Improved Channel Modeling for Performance Analysis of a DS-CDMA Link with Exponential Delay Profile

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Abstract: This paper developes a refined model for direct sequence (DS) spread spectrum transmission over a Rician fading channel. The path gains in the channel are modeled with an exponential delay profile rather than the uniform profile prevalent in other papers. Additionally, some components of the multipath are included in the desired signal. An expression for the local-mean bit error rate (BER) is derived from this model, which is extended to an analysis of the packet erasure rate (PER).

I. INTRODUCTION

The performance of wireless transmission in a multiuser indoor environment is fundamentally limited by interference due to the presence of signals from other users as well as multipath. Such an environment has been envisioned by many in the personal communications industry, who have proposed picocellular indoor communications networks which deliver integrated services to users via wireless CDMA transmission. A high speed network would be used to perform database retrieval and number crunching, delivering results to any number of portable personal terminals over a full-duplex wireless link. It is the popularity of such proposals which necessitates a robust model by which the performance of the multiple access scheme can be characterized. In this paper, we analyze only the downlink for such a system. We will examine performance assuming that each portable terminal receives a direct signal from the base station consisting of its own intended signal superimposed with those signals intended for other users. Multipath is also present, adding additional interference in the form of delayed and attenuated versions of these signals. A common interpretation of this received signal is that only the direct signal consisting of the reference user's data is desired. In this paper, we propose what we consider to be an improved model, in which a portion of the delayed versions of the reference user's signal is also desired. Additionally, we choose our channel model to consist of a dominant line-of-sight path with a decaying tail of delayed reflections, whereas in other papers the use of a uniform delay profile is widespread. In section II of this paper, we characterize the channel and power considerations. Next, the new model is applied to the received signal, and the local mean BER is obtained in sections III and IV. In section V we analyze the implications of our model, and a summary is presented in section VI.

II. CHANNEL

The system under consideration employs direct sequence spread spectrum for transmissions. Since we can resolve the fading channel only at multiples of the chip time T_c, we employ a discrete-time impulse response model for the channel [9]:

$$h(t) = \sum_{l=0}^{L-1} \beta_l \delta(t - lT_c) e^{j\theta_l} \tag{1}$$

where L is the number of resolvable multipaths, $\{\beta_l\}$ are random amplitudes (not necessarily statistically independent), and $\{\theta_l\}$ are random phases. Because of the presence of a strong line of sight between the base station and user, the 0th resolvable path is modeled as a Rician distributed random variable. It can be broken down into the sum of a deterministic line of sight component β_{LOS} plus early reflections adding coherently to the line-of-sight (LOS) component, which make up a zero-mean Gaussian scattering component $\beta_S e^{j\theta_S}$ [2]. That is,

$$\beta_0 = \beta_{LOS} + \beta_S e^{j\theta_S}. \tag{2}$$

Assuming power attenuation according to the inverse square law, β_{LOS} is determined by free space loss. Hence, for a signal transmitted at frequency f_o, the power received at a distance d from the base station is

$$\beta_{LOS} = \frac{\mathbf{c}}{4\pi d f_o} \tag{3}$$

where **c** is the speed of light. We may assume the remaining, delayed terms have amplitudes $\beta_l, 1 \leq l \leq L-1$, which are Rayleigh distributed. The phase offset $\theta_l, 1 \leq l \leq L-1$, is assumed uniform over $[0,2\pi]$. Following [8], we ascribe an exponential delay profile $f_{T_d}(t)$ to the reflected paths, where T_d is the rms delay spread of the profile. This contrasts with other work (e.g. [10]) in which a uniform delay profile with independent, identically distributed (i.i.d.) path gains is assumed. The normalized delay spread is thus

$$f_{T_d}(t) = \frac{1}{T_d} e^{-t/T_d}. \tag{4}$$

By integrating $f_{T_d}(t)$ over bins of chip intervals, we obtain the local mean power in the lth path:

