Abstract—This paper deals with the problem of efficient transmission of video signals over generalized fading channels in direct sequence-spread-spectrum (DS-SS) code division multiple access (CDMA) systems. We first propose a modified version of the H.263 video codec incorporating a selective forward error correction (FEC) coding scheme combined with a forced intra-frame update mechanism. The modified codec results in the improvement of the average video and frame-to-frame performance. We further consider a coherent DS-CDMA system for the forward link (base-to-mobile) in both single-cell and multiple-cell environments. We provide performance evaluation results by both analysis, employing the Gaussian approximation, and computer simulations, using Monte Carlo error counting techniques. By integrating the proposed video codec with a coherent DS-CDMA system based upon the IS-95 standard, we investigate the performance of the video transmission over frequency-selective, correlated Nakagami fading forward-link channels employing a RAKE receiver. To simulate the fading channel, we have implemented in software a correlated Nakagami fading simulator based upon the principle of superposition of complex partial waves, an approach which replicates the wave propagation process in actual physical situation. A variety of performance evaluation results, both in single-cell and multiple-cell environment, are presented for a different number of resolving paths, cell user capacity, signal propagation characteristics, as well as for the presence of channel estimation errors. In all cases, heuristic explanations and interpretations of the trend of the obtained results are also given.

Index Terms—CDMA, H.263 encoder, IS-95, Nakagami fading, RAKE receiver, video transmission.

I. INTRODUCTION

In recent years, code division multiple access (CDMA) schemes have gained a lot of popularity in cellular telecommunication systems, especially through the development of the IS-95 standard [1]. At the same time, the demand for wireless multimedia services has grown rapidly. Among these services, wireless digital video transmission has probably the most profound impact on the development of future wireless telecommunication applications [2]. Although in the recent past there have been some papers published in the open technical literature investigating the end-to-end video transmission over wireless networks (e.g., [4]–[6], [17]), only a few of them considered CDMA systems [4], [17]. In these papers, the mobile channel models they employed assume either Rayleigh or Rician fading characteristics. However, when compared to Rayleigh and Rician distributions, the Nakagami-η distribution [14] provides a more general and versatile way to model wireless channels [9]. Furthermore, it fits more accurately to experimental data for many physical propagation channels [7]. Among the above-cited papers, none of them has investigated video transmission performance in multiple-cell environments.

Motivated by the above, the objective of this paper is to investigate the video transmission performance over cellular CDMA systems in Nakagami fading channels. In particular, we propose a new low bit-rate H.263 [3] video codec for improved transmission performance in error-prone mobile environment. A correlated Nakagami fading simulator is implemented in software for more general and accurate fading channel modeling. We also perform BER performance analysis employing the Gaussian approximation and computer simulation for the direct-sequence spread-spectrum (DS-SS) CDMA forward link in both the single- and multiple-cell environment. Further, we integrate the proposed video codec with a IS-95 based CDMA system and evaluate the end-to-end video transmission performance of the forward link over correlated Nakagami fading channels, using computer simulation methods for both single- and multiple-cell environments under a wide variety of system and channel conditions.

The following is the organization of this paper. In Section II, a detailed description of the proposed video codec structure as well as evaluation results in additive white Gaussian noise (AWGN) and single-cell CDMA channels are presented. This is followed by Section III with the description of the CDMA forward link system and the mobile channel model. The system BER performance analysis is carried out in Section IV. In Section V, we present a variety of performance evaluation results in terms of peak signal-to-noise ratio (PSNR) for the proposed H.263 codec integrated with a IS-95 based CDMA system under various correlated Nakagami fading channel environments. Finally, in Section VI, the conclusions of the paper are given.

II. A MODIFIED H.263 VIDEO CODEC

Generally, the severity of bit errors to H.263 video quality depends on the spatial and temporal location of the errors. Huffman coding causes spatial error propagation problem, whereas motion compensation causes temporal error propagation problem. Based on some recently reported results [6], [21], [22], some important effects of the transmission errors on H.263 video can be summarized as follows. 1) Errors in
video headers can cause major damage, especially for headers in higher hierarchical layers of H.263 coding such as the Picture Layer and the group of blocks (GOB) Layer. 2) Errors propagate in the spatial domain due to improper decoding of variable length codes; Picture and GOB layer headers stop error propagation in spatial domain by providing start codes. 3) Errors propagate among Prediction frames in the temporal domain; intra-frames stop the propagation by coding the picture independently. To maximize video quality, while avoiding time delay for real-time applications and minimizing channel coding redundancy and complexity, we propose here a selective FEC coding scheme combined with an intra-frame forced update mechanism.

