the individual transmissions.

III. PERFORMANCE ANALYSIS

In this section, we analyze the performance of the proposed concatenated hybrid ARQ scheme over a fading radio channel. We model the fading radio channel as a multistate Markov chain (MSMC) because the radio channel is time varying. MSMC as a bit level model for radio channels has been studied, and shown to be much more accurate than the two-state Gilbert–Elliott model [15], [16]. In our communication system, data are transmitted in packets that contain many bits/symbols. Error correction/detection, interleaving, and retransmission schemes are employed. In fact, only residual errors are relevant to the quality of video transmission and the effective design of higher layer video encoding algorithms. A complete statistical analysis of the bit/symbol error process can be complex, and often, because of coding and interleaving, quite some approximations are needed to obtain results. In this paper, we use MSMC as a data packet level model of the radio channel, and propose a method to partition the range of the received SNR into a set of states for constructing the model so that the difference between the error rate of the MSMC model and that of the real radio channel is minimized. Based on the model, we analyze the performance of the concatenated hybrid ARQ scheme. The effects of the error protection capability in each packet, the mobile speed, and the maximum allowed number of retransmissions are investigated. We describe the channel model in Section III-A, and in Section III-B, the analysis is presented. The numerical results are given in Section III-C.

A. Multistate Markov Channel Model

The bit-error rate (BER) or SNR of a radio channel changes over time due to mobile movement and fading. The channel quality at any instant depends on the previous channel condition. This kind of channel can be modeled using a multistate Markov chain. The MSMC model is constructed by partitioning the range of the received SNR into multiple intervals. Each state of the channel corresponds to one of these intervals whose quality is characterized by a particular BER value. The methodology used to map the physical channels to the multistate Markov chain model is important.

In this paper, we consider the Rayleigh fading channel. The probability density function of the received SNR, γ, is

$$p_\gamma(\Upsilon) = \frac{1}{\gamma_0}\exp\left(-\frac{\gamma}{\gamma_0}\right) \tag{1}$$

where γ_0 is the average SNR. The maximum Doppler frequency of the system is given by $f_m = f_c v/c$, where v is the mobile speed, c is the speed of the electromagnetic wave, and f_c is the carrier frequency. For the Rayleigh fading channel, the average level crossing rate at the SNR value γ, which is defined as the expected number of times per second that the SNR passes through the given level γ in the positive or negative direction, can be expressed as [17]

$$N_\Upsilon = \sqrt{\frac{2\pi\gamma}{\gamma_0}} f_m \exp\left(-\frac{\gamma}{\gamma_0}\right). \tag{2}$$

In order to build the MSMC model, we assume that the statistical parameters defining the Rayleigh fading channel vary slowly with respect to the transmission time of a channel symbol, and that the channel state transitions occur only at the end of the packet transmission. An MSMC model can be defined by a state transition matrix, the steady-state probability, and the BER for each state. To characterize the Rayleigh fading channel as a discrete MSMC, the received SNR value is partitioned into a finite set of ranges Φ. The ranges are specified by the endpoints, $0 = x_1 < x_2 < x_3 < \cdots < x_{\Phi+1} = \infty$. If its SNR value is in the range $[x_k, x_{k+1})$, the channel is mapped to the state $s(k)$, which corresponds to a BER value e_k. This mapping process results in some distortion. We therefore need to define a distortion measure for the mapping process, and minimize the difference between the model and the real channel. We wish to minimize the mean-square error (MSE) of the BER value of a state for the MSMC model and the BER for the Rayleigh fading channel within the interval corresponding to that state. The MSE is defined as

$$\mathrm{MSE} = \sum_{k=1}^{\Phi} \int_{x_k}^{x_{k+1}} (e_k - p_b(\gamma))^2 p_\gamma(\gamma)\, d\gamma. \tag{3}$$

The necessary conditions to minimize the MSE can be obtained by differentiating MSE with respect to the e_k's and x_k's and setting derivatives equal to zero:

$$\frac{\partial \mathrm{MSE}}{\partial e_k} = \int_{x_k}^{x_{k+1}} 2(e_k - p_b(\gamma))p_\gamma(\gamma)\, d\gamma = 0 \tag{4}$$

and

$$\begin{aligned}\frac{\partial \mathrm{MSE}}{\partial x_k} &= (e_{k-1} - p_b(x_k))^2 p_\gamma(x_k) \\ &\quad - (e_k - p_b(x_k))^2 p_\gamma(x_k) = 0\end{aligned} \tag{5}$$

where $p_b(\gamma)$ is the BER value when the SNR is γ. Thus, we get that

$$e_k = \frac{\int_{x_k}^{x_{k+1}} p_b(\gamma) p_\gamma(\gamma)\, d\gamma}{\int_{x_k}^{x_{k+1}} p_\gamma(\gamma)\, d\gamma} \tag{6}$$

and

$$p_b(x_k) = \frac{e_{k-1} + e_k}{2}. \tag{7}$$

When the number of states Φ and $p_b(\gamma)$ for a modulation scheme is determined, x_k and e_k can be obtained from (6) and (7) with the iterative algorithm developed in [18]. Note that e_k in (6) is just the average BER, while the received SNR is within the interval corresponding to the MSMC state. In this paper, DPSK modulation is considered, and then the BER is equal to [19]

$$p_b(\gamma) = \frac{\exp(-\gamma)}{2}. \tag{8}$$

Let $\{s(k)\}$ $k = 1, 2, \cdots, \Phi$ denote the set of states of the Φ-state MSMC. After $\{x_k\}$ is determined, the steady-state probability in the state $s(k)$ of the MSMC model is

$$\pi_k = \int_{x_k}^{x_{k+1}} p_\gamma(\gamma)\, d\gamma = \exp\left(-\frac{x_k}{\gamma_0}\right) - \exp\left(-\frac{x_{k+1}}{\gamma_0}\right). \tag{9}$$

We also wish to obtain the transition matrix Λ. Consider a communication system with a transmission rate of R_p packets per second. There are, on average

$$R_p^{(k)} = R_p \pi_k \tag{10}$$

packets per second transmitted while the channel is in state $s(k)$. It is assumed that the channel state can only change to one of the two neighboring states at the end of a packet transmission, i.e., the transitions only occur between neighboring states. The Markov transition probabilities for nonneighboring states are equal to zero, that is

$$\Lambda_{i,j} = 0, \quad |i - j| > 1 \text{ and } i, j = 1, 2, \cdots, \Phi. \tag{11}$$

The transition probabilities between neighboring states can be approximated by [16]

$$\Lambda_{k,k+1} \approx \frac{N_{k+1}}{R_p^{(k)}}, \qquad k = 1, 2, \cdots, \Phi - 1 \tag{12}$$

and

$$\Lambda_{k,k-1} \approx \frac{N_k}{R_p^{(k)}}, \qquad k = 2, \cdots, \Phi \tag{13}$$

and the values of Λ_{kk} are given by

$$\begin{aligned} \Lambda_{k,k} &= 1 - \Lambda_{k,k-1} - \Lambda_{k,k+1}, \qquad k = 2, \cdots, \Phi - 1 \\ \Lambda_{1,1} &= 1 - \Lambda_{1,2} \\ \Lambda_{\Phi\Phi} &= 1 - \Lambda_{\Phi,\Phi-1} \end{aligned} \tag{14}$$

where N_k is the level crossing rate when the value of the received SNR is equal to x_k, which is given by

$$N_k = \sqrt{\frac{2\pi x_k}{\gamma_0}} f_m \exp\left(-\frac{x_k}{\gamma_0}\right), \qquad k = 1, 2, \cdots, \Phi + 1. \tag{15}$$

Note that $N_1 = 0$ for $x_1 = 0$ and $N_{\Phi+1} = 0$ for $x_{\Phi+1} = \infty$. We now summarize the procedures to construct the MSMC model.