$$E(\beta_l^2) = P_d e^{-lT_c/T_d}(1 - e^{-T_c/T_d}),\ l \geq 1 \tag{5}$$

where P_d is the local mean scattered and delayed power. Hence,

$$P_d = E(\beta_S^2) + \sum_{l=0}^{L-1} E(\beta_l^2). \tag{6}$$

Note that equation (6) implies that

$$E(\beta_S^2) = P_d(1 - e^{-T_c/T_d}). \tag{7}$$

III. MULTIPLE USERS

Consider now a base station transmitting a direct sequence (DS) spread spectrum BPSK signal to K mobile units in parallel. Then the passband signal $a(t)$ transmitted by the base station is the superposition of K signals intended for the different users:

$$a(t) = \sum_{k=1}^{K} d_k(t-\tau_k) p_k(t-\tau_k) \cos(2\pi f_o t + \phi_k) \tag{8}$$

where $d_k(t)$ is the kth user's data signal, $p_k(t)$ is the spreading sequence for the kth user, ϕ_k is the random phase of the kth carrier, and τ_k is the random time delay. We will assume the base station transmits all signals bit and carrier synchronously so that the delay τ_k and phase ϕ_k are identically 0 for each of the K users' signals. Each data signal is modeled as

$$d_k(t) = \sum_{m=-\infty}^{\infty} b_m^k P_T(t - mT) \tag{9}$$

where b_m^k represents the kth user's data symbol, chosen from the set $\{\pm 1\}$. T is the bit time and $P_T(\cdot)$ denotes a rectangular pulse with amplitude 1, width T. The spreading sequence consists of a sequence of rectangular chips of duration T_c taking on values $x_m^k \in \{\pm 1\}$:

$$p_k(t) = \sum_{m=-\infty}^{\infty} x_m^k P_{T_c}(t - mT_c) \tag{10}$$

where $P_{T_c}(\cdot)$ is a unit amplitude pulse with width T_c. We assume the code sequence x_m^k is periodic with period $N = T/T_c$. In the indoor environment under consideration, the BER is essentially a function of multiuser interference, with channel noise having very little consequence. Hence, channel noise is ignored in what follows. The superimposed signals all fade in unison over the same channel, so the signal received at the mobile is

$$\begin{aligned} r(t) &= a(t) * h(t) \\ &= \sum_{l=0}^{L-1} \sum_{k=1}^{K} \beta_l d_k(t - lT_c) p_k(t - lT_c) \\ &\qquad \cdot \cos\big(2\pi f_o(t - lT_c) + \theta_l\big) \end{aligned} \tag{11}$$

with $h(t)$ given by (1). This signal may be decomposed into the superposition of signals representing the reference user's signal with self-interference due to multipath, and a second term representing the combined interference of the signals transmitted to the other mobile users. With the reference user arbitrarily assigned to user 1, this decomposition is

$$\begin{aligned} r(t) &= \sum_{l=0}^{L-1} \beta_l d_1(t - lT_c) p_1(t - lT_c) \\ &\qquad \cdot \cos\big(2\pi f_o(t - lT_c) + \theta_l\big) \\ &+ \sum_{l=0}^{L-1} \sum_{k=2}^{K} \beta_l d_k(t - lT_c) p_k(t - lT_c) \\ &\qquad \cdot \cos(2\pi f_o(t - lT_c) + \theta_l). \end{aligned} \tag{12}$$

The base station will provide a pilot signal to allow synchronous reception by the reference user. Therefore, no generality is lost by setting the phase θ_0 of the first resolvable path to zero. The output of the integrate-and-dump receiver is then

$$g(nT) = \frac{2}{T} \int_{(n-1)T}^{nT} r(t) p_1(t) \cos(2\pi f_o t)\, dt. \tag{13}$$

