Multicarrier DS/SFH-CDMA Systems

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Abstract—In this paper, multicarrier direct-sequence/slow-frequency-hopping (MC DS/SFH) code-division multiple-access (CDMA) systems are proposed, in which multiple carriers are modulated by the same DS waveform and hopped in frequency according to a random hopping pattern. The receiver dehops the received signal with the same pattern, provides RAKE receivers for each carrier, and combines the outputs with a maximal ratio combiner (MRC). The performance of the proposed system is investigated over a frequency-selective Rayleigh-fading channel and compared to that of the MC DS-CDMA systems. It is shown that for the same diversity order, the MC DS/SFH-CDMA systems are superior in reducing multiple-access interference (MAI) while preserving the good capability of narrow-band interference suppression, when the system parameters are selected properly.

Index Terms—Code-division multiple access (CDMA), multicarrier (MC) systems, narrow-band interference suppression.

I. INTRODUCTION

IN ORDER to increase the spectral efficiency of wireless communications, it has been proposed to overlay a spread spectrum (SS) system on existing narrow-band systems [1], [2]. The SS signal must be transmitted at a much lower power level to avoid adversely affecting the narrow-band systems, which in turn may raise significant interference to the SS systems. A pure direct-sequence (DS) system without suppression filters usually does not work well in this situation.

Multicarrier (MC) code-division multiple-access (CDMA) systems are ingenious in suppressing narrow-band interference [3]–[7], where the entire frequency band is subdivided into multiple frequency slots (subchannels) of equal width. A replica of a DS signal is transmitted in each subchannel and the outputs of all subchannels are combined in a way that the distorted one contributes less to the final decision. Since the interference concerned is narrow-band (i.e., it does not overlay all of the subchannels), the decision can be made on the basis of the unaffected subchannels and interference suppression is achieved. In addition, the MC systems require a lower chip rate, which reduces the implementation complexity.

An adaptive algorithm has been proposed [4] to exploit the correlation between the received signals on different carriers to cancel multiple-access interference (MAI) and make the MC DS system robust to the near–far problems. However, the algorithm usually needs 10 to 20 symbol intervals to converge. An alternative resort is the frequency hopping (FH) technique [8], [9], in which the signal occupies only one subchannel at any interval and hops in frequency. By reducing the number of simultaneous

cochannel users, FH systems have a natural guard against the near–far problem. However, their performance in the presence of narrow-band interference is not satisfactory without coding [10].

The systems combining these two techniques, called multicarrier direct-sequence/slow-frequency-hopping CDMA (MC DS/SFH-CDMA) systems, in which multiple subchannels are used to transmit the same information during each hop period, are expected to show good capability in combating both narrow-band interference and MAI. In this paper, the uplink performance of the proposed system is studied in a multipath Rayleigh-fading channel and compared to that of MC DS-CDMA systems, given the same information rate, the same diversity order, and same system bandwidth. One possible application of this proposed scheme is ultra-wide-band communications with bandwidth as large as 3 GHz.

The remainder of the paper is organized as follows. In Section II, the system models are described. In Section III, the performance of the MC DS/SFH-CDMA systems in terms of average bit-error rate (BER) is analyzed. Numerical results are given in Section IV. Finally, the conclusions are presented in Section V.

II. SYSTEM MODELS

In an MC DS/SFH system, the total system bandwidth B_T is divided into $S = MQ$ contiguous subchannels, where M is the number of simultaneous carriers at any instant and Q is the number of available hopping frequencies. Each subchannel has a bandwidth $B_s = B_T/S$, and the central frequency of the sth subchannel is denoted as f_s ($s = q + mQ$, $q = 1, \ldots, Q$, and $m = 0, \ldots, M - 1$). The transmitter for the kth user is shown in Fig. 1, where $b_k(t)$ is the data sequence with bit duration T and bit rate $R_b = 1/T$, $a_k(t)$ is the random signature sequence with chip duration $T_c = T/N$, where N is the processing gain of one subchannel. The bit and chip waveforms are all rectangular, and the spread signal is band-limited by its null-to-null bandwidth $2/T_c$, which is equal to B_s. The total processing gain is $N_T = B_T/(2/T) = SN$. The hopping pattern $f_k(t)$ takes values randomly from $\{f_1, f_2, \ldots, f_Q\}$ at any hop interval, and the frequency separation between the nearest two carriers $B_H = QB_s$. Therefore, if two users take different hopping frequencies, there is no overlap between their transmitted signals in frequency.