Since the IS-95 CDMA system uses half-rate convolutional coding on all transmitted data, it provides the base layer FEC protection. Considering the impact of errors on the H.263 video bitstream, and in order to minimize the redundancy introduced by the coding, we have chosen to provide two extra FEC protection codes, termed as FEC1 and FEC2, to selected bits of the upper two hierarchical layers of the H.263 bitstream syntax, namely, the Picture and the GOB Layers [19]. Assuming that a Quarter Common Intermediate Format (QCIF) sequence is transmitted, FEC1 provides error protection for the important header bits in Picture Layer of both the intra- and Prediction frame, which include PSC, GN, TR, PTYPE, PQUANT, CPM, PEI [3]. These header bits add up to a total of 50 bits, and a (71, 50, 3) BCH codeword is assigned. The 21 bits of redundancy which are introduced can correct up to three errors [10]. FEC2 provides error protection for header bits in the GOB Layer which include GBSC, GN, GID, GQUANT [3]. These 29 bits are protected by a (41, 29, 2) BCH codeword, and the 12 bits of redundancy introduced can correct up to two errors [10]. For the Macroblock and Block Layers, because of the lack of start code for synchronization and of the variable nature of the sequence lengths, the positions of their bits are unknown. Thus, the effort to apply selective FEC protection to these lower two hierarchical layers may not be worthwhile because it unnecessarily introduces extra and equally error-sensitive information bits for location and length indication during the encoding process.

The major compression achieved by low bit-rate encoders, such as the H.263, is mostly due to the removal of temporal redundancy through motion compensation. Therefore, it is necessary for the encoder to emulate the decoder at the transmitter. If the information available to this encoder and the decoder used at the receiver is not the same, the quality of the reconstructed picture-frame can degrade considerably. Moreover, temporal mitigation of these reconstruction errors can affect the quality of the subsequently reconstructed picture-frames. If the reconstructed signal is degraded due to some channel errors, subsequent reconstruction of error-free transmissions may also be incorrect. Therefore, it is apparent that in order to ensure high-quality transmission, the effects of errors must not propagate too far beyond the temporal interval of the channel errors. To solve this problem, we apply periodically forced updates of the Prediction frames by intra-frames. Considering the tradeoff between bit usage and overall video quality, we have chosen to apply one intra-frame update in every 10 Prediction frames.

### III. System Model Description

Fig. 1 illustrates the block diagram of the direct-sequence CDMA video transceiver system model considered in this paper. QCIF format video is first compressed by the modified H.263 coder. As per the IS-95 standard, the encoded video data are half-rate convolutionally encoded and block interleaved. The symbols are then spread by a concatenated Walsh/PN sequence. Each symbol is first replaced by 1 of the 63 orthogonal Walsh codes of length 64 chips, which is uniquely assigned to each user. Then, modulo-2 addition (concatenation) is performed on each chip with a PN sequence of period $2^{15} - 1$ chips. The resulting concatenated codes retain orthogonality among users, while reducing cross-correlation surges among nonconcatenated Walsh codes. It is assumed that the forward link signals are chip-synchronized, i.e., the signal delay $\tau = kT_c$, for some integer $k$ with $T_c$ denoting the chip duration, so that the multiple-user signals can take advantage

1Note that the Walsh 0 code is reserved for power control.
of the orthogonality offered by the Walsh sequences. For a synchronous multiple-cell CDMA system, the same PN sequence is shared by all base station (BS). Each BS, however, uses a different time-shifted version of this common PN sequence for Walsh code concatenation [20]. This PN code time shift is used by the mobile receivers to distinguish each BS from the others. At the output of the CDMA transmitter, the data are modulated by a Binary Phase Shift Keying (BPSK) signal before being transmitted into the channel.

It is assumed that the mobile CDMA channel is a frequency-selective fading channel with fading statistics following the Nakagami-$m$ distribution. As such, the channel can then be modeled by a tapped delay line illustrated in Fig. 2 [10]. The complex-valued tap weights $\{c_n(t); n = 0, 1, \cdots, L - 1\}$ are statistically independent stationary random processes with magnitudes $|c_n(t)| \equiv \beta_n(t)$ following the Nakagami-$m$ probability density function (PDF) [14]:

$$p(r) = \frac{2m^m \Gamma(m) \beta^{2m-1}}{\Gamma(m) \Omega^m} e^{-\left(m/\Omega\right) \beta^2}$$

where

$\Gamma(m)$ is the Gamma function,

$m$ is the fading figure, and

$\Omega$ is the second moment of $r$.

For $m = 1$, the Nakagami distribution is identical to the Rayleigh distribution. In the presence of a direct signal component, the Nakagami distribution approximates the Rician distribution with $m = \{1 - (\kappa)/(1 + \kappa)\}^{-2}$ for $m > 1$, where $\kappa$ is the Rician $\kappa$-factor [15]. Similarly to the Rician fading channel, the higher the value of $m$ becomes, the less severe is the fading. The phases $\angle c_n(t) \equiv \theta_n(t)$ are assumed to be uniformly distributed over $[0, 2\pi)$ and independent of $\{\theta_n(t)\}$ [8], [11]. The time delay blocks of $1/W$ in Fig. 2 represent the resolution of the multipath delay profile where $W$ is the bandwidth occupied by the transmitted signal $s(t)$. $n(t)$ is the complex-valued zero-mean AWGN process with two-sided power spectral density of $\eta_0/2$. The number of multipaths $L$, which may be a random number, is bounded by $\lfloor T_m/T_c \rfloor + 1$, where $\lfloor x \rfloor$ is the floor function and $T_m$ is the maximum multipath spread of the channel. $T_m$ is assumed to be less than the bit interval $T$ so that intersymbol interference (ISI) may be neglected. In addition to the fading and AWGN, the transmitted signal is also corrupted by self-noise interference (SI) and multiple-access interference (MAI). SI originates from the sidelobes of the autocorrelation function of the reference user’s spread code. MAI is caused by the cross-correlation of the spread codes between the reference user and other users in a multipath signal environment.