So the first bit of the message is detected as

$$\begin{aligned} g(T) &= \beta_0 b_0^1 \\ &+ \frac{1}{T} \sum_{l=1}^{L-1} \beta_l \cos(2\pi f_o l T_c - \theta_l) \\ &\qquad \cdot \int_0^T d_1(t - lT_c) p_1(t - lT_c) p_1(t)\, dt \end{aligned}$$

$$+\frac{1}{T}\sum_{l=0}^{L-1}\sum_{k=2}^{K}\beta_l\cos(2\pi f_o lT_c-\theta_l)$$
$$\cdot\int_0^T d_k(t-lT_c)p_k(t-lT_c)p_1(t)\,dt \quad (14)$$

where the double frequency terms have been filtered. Note we have used the fact that

$$\frac{2}{T}\int_0^T\beta_0 d_1(t)p_1^2(t)\cos^2(2\pi f_o t)\,dt=\beta_0 b_0^1, \qquad (15)$$

and this is the desired signal.

The second and third terms contain a cross correlation multiplied by data function $d_k(t-lT_c), 0\le k\le K$. Using the fact that $d_k(t-lT_c)$ will change value from b_{-1}^k to b_0^k at time lT_c, we can write the correlation as a sum of two partial correlations:

$$g(T)=\beta_0 b_0^1$$
$$+\sum_{l=1}^{L-1}\beta_l\cos(2\pi f_o lT_c-\theta_l)[b_{-1}^1R_{11}(lT_c)+b_0^1\hat{R}_{11}(lT_c)]$$
$$+\sum_{l=0}^{L-1}\sum_{k=2}^{K}\beta_l\cos(2\pi f_o lT_c-\theta_l)$$
$$\cdot[b_{-1}^kR_{k1}(lT_c)+b_0^k\hat{R}_{k1}(lT_c)] \qquad (16)$$

where the partial correlation functions are

$$R_{k1}(\tau):=\frac{1}{T}\int_0^\tau p_k(t-\tau)p_1(t)\,dt \qquad (17)$$
$$\hat{R}_{k1}(\tau):=\frac{1}{T}\int_\tau^T p_k(t-\tau)p_1(t)\,dt. \qquad (18)$$

In equation (16) we have assumed that $L\le N$ so that we get no delayed signal transmitting bit b_{-2}^k. This is equivalent to assuming that the maximum delay spread of the fading channel is less than the bit time T, which certainly is reasonable in a LOS transmission.

The first term $\beta_0 b_0^1$ in (16) is clearly the desired signal from the reference user. In previous work, all remaining terms in (16) have been presumed to interfere with the transmitted signal (see for example [4]). However, here we adopt the posture that not all of these terms constitute interference. In particular, we focus on the second term,

$$\sum_{l=1}^{L-1}\beta_l\cos(2\pi f_o lT_c-\theta_l)[b_{-1}^1R_{11}(lT_c)+b_0^1\hat{R}_{11}(lT_c)],$$

which represents the reference user's self-interference due to multipath. The second half of the summation with $b_0^1\hat{R}_{11}(lT_c)$ results from overlap of bit b_0^1 with itself, simply delayed by lT_c. In this sense, it adds coherently to the desired signal $\beta_0 b_0^1$. It can appropriately be considered part of the wanted signal since it contains the right modulation, but merely has a phase offset. This suggests that it is logical to include the second term as part of the Rician fading of the desired signal.

We would now like to obtain the local mean BER in this environment. We consider the ensemble of all possible channel realizations of the delayed resolvable paths. In this case, the delayed path amplitudes become time-constant Rayleigh random variables. Since the Rayleigh random variable is complex Gaussian in nature, the in-phase components of the interference signals are all independent Gaussian random variables. Thus, the error analysis is exactly as in a radio channel with AWGN. To summarize the discussion above, we have that the desired signal is

$$S=[\beta_0+\gamma]b_0^1 \qquad (19)$$

where γ is the coherent self-interference caused by delayed paths:

$$\gamma=\sum_{l=1}^{L-1}\beta_l\cos(2\pi f_o lT_c-\theta_l)\hat{R}_{11}(lT_c). \qquad (20)$$