The transmitted signal of the kth user can be written as

$$S_k(t) = \mathrm{Re}\left[\sqrt{2P} \sum_{m=0}^{M-1} C_{km}(t)\right] \quad \text{(1a)}$$

where P is the average transmitted power of each carrier and

$$C_{km}(t) = b_k(t)a_k(t)\exp\{j2\pi[f_k(t)+mB_H]t+j\theta_{k,m}\} \quad \text{(1b)}$$

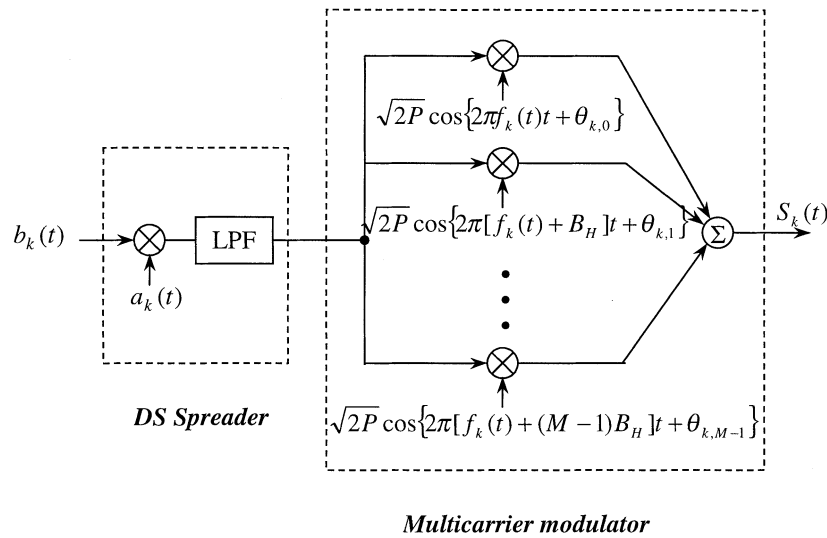where $\theta_{k,m}$ is the random phase introduced.

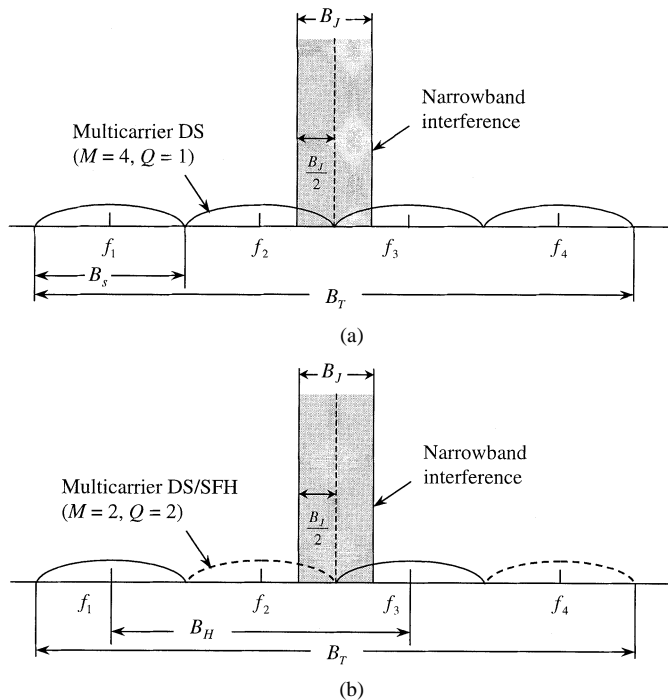**Fig. 1.** The transmitter structure of the MC DS/SFH system.

Fig. 2. Power spectrum densities of transmitted signals and the narrow-band interference.

Note that $M = S$ and $Q = 1$ corresponds to the MC DS system proposed in [3]. The power spectrum densities of the transmitted signals of the MC DS and the MC DS/SFH system are shown in Fig. 2, where the number of total subchannels $S = 4$. Each carrier has the same transmitted power in both the MC DS and MC DS/SFH systems. Therefore, the total transmitted power of the MC DS/SFH system is only $1/Q$ of that of the MC DS system, due to a smaller number of carriers. At the receiver end, multiple antennas are needed to ensure the same order of diversity and the same total received energy per bit for the MC DS/SFH system.

The mobile radio channel is assumed to be frequency selective fading with delay spread of T_m. The sth subchannel for the kth user and the vth antenna can be well approximated by a

tapped-delay-line model [10] with complex low-pass equivalent impulse response given by

$$h_k^{(s,\,v)}(t) = \sum_{l=0}^{L-1} \beta_{kl}^{(s,\,v)} \exp\left[j\mu_{kl}^{(s,\,v)}\right] \delta(t - lT_c - \tau_k) \quad (2)$$

where L is the number of resolvable paths and given by

$$L = \lfloor T_m/T_c \rfloor + 1 = \lfloor T_m B_T/(2S) \rfloor + 1.$$

Letting $L_T = \lfloor T_m B_T/2 \rfloor + 1$ denote the number of resolvable paths (also the order of diversity available) in a pure DS system, one obtains

$$L = \begin{cases} \lfloor (L_T - 1)/S \rfloor + 1 \approx L_T/S, & S \leq L_T \\ 1, & S > L_T. \end{cases} \quad (3)$$

The fading gains $\{\beta_{kl}^{(s,\,v)}\}$ are assumed to be identically Rayleigh-distributed and mutually independent, provided that B_s is wider than the channel coherence bandwidth $(1/T_m)$. It is assumed that the delay power profile is uniform and perfect power control is achieved, i.e., $E[(\beta_{kl}^{(s,\,v)})^2] = 2\rho$ for all l and k. The phases $\{\mu_{kl}^{(s,\,v)}\}$ are assumed to be uniformly and independently distributed in $[-\pi,\,\pi)$. τ_k is the relative delay of the first path signal of the kth user and is randomly distributed in $[0,\,T - (L-1)T_c)$.

