CDMA Overlay Situations for Microcellular Mobile Communications

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Abstract--**Direct sequence code division multiple access communications is a promising approach to cellular mobile communications, which operates in an environment characterized by multipath Rician fading. In this paper, the CDMA network is assumed to share common spectrum with a narrowband microwave user. Because of the presence of the narrowband waveform, an interference suppression filter at each CDMA receiver is employed to reject the narrowband interference. The problem of interference from adjacent cells is also considered. Average power control is assumed to combat the near/far problem, and multipath diversity, in conjunction with simple interleaved channel coding, is considered for improving the performance of the CDMA system.**

I. INTRODUCTION

Code division multiple access (CDMA) is becoming a very attractive technique for personal communications networks (PCN) and microcellular mobile communications [1-7]. One reason is that the CDMA code sequences inherently combat multipath which is a major concern in mobile radio communications. Also, since the CDMA signals spread their power over a large bandwidth, the effect of such a transmission on a narrowband receiver is often just an imperceptible rise in its noise level. Hence, it may be possible to overlay a CDMA system on an existing set of microwave narrowband users without adversely affecting either. Moreover, because of the broadband noise-like character of direct sequence (DS) CDMA signals, it is possible to use newly developed signal processing techniques to perform narrowband interference suppression [14]. This will clearly reduce the narrowband interference for CDMA users, but it will also help the narrowband users, since it is then possible to operate the CDMA transmissions at a reduced power level.

As is well known, microcellular mobile radio can increase the capacity of future cellular systems. The chief advantages of the microcells are the enhanced spectrum efficiency and low power consumption of the handsets. Microcells, as the name indicates, are relatively much smaller in size (0.2-1 km radius) than the conventional cells (1 - 10 km radius). Also, microcells operate at lower power and have their antennas at streetlamp elevations. In [3], it is shown that in multipath models the propagation path (per path) loss exponent for close-in cells can be between two and three in microcellular radio, instead of four in conventional cellular and terrestrial radio. Also, the propagation measurements in [4] characterize the microcellular environments as a multipath Rician (rather than Rayleigh) fading channel.

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Because of the presence of the narrowband waveform, a suppression filter at each CDMA receiver is employed to reject the interference. The problem of interference from adjacent cells is considered. Also, diversity is used in conjunction with block interleaved coding for improving the BER performance. The outline of the paper is as follows. The system and channel models are described in Section II. Section III derives an approximation to the bit error rate (BER) performance of the DS-CDMA overlay system, and this is followed by both representative numerical results and a discussion of these results in Section IV. Finally, conclusions are presented in Section V.

II. SYSTEM AND CHANNEL MODELS

A. Transmitter Model

The transmitter model of a DS system with interleaved coding and PSK modulation consists of a channel encoder, an interleaver and a DS spreader (or modulator). The transmitted signal of the kth user in the CDMA system takes the form

$$S_k(t)=\mathrm{Re}\left\{\sqrt{2P_k}\, b_k(t)\, a_k(t) \exp\left[j(2\pi f_0 t+\theta_k)\right]\right\}, \quad (1)$$

where $\mathrm{Re}\{\cdot\}$ stands for real part, f_0 denotes the carrier frequency, which is common to all users, and P_k and θ_k are the transmitted power and phase of the kth user, respectively. The symbol $b_k(t)$ is the kth interleaved and coded information bit; the bit rate $R_b = 1/T_b$, where T_b denotes the duration of a coded symbol. The information bit rate R_i (or the rate of source bits) and information bit duration T_i are given by $R_i = 1/T_i = (m/n)R_b$ and $T_i = T_b/(m/n)$, respectively, where m/n is the rate of the code. The λth square pulse of $b_k(t)$ has amplitude $b_k^{(\lambda)}$, taking values from $\{+1,-1\}$. The random signature sequence waveform $a_k(t)$ has a chip rate $1/T_c$, and the jth chip of $a_k(t)$ is denoted by $a_j^{(k)}$. We assume that there are N chips (processing gain) of the random signature sequence in each coded bit ($T_b = NT_c$). Therefore, the spread spectrum bandwidth (B_s) is approximately given by

$$B_s = 2T_c^{-1} = 2N(T_i m/n)^{-1} . \quad (2)$$

B. Channel Model

The mobile radio channel can be modeled as a discrete multipath fading channel. "Discrete multipath" means that the received signal is the sum of a finite number of delayed versions of the transmitted signal, whereas "fading" refers to the fact that, for a given transmitted power, the received power fluctuates in a random fashion. Because of the multipath nature of the channel, the fading is frequency-selective.

The propagation measurements in [4] characterize the microcellular environments as a multipath Rician fading channel, where the received signal consists of several specular components plus several Rayleigh-fading components. The multipath Rician fading channel between the kth user and the receiver of interest (namely the receiver in the base station of what we refer to as the first cell) is modeled by the complex lowpass equivalent impulse response

$$h_k(t) = \frac{1}{(d_{1,k})^{\gamma/2}} \sum_{l=1}^{L} \left[A_{kl}\exp(j\eta_{kl}) + \beta_{kl}\exp(j\mu_{kl}) \right] \delta(t-\tau_{kl}), \quad (3)$$

where $d_{1,k}$ ($d_{1,k} \neq 0$) is the distance between the kth user and the first cell base station, and γ is the propagation path loss exponent. In (3), A_{kl} ($0 \leq A_{kl} \leq 1$) and η_{kl} are gain and phase of the specular component of the lth path from the kth user, respectively; they are assumed to be deterministic and nearly constant over the duration of at least one bit. The random gain β_{kl} and random phase μ_{kl} of the fading component of the lth path of the kth user have a Rayleigh distribution with $E[\beta_{kl}^2] = 2\rho_{kl}$, and a uniform distribution in $[0, 2\pi]$, respectively. The path delay, τ_{kl}, is uniformly distributed in $[0, T_b]$. We assume that there are L paths associated with each user. The gains, delays and phases of different paths and/or of different users are all statistically independent. We define the quantity H_{kl}, as the ratio of the specular component power to the Rayleigh fading power. That is,

$$H_{kl} \equiv A_{kl}^2 / E[\beta_{kl}^2] = A_{kl}^2/(2\rho_{kl}) . \quad (4)$$

The microcellular (outdoor) channel measurements from [4] show that Rician distributions occur with H_{kl} values ranging from 7 dB to 12 dB, instead of 2dB to 7dB for an indoor radio channel [7].

The interference in the channel is assumed to be a non-fading narrowband BPSK signal, and is given by

$$J(t) = \mathrm{Re}\left\{ \sqrt{2J}\, d(t) \exp\left[j\, [2\pi(f_0+\Delta)t + \Theta] \right] \right\}, \quad (5)$$

where Δ stands for the offset of the interference carrier frequency from the carrier frequency (f_0) of the CDMA signals. The parameters J and Θ denote the received interference power and phase, respectively. The information sequence $d(t)$ has the bit rate T_j^{-1}, where T_j denotes the duration of one bit. Therefore, the interference bandwidth is approximately $B_j = 2T_j^{-1}$ (we assume $B_j < B_s$). The ratio (p) of the interference bandwidth to the spread spectrum bandwidth and the ratio (q) of the offset of the interference carrier frequency to half of the spread spectrum bandwidth are defined as, respectively,