At the receiver end, to take advantage of the wideband characteristics of spread spectrum signals, a coherent maximal-ratio combing RAKE receiver is employed in order to provide multipath diversity. It is assumed that the BS continuously transmits a pilot signal which is used by the mobile receiver to acquire synchronization as well as to estimate the mobile channel impulse response. Fig. 3 illustrates the RAKE receiver structure for the reference user where the number of fingers, $L_p$, is a variable parameter which takes values less than or equal to the total number of multipaths $L$. The matched filter is matched to the reference user’s CDMA spreading code and is assumed to have achieved time synchronization with the initial path of the reference signal. To simplify the mathematical analysis, we will be assuming that the tap weights $\{\beta_i\}$ and phases $\{\theta_i\}$, $i = 0, \cdots, L_p - 1$, to be the perfect estimates of the channel parameters. However, for computer simulation, perfect knowledge of channel amplitude is not required. The sampling times of the receiver are $t_n = nT + (L_r - 1)T_c$, where $n$ is an integer index. The sampled output $U_n$ is then used in the decision device for detection. As illustrated in Fig. 1, the detected symbols from the RAKE receiver output are deinterleaved and fed into the hard-decision Viterbi decoder to recover the transmitted video data bits. Finally, the modified H.263 decoder is used to reconstruct the video sequences.

As shown in Fig. 4, we have considered a multiple-cell cellular system which takes into account the MAI originated from the 18 surrounding cells of the first two tiers, with number 0 denoting the reference cell, and 1 to 18 denoting the surrounding cells. To simplify the mathematical analysis, all cells are assumed to be circular with equal radius, and each BS is located at the center of all cells. Also, the reference mobile user is assumed to be equally likely located anywhere within the reference cell. The employed large-scale fading model assumes that the average signal attenuation is the product of the $\gamma$th power of distance with a log-normal random variable. Thus, the transmitted signal average power $P_t$ and the received signal average...
power $P_q$ from the $q$th BS to the reference mobile user is given by

$$P_q \propto P_q d_q^{-\gamma}10^{\gamma/10} \tag{2}$$

where
- $d_q$ is the distance between the reference mobile and the $q$th BS,
- $\gamma$ is the path loss exponent, and
- $\zeta_q$ is a Gaussian random variable with zero mean and standard deviation $\sigma_q$.

As suggested in [24], we will use $\gamma = 4$ for the power law and $\sigma_q = 8$ dB for the standard deviation of the log-normal shadowing random variable $\zeta_q$.

IV. FORWARD LINK BER PERFORMANCE ANALYSIS

A. Single-Cell CDMA System

In a single-cell environment for the CDMA forward link, the transmitted signal of the $k$th user is a phase-modulated carrier expressed as

$$s^{(k)}(t) = \sqrt{2P_d^{(k)}(t)}b^{(k)}(t)\cos(\omega t + \phi) \tag{3}$$
where \( a^{(k)}(t) \) is the spreading code of the \( k \)th user
\[
a^{(k)}(t) = \sum_{j=-\infty}^{\infty} a^{(k)}_j \cdot p_R(t-jT_c), \quad a^{(k)}_j \in \{-1, 1\}
\]
and \( b^{(k)}(t) \) is the data waveform given by
\[
b^{(k)}(t) = \sum_{j=-\infty}^{\infty} b_j^{(k)} \cdot p_D(t-jT_d), \quad b_j^{(k)} \in \{-1, 1\}.
\]
In (3), \( P = E_b/T \) is the transmitted average power, assumed to be equal for each user, \( E_b \) is the bit energy, \( \omega_c \) is the carrier radian frequency, and \( \phi_c \) is the initial phase of the modulator uniformly distributed over \([0, 2\pi)\). \( p_R(t) \) and \( p_D(t) \) are rectangular pulses of unit height and duration of \( T_c \) and \( T_d \), respectively. For a single cell multiple-user model, the received signal after the channel is [11]
\[
r(t) = \sqrt{2P} \cdot \sum_{k=0}^{K-1} \sum_{l=0}^{L-1} \alpha^{(k)}_l \cdot \cos(\omega_c t + \varphi_l) + n(t)
\]
\[
\cdot b^{(k)}(t) \cdot \cos(\omega_c t + \varphi_l) + n(t)
\]
\[
\cdot \exp(j2\pi \cdot \tau_l \cdot \cos(\omega_c t + \varphi_l)) + n(t)
\]
where \( K \) is the total number of users in the cell, and \( L \) is the number of multipaths. \( \varphi_l = \varphi_c + \theta_l - \omega_c \cdot \tau_l \) is the phase of the \( l \)th path, where \( \theta_l \) is the channel phase shift and \( \tau_l \) is the multipath time delay for the \( l \)th path. As previously explained, each path is assumed to fade independently with fading coefficient \( \beta_l \), of which the amplitude \( \beta_l \) follows the Nakagami-\( m \) distribution and the phase \( \theta_l \) follows the uniform distribution.