1) Based on (6) and (7), the iterative algorithm [18] is used to obtain the x_k's and the e_k's, $k = 1, 2, 3, \cdots, \Phi$.
2) The steady-state probabilities are obtained from (9).
3) Equations (10)–(15) are used to determine the MSMC transition matrix.

It is clear that (12) and (13) only hold when the fading is slow enough, i.e., the level crossing rate N_k and/or N_{k+1} is much smaller than the value $R_p^{(k)}$:

$$\left(C = \frac{N_k + N_{k+1}}{R_p^{(k)}}\right) < 1, \qquad k = 1, 2, \cdots, \Phi. \tag{16}$$

When this condition cannot be satisfied, we can reduce the number of states by combining the states until the condition is satisfied. Errors can be perfectly randomized in a packet through interleaving when the fading is very fast. As a matter of fact, this perfectly interleaved channel is a special case of the MSMC model with the number of states reduced to one.

Finally, we also want to derive the channel δ-step transition probability. The channel δ-step transition probability $P^{(\delta)}(s(j)|s(i))$ is defined as the probability of being in state $s(j)$ if the channel was in state $s(i)$ δ-packet transmission durations back [15]. The δ-step transition probability matrix can be obtained as

$$T = \Lambda^\delta. \tag{17}$$

B. Performance Analysis of the Concatenated Hybrid ARQ Scheme

In this subsection, the throughput, the RPER, and the transmission delay of the CH-ARQ scheme are analyzed. The channel is modeled by the MSMC channel defined in the previous subsection.

For an RCPC code with Viterbi decoding, given that the channel state is s_j when a packet is transmitted, the event error probability at any time unit $p_{ev}(s_j)$ and the postdecoding bit-error probability $p_{\text{bit}}(s_j)$ are bounded, respectively, by [3], [4]

$$p_{ev}(s_j) \le \sum_{d=d_{\text{free}}}^{\infty} a_d p_d(s_j) \tag{18}$$

and

$$p_{\text{bit}}(s_j) \le \frac{1}{p} \sum_{d=d_{\text{free}}}^{\infty} c_d p_d(s_j) \tag{19}$$

where d_{free} is the free distance of the code; p is the number of information bits per time unit, i.e., the puncturing period; a_d represents the number of incorrect paths at distance d; c_d is the number of erroneous information bits on these distance

d paths; and $p_d(s_j)$ is the probability that a wrong path at distance d is selected when the packet is transmitted in channel state s_j.

Let $\{A_j^c|s_j\}$ and $\{A_j^e|s_j\}$ be the events that the jth retransmission of a packet, data, or parity of the RS code contains no errors and errors, respectively, after convolutional decoding and error detection with the CRC code C_0 ($j=0$ represents the initial transmission), given that the channel state for the jth retransmission is s_j. We assume that there are no undetected errors. Then, the probabilities of these events are given by [3], [4]

$$\Pr(A_j^c|s_j) = (1-p_{ev}(s_j))^{(N+M)p} \tag{20}$$

and

$$\Pr(A_j^e|s_j) = 1-(1-p_{ev}(s_j))^{(N+M)p}. \tag{21}$$

Given that the channel states for the $(j-1)$th and the jth retransmissions $(j>0)$ are s_{j-1} and s_j, respectively, $\{G_j^c|s_{j-1},s_j\}$ and $\{G_j^e|s_{j-1},s_j\}$ denote the events that the decoded data block (consisting of λ RS codes) contains no errors and errors, respectively, after the 1/2 rate RS decoder combines the $(j-1)$th and jth retransmission and decodes them. The probability of these events are given by

$$\Pr(G_j^c|s_{j-1},s_j) = P_{\mathrm{RS}}^{\lambda}(c|s_{j-1},s_j) \tag{22}$$

and

$$\Pr(G_j^e|s_{j-1},s_j) = 1-P_{\mathrm{RS}}^{\lambda}(c|s_{j-1},s_j) \tag{23}$$

where $P_{\mathrm{RS}}(c|s_{j-1},s_j)$ is the probability of a correct decoding event of $(2k,k,t)$ RS decoding, given that the channel states for the two successive transmissions are s_{j-1} and s_j, respectively. We have

$$P_{\mathrm{RS}}(c|s_{j-1},s_j) = \sum_{t_1\ge 0}\sum_{\substack{t_2\ge 0\\ t_1+t_2\le t}} q_{j-1}^{t_1}q_j^{t_2} \tag{24}$$

and

$$q_j^{t_2} = \binom{k}{t_2}\varepsilon_j^{t_2}(1-\varepsilon_j)^{k-t_2} \tag{25}$$

where ε_j is the symbol error rate before the RS decoding when the channel state is s_j. In the following analysis, we also need to know the conditional probability $\Pr(G_j^c|s_{j-1}s_jA_{j-1}^eA_j^e)$ that, given that the two individual packets are in error after the convolutional decoding, the RS decoding by combining these two packets yields a decoded data block without errors

$$\Pr(G_j^c|s_{j-1}s_jA_{j-1}^eA_j^e) = \frac{\Pr(A_{j-1}^eA_j^eG_j^c|s_{j-1}s_j)}{\Pr(A_{j-1}^e|s_{j-1})\Pr(A_j^e|s_j)} \tag{26}$$

where

$$\begin{aligned}
&\Pr(A_{j-1}^eA_j^eG_j^c|s_{j-1}s_j)\\
&\qquad = \Pr(G_j^c|s_{j-1}s_j) - \Pr(A_{j-1}^cG_j^c|s_{j-1}s_j)\\
&\qquad\quad - \Pr(A_j^cG_j^c|s_{j-1}s_j) + \Pr(A_{j-1}^cA_j^cG_j^c|s_{j-1}s_j)\\
&\Pr(A_{j-1}^cG_j^c|s_{j-1}s_j) = \Pr(A_{j-1}^c|s_{j-1})\left(\sum_{t_2=0}^{t} q_j^{t_2}\right)^{\lambda}\\
&\Pr(A_j^cG_j^c|s_{j-1}s_j) = \Pr(A_j^c|s_j)\left(\sum_{t_1=0}^{t} q_{j-1}^{t_1}\right)^{\lambda}\\
&\Pr(A_{j-1}^cA_j^cG_j^c|s_{j-1}s_j) = \Pr(A_{j-1}^c|s_{j-1})\Pr(A_j^c|s_j)
\end{aligned} \tag{27}$$

and also

$$\Pr(G_j^e|s_{j-1}s_jA_{j-1}^eA_j^e) = 1-\Pr(G_j^c|s_{j-1}s_jA_{j-1}^eA_j^e). \tag{28}$$