$$p = B_j/B_s = T_c/T_j \quad (6)$$

and

$$q = \Delta/(B_s/2) = \Delta T_c . \quad (7)$$

Therefore, the received signal $r(t)$ can be represented as

$$r(t) = \mathrm{Re}\left\{ \sqrt{2P} \sum_{k=1}^{CK} \sqrt{\varepsilon(\gamma, c_k, k)} \sum_{l=1}^{L} [A_{kl} exp(j\phi_{kl}) + \beta_{kl} exp(j\psi_{kl})] \cdot b_k(t-\tau_{kl}) a_k(t-\tau_{kl}) \exp(j2\pi f_0 t) \right\} + J(t) + n(t), \quad (8)$$

where $n(t)$ is an AWGN process with two-sided power spectral density $N_0/2$, $\phi_{kl} \equiv \theta_k + \eta_{kl} - 2\pi f_0 \tau_{kl}$, $\psi_{kl} \equiv \theta_k + \mu_{kl} - 2\pi f_0 \tau_{kl}$, C stands for the number of cells, each one containing K users, and c_k denotes the c_kth cell (the integer portion of $1+(k-1)/K$, $c_k = 1, \cdots, C$). The first cell ($c_k = 1$) is defined as the cell of interest, and P and $\varepsilon(\gamma, c_k, k)$ are defined as

$$P = P_k/(d_{c_k,k})^{\gamma} = \text{constant} \quad (9)$$

and

$$\varepsilon(\gamma, c_k, k) = (d_{c_k,k}/d_{1,k})^{\gamma} , \quad (10)$$

respectively, where $d_{c_k,k}$ is the distance of the kth mobile user to its own base station (the c_kth cell), and $d_{c_k,k} \neq 0$. The constant P means each base station provides adaptive power control to all K users of its own cell so that the received signals from its cell have approximately equal power to avoid the near/far problem.

C. Receiver Model

As shown in Fig. 1(a), the receiver contains a suppression filter, a bank of bandpass (BP) matched filters, a bank of U parallel PSK demodulators and diversity circuits a hard decision device, a deinterleaver and a hard decision decoder. A detailed model of the PSK demodulators and diversity circuit is shown in Fig. 1(b).

Notice that, for simplicity, we have ignored any bandpass filtering at the front-end of the suppression filter. We take the suppression filter to be a double-sided Wiener filter with M taps on each side. Its impulse response is $\sum_{m=-M}^{M} \alpha_m \delta(t - mT_c)$, where $\alpha_0 = 1$ and $\alpha_m = \alpha_{-m}$. The resulting suppression filter output is given by

$$r_f(t) = \sum_{m=-M}^{M} \alpha_m r(t - mT_c) , \quad (11)$$

where $f_0 T_c$ is taken to be an integer.

The signal $r_f(t)$ enters a bank of BP matched filters, whose impulse responses are matched to consecutive T_b-second segments of $2a_i(t)\cos(2\pi f_0 t)\ P_{T_b}(t - vT_b)$, where $P_{T_b}(t) = 1$ for $0 \leq t < T_b$ and is 0 otherwise, $v = 1, \cdots, V$,

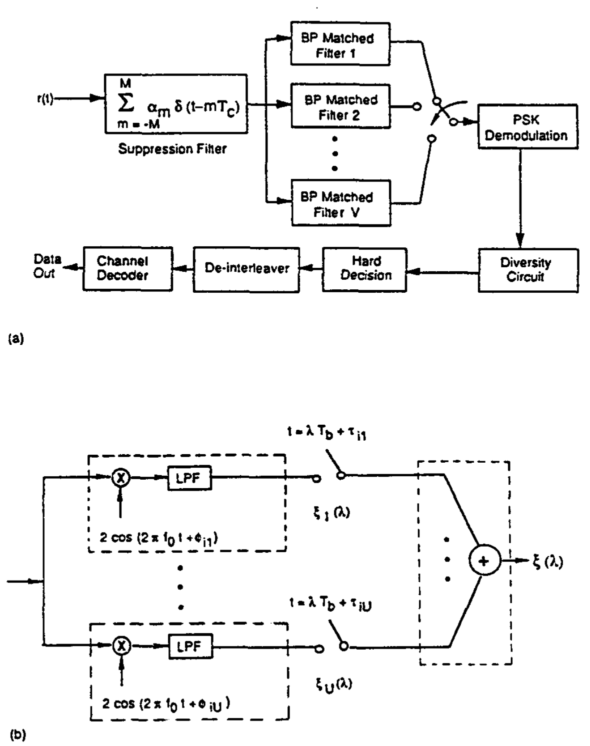

Fig. 1. (a) Receiver model of a CDMA overlay system. (b) A detailed model of the PSK demodulators and the diversity circuit.

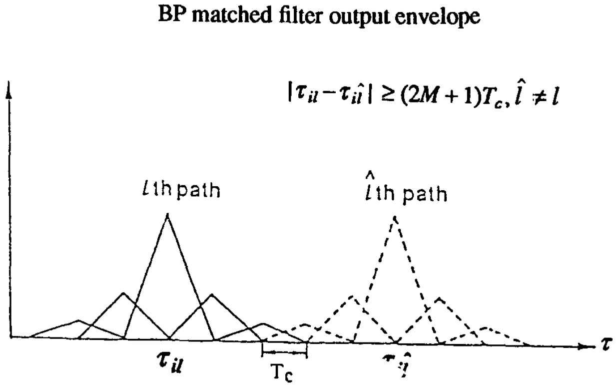

Fig. 2. The matched filter output envelope in CDMA overlay situation with suppression filters (2M+1=5) over a multipath channel.

and V is the number of matched filters. Also it is assumed that the ith user of the first cell is the reference user. During each bit duration, the BP signal at the output of the correct matched filter contains $2M+1$ narrow pulses, each of width $2T_c$, centered at instants $\lambda T_b+\tau_{il}+mT_c$ for $m=-M, \cdots, 0, \cdots, M$, which are caused by the $2M+1$ taps of the suppression filter and the lth $(l=1, 2, \cdots, L)$ path of the reference user. Of the $2M+1$ peaks, the middle peak (m=0) is the largest (or main) peak, due to the zero-th tap of the suppression filter, whereas sidelobe peaks (or earlier and later peaks) are due to the taps excluding the zero-th tap. We assume that $|\tau_{il}-\tau_{i\hat{l}}| \geq (2M+1)T_c$ for $\hat{l}=1, \cdots, L$, $\hat{l} \neq l$ (see Fig. 2), since if $|\tau_{il}-\tau_{i\hat{l}}| < (2M+1)T_c$, the overlay of the outputs due to different paths might preclude the paths from being resolvable. We further assume that the receiver only uses the combination of the main peaks of the resolvable paths of the reference user to form the diversity. Paths from non-reference users do not give rise to regular peaks, because their spreading sequences are different from $a_i(t)$.

In any given symbol interval, the output of the appropriate BP matched filter enters U conventional PSK demodulators, $1 \leq U \leq L$, (Fig. 1(b)), corresponding to U different resolvable paths of the reference user. The coherent receiver makes use of carrier reference signals $\{2\cos(2\pi f_0 t+\phi_{il})\}$, which are in phase with the specular components of the different signals from the reference user. These carrier reference signals can be derived from the BP matched filter outputs by means of carrier synchronization circuits (one circuit per considered path). The output of the lth, $1 \leq l \leq U$, demodulator is a lowpass signal, which has $2M+1$ peaks at the instants $\lambda T_b+\tau_{il}+mT_c$, $m=-M, \cdots, 0, \cdots, M$. The desired component of the lowpass output signal (i.e., the main peak, m=0) of the lth demodulator is sampled at the instant $t=\lambda T_b+\tau_{il}$, and this gives rise to the random variable $\xi_l(\lambda)$. Note that the time difference between any two samples is at least equal to $(2M+1)T_c$. The U random variables, $\xi_l(\lambda)$, $l=1,\cdots,U$, are directly summed to form the decision variable $\xi(\lambda)$.