The narrow-band interference is characterized as band-limited white Gaussian noise with the double-sided power spectral density $\eta_J/2$ and bandwidth B_J (see Fig. 2). For simple analysis, it is assumed that the interference is located at the center of the system bandwidth. The ratio of the interference bandwidth to the system bandwidth is defined as

$$p = \frac{B_J}{B_T} \quad (4)$$

and the ratio of the interference power to the signal power is defined as

$$w = \frac{P_J}{P_s} = \frac{\eta_J B_J}{E_b/T}, \quad (5)$$

where E_b is the average total received signal energy per bit.

Considering the case when $B_J \leq B_s$ and S is even, the interference is limited in the two middle subchannels and causes
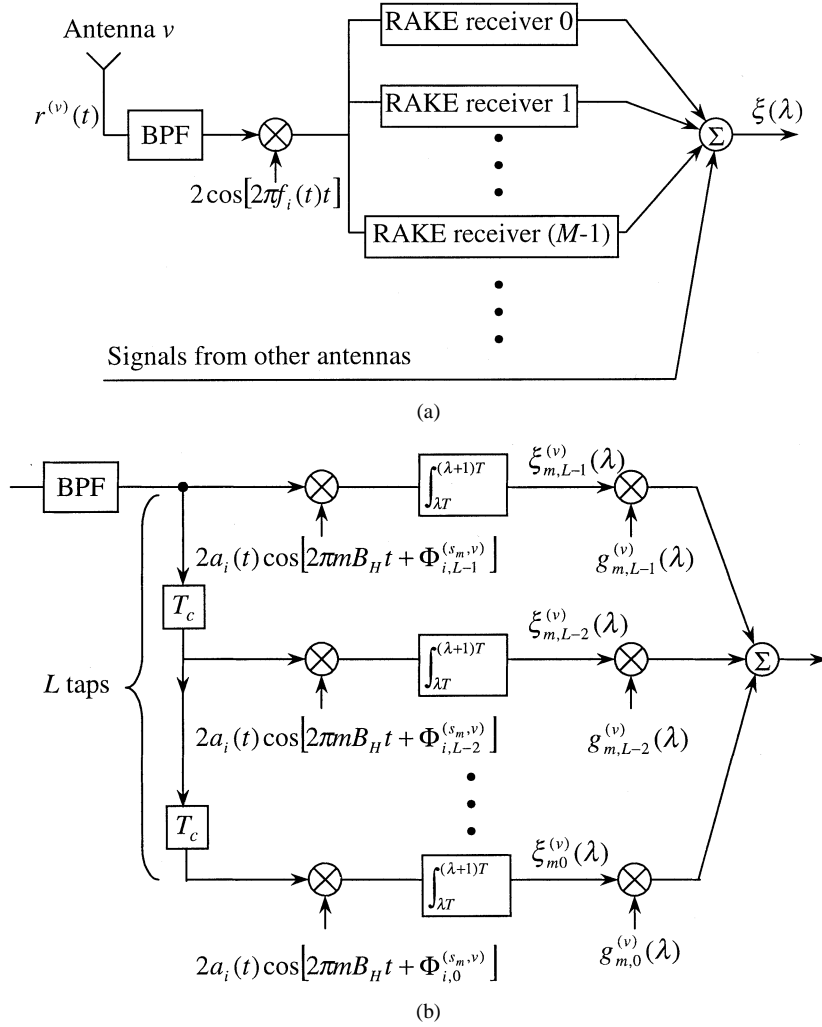
Fig. 3. The receiver structures (a) MC DS/SFH system and (b) detailed block diagram of RAKE receiver #m for user i.

no impairment to others. Assuming that each subchannel employs an ideal bandpass filter, the interference can be viewed as two independent interferers with bandwidth $B_J/2$, one in each of the two subchannels. The interference in the sth subchannel received by the vth antenna $J^{(s,v)}(t)$ can then be written as

$$J^{(s,v)}(t) = \Delta(s) \cdot \mathrm{Re} \left\{ \left[J_I^{(v)}(t) + j J_Q^{(v)}(t) \right] \right.$$
$$\left. \cdot \exp\{j 2\pi [f_s \pm (B_s/2 - B_J/4)]t\} \right\} \quad (6a)$$

where

$$\Delta(s) = [\delta(s, S/2) + \delta(s, S/2+1)] \quad (6b)$$

$\pm$ takes $+$ for $s = S/2$ and $-$ for $s = S/2+1$, respectively. $\delta(x, y) = 1$ and 0 for $x = y$ and $x \neq y$, respectively. Both $J_I^{(v)}(t)$ and $J_Q^{(v)}(t)$ are independent lowpass processes with

$$E[(J_I^{(v)}(t))^2] = E[(J_Q^{(v)}(t))^2] = P_J/2.$$

The autocorrelation function of $J^{(s,v)}(t)$ is given by

$$R_J^{(s,v)}(\tau) = \Delta(s) P_J/2 \cdot \mathrm{sinc}(\pi B_J \tau/2)$$
$$\cdot \cos\{2\pi [f_s \pm (B_s/2 - B_J/4)]\tau\} \quad (7)$$

where $\mathrm{sinc}(x) = \sin(x)/x$.