After receiving the jth retransmission of a packet $(j>0)$, the receiver tries to recover the information from the jth retransmission itself by decoding the convolutional code. If this fails, it combines the jth retransmission and $(j-1)$th retransmission to recover the information using the rate 1/2 RS code. Let E_j^c and E_j^e be the events that, after receiving the jth retransmission of a packet, the receiver correctly recovers the information or fails to recover the information, respectively. It is assumed that the undetected error probability is negligible. Then,

$$\Pr(E_j^c|s_{j-1}s_j) = \Pr(A_j^c|s_j) + \Pr(A_j^eG_j^c|s_{j-1}s_j) \tag{29}$$

and

$$\Pr(E_j^e|s_{j-1}s_j) = \Pr(A_j^eG_j^e|s_{j-1}s_j). \tag{30}$$

If the allowed maximum number of retransmissions is L $(L>0)$, the packet will definitely be accepted after L retransmissions whether it is correct or not. Given the channel states $s_0,s_1,s_2,\cdots,s_L$ for the $L+1$ successive transmissions of a packet, respectively, the average number of transmissions is equal to [14]

$$\begin{aligned}
&E[\mathrm{Tr}|s_0,s_1,s_2,\cdots,s_L]\\
&\quad = \Pr(A_0^c|s_0) + 2\Pr(A_0^eE_1^c|s_0,s_1)\\
&\qquad + \cdots + L\Pr(A_0^eE_1^eE_2^e\cdots E_{L-2}^eE_{L-1}^c|s_0,s_1\cdots s_{L-1})\\
&\qquad + (L+1)\Pr(A_0^eE_1^eE_2^e\cdots E_{L-2}^eE_{L-1}^e|s_0,s_1\cdots s_{L-1})\\
&\quad = 1+\Pr(A_0^e|s_0) + \Pr(A_0^eE_1^e|s_0,s_1)\\
&\qquad + \cdots + \Pr(A_0^eE_1^eE_2^e\cdots E_{L-2}^eE_{L-1}^e|s_0,s_1\cdots s_{L-1}).
\end{aligned} \tag{31}$$

For the Φ-state MSMC model, the average number of transmissions $E[\mathrm{Tr}]$ is obtained by averaging (31) over all possible transmission channel states

$$\begin{aligned}
E[\mathrm{Tr}] = 1 &+ \sum_{s_0=s(1)}^{s(\Phi)} P(s_0)\Pr(A_0^e|s_0)\\
&+ \sum_{s_0=s(1)}^{s(\Phi)}\sum_{s_1=s(1)}^{s(\Phi)} P(s_0)P^{(\delta)}(s_1|s_0)\Pr(A_0^eE_1^e|s_0,s_1)\\
&+ \cdots + \sum_{s_0=s(1)}^{s(\Phi)}\sum_{s_1=s(1)}^{s(\Phi)}\cdots\sum_{s_{L-1}=s(1)}^{s(\Phi)} P(s_0)P^{(\delta)}(s_1|s_0)\\
&\cdot P^{(\delta)}(s_{L-1}|s_{L-2})\\
&\cdot \Pr(A_0^eE_1^eE_2^e\cdots E_{L-2}^eE_{L-1}^e|s_0,s_1\cdots s_{L-1})
\end{aligned} \tag{32}$$

where $P(s_0)$ is the probability that the channel starts in state s_0, i.e., the steady-state probability of the channel state s_0; $P^{(\delta)}(S_j|S_{j-1})$ $(j=1,2,\cdots,L)$ is the channel δ-step transition probability from state s_{j-1} to state s_j, which is obtained from the elements of the δ-step transition matrix in (17). δ represents the number of packets that can be transmitted during one channel round-trip delay, i.e., the interval between the two successive transmissions of a packet containing the same information (a data packet and its RS code parity packet are considered to contain the same information) in units of packet transmission duration. In order to obtain the analysis solutions, we approximate the analysis by assuming that each RS decoding attempt is statistically independent (actually, this gives the lower bound). Then we have

$$\begin{aligned}\Pr(A_0^e E_1^e E_2^e \cdots E_j^e | s_0, s_1 \cdots s_j) &= \Pr(A_0^e A_1^e G_1^e A_2^e G_2^e \cdots A_j^e G_j^e | s_0 s_1 \cdots s_j) \\ &= \Pr(A_0^e | s_0) \Pr(A_1^e | s_1) \cdots \\ &\quad \cdot \Pr(A_j^e | s_j) \Pr(G_1^e | s_0 s_1 A_0^e A_1^e) \Pr(G_2^e | s_1 s_2 A_1^e A_2^e) \\ &\quad \cdot \Pr(G_3^e | s_2 s_3 A_2^e A_3^e) \cdots \Pr(G_j^e | s_{j-1} s_j A_{j-1}^e A_j^e). \qquad (33)\end{aligned}$$

Following the steps in [14], we define a matrix such that $H_{ij} = T_{ji} \Pr(A^e | s(i)) \Pr(G^e | s(j) s(i) A_j^e A_i^e)$, where $T = \Lambda^\delta$ denotes the δ-step transition probability matrix of the channel (the first and the second subscripts of the matrix element denote the row and column indexes, respectively), and A_i^e is the event that the packet contains errors after the convolutional decoding, given that the channel state is $s(i)$. That is, we have (34), shown at the bottom of the page, and a column vector

$$V = [\pi_1 \Pr(A^e | s(1)) \pi_2 \Pr(A^e | s(2)) \cdots \pi_\Phi \Pr(A^e | s(\Phi))]' \qquad (35)$$

where π_j is the steady-state probability of the channel state $s(j)$ given in (9). Based on (32)–(35), we have

$$E[\mathrm{Tr}] = 1 + \sum_{j=1}^{L} U H^{j-1} V \qquad (36)$$

where U is a Φ-dimension unit row vector, i.e., $U = [1\,1 \cdots 1]$.

The throughput, defined as the average number of accepted information bits by the receiver per transmitted channel bit [4], [14], can be expressed as

$$\eta = \frac{\alpha}{E[\mathrm{Tr}]} \cdot \frac{K}{N+M} \qquad (37)$$

where α is the RCPC code rate; the factor $K/(N+M)$ is the loss in throughput due to the added CRC bits for error detection and due to the M tail bits appended to each transmitted sequence for clearing the memory of the convolutional encoder, i.e., the coding overhead is excluded when calculating the throughput.