III. SYSTEM PERFORMANCE

The lth random variable, $\xi_l(\lambda)$, for detection can be written as

$$\xi_l(\lambda) = \int_{\lambda T_b+\tau_{il}}^{(\lambda+1)T_b+\tau_{il}} r_f(t) a_i(t-\tau_{il}) \tag{12}$$

where $2\cos(2\pi f_0 t+\phi_{il})$ is the recovered carrier of the specular component of the lth resolvable path of the reference user and $a_i(t-\tau_{il})$ is the random spreading sequence of the lth path of the reference user. Since the high frequency terms are removed by the lowpass filter following the mixer of the demodulator, the above expression reduces to

$$\xi_l(\lambda) = \sqrt{2P}\, A_{il} T_b b_i^{(\lambda)} + D_l(\lambda) + F_l(\lambda) + N_l(\lambda) + J_l(\lambda) + \sum_{\substack{\hat{l}=1 \\ \hat{l}\neq l}}^{L} I_{i,l,\hat{l}}$$
$$+ \sum_{\substack{k=1 \\ k\neq i}}^{K} \sum_{\hat{l}=1}^{L} I_{k,l,\hat{l}} + \sum_{k=K+1}^{CK} \sum_{\hat{l}=1}^{L} W_{k,l,\hat{l}}, \tag{13}$$

where the terms on the right hand side of (13) are described in detail below.

--- The first term is the desired signal, corresponding to the specular component of the lth path of the reference user and the zero-th tap of the suppression filter. The power of this term equals

$$S_c = 2PA^2T_b^2, \tag{14}$$

where it is assumed that $A_{il}=A$ for all i and l. For most of the remaining terms, we are interested in their conditional variances, conditioned on $b_i^{(\lambda)}$ and $a_i(t)$. Let us consider them one at a time.

--- $D_l(\lambda)$ is an interference term due to the fading component of the lth path of the reference user and the zero-th tap of the suppression filter, and is given by

$$D_l(\lambda) = \sqrt{2P}\beta_{il}T_b b_i^{(\lambda)}\cos(\mu_{il}-\eta_{il}) . \quad (15)$$

Note that this term is a conditional zero-mean Gaussian random variable with variance $2P\rho T_b^2$, where it is assumed that $\rho_{il}=E(\beta_{il}^2)/2=\rho$ for all i and l.

--- $F_l(\lambda)$ is a self interference term due to the lth path of the reference user which is caused by the non-reference taps of the suppression filter; it is given by

$$F_l(\lambda) = \sqrt{2P}\,[A_{il}+\beta_{il}\cos(\mu_{il}-\eta_{il})] \sum_{\substack{m=-M \\ m\neq 0}}^{M} \alpha_m \cdot \int_{\lambda T_b+\tau_{il}}^{(\lambda+1)T_b+\tau_{il}} b_i(t-mT_c-\tau_{il})a_i(t-mT_c-\tau_{il})a_i(t-\tau_{il})dt . \quad (16)$$

For relatively small narrowband interference bandwidths (relative to the spread bandwidth of the CDMA signals), this term does not have a significant effect, and thus it will be ignored in what follows.

--- $N(\lambda)$ is due to the thermal noise; its conditional variance equals $N_0T_b \sum_{m=-M}^{M} \alpha_m^2$.

--- $J_l(\lambda)$ is due to the BPSK narrowband interference, and is given by

$$J_l(\lambda) = \sum_{m=-M}^{M} \alpha_m \int_{\lambda T_b+\tau_{il}}^{(\lambda+1)T_b+\tau_{il}} \sqrt{2J}\,d(t-mT_c) \cdot \cos\,[2\pi\Delta(t-mT_c)+\hat{\Theta}]a_i(t-\tau_{il})dt , \quad (17)$$

where $\hat{\Theta}\equiv\Theta+\phi_{il}$ and f_0T_c is assumed to be an integer. This term is very significant, and will be treated separately.

--- $I_{i,l,\hat{l}}\ (\hat{l}\neq l)$ is a multipath interference term due to the $\hat{l}$th path of the reference user and is given by

$$I_{i,l,\hat{l}} = \sqrt{2P}\,[A_{i\hat{l}}cos\,(\phi_{i\hat{l}}-\phi_{il})+\beta_{i\hat{l}}\cos(\psi_{i\hat{l}}-\phi_{il})] \sum_{m=-M}^{M} \alpha_m \cdot \int_{\lambda T_b+\tau_{il}}^{(\lambda+1)T_b+\tau_{il}} b_i(t-\tau_{i\hat{l}}-mT_c)a_i(t-\tau_{i\hat{l}}-mT_c)a_i(t-\tau_{il})dt \quad (18)$$

For a large number of CDMA users (i.e., for $K \gg 1$), the effect of the multipath of the desired user is very small, since it roughly acts as one additional user. However, its inclusion in the analysis greatly complicates that analysis, and so, as was the case with $F_l(\lambda)$, it will be ignored.

--- $I_{k,l,\hat{l}}\ (k\neq i)$ is a multi-access interference term due to the $\hat{l}$th path of the kth user in the first cell (i.e., c=1 and k=1, 2, ⋯, K; $k\neq i$) and is given by

$$I_{k,l,\hat{l}} = \sqrt{2P}\,[A_{k\hat{l}}\cos(\phi_{k\hat{l}}-\phi_{il})+\beta_{k\hat{l}}\cos(\psi_{k\hat{l}}-\phi_{il})] \sum_{m=-M}^{M} \alpha_m \cdot \int_{\lambda T_b+\tau_{il}}^{(\lambda+1)T_b+\tau_{il}} b_k(t-\tau_{k\hat{l}}-mT_c)a_k(t-\tau_{k\hat{l}}-mT_c)a_i(t-\tau_{il})dt . \quad (19)$$

The random variable $I_{k,l,\hat{l}}$ has a conditional variance equal to approximately σ_0^2, where

$$\sigma_0^2 = \frac{P(A^2+2\rho)}{3N}\left[2\sum_{m=-M}^{M}\alpha_m^2+\sum_{m=-M}^{M}\alpha_m\alpha_{m+1}\right]T_b^2 , \quad (20)$$

where $\alpha_m = 0$ for $|\,m\,| > M$. Therefore, the conditional variance of the total multi-access interference term in (13) equals $(K-1)L\sigma_0^2$.