Each signal path which matches with the \( n \)th RAKE finger gives a desired signal component \( S^{(n)} \). In addition, at the \( n \)th finger of the RAKE receiver, there are also MAI, SI, and the noise interference terms denoted as \( I^{(n)}_{\text{ma}}, I^{(n)}_{\text{si}}, \) and \( I^{(n)}_{\text{ni}}, \) respectively. In [11], the coherent RAKE receiver output for an asynchronous (i.e., chip-misaligned) CDMA system is given. To adapt the receiver output for the synchronous forward link CDMA, we replace the independent fading amplitudes and signal phases for each user’s multipath signal by the common values which are shared by intracell users. Thus, the sampled output of the coherent RAKE receiver with \( L_r \) fingers can be expressed as (see Fig. 3)
\[
U_n = \sum_{n=0}^{L-1} \left[ S^{(n)} + I^{(n)}_{\text{ma}} + I^{(n)}_{\text{si}} + I^{(n)}_{\text{ni}} \right]
\]
\[
\text{where the four signal components can be mathematically expressed as [19]}
\]
\[
S^{(n)} = \sqrt{\frac{P}{2}} \cdot b^{(0)} \cdot T \cdot \{ \beta_n \}^2
\]
\[
I^{(n)}_{\text{ma}} = \sqrt{\frac{P}{2}} \cdot \sum_{k=1}^{K-1} \sum_{l=0}^{L-1} \beta_k \cdot \{ \alpha^{(k)}_l \cdot R_k^{(l)} \cdot \cos(\varphi_{nl}) \}
\]
\[
I^{(n)}_{\text{si}} = \sqrt{\frac{P}{2}} \cdot \sum_{l=0}^{L-1} \beta_l \cdot \{ \alpha^{(0)}_l \cdot R_0^{(l)} \cdot \cos(\varphi_{n0}) \}
\]
\[
I^{(n)}_{\text{ni}} = \int_{nT_c}^{n+1T_c} n(t) \cdot \beta_n \cdot \alpha^{(0)}(t-nT_c) \cos(\omega_c t + \varphi_n) dt.
\]
In the above equations, \( b^{(0)} \) is the information bit (to be detected) of the reference users, whereas \( b^{(0)} \) is the preceding bit. \( \tau_{nl} = \tau_l - \tau_n \) is the time delay difference and \( \varphi_{nl} = \varphi_l - \varphi_n \) is the phase difference between the \( l \)th and the \( n \)th multipath, respectively. \( R_k^{(l)} \) and \( R_0^{(l)} \) are the continuous-time partial cross-correlation functions between the \( k \)th and the reference user as defined in [13].

Similar to [8], in order to simplify the BER performance analysis of the proposed system, we apply the Gaussian approximation on the MAI and SI terms. To obtain the variance of MAI conditioned on \( \beta_n \), we compensate for the chip-synchronization with a 3/2 factor (see, for example [8] and [11]) to the MAI variance of an asynchronous system in [11] so that
\[
\sigma^2_{\text{MAI}_n} = \frac{P \cdot \delta^2}{4N} \cdot \{ \beta_n \}^2 \cdot \sum_{l=1}^{K-1} \sum_{n=0}^{L-1} \Omega_l
\]
where \( \Omega_l = E[|\beta_l|^2] = \Omega_0 e^{-\delta \cdot l}, \delta \geq 0 \) is the average power for the signal with a logarithmic multipath intensity profile (MIP) of average power decay rate \( \delta \) [16]. SI may be considered as an additional multiple-access user where, instead of \( L \) paths, there would only be \( L - 1 \) paths at the input to the receiver because one path contributes to the desired signal component \( U_{n,s}^2 \) [8]. Thus, the conditional variance of \( I^{(n)}_{\text{si}} \) can be approximated as
\[
\sigma^2_{\text{SI}_n} = \frac{P \cdot \delta^2}{4N} \cdot \{ \beta_n \}^2 \cdot \sum_{l=1}^{K-1} \sum_{n=0}^{L-1} \Omega_l
\]
\[
\text{The variance of the AWGN term conditioned on } \beta_n \text{ is [11]}
\]
\[
\sigma^2_{\text{AWG}_n} = \frac{T \cdot \delta^2}{4} \cdot \{ \beta_n \}^2
\]
\[
\text{Therefore, the sampled output of the RAKE receiver } U_n^s \text{ can now be modeled as a conditional complex Gaussian random variable with the conditional mean of the desired signal component given by}
\]
\[
U_{n,s}^s = \sqrt{\frac{E_b T \cdot \delta^2}{4N}} \cdot \sum_{n=0}^{L-1} \{ \beta_n \}^2
\]
\[
\text{and the conditional variance } \sigma^2_{S} \text{ equal to the sum of the variances of all three interferences. Using (12)–(14), and after some straightforward mathematical manipulations, } \sigma^2_{S} \text{ becomes [19]}
\]
\[
\sigma^2_{S} = \sum_{n=0}^{L-1} \left[ \sigma^2_{\text{MAI}_n} + \sigma^2_{\text{SI}_n} + \sigma^2_{\text{AWG}_n} \right]
\]
\[
= \left( \frac{E_b T \delta^2}{4N} \right) \cdot \left( \frac{(K-1)(L, \delta) - q(L, \delta) - 1}{N} + \frac{\eta_0}{E_b \delta \lambda_0} \right)
\]
\[
\cdot \sum_{n=0}^{L-1} \{ \beta_n \}^2
\]
\[
\text{where } q(L, \delta) = \sum_{l=0}^{L-1} e^{-\delta l}. \text{ The received signal-to-noise ratio at the output of the RAKE receiver is } \text{SNR} = U_{n,s}^2 / \sigma^2_{S}. \text{ By denoting the random component of the SNR as}
\]
\[
S = \frac{1}{\Omega_0} \cdot \sum_{n=0}^{L-1} \{ \beta_n \}^2
\]
and the deterministic component as