The average RPER in the packets accepted by the receiver, after the maximum number of retransmissions L has been reached, is equal to

$$\begin{aligned}\mathrm{RPER} = &\sum_{s_0=s(1)}^{s(\Phi)} \sum_{s_1=s(1)}^{s(\Phi)} \cdots \sum_{s_L=s(1)}^{s(\Phi)} P(s_0) P^{(\delta)}(s_1|s_0) \\ &\cdot P^{(\delta)}(s_L | s_{L-1}) \\ &\cdot \Pr(A_0^e E_1^e E_2^e \cdots E_L^e | s_0, s_1, \cdots, s_L) = U H^L V. \qquad (38)\end{aligned}$$

Following the same approach to get the average number of transmissions, the average transmission delay can be given by

$$E[D] = t_o + t_r \sum_{j=1}^{L} U H^{j-1} V \qquad (39)$$

where t_o is the one-way delay and t_r is the ARQ time out. If we assume that the ARQ time out is the round-trip delay and the one-way delay is half of the round-trip delay, i.e., $t_r = 2t_o$, we have

$$E[D] = 2t_o E[\mathrm{Tr}] - t_o. \qquad (40)$$

C. Numerical Results

In this subsection, we present some numerical results based on the analysis in the last subsection. For the sake of illustration, we consider the following values for the parameters: the length of each packet is 420 bits. Three RCPC code rates 1, 4/5, 1/2 are used: 1) for rate 1 RCPC code, the packet consists of 400 data bits and 20 CRC bits; 2) for rate 4/5 RCPC code, the packet consists of 320 data bits, 12 CRC bits, and four tail bits which are encoded by the rate 4/5 RCPC code before transmission; 3) for rate 1/2 RCPC code, the packet contains 192 data bits, 12 CRC bits, four tail bits, and two stuffing bits which are encoded by the rate 1/2 RCPC code for the transmission. The data bits may be the original information data or the parity generated by the RS code. We consider a (2, 1, 4) convolutional code as the mother code of the RCPC code family with Viterbi decoding [3]. An invertible shortened RS code (8, 4, 2) with 4 bits per symbol is chosen as code C_2. This RS code is very easy to decode. The packets are transmitted over a TDMA radio channel. For the TDMA channel, the time is divided into slots, where each slot is equal to the packet transmission duration. The transmitter and the receiver are synchronized with the slotted channel. In our example, the user data rate is 32 kbit/s. The packet transmission duration (a slot time) is 0.21 ms, and the interval between the two successive transmissions from a user (packet transmission interval) is 13.125 ms. The round-trip delay is assumed to be the interval between the two consecutive transmissions from a user. The channel is modeled by the MSMC, and the number of states is initially chosen to be 50. If the condition in (16) is not satisfied, the number of states is reduced until the condition is satisfied. The modulation scheme is DPSK with a carrier frequency of 1.9 GHz.

Fig. 1 shows the RPER for the RCPC code rates of 1 (i.e., no RCPC code; this scheme is equivalent to a conventional type-II hybrid ARQ scheme using a rate 1/2 RS code [5]), 4/5, and 4/8 (i.e., 1/2) in the proposed CH-ARQ scheme, respectively, when the mobile speed is 2 km/h and the maximum allowed number of retransmissions L is 1. The corresponding throughputs are plotted in Fig. 2. For comparison, the simulation results are

$$H = \begin{bmatrix} T_{11} \Pr(A^e|s(1)) \Pr(G^e|s(1)s(1)A_1^e A_1^e) & \cdots & T_{\Phi 1} \Pr(A^e|s(1)) \Pr(G^e|s(\Phi)s(1)A_\Phi^e A_1^e) \\ T_{12} \Pr(A^e|s(2)) P(G^e|s(1)s(2)A_1^e A_2^e) & \cdots & T_{\Phi 2} \Pr(A^e|s(2)) \Pr\big(G^e|s(\Phi)s(2)A_\Phi^e A_2^e\big) \\ \cdots & \cdots & \cdots \\ T_{1\Phi} \Pr(A^e|s(\Phi)) \Pr(G^e|s(1)s(\Phi)A_1^e A_\Phi^e) & \cdots & T_{\Phi\Phi} \Pr(A^e|s(\Phi)) \Pr(G^e|s(\Phi)s(\Phi)A_\Phi^e A_\Phi^e) \end{bmatrix} \qquad (34)$$

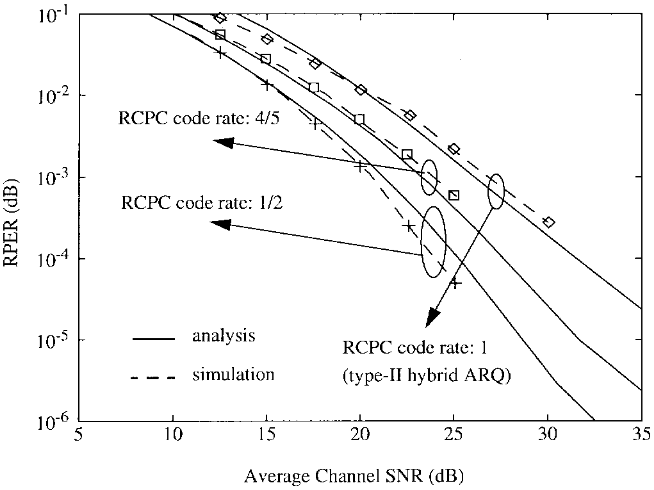

Fig. 1. RPER for different rates of the RCPC code in the CH-ARQ scheme when the mobile speed is 2 km/h.

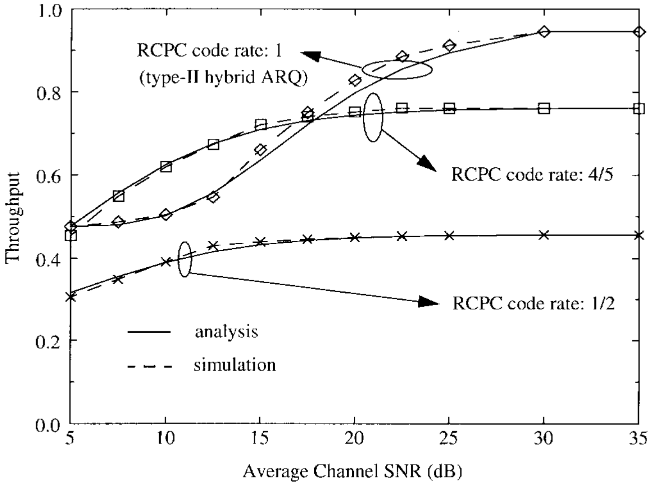

Fig. 2. Throughput for different rates of the RCPC codes in the CH-ARQ when the mobile speed is 2 km/h.

also shown. We see that the theoretical values agree reasonably well with the simulation results. A lower rate RCPC code generally results in better RPER performance due to more powerful error correction capabilities, but the optimal value of the throughput depends on both the rate of the RCPC code and the channel conditions. An appropriate rate of RCPC code can enhance the throughput compared to no RCPC code (i.e., a RCPC code rate of 1) within a certain range of the channel SNR. This is because the error correction capability of the RCPC code corrects the vast majority of packets, thereby greatly reducing the frequency of retransmissions. Under a certain channel condition, it is possible to find an appropriate code rate which yields low RPER and high throughput. This also indicates that an adaptive scheme can be used to achieve the optimal performance for the slow fading channel. In this paper, we do not consider the design of the adaptive algorithm. The adaptive algorithms developed for type-I hybrid ARQ schemes can be used with minor modifications [7]–[9].