--- $W_{k,l,\hat{l}}$ is the interference term due to the $\hat{l}$th path of the kth user in the c_kth adjacent cell (k=K+1, ⋯, CK and $c_k\neq 1$), given by

$$W_{k,l,\hat{l}}=\sqrt{\varepsilon(\gamma,c_k,k)}\,I_{k,l,\hat{l}} , \quad (21)$$

where $I_{k,l,\hat{l}}$ is given by (19). $W_{k,l,\hat{l}}$ has the conditional variance, conditioned on $\varepsilon(\gamma,c_k,k)$, given by

$$E[W_{k,l,\hat{l}}^2]\,|\,_{\varepsilon(\gamma,c_k,k)} = \varepsilon(\gamma,c_k,k)\sigma_0^2 . \quad (22)$$

An exact description of the adjacent-cell interference in the hexagonal cellular model, shown in Fig. 3(a), is difficult, because the value of (22) depends on the position of the kth mobile user in the c_kth cell ($c_k\neq 1$). In order to avoid the dependence on the position, we average the result of (22) over the area of c_kth cell. Also, we approximate the hexagonal cell with a circular cell of equal area as in [5] (a hexagonal cell inscribed in a circular cell of radius 1.035r is equal in area to a circular cell of radius r), as shown in Fig. 3(b). (Note that there appears to be an error in the calculation of the radius of the circular cell in [5].) The factor $\varepsilon(\gamma,c_k,k)$, for c_k=2,⋯,C, is given by

$$\varepsilon(\gamma,c_k,k)=\left[\frac{x^2}{(x\cos\theta)^2+(y+x\sin\theta)^2}\right]^{\gamma/2} , \quad (23)$$

where y is the distance between the c_kth cell base station and the first cell base station. As shown in Fig. 3(a), the distance between any cell base station in the first layer of cells and the first cell base station is identical ($y=\sqrt{3}*1.035r=1.793r$), where r is the radius of the circular cell, and the first layer contains six cells. The second layer of cells contains twelve cells, of which six have the distance $y=2\sqrt{3}*1.035r=3.586r$ from the first cell base station and the other six have the distance $y=3*1.035r=3.105r$ from the first cell base station. Finally, the third layer of cells contains eighteen cells, of which six have the distance $y= 3\sqrt{3}*1.035r=5.379r$ and the other twelve have the distance $y=\sqrt{4.5^2+0.75^2}*1.035r\approx 4.74r$ from the first cell base station. Referring to Fig. 3(b), the average of (23) over the c_kth cell area is given by

$$\begin{aligned}\varepsilon(\gamma,c_k) &= E\,[\varepsilon(\gamma,c_k,k)] \\ &= \frac{1}{\pi r^2}\int_0^r\int_0^{2\pi}\left[\frac{x^2}{(xcos\theta)^2+(y+x\sin\theta)^2}\right]^{\gamma/2} xdxd\theta \\ &= \frac{1}{\pi}\int_0^1\int_0^{2\pi}\left[\frac{\hat{x}^2}{(\hat{x}\cos\theta)^2+(y/r+\hat{x}\sin\theta)^2}\right]^{\gamma/2} \hat{x}d\hat{x}d\theta . \end{aligned}\quad (24)$$

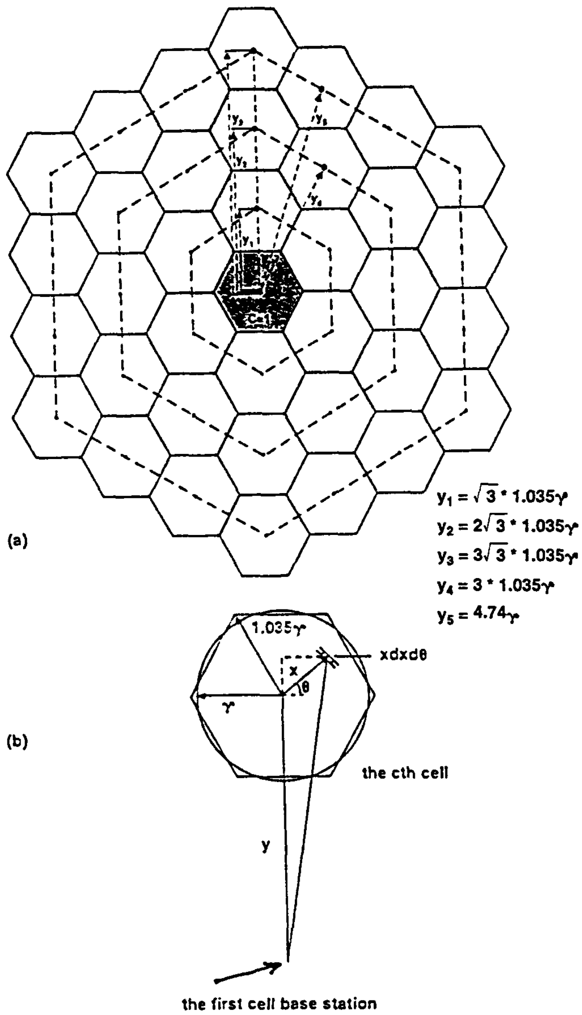

Fig. 3. (a) The microcellular model. (b) The hexagonal cell approximated by a circular cell of equal area.

Table I shows the result of (24) for γ=2,3,4 and for the various values of y. Therefore, the conditional variance of the total adjacent-cell interference involved with G layers of cells is approximated by

$$\sigma^2=\zeta(\gamma)KL\sigma_0^2 . \tag{25}$$

In (25), the factor $\zeta(\gamma)$ is due to all adjacent cells and is given by

$$\zeta(\gamma)=\sum_{j=1}^{G}\zeta_j(\gamma) , \tag{26}$$

where $\zeta_j(\gamma)$ is defined as

$$\zeta_j(\gamma)=\sum_{c_k\in s_j}\varepsilon(\gamma,c_k) , \tag{27}$$

and where s_j is the set of all cells in the jth layer.

Note that the factors $\varepsilon(\gamma,c_k)$, given by (24), are independent of cell size because we normalize distance to the cell radius. Thus, the total factor $\zeta(\gamma)$, given by (26), is due only to the number of layers of cells (or, equivalently, the number of cells).

A. *Average Probability of Error.*

To obtain an approximation to the average probability of error, we will use the general approach of [8] - [11], although the presence of the narrowband interference complicates the derivation. Before proceeding, let us first simplify the notation. The final test statistic is given by

$$\psi(\lambda) = \sum_{l=1}^{U}\xi_l(\lambda) , \tag{28}$$

where $\xi_l(\lambda)$ is given in (13). Consider rewriting (13) (after neglecting $F_l(\lambda)$ and $\sum_{\substack{l=1\\ l\neq 1}}^{L} I_{i,l,\hat{i}}$) as

γ	First Layer of cells		Second Layer of cells			Third Layer of cells			$\zeta(\gamma)$		
	$\varepsilon(\gamma,C)$	$\zeta_1(\gamma)$	$\varepsilon(\gamma,C)$		$\zeta_2(\gamma)$	$\varepsilon(\gamma,C)$		$\zeta_3(\gamma)$	$\zeta_1(\gamma)$	$\sum_{j=1}^{2}\zeta_j(\gamma)$	$\sum_{j=1}^{3}\zeta_j(\gamma)$
	y=1.793r		y=3.105r	y=3.586r		y=4.74r	y=5.379r				
2	0.1986	1.192	0.0559	0.0411	0.582	0.023	0.0177	0.382	1.192	1.774	2.16
3	0.125	0.75	0.016	0.0099	0.155	0.0041	0.0027	0.065	0.75	0.905	0.97
4	0.0935	0.561	0.005	0.0026	0.046	0.0008	0.0005	0.013	0.561	0.607	0.62

Table I. The result of Evaluating Eq. (24).

$$\xi_l(\lambda) = s_l + n_l + j_l, \tag{29}$$

where

$$s_l \triangleq \sqrt{2P}\, A\, T_b\, b_l^{(\lambda)}, \tag{30}$$

$$n_l \triangleq D_l(\lambda) + N_l(\lambda) + \sum_{\substack{k=1 \\ k\neq i}}^{K} \sum_{l=1}^{L} I_{k,l,\hat{l}} + \sum_{k=K+1}^{CK} \sum_{l=1}^{L} W_{k,l,\hat{l}}, \tag{31}$$

and

$$j_l \triangleq J_l(\lambda). \tag{32}$$

Also, define

$$S_T \triangleq \sum_{l=1}^{U} s_l, \tag{33}$$

$$N_T \triangleq \sum_{l=1}^{U} n_l \tag{34}$$

and

$$J_T \triangleq \sum_{l=1}^{U} j_l. \tag{35}$$

Note that, following the same arguments as in [7], conditioned on $b_l^{(\lambda)}$ and $a_i(t)$, N_T is asymptotically normal as $K \to \infty$, with zero conditional mean and conditional variance approximately given by

$$\mathrm{var}(N_T) \triangleq \sigma_T^2 = U\Bigg\{N_0 T_b \sum_{m=-M}^{M} \alpha_m^2 + P\rho T_b^2 + P(A^2+2\rho)T_b^2 \frac{[1+\zeta(\gamma)]KL-L}{3N} \sum_{m=-M}^{M} (2\alpha_m^2 + \alpha_m \alpha_{m+1})\Bigg\} \tag{36}$$