\[ \gamma = \left( \frac{1}{2} \cdot \left( \frac{q(L_s \delta) - 1}{N} + \frac{\gamma_0}{E_0 L_s} \right) \right)^{-1} \]  

(18)

the received SNR can be written in compact form as \( \gamma S \). Furthermore, the conditional bit error probability can be expressed as

\[ P(S) = Q(\sqrt{\gamma} S) \]  

(19)

where \( Q(\cdot) \) is the well-known \( Q \)-function [10]. The average error probability \( P_e \) is obtained by averaging \( P(S) \) over the PDF of \( S \). Using the method suggested in [8], we have computed \( P_e \) under different channel conditions [19]. Some of the obtained numerical results of \( P_e \) will be compared to equivalent computer simulation results in Section V.

B. Multiple-Cell CDMA System

For multiple-cell systems, there is an additional amount of MAI contributed by the extra \( K_q \) number of users from the \( Q = 18 \) surrounding cells. Since these signals come from different cells, they have different attenuation and MIP. To account for all these factors, the single-cell MAI term \( \sigma^2_{\text{MAL},n} \) in (12) is modified by adding a second term which represents the combined MAI from cell 1 to cell \( Q = 1 \), and replacing the common power \( P \) of the reference cell by the attenuated signal power \( P_q \) for the \( q \)th cell as given by (2). In addition, the single-cell MIP term \( \Omega_q \) is replaced by a multiple-cell MIP term \( \Omega_{qf} \) denoting the \( q \)th-channel cell which is defined as

\[ \Omega_{qf} = \Omega_q e^{-\delta_q}, \quad \delta_q \geq 0 \]  

(20)

where \( \Omega_q \) is the first path average signal strength, and \( \delta_q \) is the rate of average power attenuation for the \( q \)th cell signal propagation, respectively. Thus, the resulting conditional MAI variance at the \( q \)th RAKE finger for a multiple-cell system can be mathematically expressed as

\[
\sigma^2_{\text{MAL},n} = \frac{\hat{P}_0 T^2}{4N} \left\{ \frac{\Omega_{qf}^2}{2} \cdot \sum_{k=1}^{K_q-1} \sum_{l=0}^{L_q-1} \Omega_q \right\} \]

\[
+ \sum_{q=1}^{Q-1} \frac{\hat{P}_q T^2}{4N} \left\{ \frac{\Omega_{qf}^2}{2} \cdot \sum_{k=0}^{K_q-1} \sum_{l=0}^{L_q-1} \Omega_q \right\} \]

(21)

where the first term represents MAI from \textit{within} the reference cell, and the second term represents MAI from the \( Q \) surrounding cells. Clearly, in the above equation, in order to account for the additional MAI interfering users from each of the surrounding cells, the summation sign of users changes from \( k = 1, \ldots, K_0 \) for the reference cell user MAI term to \( k = 0, \ldots, K_q - 1 \) for the surrounding cell MAI term. The SI and AWGN terms for the multiple-cell systems \( \sigma^2_{\text{SI},n} \) and \( \sigma^2_{\text{AWGN},n} \) are obtained in a similar manner to give

\[
\sigma^2_{\text{SI},n} = \frac{\hat{P}_0 T^2}{4N} \left\{ \frac{\Omega_{qf}^2}{2} \cdot \sum_{l=1}^{L_q-1} \Omega_q \right\} \]

\[
\sigma^2_{\text{AWGN},n} = \frac{T_{\text{AWGN}}}{4} \cdot \left\{ \frac{\Omega_{qf}^2}{2} \right\} \]

(22)

(23)