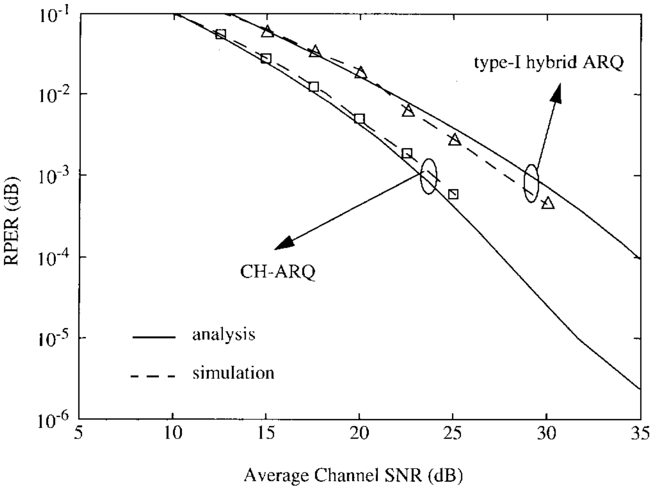

Fig. 3. RPER of CH-ARQ and type-I hybrid ARQ.

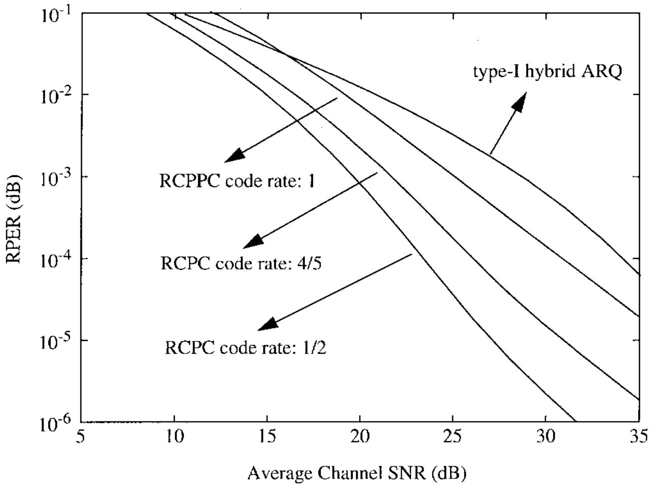

Fig. 4. RPER for different schemes when the mobile speed is 100 km/h.

Fig. 3 shows the RPER of the CH-ARQ and type-I hybrid ARQ when the RCPC code rate is 4/5 and L is equal to 1. Once again, the numerical results are matched to the simulation results. The CH-ARQ has the same throughput as the corresponding type-I hybrid ARQ (with the same rate of RCPC code) when the maximum allowed number of retransmissions is one; however, its RPER is always lower than the corresponding type-I hybrid ARQ, that is, the CH-ARQ always outperforms the corresponding type-I hybrid ARQ. This is because the retransmitted parity packets can be used to recover the information alone or can be combined with previously transmitted information data to form an RS-RCPC concatenated code to provide the CH-ARQ scheme with more powerful error correction capability.

Fig. 4 shows the RPER and the throughput of the CH-ARQ scheme with RCPC code rates of 1, 4/5, and 4/8, respectively when the mobile speed is 100 km/h and the maximum allowed number of retransmission is one. The RPER of the type-I hybrid ARQ using a rate 4/5 RCPC code is also shown for comparison. The corresponding throughputs are plotted in Fig. 5. We notice that the RCPC code improves the relative performance of the RPER and the throughput when the mobile speed become high. This is because higher speed results in more random error patterns, and the error correction codes are more beneficial.

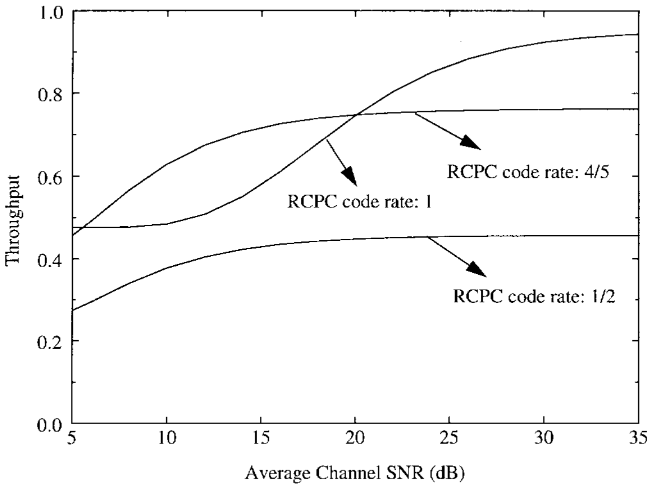

Fig. 5. Throughput for different schemes when the mobile speed is 100 km/h.

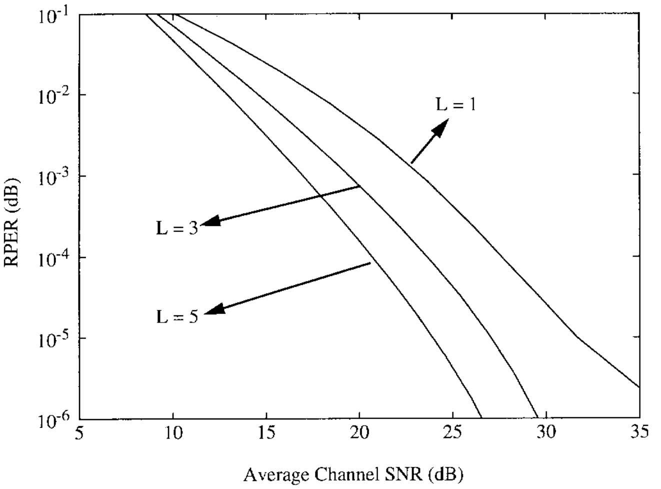

Fig. 6. RPER for different maximum allowed number of retransmissions.

We also would like to investigate the impact of the number of retransmissions on the performance. Fig. 6 shows the RPER of the CH-ARQ scheme for the maximum allowed number of retransmissions L equal to 1, 3, 5, and infinity, respectively. The mobile speed is 2 km/h and the rate of the RCPC code is 4/5. The corresponding throughputs and the transmission delays are depicted in Figs. 7 and 8, respectively.

We notice an improvement in RPER performance for increasing the allowed maximum number of retransmissions. However, the tradeoff is lower throughput and higher delay. For $L =$ infinity, the throughput may be very low when the channel conditions are very poor. Unlike pure data, video can tolerate certain error probability. Therefore, we can choose the number of retransmissions based on the channel conditions to satisfy the requirement of RPER, throughput, and delay. In the next section, we study the transmission of H.263 video with error protection provided by the CH-ARQ scheme.