The interference terms which add incoherently are grouped as

$$I=\sum_{l=1}^{L-1}\beta_l\cos(2\pi f_o lT_c-\theta_l)[b_{-1}^1R_{11}(lT_c)]$$
$$+\sum_{l=1}^{L-1}\sum_{k=2}^{K}\beta_l\cos(2\pi f_o lT_c-\theta_l)$$
$$\cdot[b_{-1}^kR_{k1}(lT_c)+b_0^k\hat{R}_{k1}(lT_c)]$$
$$+\sum_{k=2}^{K}\beta_0[b_{-1}^kR_{k1}(0)+b_0^k\hat{R}_{k1}(0)]. \qquad (21)$$

The last term corresponds to the $l=0$ case, in which other users' signals are arriving with Rician distributed amplitudes, and $\theta_0=0$. It is fairly simple to choose spreading codes which are orthogonal or very nearly so at zero offset. Hence we may take $R_{k1}(0)=R_{k1}(0)$ and thereby discard the final term of (21).

By applying the Gaussian assumption used in [5], we can simplify the analysis of the BER. The idea here is that in general, $KL\gg 1$, and it follows that we can use the central limit theorem to model all the self-interference and multiuser interference terms of I as random Gaussian noise. The two (remaining) terms of I are mutually independent, and thus I can be modeled as zero-mean Gaussian noise with variance equal to the sum of the variances of the two individual terms. Hence, the total power of the interference term is

$$E(I^2)=\frac{1}{2}\sum_{l=1}^{L-1}E(\beta_l^2)E\left([b_{-1}^1R_{11}(lT_c)]^2\right)$$

$$+\frac{1}{2}\sum_{l=1}^{L-1}\sum_{k=1}^{K}E(\beta_l^2) \cdot E\left([b_{-1}^k R_{k1}(lT_c)+b_0^k\hat{R}_{k1}(lT_c)]^2\right). \quad (22)$$

Note we have used the fact that θ_l, β_l, and the bit sequence for each user are independent of one another. The factor of $\frac{1}{2}$ falls out of the $\cos(\cdot)$ term as a result of the fact that θ_l is uniform on $[0, 2\pi]$. The conditional probability of error is then

$$P(e|\beta_0,\gamma)=Q\left(\frac{\beta_0+\gamma}{\sqrt{Var(I)}}\right) \quad (23)$$

where $Var(I)$ is given by (22) and

$$Q(y)=\frac{1}{\sqrt{2\pi}}\int_y^\infty e^{-\frac{x^2}{2}}\,dx. \quad (24)$$

IV. INTERFERENCE POWER

In order to evaluate (22), the variance of the partial correlation terms of the form $E[(R_{kl})^2]$ must be determined. A good approximation for this variance can be obtained by modeling the spreading sequence x_m^k as a purely random binary sequence [7]. To calculate

$$E\left([b_{-1}^k R_{k1}(lT_c)+b_0^k\hat{R}_{k1}(lT_c)]^2\right)$$

we note that bit reversal will occur $\frac{1}{2}$ of the time. If there is no bit reversal, the expectation becomes, by making use of (17), (18), and (10),

$$\begin{aligned}
&E\left([b_{-1}^k R_{k1}(lT_c)+b_0^k\hat{R}_{k1}(lT_c)]^2\,\middle|b_{-1}^k=b_0^k\right)\\
&=E\left([b^k]^2\right)E\left(\left[\frac{1}{T}\int_0^T p_k(t-lT_c)p_1(t)\,dt\right]^2\right)\\
&=E\left(\frac{1}{T^2}\int_0^T p_k(t-lT_c)p_1(t)\,dt\int_0^T p_k(\tau-lT_c)p_1(\tau)\,d\tau\right)\\
&=E\left(\frac{1}{T^2}\sum_{i=0}^{N-1}T_c x_{i-l}^k x_i^1\sum_{j=0}^{N-1}T_c x_{j-l}^k x_j^1\right)\\
&=E\left(\frac{T_c^2}{T^2}\sum_{i=0}^{N-1}\sum_{j=0}^{N-1}X_iX_j\right)
\end{aligned}$$