The received signal by antenna v can be written as [the local clock is synchronized with the $(L-1)$th path signal]

$$r^{(v)}(t)$$
$$= \mathrm{Re} \left[\sqrt{2P} \sum_{k=1}^{K} \sum_{m=0}^{M-1} C_{km}(t) * h_k^{(s_{km}, v)}[t + (L-1)T_c] \right]$$
$$+ n^{(v)}(t) + \sum_{s=1}^{S} J^{(s,v)}(t)$$
$$= \mathrm{Re} \left\{ \sqrt{2P} \sum_{k=1}^{K} \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} \beta_{kl}^{(s_{km}, v)} \right.$$
$$\left. \cdot \exp\left[j \mu_{kl}^{(s_{km}, v)} \right] C_{km}[t - lT_c - \tau_k + (L-1)T_c] \right\}$$
$$+ n^{(v)}(t) + \sum_{s=1}^{S} J^{(s,v)}(t) \quad (8)$$

where K stands for the number of users, s_{km} is the subchannel occupied by the mth carrier of the kth user. $n^{(v)}(t)$ is the additive white Gaussian noise (AWGN) with double-sided spectral density $N_0/2$. The receiver for the reference user (the ith user) is shown in Fig. 3(a). The received signal is first dehopped according to the hopping pattern of the reference user, and the

output is fed to M RAKE receivers. The outputs of all RAKE receivers are combined to obtain the final decision variable.

The detailed structure of the mth RAKE receiver is shown in Fig. 3(b). A single delay line is employed through which the input signal is passed. The lth resolvable path signal is collected by the lth tap, which correlates the delayed input signal with the spreading sequence. The correlator outputs are combined by a maximal ratio combiner. For simplicity, we do not consider the cross correlations between the delayed code sequences. Consequently, a RAKE receiver with perfect estimates of the tap gains is equivalent to a maximal ratio combiner with Lth-order diversity. When V antennas are employed and M RAKE receivers for each of them, the whole receiver is equivalent to a maximal ratio combiner with diversity order equal to VML. Note that $ML = L_T/Q$ when $S \leq L_T$. If we choose $V = Q$, all systems have the same diversity order equal to L_T.

III. PERFORMANCE ANALYSIS

In this paper, we aim to compare the capabilities of different systems in suppressing MAI and narrow-band interference. Perfect tracking of all path signals of the reference user is assumed, as well as perfect power control and channel estimation. Therefore, the performance presented in what follows should be considered as the lower bounds of the practical systems.

In the presence of narrow-band interference, the performance of an SFH system depends on the instantaneous value of its hopping frequency. Conditioned on $f_i(\lambda T) = f_q$, the output of the correlator of the lth tap of the mth RAKE receiver in the λth symbol interval is given by

$$\xi_{ml}^{(v)}(\lambda) = \int_{\lambda T}^{(\lambda+1)T} 2r^{(v)}[t - (L-1-l)T_c] \\ \cdot \cos\{2\pi f_q[t - (L-1-l)T_c]\} \\ \cdot 2a_i(t)\cos\left[2\pi m B_H t + \Phi_{i,l}^{(s_m, v)}\right] dt \quad (9)$$

where $s_m = q + mQ$ denotes the subchannel occupied by the mth carrier of the ith user's signal ($f_{s_m} = f_q + mB_H$) and

$$\Phi_{i,l}^{(s_m, v)} = \theta_{i,m} + \mu_{il}^{(s_m, v)} + 2\pi f_q(L-1-l)T_c.$$

Neglecting the high-frequency terms, (9) can be written as

$$\xi_{ml}^{(v)}(\lambda) = D_{ml}^{(v)} + N_{ml}^{(v)} + J_{ml}^{(v)} + P_{ml}^{(v)} + I_{ml}^{(v)} \quad (10)$$

where

$$D_{ml}^{(v)} = \sqrt{2P}T\beta_{il}^{(s_m, v)}b_i^{(\lambda)}$$

is the desired signal component, where $b_i^{(\lambda)}$ stands for the λth bit of the reference user with $b_i^{(\lambda)} = 1$ or -1.

$N_{ml}^{(v)}$ is a Gaussian random variable with zero mean and variance equal to $N_0 T$.

$J_{ml}^{(v)}$ is due to the narrow-band interference in the s_mth subchannel and can be written as

$$J_{ml}^{(v)} = \int_{\lambda T}^{(\lambda+1)T} J^{(s_m, v)}[t - (L-1-l)T_c]2 \\ \cdot \cos[2\pi f_q[t - (L-1-l)T_c]] \\ \cdot 2a_i(t)\cos\left[2\pi m B_H t + \Phi_{i,l}^{(s_m, v)}\right] dt. \quad (11)$$

Since $J^{(s_m, v)}(t)$ is a Gaussian process, $J_{ml}^{(v)}$ is also a zero-mean Gaussian random variable. Its variance $\sigma_J^2(s_m)$, conditioned on s_m, is derived in Appendix A as

$$\sigma_J^2(s_m) = \sigma_J^2(q + mQ) = \Delta(q + mQ)\frac{P_J T^2}{N}\hat{\sigma}_J^2 \quad (12a)$$

where $\Delta(s)$ is given by (6b), P_J is the interference power, and $\hat{\sigma}_J^2$ is given by

$$\hat{\sigma}_J^2 = \int_0^1 \sin c(\pi p S\tau)\cos[2\pi(1 - pS/2)\tau](1-\tau)(1-\tau/N)\,d\tau \quad (12b)$$

where p is given by (4) and S is the number of subchannels.