IV. SIMULATION OF H.263 VIDEO TRANSMISSION

Fig. 9 shows the block diagram of the video transmission system under investigation using the proposed CH-ARQ error control scheme. We assume that the selective repeat protocol is used and the feedback channel is error free. For the simulation, the video sequence is first encoded into a bit stream by an H.263 encoder, and is then transmitted using the proposed CH-ARQ scheme. The interleaver following the RCPC encoder randomizes the error bursts in a transmitted packet so that the RCPC codes, which are ideally suited to correct uncorrelated errors, can handle them better. The packet is sent over the wireless channel. At the receiver, the video sequence is reconstructed, possibly distorted by transmission errors.

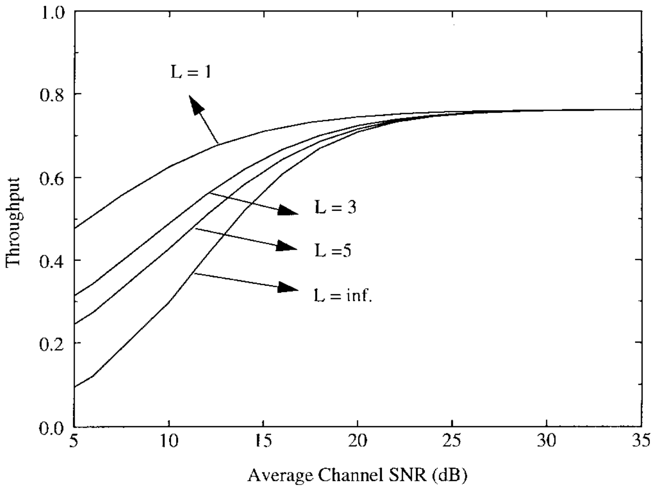

Fig. 7. Throughput for different maximum allowed number of retransmissions.

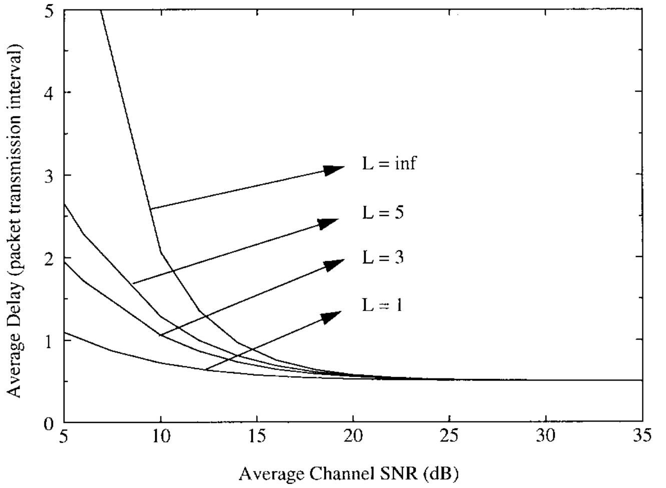

Fig. 8. Transmission delay for different maximum allowed number of retransmissions.

Simulation was carried out on the QCIF "Mother and Daughter" sequence which contains typical video telephony-like images. The original (Y, Cb, Cr) 4:1:1 video sequence is encoded with 15 frames/s. Video encoding and decoding are performed with the modified Telenor R&D H.263 software. A Rayleigh fading simulator is used to simulate the TDMA radio channel [17]. The simulator generates the Rayleigh distributed envelope of the received signal summing the output of several low-frequency oscillators with phases selected to be

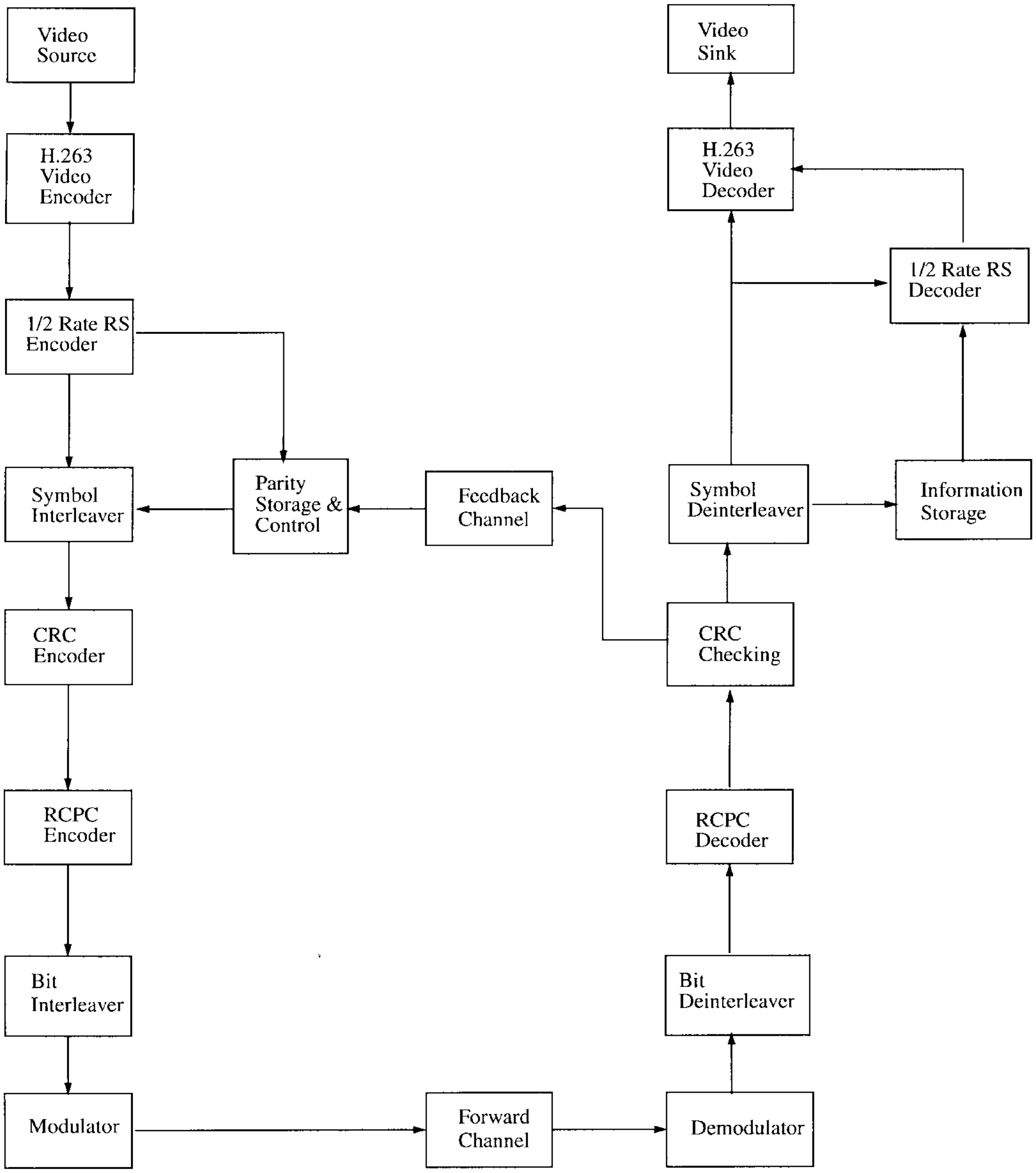

Fig. 9. Block diagram of the video transmission system using the proposed CH-ARQ scheme.

uniformly distributed about the unit circle. For the simulation, 30 oscillators are used to obtain the Rayleigh fade profile. The mobile speed is assumed to be 2 km/h. The round-trip delay is the packet transmission interval, and the allowed maximum number of retransmissions L is one. The modulation scheme is DPSK with a carrier frequency of 1.9 GHz, which is in the frequency band of the emerging personal communication system (PCS).