Therefore, we have for the conditional bit error rate, assuming $b_l^{(\lambda)} = +1$,

$$P(e \mid a_i(t), J_T) = \phi\left[-\frac{S_T + J_T}{\sigma_T}\right], \tag{37}$$

where

$$\phi(x) \triangleq \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{x} \exp[-y^2/2]\, dy. \tag{38}$$

Now let us consider J_T. For future use, let us tie the processing gain, N, and number of CDMA users, K, together, in the sense of letting $N = N(K)$ in such a manner that N increases monotonically with K. This can be justified physically by noting that any CDMA system must increase its processing gain as the number of users increases in order to maintain a satisfactory performance. However, the consequence of increasing processing gain is that either RF bandwidth increases or information rate decreases (or both). For the purposes of this paper, we assume T_b is constant and that T_c decreases as N increases (i.e., we assume constant information rate but we allow the RF bandwidth to increase). For simplicity, assume that there are an integer number of bits of narrowband waveform in T_b seconds, and ignore any timing offset between the bits of the BPSK signal and the bits of the reference CDMA signal. From (17), it is straightforward to show that

$$j_l = J_l(\lambda) = \sqrt{2J} \sum_{m=-M}^{M} \alpha_m \sum_{j=0}^{N-1} a_{j+m}^{(i)}\, d_{[pj]} \cdot \left[\frac{\sin[2\pi\Delta(\lambda T_b + (j+1)T_c) - 2\pi\Delta m T_c + \hat{\theta}_l]}{2\pi\Delta} - \frac{\sin[2\pi\Delta(\lambda T_b + jT_c) - 2\pi\Delta m T_c + \hat{\theta}_l]}{2\pi\Delta}\right], \tag{39}$$

where $\{d_j\}$ is the data sequence of the narrowband BPSK waveform, $[x]$ is the largest integer less than x, $p \triangleq T_c/T_j = N_1/N$, and N_1 is the number of bits of the narrowband BPSK signal per bit of any of the CDMA waveforms. If we assume $\Delta T_b = Nq$ is an integer, where $q \triangleq \Delta T_c$, we can ignore the term λT_b in the arguments of the sine functions. Further, if we define

$$\beta(j,m,\hat{\theta}_l) \triangleq \frac{\sin[2\pi q(j+1-m) + \hat{\theta}_l] - \sin[2\pi q(j-m) + \hat{\theta}_l]}{2\pi q}, \tag{40}$$

then

$$j_l = \sqrt{2J}\, T_c \sum_{j=0}^{N-1} x_j, \tag{41}$$

where

$$x_j \triangleq \sum_{m=-M}^{M} \alpha_m a_{j+m}^{(i)}\, d_{[pj]}\, \beta(j, m, \hat{\theta}_l). \tag{42}$$

Recall that

$$\hat{\theta}_l = \theta + \phi_{il}. \tag{43}$$

Conditioned on $\theta + \phi_{il}$, x_j is a function of $d_{[pj]}$ and $\{a_{j-M}^{(i)}, \cdots, a_{j+M}^{(i)}\}$. It can be shown that the sequence $\{x_j\}$ is a 2M - dependent sequence ([12]). For example, consider the simple case of $\Delta = 0$, $M = 1$ and $1/p = \frac{N}{N_1} = 5$. Then, assuming $a_j^{(i)} = d_j = 0$ for $j < 0$, we have the following sequence:

$$x_0 = \alpha_0\, d_0\, a_0 + \alpha_1\, d_0\, a_1$$
$$x_1 = \alpha_{-1}\, d_0\, a_0 + \alpha_0\, d_0\, a_1 + \alpha_1\, d_0\, a_2$$
$$x_2 = \alpha_{-1}\, d_0\, a_1 + \alpha_0\, d_0\, a_2 + \alpha_1\, d_0\, a_3$$
$$x_3 = \alpha_{-1}\, d_0\, a_2 + \alpha_0$$
$$x_4 = \alpha_{-1}\, d_0\, a_3 + \alpha_0$$
$$x_5 = \alpha_{-1}\, d_1\, a_4 + \alpha_0$$
$$x_6 = \alpha_{-1}\, d_1\, a_5 + \alpha_0$$
$$x_7 = \alpha_{-1}\, d_1\, a_6 + \alpha_0$$
$$\cdots$$

Note that in the sequence $\{x_0, x_1, x_2, \cdots\}$, if any $2M = 2$ consecutive x_i's are removed, the remaining two subsequences (the one to the left of the removed x_i's and the one to the right) are statistically independent.

Further, if $2\pi q = \frac{k_1 \pi}{k_2}$, where k_1 and k_2 are integers, then as $N \to \infty$, the sequence $\{x_j\}$ satisfies the conditions of the Hoeffding-Robbins version of the central limit theorem for dependent random variables ([12]). To illustrate this, consider again the above example. From [12], for a sum of m-dependent random variables to be asymptotically normal, it is necessary that

$$\lim_{P \to \infty} \frac{1}{P} \sum_{h=1}^{P} A_{i+h} = A \tag{44}$$

exist uniformly for all i, where

$$A_i \triangleq 2 \sum_{j=0}^{m-1} cov\{x_{i+j}, x_{i+m}\} + \mathrm{var}\{x_{i+m}\} . \tag{45}$$

In our example, $m = 2M = 2$ and

$$\mathrm{var}(x_i) = \sum_{j=-M}^{M} \alpha_j^2 \tag{46}$$

for all $i \geq 1$ (i.e., the variance term is independent of i as long as $i \geq 1$). The covariance term can be shown to take on one of three possible values as long as $i > 2$.

$$\sum_{j=0}^{1} \mathrm{cov}\{x_{i+j}, x_{i+2}\} = \begin{cases} 0 , & i = 3+5k \\ \alpha_{-1}\alpha_0 + \alpha_0\alpha_1 , & i = 4+5k \\ \alpha_{-1}\alpha_1 + \alpha_{-1}\alpha_0 + \alpha_0\alpha_1 , & \text{otherwise} \end{cases} \tag{47}$$

where $k = 1, 2, 3, \cdots$. Since the variance term is independent of i for $i \geq 1$, and the sum of covariances is periodic in i for $i \geq 2$, it can be seen that $\frac{1}{P}\sum_{h=1}^{P} A_{i+h}$ equals the same value, independent of i, as $P \to \infty$.