After some straightforward mathematical simplifications, the SNR \( \hat{U}_s(\sigma^2_{\text{MAL},n} + \sigma^2_{\text{SI},n} + \sigma^2_{\text{AWGN},n}) \) for the multiple-cell CDMA system forward link can be mathematically expressed as shown in (24) at the bottom of the page [19]. For an interference-limited CDMA system, the AWGN term \( \sigma^2_{\text{AWGN},n} \) is relatively small and thus, in order to simplify the mathematical analysis, it could be neglected. As illustrated in Fig. 5, since we are only interested in the relative received power from the \( q \)th BS to that of the reference BS, we introduce the attenuation factor of the signal power from the \( q \)th BS relative to that of the reference BS \( (q = 0) \) as

\[ \rho_q = \frac{10^{\eta_q/10}}{10^{\eta_0/10}} \]

(25)

In this figure, cell \( q \) represents any of the 18 surrounding cells in the first two tiers, \( R \) is the cell radius, and \( \phi \) is the mobile location angle with respect to the reference BS. In addition, assuming identical MIP and uniform average number of users among the different BS, i.e., \( \Omega_q = \Omega_q = \Omega_q \) and \( K_q = K_0 = K_q \), (24) can be rewritten as

\[
\text{SNR} = \frac{1}{2N} \left( \sum_{k=1}^{K_q} \sum_{l=0}^{L_q} \Omega_q + \sum_{q=1}^{Q-1} \sum_{k=0}^{K_q} \sum_{l=0}^{L_q} \Omega_q \right) \]

(26)

\[
= \frac{1}{2N} \left( \sum_{n=0}^{L_q-1} \left\{ \frac{\hat{P}_q}{N} \cdot \left\{ \frac{\Omega_{qf}^2}{2} \cdot \sum_{k=0}^{K_q} \sum_{l=0}^{L_q} \Omega_q \right\} \right\} \right) \]

\[
= 2\hat{P}_0 \cdot \left\{ \sum_{n=0}^{L_q-1} \left\{ \frac{\hat{P}_q}{N} \cdot \left\{ \frac{\Omega_{qf}^2}{2} \cdot \sum_{k=0}^{K_q} \sum_{l=0}^{L_q} \Omega_q \right\} \right\} \right\} \]

(24)
Fig. 6. Video performance of modified H.263 codec in AWGN channel.

Similar to the single-cell case, the SNR at the output of the receiver may be rewritten in a more compact form as $ \Upsilon_m S_m $, where

$$
\Upsilon_m = 2N \left( (K - 1)q(L, \delta) + \sum_{q=1}^{Q-1} \rho_q K \cdot q(L, \delta) \right.
$$

$$
+ \left( q(L, \delta) - 1 \right) \right)
$$

To obtain the area-averaged BER, mean values of the attenuation factor $ \rho_q $ in (27) for different reference mobile locations evaluated by Fong et al. [25] were used. Therefore, $ \overline{P_e} $ can be obtained by replacing in (19) $ \Upsilon $ and $ S $ by $ \Upsilon_m $ and $ S_m $, respectively. As with the single-cell case, numerical results from the analysis as well as comparisons to those obtained from the Monte Carlo simulation will be presented in Section V.

V. PERFORMANCE EVALUATION RESULTS

A. Modified H.263 Codec Performance

We have considered the transmission of 100 QCIF video frames of Miss America at a nominal rate of 64 kbits/s with FEC overhead of 1.1%. In order to provide a better evaluation for the full-color video quality, we have proposed a weighted luminance and chrominance average PSNR values such that $ \text{PSNR} = (\text{PSNR})_L + 0.3(\text{PSNR})_{C_b} + 0.3(\text{PSNR})_{C_r} $, with (PSNR)$_L$, (PSNR)$_{C_b}$, and (PSNR)$_{C_r}$ corresponding to the various PSNR values of the luminance and the two chrominance components, respectively [19]. The proposed method of PSNR calculation takes into account both brightness and colors of the video frames, thus providing a more objective measure for the full-color video quality. In our results, each data point is generated by taking the ensemble average of 20 simulation runs, while the PSNR value of each run is obtained by taking the mean of the PSNR values of all received frames. No special treatment or assumptions are associated with the start frame or the intra-frame. As illustrated in Fig. 6, the modified H.263 video codec, as compared to the original H.263 codec, offers significant PSNR improvements for transmission over AWGN channels with gains ranging from 3 to 10 dB, depending upon the values of bit error probability $ P_e $. As can be observed from the same figure, the performance gains decrease for lower values of $ P_e $. This happens because the FEC redundancy is greater than the gain achieved from the protected bits being in error, and thus the average video quality deteriorates. For high values of $ P_e $, the performance gains again decrease due to the severe error conditions, i.e., the video performance degrades to a point where the FEC can provide very little help to recover the damage done to the received video bits.

Fig. 7 shows the frame performance of both coding schemes (at $ P_e = 5 \times 10^{-3} $) as compared to the error-free condition. Again, it can be observed that the modified coding scheme offers significantly better performance in most of the frames, thus resulting in better overall PSNR. The step-like shape of the modified H.263 codec curve (solid line) is because of the forced intra-frame refresh mechanism which occurs every ten frames.