$P_{ml}^{(v)}$ ($l' \neq l$) is the total multipath interference term from the reference user. Since the number of resolvable paths in MC systems has been greatly reduced, as the result of lower chip rate, $P_{ml}^{(v)}$ is insignificant and can be neglected for simple analysis.

$I_{ml}^{(v)} = \sum_{k=1}^{K_h}\sum_{l'=0}^{L-1} I_{kl'}^{(m, v)}$ is the total MAI term, where K_h is the number of nonreference users that collide with the reference user in frequency ($f_k(\lambda T) = f_q$). Assuming the hopping patterns of different users are mutually independent, K_h has a binomial distribution with probability density function (pdf) given by

$$P_{K_h}(K_h) = \binom{K-1}{K_h}(P_h)^{K_h}(1 - P_h)^{K-K_h-1} \quad (13)$$

where P_h is the probability that a nonreference user hits the reference user and approximates to $1/Q$ for SFH [8]. Note that (13) also applies to the pure DS and the MC DS systems, i.e., $P_{K_h}(K_h) = \delta(K_h, K - 1)$, due to $P_h = 1$ for these two systems.

$I_{kl'}^{(m, v)}$ is due to the l'th path signal of the kth user ($k \neq i$) and is given by

$$I_{kl'}^{(m, v)} = \sqrt{2P}\beta_{kl'}^{(s_m, v)}\cos\left[\Phi_{k,l'}^{(s_m, v)} - \Phi_{i,l}^{(s_m, v)}\right] \\ \cdot \int_{\lambda T}^{(\lambda+1)T} b_k\left(t - \Gamma_{l'l}^k\right)a_k\left(t - \Gamma_{l'l}^k\right)a_i(t)\,dt \\ = \sqrt{2P}\beta_{kl'}^{(s_m, v)}\cos\left[\Phi_{k,l'}^{(s_m, v)} - \Phi_{i,l}^{(s_m, v)}\right] \\ \cdot \left[b_k^{(\lambda-1)}R_{ki}\left(\Gamma_{l'l}^k\right) + b_k^{(\lambda)}\hat{R}_{ki}\left(\Gamma_{l'l}^k\right)\right] \quad (14)$$

where

$$\Phi_{k,l'}^{(s_m, v)} = \theta_{k,m} + \mu_{kl'}^{(s_m, v)} - 2\pi f_{s_m}\Gamma_{l'l}^k + 2\pi f_q(L-1-l)T_c$$

and $\Gamma_{l'l}^k = \tau_k + \{l' - l\}T_c$. $b_k^{(\lambda-1)}$ and $b_k^{(\lambda)}$ stand for the $(\lambda-1)$th and the λth bit of the kth user, respectively, which are assumed to be uncorrelated and take values randomly from $\{+1, -1\}$. The partial cross-correlation functions $R_{ki}(\tau)$ and $\hat{R}_{ki}(\tau)$ are defined as

$$R_{ki}(\tau) = \int_0^\tau a_k(t - \tau)a_i(t)\,dt$$

$$\hat{R}_{ki}(\tau) = \int_\tau^T a_k(t - \tau)a_i(t)\,dt. \quad (15)$$

It is easy to prove that $\{I_{kl'}^{(m, v)}\}$ are identically distributed but uncorrelated for different k and l. When $K_h L$ is large, one can

use the Gaussian approximation [11], and treat $I_{ml}^{(v)}$ as a random Gaussian variable with variance, conditioned on K_h, given by

$$\sigma_I^2(K_h) = \sum_{k=1}^{K_h} \sum_{l'=0}^{L-1} P \cdot E\left[\left(\beta_{kl'}^{(s_m, v)}\right)^2\right]$$
$$\cdot E\left[R_{ki}^2\left(\Gamma_{l'l}^k\right) + \hat{R}_{ki}^2\left(\Gamma_{l'l}^k\right)\right]$$
$$= 4K_h L P \rho T^2 / (3N) \tag{16}$$

where

$$E[R_{ki}^2(\Gamma_{l'l}^k)] = E[\hat{R}_{ki}^2(\Gamma_{l'l}^k)] = T^2/(3N).$$

The signal-to-noise ratio (SNR) of $\xi_{ml}^{(v)}(\lambda)$, conditioned on $\beta_{il}^{(s_m, v)}$, K_h, and q is then defined as

$$\gamma_{ml}^{(v)}\Big|_{\beta_{il}^{(s_m, v)}, K_h, q}$$
$$= \frac{E^2\left[\xi_{ml}^{(v)}(\lambda) \Big| \beta_{il}^{(s_m, v)}\right]}{2 \operatorname{var}\left[\xi_{ml}^{(v)}(\lambda) \Big| K_h, q\right]}$$
$$= \frac{2PT^2\left(\beta_{il}^{(s_m, v)}\right)^2}{2\left[\sigma_I^2(K_h) + \sigma_J^2(q + mQ) + \sigma_N^2\right]}$$
$$= \frac{\left(\beta_{il}^{(s_m, v)}\right)^2}{2\rho}\left[2K_h L/(3N) + \Delta(q + mQ)\frac{P_J T}{E_c}\frac{\hat{\sigma}_J^2}{N}\right.$$
$$\left. + (E_c/N_0)^{-1}\right]^{-1} \tag{17}$$

where $E_c = 2\rho PT$ is the average received signal energy per diversity branch.