If we call the conditional variance of j_l, conditioned on $\hat{\theta}_l$, $\sigma_j^2(\hat{\theta}_l)$, then

$$\sigma_j^2(\hat{\theta}_l) = (2J\, T_b^2/N^2) \sum_{m_1=-M}^{M} \sum_{m_2=-M}^{M} \alpha_{m_1} \alpha_{m_2} \cdot \sum_{j_1=0}^{N-1} \sum_{j_2=0}^{N-1} \beta(j_1, m_1, \hat{\theta}_l)\beta(j_2, m_2, \hat{\theta}_l) \cdot \delta(j_1+m_1, j_2+m_2)\, \delta([pj_1], [pj_2]), \tag{48}$$

where $\delta(i,j)$ is the Kronecker delta function. Further, since j_{l_1} is uncorrelated with j_{l_2} for $l_1 \neq l_2$, the variance of J_T is $\sigma_J^2(\bar{\phi}_i) \triangleq \sum_{l=1}^{U} \sigma_j^2(\hat{\theta}_l)$, where $\bar{\phi}_i \triangleq \{\hat{\theta}_1, \hat{\theta}_2, \cdots, \hat{\theta}_U\}$.

Finally, to obtain our approximation to the average BER, consider the following: From (37), we must average the conditional error rate over $a_i(t)$ and J_T. However, from (36), σ_T^2 is not a function of the $\{a_j^{(i)}\}$, and from (48), $\sigma_j^2(\hat{\theta}_l)$ is only a function of the random phase $\hat{\theta}_l$. Therefore,

$$P(e \mid a_i(t), J_T) = P(e \mid \bar{\phi}_i) \approx \frac{1}{\sqrt{2\pi}\,\sigma_J(\bar{\phi}_i)} \int_{-\infty}^{\infty} \phi\left[-\frac{S_T + J_T}{\sigma_T}\right] \cdot \exp\{-J_T^2/2\sigma_J^2(\bar{\phi}_i)\}\, dJ_T = \phi\left[-\frac{S_T}{\sqrt{\sigma_J^2(\bar{\phi}_i)+\sigma_T^2}}\right], \tag{49}$$

and

$$P_e = E_{\bar{\phi}_i}\{P(e \mid \bar{\phi}_i)\}, \tag{50}$$

where $E_{\bar{\phi}_i}\{\cdot\}$ means expectation over the random vector $\bar{\phi}_i$. The total signal-to-noise ratio, conditioned on $\bar{\phi}_i$, is approximately given by

$$SNR(\bar{\phi}_i) \triangleq \frac{1}{2}\, \frac{S_T^2}{\sigma_T^2 + \sigma_J^2} = U\left\{\left[H\frac{E_b}{N_0}\right]^{-1} \sum_{m=-M}^{M} \alpha_m^2 + \frac{1}{H} + (1+\frac{1}{H})\frac{[1+\zeta(\gamma)]KL - L}{3N} \sum_{m=-M}^{M} (2\alpha_m^2 + \alpha_m \alpha_{m+1}) + \frac{J/S}{N^2}\cdot\frac{2}{U}\sum_{l=1}^{U}\left[\sum_{m_1=-M}^{M}\sum_{m_2=-M}^{M} \alpha_{m_1}\alpha_{m_2} \sum_{j_1=0}^{N-1}\sum_{j_2=0}^{N-1} \cdot\beta(j_1, m_1, \hat{\theta}_l)\beta(j_2, m_2, \hat{\theta}_l) \cdot\delta(j_1+m_1, j_2+m_2)\delta([pj_1],[pj_2])\right]\right\}^{-1}, \tag{51}$$

where $\bar{E}_b = 2P\rho T_b$ is the average energy per coded bit, $J/S = J/(PA^2)$ denotes the interference power-to -useful signal power ratio, and $H = A^2/(2\rho)$.

B. Channel Coding

We are interested in the performance of two simple block codes used in conjunction with the diversity. These are the BCH (15,7) code and the Golay (23,12) code. The former corrects two channel errors, while the latter corrects the three channel errors in each codeword. Assuming perfect interleaving, (i.e., ignoring the correlation introduced by both the suppression filter and the fading channel), the resulting bit error rate (BER) after hard decision decoding is approximately given by [13]

$$P_b \approx \frac{1}{n} \sum_{i=t+1}^{n} i \binom{n}{i} (P_e)^i (1-P_e)^{n-i}, \tag{52}$$

where n is the number of bits in a codeword and t denotes the number of errors that the code can correct.

C. Determination of the Suppression (Wiener) Filter Coefficients

The coefficients of the suppression filter can be determined by solving the following Wiener-Hopf equation:

$$\sum_{\substack{m=-M \\ m\neq 0}}^{M} \alpha_m \hat{\rho}[(n-m)T_c] + \hat{\rho}(nT_c) = 0, \tag{53}$$

where $n=-M, \cdots, M, n \neq 0$, and $\hat{\rho}(\cdot)$ is the low-pass version (normalized to the received useful power PA^2 of the reference user) of the autocorrelation function of the input signal to the suppression filter and is given by

$$\hat{\rho}(lT_c) \equiv \begin{cases} (1+1/H)[1+\zeta(\gamma)]KL+2N[\bar{E}_b/N_0]^{-1}+J/S, & l=0 \\ (J/S)(1-|l|p)\cos(2\pi lq), & |l|\leq[1/p] \\ 0, & |l|\geq[1/p] \end{cases} \quad (54)$$

where p and q are given by (6) and (7), respectively.

IV. NUMERICAL RESULTS

In this section, we present numerical results for the BER of a CDMA system in the presence of the narrowband interference. Unless noted otherwise, it is assumed that the ratio of the interference bandwidth to the spread spectrum bandwidth is 10% (p=0.1), the ratio of the offset of the interference carrier frequency to half of the spread spectrum bandwidth is 20% (q=0.2), the ratio of the specular component power to the fading component power, H, is 7 dB, and the processing gain, N, is 255. Note that where the BER is plotted as a function of the ratio $\bar{E}_i/N_0$, this corresponds to the energy-per -bit to noise spectral density of an uncoded system.

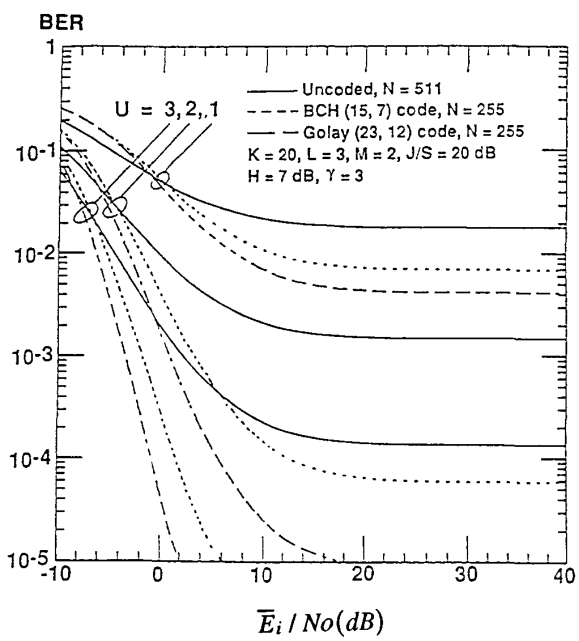

Fig. 4. The BER performance of DS-CDMA overlay system with diversity plus coding.

In Fig. 4, the decoded BER for H=7 dB is shown. Also, the uncoded BER is plotted in order to allow for comparison. Note that the uncoded and coded systems are compared for the same spread spectrum bandwidth $B=2N/(T_i m/n)$. In our example, N=255 for the coded systems, whereas N=511 for the uncoded system. As expected, both the uncoded and coded BER decreases as the order, U, of diversity increases. The performance advantage obtained by using diversity becomes larger as the error correcting capability of the code increases for the same spread bandwidth.