In order to measure the video quality, two metrics are used in this paper: average peak signal-to-noise ratio (PSNR) and an objective video quality assessment scheme based on the human visual system [20] which uses a grade point (GP). The average PSNR gives the general quality of a video sequence after transmission; however, the GP can reflect the overall viewing quality of a video sequence as perceived by a human being. The objective video quality assessment scheme gives a grade point ranging from 1.0 to 5.0, with 5.0 meaning excellent quality, 4.0 meaning good quality, 3.0 meaning acceptable quality, 2.0 meaning bad quality, and 1.0 meaning absolutely unacceptable quality.

The overall bit rate, including the video and error control overhead, is always 32 kbit/s for all simulations. As we know, the temporal and spacial location of an error in the video bit stream determines its impact on the quality of the reconstructed video sequence. The coded sequence with error protection is transmitted 20 times using different starting points in the fading simulator. The average PSNR value and GP over all of the runs are presented. This reduces the sensitivity of the simulation results to error location, and the performance of the error control schemes stands out.

Figs. 10 and 11 depict the average PSNR and GP of the reconstructed video sequence when the rates of the RCPC code in the CH-ARQ scheme are 1, 4/5, and 4/8, respectively. For each case, the video quality rapidly degrades as the channel SNR decreases below a threshold. This is because there is a dramatic drop in video quality once errors occur in the headers.

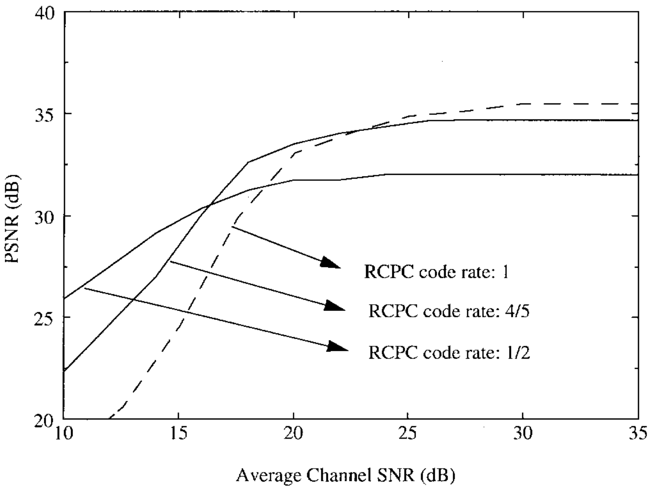

Fig. 10. Average PSNR for different rates of the RCPC code in the CH-ARQ scheme.

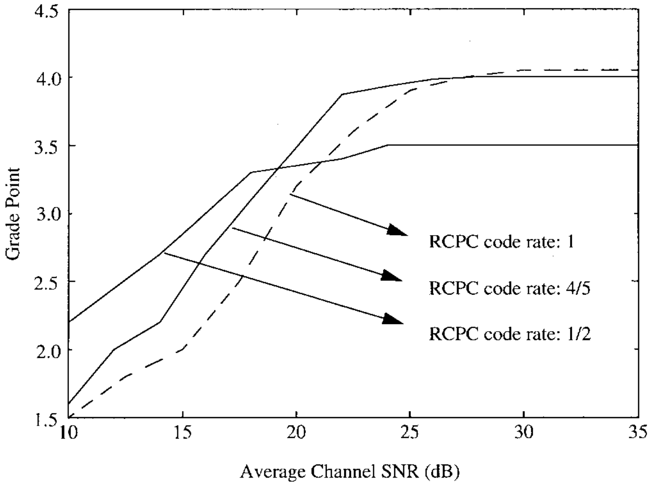

Fig. 11. Grade point for different rates of the RCPC code in the CH-ARQ scheme.

The threshold of the curve occurs at a lower channel SNR when using a lower rate RCPC code. However, the PSNR and GP are smaller for the lower rate RCPC code at high channel SNR when there is no transmission error because the higher overhead required by the lower rate RCPC code reduces the throughput, i.e., the video source rate. For each value of channel SNR, there is an optimal RCPC code rate which maximizes video quality. Once again, this indicates that an adaptive scheme can be used to achieve the optimal performance for the slow fading channel. In this paper, we do not consider the design of the adaptive algorithm. The adaptive algorithms developed for type-I hybrid ARQ schemes can be used with minor modifications [7]–[9].

Figs. 12 and 13 show the performance of the CH-ARQ and the type-I hybrid ARQ with a RCPC code rate of 4/5 for H.263 video transmission. The CH-ARQ gives better performance than the corresponding type-I hybrid ARQ schemes using only convolutional codes. This is because the retransmitted parity packets can be used to independently recover the information; if this fails, it can be combined with previously transmitted information data to form an RS-RCPC concatenated code to provide the CH-ARQ scheme with more powerful error correction capability at no cost of extra overhead.

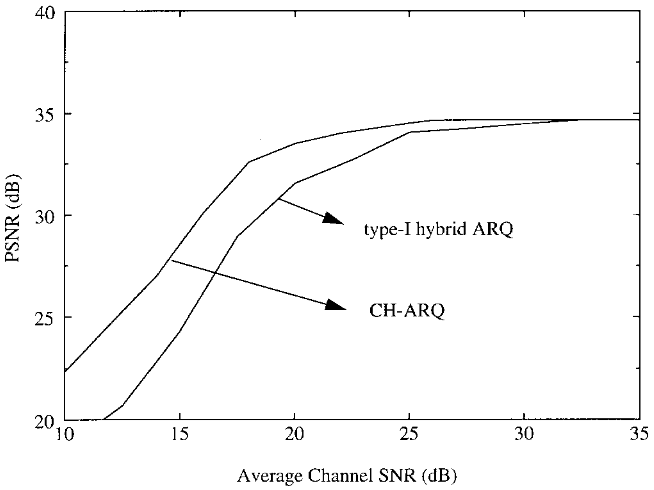

Fig. 12. Average PSNR for the CH-ARQ and the type-I hybrid ARQ.

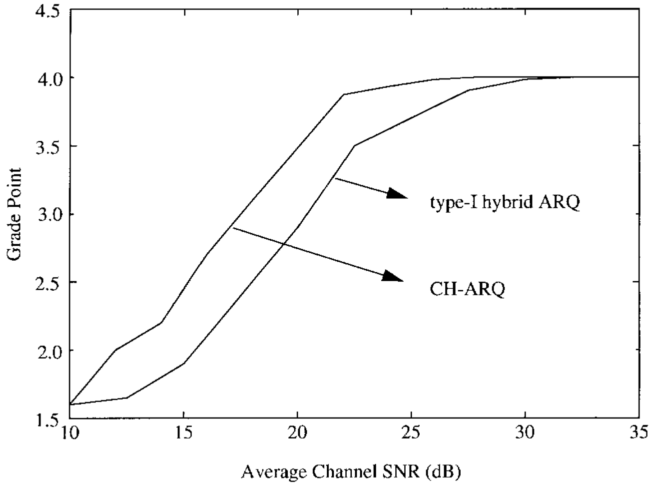

Fig. 13. Grade point for the CH-ARQ and the type-I hybrid ARQ.