B. CDMA Forward Link Performance

In this subsection, we present the numerical results of the BER performance for the CDMA forward link system without the video codec. The fading channel errors are generated by a
software Nakagami simulator which is described with details in [19]. In Figs. 8 and 9, we compare the BER performance of the single-cell and multiple-cell CDMA forward link obtained by the analytical methods presented in Section IV and Monte Carlo computer simulations, respectively. For both cases, random spreading sequences for each user are being employed with processing gain $N = 64$, $E_b/\sigma_0 = 30$ dB, $L = 5$, and uniform MIP. In Fig. 8, we consider a RAKE receiver with $L_r = 2$ and 3 in a Nakagami channel of $m = 1$ (i.e., Rayleigh fading), whereas in Fig. 9, we consider a RAKE receiver with $L_r = 3$ and 5 in a Nakagami fading channel with $m = 2$. Comparing the performance results, we observe that the discrepancies between the BER performance obtained from the Gaussian approximation analysis and computer simulation are larger for the case of single-cell system than those for the multiple-cell system. This can be explained by the fact that the Gaussian approximation is based upon the central limit theorem which requires statistical independency among random variables [10]. For the single-cell
case, since the forward link interference signals from intra-cell users arrive at the reference mobile with the same channel characteristics (i.e., same amplitude fading, phase distortions, and transmission time delays), the MAI signals lack the required statistical independency. On the other hand, for the multiple-cell system, since the MAI signals from different cells are independent from each other, they satisfy more accurately the central limit theorem. Consequently, as observed, more accurate BER performance predictions are obtained from the Gaussian approximation for the multiple-cell system.

C. Single-Cell CDMA Video Transceiver Performance

We next present the PSNR performance of the modified H.263 coded video transmitted through the proposed CDMA system described in Section III. If not otherwise stated, the following parameter values have been used for all the results presented here: $L = 3$, $L_r = 2$, $m = 1$, $\delta = 0$, and mobile velocity $v = 100$ km/h. In Fig. 10, we compare the PSNR performance of the modified video codec to its original version operating in a single-cell CDMA forward link system over Nakagami fading channels. As in the case for the AWGN
channel, we also observe here improved video transmission performance achieved by the modified video codec over the original H.263. The PSNR improvement in this case ranges from 1 to 4 dB. Alternatively, at a PSNR = 20 dB, the system capacity is increased by five users.

In Fig. 11, the video transmission performance is evaluated against different \( L_r \) values. We can see that the PSNR performance improves as \( L_r \) increases from 1 to 3. This coincides with the theoretical optimal value of \( L_r = L \) for coherent demodulation because a coherent \( L_r \)-finger RAKE with perfect estimates of the channel tap weights is equivalent to a maximal-ratio combiner with \( L_r \)-th-order diversity [10]. For the case of \( L_r = 1 \), the recovered video data are so severely corrupted that both FEC provided by the proposed modified video codec and convolutional coder are unable to provide adequate error recovery, thus resulting in relatively poor PSNR performance. However, for \( L_r = 2 \), the quality of the video is rather acceptable at PSNR = 40 dB for up to about 25 users, and for \( L_r = L = 3 \), there is virtually no degradation of the video sequence quality for up to 60 users.

Fig. 12 illustrates the PSNR performance of the modified H.263 video transmission under different channel conditions as
a function of the fading figure $m$ (with $m = 1, 2, \text{and } 3$). As the value of $m$ increases, there is significant improvement in the video quality in terms of PSNR because of the diminishing severity of the fading. The gaps in PSNR among the different $m$-value cases become smaller when the number of users $K$ gets larger. This happens because the increasing MAI has become the dominant factor in the degradation of the video transmission over the effects of fading of the channels.

Fig. 13 illustrates the effects of different MIP of the signal on the PSNR performance where we consider $\delta = 0, 0.2, 0.4$. The decreasing power in the multipath signals has two major effects when $L_r < L$. First, there is less interference arriving at the receiver. This is especially true for the SI due to the decreased power of delayed versions of the reference user signal. At the same time, the sum of desired signal power decreases as well. As shown in the graphs, in this case, the average PSNR performance improves for increased values of $\delta$. This happens because the reduced MAI and SI have more impact on the system performance than the reduced desired signal power does.

In Fig. 14, we present results showing the impact of Doppler spread on the system performance. We have considered cases of $B_d T = 2.8 \times 10^{-3}$, $1.4 \times 10^{-3}$, and $1.6 \times 10^{-3}$ which
correspond to typical mobile speeds of 100, 50, and 5 km/h, respectively. The term $B_d$ denotes the maximum Doppler spread. It is well known that smaller $B_d T$ products result in longer burst error due to their longer fade duration, and vice versa [26]. Thus, for a fixed degree of interleaving, channels with larger $B_d T$ values produce shorter error bursts, which in turn facilitate error correction by the FEC channel coding. This is the main reason why, in Fig. 14, the higher $B_d T$ products yield better average PSNR performance than the lower $B_d T$ products.

We have also investigated the impact of imperfect estimation of the channel tap weights on the video transmission quality. Channel tap weight estimations are used at the RAKE receiver to achieve maximal-ratio combining. In practical situations when fading is relatively fast, there could be inaccuracies in the estimation process. In Fig. 15, we simulate such estimation errors by deviating the exact channel amplitudes by Gaussian variance of different values. As seen from the results, imperfect channel tap weight estimations could degrade the video transmission performance of a CDMA system significantly.