By definition, $\{N_{ml}^{(v)}\}$ are independent and identically distributed (i.i.d.). It is also easy to verify that $\{I_{ml}^{(v)}\}$ and $\{J_{ml}^{(v)}\}$ are mutually uncorrelated. Therefore, $\{\xi_{ml}^{(v)}(\lambda)\}$ are uncorrelated for different $m, l,$ and v. To maximize the SNR of the final decision variable, each $\xi_{ml}^{(v)}(\lambda)$ is weighted by a tap gain [3]

$$g_{m, l}^{(v)}(\lambda) = \frac{E\left[b_i^{(\lambda)}\xi_{ml}^{(v)}(\lambda) \Big| \beta_{il}^{(s_m, v)}\right]}{\operatorname{var}\left[\xi_{ml}^{(v)}(\lambda) \Big| K_h, q\right]} \tag{18}$$

and the final decision variable is given by

$$\xi(\lambda) = \sum_{v=1}^{V} \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} g_{m, l}^{(v)}(\lambda)\xi_{ml}^{(v)}(\lambda). \tag{19}$$

The instantaneous SNR of $\xi(\lambda)$, conditioned on $\{\beta_{il}^{(s_m, v)}\}$, K_h, and q is given by

$$\gamma_b\Big|_{\{\beta_{il}^{(s_m, v)}\}, K_h, q}$$
$$= \frac{E^2\left[\xi(\lambda) \Big| \{\beta_{il}^{(s_m, v)}\}\right]}{2 \operatorname{var}[\xi(\lambda)|K_h, q]}$$
$$= \sum_{v=1}^{V} \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} \gamma_{ml}^{(v)}\Big|_{\beta_{il}^{(s_m, v)}, K_h, q}$$
$$= \sum_{v=1}^{V} \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} \frac{\left(\beta_{il}^{(s_m v)}\right)^2}{2\rho}\{2K_h L/(3N)$$
$$+ D\left[\Delta(q + mQ)w\hat{\sigma}_J^2/N + (E_b/N_0)^{-1}\right]\}^{-1} \tag{20}$$

where $D = VML$ is the total order of diversity and $E_b = DE_c$ is the total received signal energy per bit, and w is the interference power to the signal power ratio defined in (5).

The conditional BER takes the simple form

$$P_e(\gamma_b) = \frac{1}{\sqrt{2\pi}} \int_{\sqrt{2\gamma_b}}^{\infty} e^{-x^2/2}\, dx. \tag{21}$$

To get the final average BER P_e, we first find the BER $P_e|_{K_h, q}$, conditioned on K_h and q, which can be obtained by averaging $P_e(\gamma_b)$ over the statistics of $\{\beta_{il}^{(s_m, v)}\}$. The detailed derivation of $P_e|_{K_h, q}$ is shown in Appendix B. Since all hopping frequencies appear with equal probability, the final average BER is given by

$$P_e = \frac{1}{Q} \sum_{q=1}^{Q}\left[\sum_{K_h=0}^{K-1} P_e|_{K_h, q} \cdot P_{K_h}(K_h)\right] \tag{22}$$

where $P_{K_h}(K_h)$ is the pdf of K_h and given by (13), and $P_e|_{K_h, q}$ is given by (B5) or (B8).

In case K_h is small, the Gaussian approximation may be not accurate enough to study the BER performance. However, small values of K_h appear with small hit probability and small conditional BER. Therefore, their contributions to the final average BER are not significant. In the numerical section, the results of Gaussian approximation are verified by Monte Carlo simulations, which reveal the applicability of the Gaussian approximation in various situations. The description of the simulation model is given in Appendix C.

IV. NUMERICAL RESULTS

In our comparison, all systems have the same order of diversity ($D = L_T = 4$), the same processing gain ($N_T = SN = 512$), and the same total average received energy ($E_b = D \cdot E_c$). Pseudonoise (PN) sequences applied in Monte Carlo simulations are Gold sequences of period 127, which are obtained from multiplying two primitive polynomials

$$h_1(x) = x^7 + x^3 + x^2 + x + 1 \quad \text{and} \quad h_2(x) = x^7 + x^3 + 1$$

respectively. The initial loadings are given in [12] to generate a class of codes known as auto-optimal with least sidelobe energy (AO/LSE).