where X_n are i.i.d., taking on ± 1 with equal probability (note from (21) that $l \geq 1$ here). Continuing,

$$\begin{aligned}
E\left([b_{-1}^k R_{j1}(lT_c)+b_0^k\hat{R}_{k1}(lT_c)]^2\,\middle|b_{-1}^k=b_0^k\right)
&=E\left(\frac{1}{N^2}\sum_{i=0}^{N-1}X_i^2\right)\\
&=\frac{N}{N^2}=\frac{1}{N}. \quad (25)
\end{aligned}$$

The second case, also occurring with probability $\frac{1}{2}$, is when the bit switches sign between b_{-1}^k and b_0^k. However, the reversal of the bit does not effect the randomness of the binary sequence x_m^k. Hence, the variance works out to the same value as that without bit reversal, $\frac{1}{N}$. Therefore, the overall variance is

$$E\left([b_{-1}^k R_{k1}(lT_c)+b_0^k\hat{R}_{k1}(lT_c)]^2\right)=\frac{1}{2}\frac{1}{N}+\frac{1}{2}\frac{1}{N}=\frac{1}{N}. \quad (26)$$

Similar calculations show that

$$E\left([b_{-1}^1 R_{11}(lT_c)]^2|l\right)=\frac{l}{N^2}. \quad (27)$$

Hence, we arrive at

$$E(I^2)=\frac{1}{2}\sum_{l=1}^{L-1}E(\beta_l^2)\frac{l}{N^2}+\frac{K-1}{2}\sum_{l=1}^{L-1}E(\beta_l^2)\frac{1}{N} \quad (28)$$

where $E(\beta_l^2)$ can be determined from (5).

We can now return to the probability of bit detection error, as given in equation (23). To remove the conditioning on β_0 and γ, we use the argument previously set forward that γ, being representative of the interference between the modulated data bit and a delayed version of itself, is part of the Rician fading of the signal. In this context, we can model the signal $(\beta_0+\gamma)$ as a single Rician fading random variable R_0 with pdf

$$f_{R_0}(r)=\frac{r}{\sigma_{R_0}^2}e^{-(r^2+\beta_{LOS}^2)/2\sigma_{R_0}^2}I_0\left(\frac{r\beta_{LOS}}{\sigma_{R_0}^2}\right) \quad (29)$$

where $\sigma_{R_0}^2=\sigma_{\beta_0}^2+\sigma_\gamma^2$ is the local mean power of the scattered component of R_0. Note that by symmetry with (27),

$$\begin{aligned}
\sigma_\gamma^2=E(\gamma^2)&=\frac{1}{2}\sum_{l=1}^{L-1}E(\beta_l^2)E([b_0^1\hat{R}_{11}(lT_c)]^2)\\
&=\frac{1}{2}\sum_{l=1}^{L-1}E(\beta_l^2)\frac{l}{N^2}. \quad (30)
\end{aligned}$$

Using this model, together with equation (5), we integrate over the conditional probability of (23) to get

$$P(e)=\int_0^\infty Q\left(\frac{r}{\sqrt{Var(I)}}\right)f_{R_0}(r)\,dr. \quad (31)$$

Thus, with (31) we have obtained an expression for the local-mean BER for the downlink in a DS-CDMA transmission over a Rician multipath channel.

V. DISCUSSION OF RESULTS

Our primary result in the last section was equation (31), which gives the local mean BER for the base station-to-user downlink. This expression was obtained by setting the desired portion of the received signal to

$$S = [\beta_0 + \gamma] b_0^1$$

with

$$\gamma = \sum_{l=1}^{L-1} \beta_l \cos(2\pi f_o l T_c - \theta_l) \hat{R}_{11}(lT_c)$$

while the remaining portion of the received signal was modeled as interference with power

$$E(I^2) = \frac{1}{2}\sum_{l=1}^{L-1} E(\beta_l^2)\frac{l}{N^2} + \frac{K-1}{2}\sum_{l=1}^{L-1} E(\beta_l^2)\frac{1}{N}.$$