In order to investigate the relative effect of the narrowband interference on the performance of the CDMA system with respect to the multiple-access and adjacent-cell interference, Fig. 5 illustrates the asymptotic $(\bar{E}_i/N_0 \rightarrow \infty)$

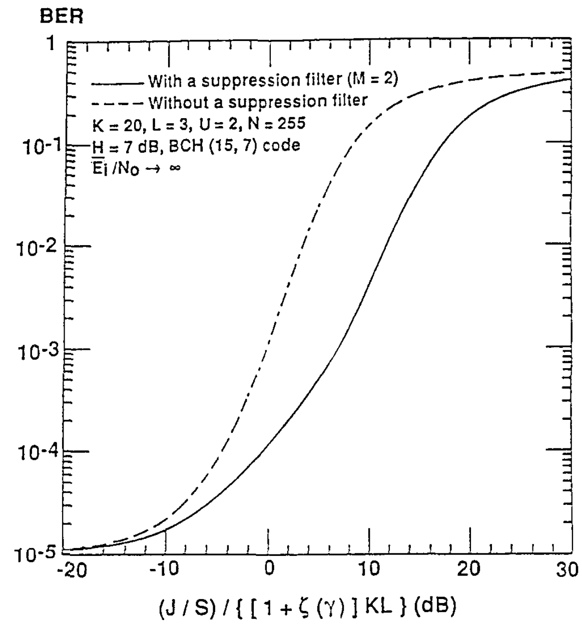

Fig. 5. The asymptotic $(\bar{E}_i/N_0 \rightarrow \infty)$ decoded BER of DS-CDMA overlay system as a function of $(J/S)/[(1+\zeta(\gamma))KL]$.

BER of the CDMA system both with or without the suppression filter as a function of $(J/S)/[(1+\zeta(\gamma))KL]$. Note that the ratio of the narrowband interference to the multiple-access and adjacent-cell interference can be roughly represented by $(J/S)/[(1+\zeta(\gamma))KL]$. As expected, the BERs with or without suppression filters increase as J/S increases. When J/S is small, the multiple-access and adjacent-cell interference dominates; in this case, the BERs of the systems both with and without a suppression filter are very close, so that the suppression filter is unnecessary. When J/S is large, the narrowband interference is on the order of the multiple-access and adjacent-cell interference; in this case, the system without a suppression filter degrades significantly, whereas the system is much more tolerant of the interference when the suppression filter is present.

In Fig. 6, the asymptotic BER of the system with and without the suppression filter is shown as a function of the ratio (p) of the interference bandwidth to the spread spectrum bandwidth. It is seen from the figure that the CDMA system with a double-sided Wiener filter can suppress the narrowband interference very effectively (assuming, of course, that p does not become too large).

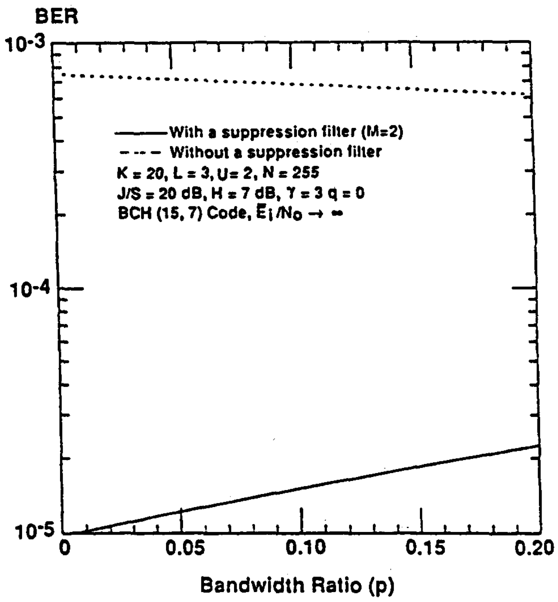

Fig. 6. The asymptotic ($\bar{E}_i/N_0 \leftarrow \infty$) BER of DS-CDMA system with and without suppression filters as a function of the bandwidth ratio (p) of the interference BW to the spread spectrum BW.

Fig. 7 illustrates the asymptotic BER for the system employing the suppression filter as a function of the number of taps on each side, M. It is seen from the figure that the BER decreases first when $M = 1$ and then keeps a steady value with increasing M. For $M \geq 1$, the narrow-band interference is reduced by the suppression filter, and increasing M beyond 1 does not noticeably reduce the narrowband interfer-

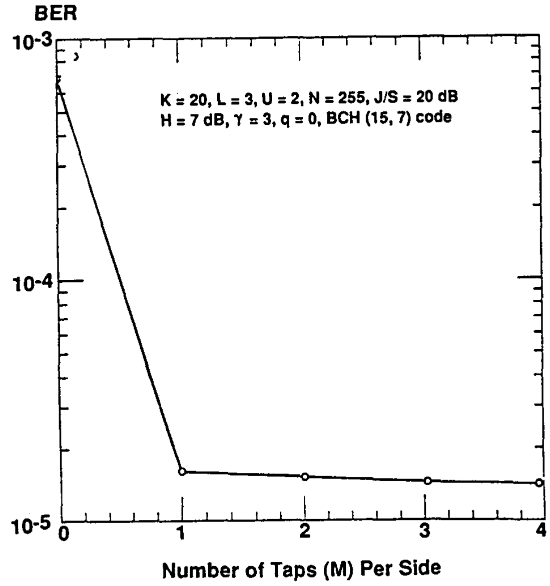

Fig. 7. The asymptotic ($\bar{E}_i/N_0 \to \infty$) BER of DS-CDMA systems with a suppression filter as a function of number of taps on each side.

Table II (K=20, L=3 U=2, N=255, M=2, q=0, H=7 dB, γ=3, J/S=20 dB, BCH (15, 7) code)

Number of taps per side (M)	Coefficients (α_m)									Q	W
	α_{-4}	α_{-3}	α_{-2}	α_{-1}	α_0	α_1	α_2	α_3	α_4		
0	0	0	0	0	1	0	0	0	0	0.379	0.36
1	0	0	0	-0.302	1	-0.302	0	0	0	0.079	0.33
2	0	0	-0.165	-0.213	1	-0.213	-0.165	0	0	0.056	0.35
3	0	-0.1	-0.131	-0.185	1	-0.185	-0.131	-0.1	0	0.053	0.35
4	-0.063	-0.083	-0.119	-0.174	1	-0.174	-0.119	-0.083	-0.063	0.054	0.35

$$Q = \frac{2J/S}{N^2} \sum_{m_1=-M}^{M} \sum_{m_2=-M}^{M} \alpha_{m_1} \alpha_{m_2} \sum_{j_1=0}^{N-1} \sum_{j_2=0}^{N-1} E_{\Phi_l} \left[\beta(j_1, m_1, \hat{\theta}_l)\, \beta(j_2, m_2, \hat{\theta}_l)\right] \delta(j_1+m_1, j_2+m_2)\, \delta\left\{[pj_1], [pj_2]\right\}$$

$$W = (1+H^{-1}) \frac{[1+\zeta(\gamma)]\, KL-L}{3N} \sum_{m=-M}^{M} \alpha_m (2\alpha_m + \alpha_{m+1})$$

Table II. Tap Coefficients and Their Effects on the Interference

ence, because $\hat{\rho}(T_c)$ and $\hat{\rho}(-T_c)$ are almost identical to $\hat{\rho}(0)$ in (54). Thus, the suppression filter with M=1 is sufficient to reject the interference in this case (small p and q=0), and more taps than one on each side are unnecessary. Table II shows the tap coefficients of the suppression filters for different numbers of taps each side ($M = 0, 1, 2, 3$). As shown in Table II, for $M = 0$, $Q \approx W$, where Q roughly represents the effect of the narrowband interference and W represents the effect of the multiple access interference; in this case, both the narrowband interference and multiple-access interference have comparable effects. However, for $M \geq 1$, $Q \ll W$ and W does not vary very much. Therefore, we conclude that when the carrier frequencies of the narrowband interference and CDMA signals are the same, the suppression filter with three total taps is sufficient to reduce the narrowband interference, and increasing the number of total taps beyond three is not necessary.