To find the optimum channel division, the average asymptotic (without AWGN and narrow-band interference) SNR per diversity branch $E[\gamma_{ml}^{(v)}]$ is plotted for different subchannel numbers (S) in Fig. 4. From (17) one obtains that

$$E\left[\gamma_{ml}^{(v)}\right] = \frac{3N}{2(K-1)/Q \cdot L} = \begin{cases} \dfrac{3N_T}{2(K-1)L_T}Q, & S \leq L_T \\ \dfrac{3N_T}{2(K-1)}\dfrac{1}{M}, & S > L_T. \end{cases} \tag{23}$$

For the MC DS system ($M = S, Q = 1$), the SNR per diversity branch remains constant as long as $S \leq L_T$, but decreases after S exceeds L_T. For the hybrid DS/SFH system, the SNR increases with S before S reaches L_T, and remains constant afterwards. The SNRs of the MC DS/SFH systems fall between those of the MC DS and the hybrid DS/SFH systems. It can be seen from Fig. 4 that the average asymptotic SNRs per diversity branch of all systems reach their highest when $S = L_T$. Since
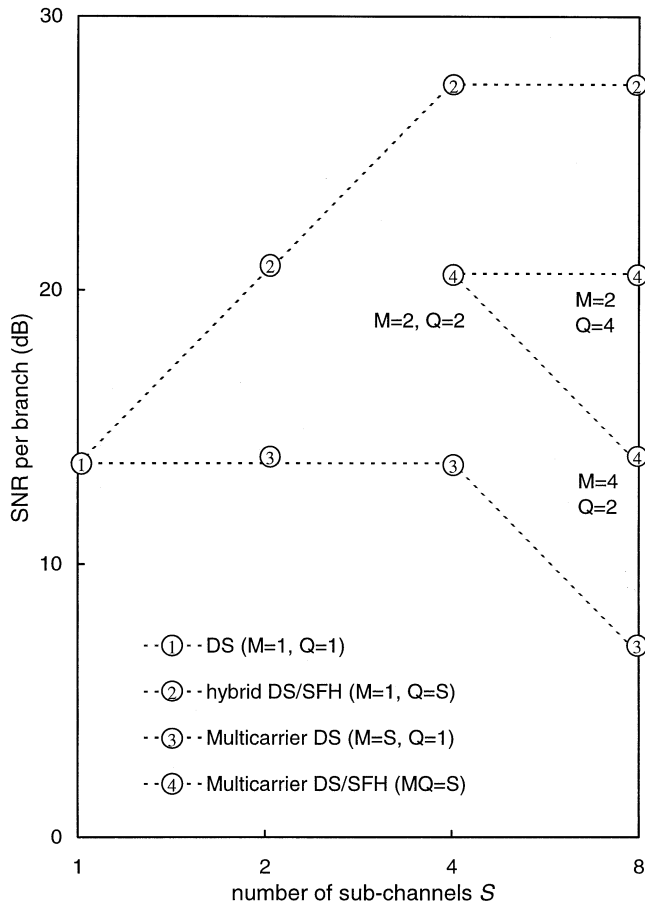
Fig. 4. Asymptotic SNR per branch with different number of subchannels S ($K = 50$, $N_T = 512$).

Fig. 5. BER performance with respect to E_b/N_0 in the absence of narrow-band interference ($K = 50$, $w = -\infty$ dB).

the conclusion above is not dependent on the values of K, it applies for all system loads.

In the following part, we set $S = L_T = 4$ and consider three systems: the MC DS/SFH ($M = 2$, $Q = 2$), the MC DS ($M = 4$, $Q = 1$) systems, and the hybrid DS/SFH system ($M = 1$, $Q = 4$). The number of antennas V is chosen to be equal to Q so that all systems have a diversity order of four. The BER performance of the three systems is compared along with that of the pure DS system ($S = 1$).

In the absence of narrow-band interference, $\overline{\gamma}_m|_{K_h, q}$ in (B3) are identical for all m. Therefore, $P_e|_{K_h, q}$ is given by (B5)

$$P_e|_{K_h, q} = \mu^4 \sum_{i=0}^{3} \binom{3+i}{i} (1 - \mu)^i \tag{24}$$

where $\mu = [1 - \sqrt{r_0/(1 + r_0)}]/2$, while r_0 is given by (B2) with $w = 0$

$$r_0 = \left[2K_h L/(3N) + 4(E_b/N_0)^{-1} \right]^{-1} \tag{25}$$

where $L = 4$ for pure DS system and $L = 1$ for other systems.