Prior models differ from that presented in this paper in two respects. First, they most often assume a uniform delay profile in which all path amplitudes are i.i.d. Rayleigh or Rician distributed. This gives rise to a different channel impulse response

$$h'(t) = \sum_{l=0}^{L-1} \beta_l' \delta(t - lT_c) e^{j\theta_l'}. \qquad (32)$$

Secondly, prior models assume the second term γb_0^1 in S to be part of the interference I. Hence, in the prior model, the desired portion of the received signal is

$$S' = \beta_0' b_0^1. \qquad (33)$$

The interference now includes γ, and so we have

$$E[(I')^2] = \frac{1}{2}\sum_{l=1}^{L-1} E[(\beta_l')^2] E\left([b_{-1}^1 R_{11}(lT_c) b_0^1 + \hat{R}_{11}(lT_c)]^2\right) + \frac{1}{2}\sum_{l=1}^{L-1}\sum_{k=2}^{K} E[(\beta_l')^2] \cdot E\left([b_{-1}^k R_{k1}(lT_c) + b_0^k \hat{R}_{k1}(lT_c)]^2\right) \qquad (34)$$

which is a modified version of (22). Taking expectations of the partial correlations gives us

$$E[(I')^2] = \frac{L-1}{2} E[(\beta_l')^2]\frac{1}{N} + \frac{(K-1)(L-1)}{2} E[(\beta_l')^2]\frac{1}{N}. \qquad (35)$$

The local mean BER in this model is

$$P'(e) = \int_0^\infty Q\left(\frac{r}{\sqrt{Var(I')}}\right) f_{\beta_0'}(r)\, dr \qquad (36)$$

where

$$f_{\beta_0'}(r) = \frac{r}{\sigma_{\beta_0}^2} e^{-(r^2+\beta_{LOS}^2)/2\sigma_{\beta_0}^2} I_0\left(\frac{r\beta_{LOS}}{\sigma_{\beta_0}^2}\right). \qquad (37)$$

We compare (36) to the BER implied by our model, given in (31), in Fig. 1. In this graph the spreading factor is set at $N = 64$, with the exponential delay profile consisting of $L = 30$ multipaths, and the uniform profile having $L = 6$ i.i.d. Rician paths. We assume a data rate of 2 Mb/s and rms delay spread $T_d = 50ns$. These parameters are used in the INFOPAD project at the University of California at Berkeley. The total power in both paths are equal with Rician K factors set at 6 dB. The new model clearly suggests that previous models are pessimistic in their BER estimation, although most authors enhance the i.i.d. results via maximal ratio combining (MRC) reception or selection diversity.

Our new model allows us to find the packet erasure rate, conditional on the channel. If the channel environment is fast fading, then fading is independent from bit to bit, and the local-mean PER is

$$PER = 1 - [1 - \int_0^\infty P(e|R_0) f_{R_0}(r)\, dr]^M \qquad (38)$$

where M is the number of bits in a packet. The PER for fast fading is plotted in Fig. 2, with M=63 bits, for three different Rician K_R factors. A more realistic model in the indoor environment is to assume slow fading. In this case the fading is constant from bit to bit, and the local-mean PER becomes

$$PER = 1 - \int_0^\infty [1 - P(e|R_0)]^M f_{R_0}(r)\, dr. \qquad (39)$$

A graph of the slow fading PER for M=63 bits and different values of K_R is given in Fig. 3.