D. Multiple-Cell CDMA Video Transceiver Performance

In this subsection, we present the numerical results obtained from computer simulations of the modified H.263 video transmission performance over the proposed multiple-cell CDMA forward link operating in Nakagami fading channel environments. As described in Section III, we have considered a multiple-cell system consisting of the first two tier cells, and we present the PSNR performance results of the video transceiver system with different channel characteristics and cell user load conditions. Fig. 16 illustrates the average PSNR performance of video transmission in Nakagami fading channels with nonidentical $m$ values for each multipath. The initial multipath signal suffers less severe fading than the subsequent ones because it has a lower probability of being scattered. As examples, we consider three multipath fading $m$ values in the order of signal arrivals namely, $\{1.0, 1.0, 1.0\}$, $\{2.0, 1.5, 1.0\}$, and $\{3.0, 2.0, 1.0\}$. As expected, the last case gives the best PSNR performance because it has the lowest average fading severity.

We have also evaluated the video performance for multiple-cell systems where multipath signals transmitted from BS of different tiers go through mobile fading channels modeled with nonidentical set of $m$ values. As illustrated in Fig. 17, for each case, the set of $m$ values in the first, second, and third bracket $\{\}$ correspond to the fading figures of the reference-cell channel, the 6 first-tier surrounding-cell channels, and the 12 second-tier surrounding-cell channels, respectively. The $m$ values for multipath signals in each bracket $\{\}$ are presented in the order of arrivals. In the first case, the simulation resembles the physical situation in which the reference mobile is most likely closer to the reference BS than the other BS. This results in a higher probability for the existence of direct line-of-sight signals, and thus justifies the relatively less severe channel fading for the reference-cell signal. From the same figure, we observe that the video performance of the first case is better than that of the second case where the reference cell has higher $m$ values or less fading. Next, for the second and third case, we assume the same fading figure set for the reference-cell channels but different fading figure set for the surrounding-cell channels. We observe that in the third case, where the surrounding-cell channels exhibit more severe fading, better PSNR performance is obtained. This is because the greater severity of fading in the surrounding-cell channels reduces the amount of MAI from reaching the reference mobile.

In Fig. 18, the effects of different user load of the surrounding cells to the video transmission performance are investigated. We have evaluated the PSNR values of systems with all the 18 surrounding cells being occupied by $1/2$, $1/3$, and $1/4$ of
the reference-cell user population. As shown in this figure, the PSNR performance improves as the surrounding-cell user load decreases. The gain in PSNR performance is due to the decrease in the amount of MAI generated from the neighboring-cell users. The results obtained clearly indicate that the user load of surrounding cells could have a significant impact on the video transmission performance in CDMA systems.

In Fig. 19, we investigate the CDMA video transmission performance with nonidentical MIP for channels of cells in different tiers. We have considered $\delta = 0.2$, $0.4$, and $0.8$, where the first value corresponds to the logarithmic MIP exponent of the reference-cell channel, and the second value corresponds to that of the 18 surrounding-cell channels. As shown in this figure, the last case provides the best PSNR performance. This is because greater logarithmic MIP exponent for surrounding-cell signals means these interfering multipath signals degrade faster, which translates to less MAI at the reference mobile.

Finally, we have evaluated the video transmission performance of the CDMA system with nonidentical number of
multipaths for cells at different tiers to resemble the variant nature of the physical environment. Typical performance evaluation results are presented in Fig. 20, where \( L_0, L_1, \) and \( L_2 \) correspond to the multipath number of the reference cell, first-tier cells, and second-tier cells, respectively. We observe a significant amount of PSNR degradation as the multipath number of the neighboring-cells increases, which is due to the corresponding increase of MAI from the extra multipath signals under consideration.

VI. CONCLUSION

We have proposed a modified version of the H.263 codec which provides improved transmission performance in both AWGN and CDMA channels. The BER performance of CDMA forward link was evaluated by analytical method and computer simulation. It was found that the Gaussian approximation provides a more accurate BER prediction for multiple-cell systems than for single-cell systems. We have also evaluated...
the proposed modified H.263 video codec in an IS-95 based CDMA forward link over correlated Nakagami fading channels for both single- and multiple-cell environments. It was shown that the optimal number for coherent RAKE fingers equals to the number of multipaths, i.e., it functions as a maximal-ratio combiner. In general, the PSNR performance improves as the Nakagami fading figure values of the reference-cell channel increase due to the corresponding reduction in fading severity. The system also performs better for larger logarithmic MIP exponents due to the net gain in interference reduction versus desired signal power reduction. We have also observed that the sensitivity of the video transmission performance on the channel estimation error could be significant. For multiple-cell systems, we have found that better PSNR performance is achieved when the neighboring cells have more severe fading channel conditions, lower user capacity, and greater logarithmic MIP exponent. Finally, we have illustrated that increasing the number of multipaths in surrounding-cell channels can degrade the video transmission performance in a CDMA system considerably.

REFERENCES


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