V. CONCLUSION

With the proposed CH-ARQ scheme, it is easy to reach high system reliability and throughput because of its ability to adapt to varying channel conditions and because of the powerful error correction capability of the concatenated RS and RCPC codes. The error correction power of the RCPC code reduces the frequency of retransmissions, thereby maintaining a low time delay with a high throughput when the channel is not in long error bursts. Maximum video quality can therefore be obtained. When the initial transmission encounters an unexpected slow deep fade, which causes long error bursts that are beyond the power of the RCPC, retransmission is performed. The retransmitted parity bits can be used independently to recover the information or can be combined with previously transmitted information bits to form an RS-RCPC concatenated code to provide the CH-ARQ with a powerful error correction capability.

For the accurate analysis of the performance of the hybrid ARQ scheme, we use an MSMC to model a radio channel at the data packet level, and propose a method to partition the range of the received SNR into a set of states for constructing the model so that the difference between the error rate for the real radio channel and that for the MSMC model is minimized. Based on this model, we analyze the performance of the concatenated hybrid ARQ scheme. The analysis results give valuable insight regarding the effects of the error protection capability in each packet, the mobile speed, and the number of retransmissions on the residual packet error rate, the throughput, and the transmission delay. Finally, the transmission of H.263 coded video over a wireless channel with error protection provided by the CH-ARQ scheme is studied by means of simulations.

REFERENCES

[1] ITU-T/SG15, Draft Recommendation H.263, *Video Coding for Low Bitrate Communication*, July 1995.

[2] H. Liu and M. El Zarki, "Delay bounded type-II hybrid ARQ for video transmission over wireless networks," in *Proc. Conf. Inform. Sci. Syst.*, Princeton, NJ, Mar. 1996.

[3] J. Hagenauer, "Rate-compatible punctured convolutional codes (RCPC Codes) and their applications," *IEEE Trans. Commun.*, vol. 36, pp. 389–400, Apr. 1988.

[4] S. Kallel and D. Haccoun, "Generalized type-II hybrid ARQ scheme using punctured convolutional coding," *IEEE Trans. Commun.*, vol. 38, pp. 1938–1946, Nov. 1990.

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

[6] Y. Wang and S. Lin, "A modified selective-repeat type-II hybrid ARQ system and its performance analysis," *IEEE Trans. Commun.*, vol. COM-31, pp. 593–608, May 1983.

[7] K. Wu, S. Lin, and M. Miller, "A hybrid ARQ scheme using multiple shorten cyclic codes," in *Proc. IEEE GLOBECOM*, 1982, pp. C8.61–C8.65.

[8] A. Shiozaki, K. Okuno, K. Suzuki, and T. Segawa, "A hybrid ARQ scheme with adaptive forward error correction for satellite communications," *IEEE Trans. Commun.*, vol. 39, pp. 482–484, Apr. 1991.

[9] Z. Sayeed and V. Weerackody, "Adaptive hybrid ARQ scheme for slow fading channels," in *Proc. Conf. Inform. Sci. Syst.*, Princeton, NJ, Mar. 1996.

[10] S. Kallel, "Analysis of a type-II hybrid ARQ scheme with code combining," *IEEE Trans. Commun.*, vol. 38, pp. 1133–1137, Aug. 1990.

[11] S. B. Wicker, "High-reliability data transfer over the land mobile radio channel using interleaved hybrid-ARQ error control," *IEEE Trans. Veh. Technol.*, vol. 39, pp. 48–55, Feb. 1990.

[12] Q. Yang and V. K. Bhargava, "Optimum coding design for type-I hybrid ARQ error control schemes," *Electron. Lett.*, vol. 25, pp. 1195–1196, Nov. 1989.

[13] M. Zorzi and R. Rao, "Throughput of selective-repeat ARQ with time diversity in Markov channels with unreliable feedback," *Wireless Networks*, pp. 63–75, Apr. 1996.

[14] S. Kallel, "Analysis of memory and incremental redundancy ARQ schemes over a nonstationary channel," *IEEE Trans. Commun.*, vol. 40, pp. 1474–1480, Sept. 1992.

[15] E. N. Gilbert, "Capacity of burst noise channels," *Bell Syst. Tech. J.*, vol. 39, pp. 1253–1256, 1960.

[16] H. Wang and N. Moayeri, "Finite-state Markov channel—A useful model for radio communication channels," *IEEE Trans. Commun.*, vol. 44, pp. 163–171, Feb. 1995.

[17] M. Yacoub, *Foundations of Mobile Radio Engineering*. Boca Baton, FL: CRC, 1983.

[18] S. Lloyd, "Least square quantization in PCM," *IEEE Trans. Inform. Theory*, vol. IT-28, pp. 129–137, Mar. 1982.

[19] J. Proakis, *Digital Communications*, 2nd ed. New York: McGraw-Hill, 1989.

[20] A. Webster, C. Jones, M. Pinson, S. Voran, and S. Wolf, "An objective video quality assessment system based on human perception," in *Proc. Human Vision, Visual Processing and Digital Display TV*, vol. 1913, SPIE, San Jose, CA, Feb. 1993.

[21] H. Liu and M. El Zarki, "Transmission of video telephony images over wireless channels," *Wireless Networks*, vol. 2, no. 3, pp. 219–228, 1996.

Hang Liu received the B.S. degree from Tianjin University, China, in 1985, and the M.S. degree from the University of New Orleans, New Orleans, LA, in 1992.

He is currently a Ph.D. candidate in the Department of Electrical Engineering, University of Pennsylvania. His research interests include video compression and communications, wireless networking, ATM-based networks, and digital signal processing.

Magda El Zarki (S'81–M'88) received the B.E.E. degree from Cairo University, Egypt, in 1979, and the M.S. and Ph.D. degrees in electrical engineering from Columbia University, New York, NY, in 1981 and 1987, respectively.

She worked from 1981 to 1983 as a Communication Network Planner in the Department of International Telecommunications at Citibank, New York, NY. She joined Columbia University in 1983 as a Research Assistant in the Computer Communications Research Laboratory, where she was involved in the design and development of an integrated local area network testbed called MAGNET. In 1988, she joined the faculty of the Department of Electrical Engineering, University of Pennsylvania, teaching courses and conducting research in the field of telecommunications, where she currently serves as an Associate Professor. She also holds a secondary appointment in the Department of Computer and Information Sciences. In January 1993, she was appointed a part-time Professor of Telecommunication Networks in the Faculty of Electrical Engineering, Delft University of Technology, Delft, The Netherlands.

Dr. El Zarki is a member of the ACM and Sigma Xi. She is actively involved in many of their sponsored conferences and journals. She was the Technical Program Chair of IEEE INFOCOM'94.