VI. CONCLUSION

In this paper we have examined the discrete time channel model for the synchronous downlink in an indoor CDMA system. In such systems multipath is inevitably present, causing degradation of signal quality via fading. A commonly used model assumes that for a channel with L resolvable paths, the signal amplitudes are all i.i.d. We have applied two improvements. First we introduced different mean powers for different resolvable paths according to an exponential delay profile. This was reported in [8]. Secondly, we improved the model of the effect of multipath self-interference by taking into account that delayed resolvable paths contain the same bits, and therefore add coherently (by phasor addition) rather than the incoherent (noise-like)

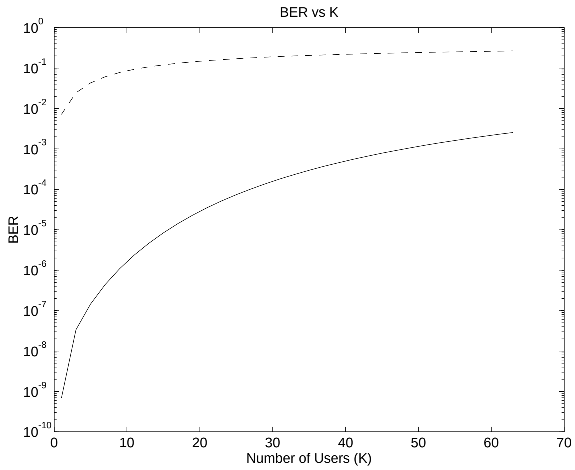

Figure 1: Local Mean BER vs. Number of Users. The dashed line is the BER $P'(e)$ obtained by (36), and the solid line shows $P(e)$ obtained by (31).

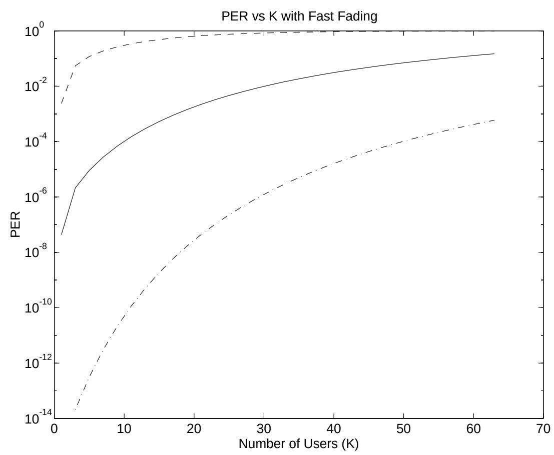

Figure 2: PER vs. Number of Users For Fast Fading $- -\ K_R = 0dB \quad - K_R = 6dB \quad - \cdot\ K_R = 10dB$

cumulation modeled previously. This improvement has been addressed in detail in this paper.

We conclude that our improved model gives significantly different local mean bit error rates, as our curves are several orders of magnitude more optimistic than the older model. We believe these improvements to be realistic in a picocellular indoor environment. It is also notable that although most papers use the fast fading model for indoor transmission, the indoor channel is actually very slowly fading and mathematically this results in improved performance, at least when no error correction scheme is applied.

Our model allows the computation of the correlation (and covariance matrix) of the samples taken in a RAKE receiver at the different resolvable paths, using Wiener filtering rather than Maximal Ratio Combining. Analysis of this system is out of the scope of the present paper, but will be conducted in the future.

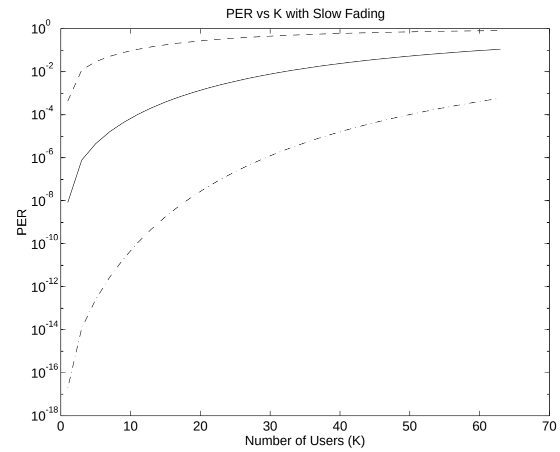

Figure 3: PER vs. Number of Users For Slow Fading $- -\ K_R = 0dB \quad - K_R = 6dB \quad - \cdot\ K_R = 10dB$

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