Fig. 8 illustrates the BER performance of the systems with the suppression filter as a function of the ratio (q) of the

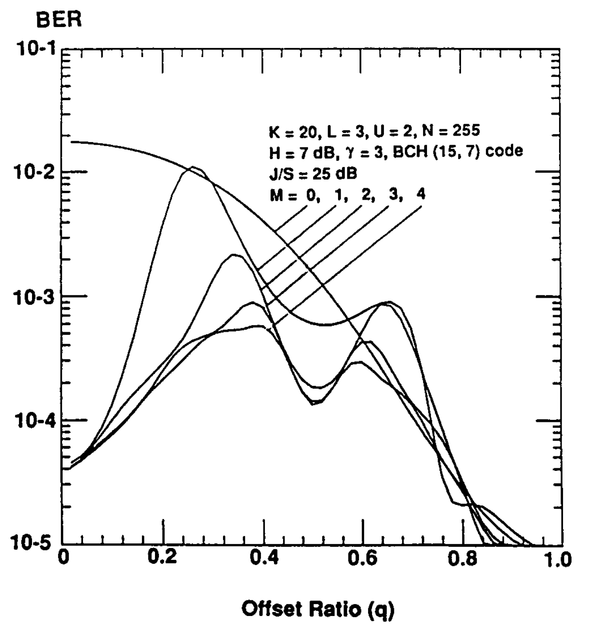

Fig. 8. The asymptotic ($\bar{E}_i/N_0 \to \infty$) BER of the system with double-sided filters as a function of the offset ratio (q) of the interference carrier frequency to the half spread spectrum BW.

offset of the interference carrier frequency to the half spread spectrum bandwidth for different number of the taps on each side. It is seen that when q is very small, the BERs of the system with double-sided filters for $M \geq 1$ are almost identical, as was the case in Fig. 7. However, the BER performance improves as M increases when q increases, and the BER for large M is also more robust to a change in q. Also, for $q \approx 0.25$, the BERs of the systems with the double-sided suppression filter and without the suppression filter are almost identical, because the low-pass version of the autocorrelation of the interference is zero, (i.e., $\rho_j(lT_c) = 0$ for $l \neq 0$), so that all off-center tap coefficients are zero.

In Fig. 9, the asymptotic BER of the DS-CDMA system with a suppression filter is plotted as a function of the number

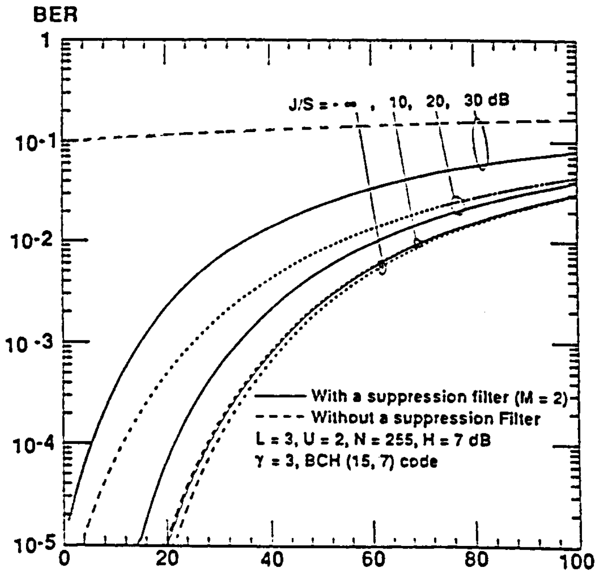

Fig. 9. The asymptotic decoded BER of DS-CDMA overlay systems as a function of the number of active users per cell.

of active users, K, for various values of J/S. Also, the BER of the system without the suppression filter is shown for comparison. It is seen that at large J/S, for a given BER, the system with the suppression filter can support many more users

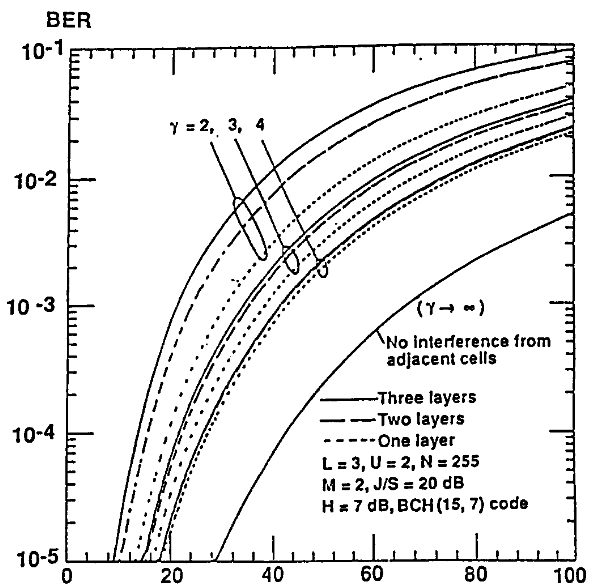

Fig. 10. The asymptotic decoded BER of DS-CDMA overlay system as a function of active users (K) per cell for various values of the propagation loss exponent (γ).

than can the system without the suppression filter. When the interference is such that $J/S \leq 10dB$, the multiple-access and adjacent-cell interference dominates so that the suppression filter is not needed, as was seen in Fig. 5.

Fig. 10 illustrates the asymptotic BER of the CDMA system versus K, the number of active users per cell, for various values of the propagation loss exponent, γ, (γ=2,3,4). Also, the BER in the absence of adjacent cell interference (*i.e.*,$\gamma \to \infty$) is shown for comparison. In order to show the effect of the adjacent-cell interference on the performance of the system for different values of γ, a single layer of cells, two layers of cells and three layers of cells are considered. It is clear from this figure that when γ =2, the BER computed by accounting for only a single layer of cells is too optimistic; alternately, if $\gamma \geq 3$, the difference in BER performance between accounting for only a single layer and accounting for more layers is insignificant. Further, for a given γ, the gap in the BER between two and three layers is much less than that between one and two layers. Indeed, these results can be seen from the $\zeta_j(\gamma)$ entries of Table I, and are consistent with the results presented in [6].

V. CONCLUSIONS

An approximate technique has been described for determining the performance of a DS-CDMA system operating over a multipath Rician fading channel and sharing common spectrum with various narrowband waveforms. It was shown that

(1) When the narrowband interference power-to-signal power ratio (J/S) is small with respect to the multiple-access and adjacent-cell interference, the suppression filter is unnecessary, whereas when J/S is on the order of the multiple-access and adjacent-cell interference, the suppression filter provides significant enhancement in performance;

(2) When the ratio (q) of the offset of the interference carrier frequency to half of the spread spectrum bandwidth is very small, that is, when the interference carrier frequency is very close to the spread spectrum carrier frequency, spread spectrum systems with double-sided suppression filters only need three total taps. However, the double-sided filter with a large number of taps (i.e., $M>1$) is preferable when the ratio is large.

(3) For large narrowband interference, the system with a suppression filter can support many more users than can the system without a suppression filter;

(4) As is well known, the number of users supported by the CDMA overlay system increases as the propagation loss exponent (γ) increases. For typical exponents, three layers of cells are sufficient for considering the adjacent cell interference.

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