The BER of all the systems with respect to E_b/N_0 in the absence of narrow-band interference is shown in Fig. 5, obtained by both Gaussian approximation and Monte Carlo simulation. The Gaussian approximation is shown to be a little optimistic in comparison with the simulation results when E_b/N_0 is large. However, the differences are not significant. It is also shown that the Gaussian approximation is more accurate for systems with
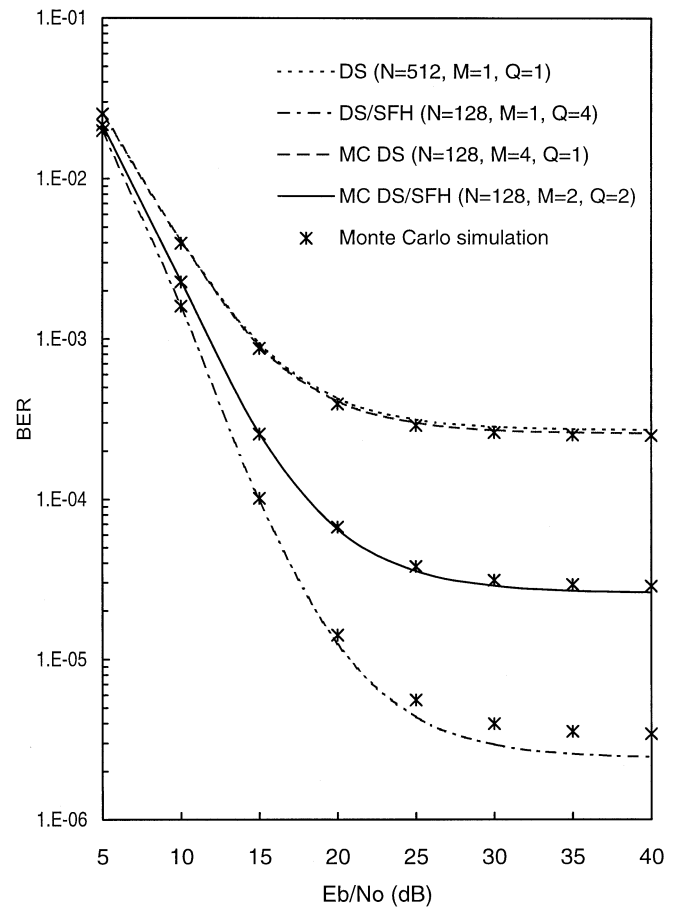
smaller Q which have a larger average hits number $E[K_h]$. It is also seen from Fig. 5 that the systems with larger Q have better performance in the AWGN channel. It has already been shown in Fig. 4 that the systems with larger Q have a higher SNR per diversity branch, therefore, with the same diversity order, the hybrid DS/SFH system has the best performance. The MC DS/SFH system, which employs fewer antennas and a shorter hopping pattern and, consequently, has lower implementation complexity, is second in performance. The MC DS and the pure DS systems exhibit almost the same performance since they have almost identical SNR per diversity branch.

The system performance in the CDMA overlay situations is of more concern in this paper. For single-carrier systems, namely, the hybrid DS/SFH and the pure DS systems, $\overline{\gamma}_m|_{K_h, q}$ are still identical because at any instant the desired signal only use one subchannel. Therefore, $P_e|_{K_h, q}$ is still given by (24). For the hybrid DS/SFH system, r_0 should be replaced by r_2 given by (B2) with $m = 0$

$$r_2 = \left[2K_h/(3N) + 4\Delta(q)w\hat{\sigma}_J^2/N + 4(E_b/N_0)^{-1} \right]^{-1}. \tag{26}$$

For a pure DS system, r_0 should be replaced by r_3 given by

$$r_3 = \left[2K_h L_T/(3N) + 4w\overline{\sigma}_J^2/N + 4(E_b/N_0)^{-1} \right]^{-1} \tag{27}$$

where $\overline{\sigma}_J^2$ is given by

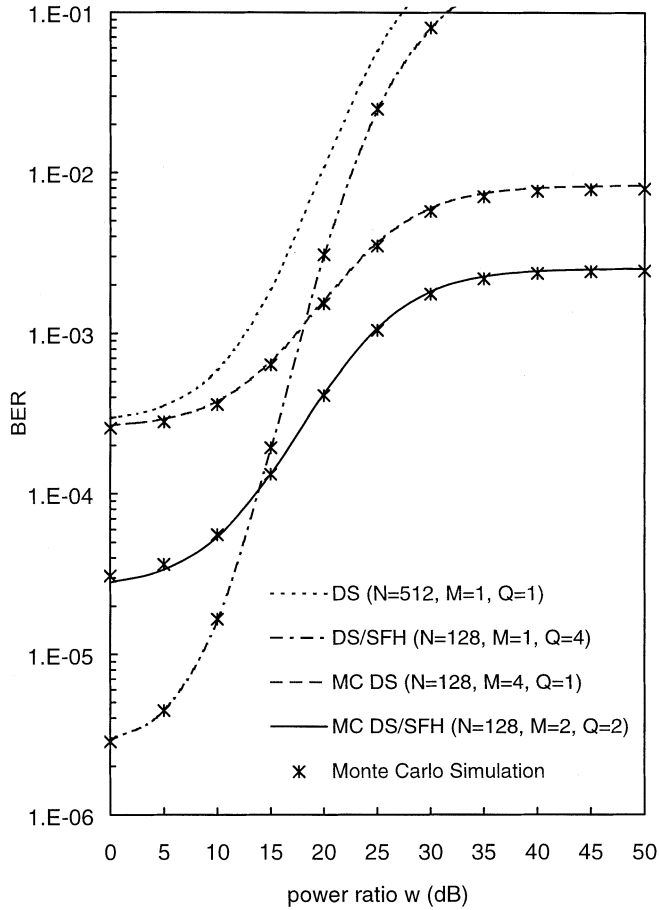$$\overline{\sigma}_J^2 = 2 \int_0^1 \text{sinc}(2\pi p\tau)(1 - \tau)(1 - \tau/N) \, d\tau. \tag{28}$$

Fig. 6. BER performance versus interference power to signal power ratio w ($K = 50$, $B_J = B_s$, $E_b/N_0 \to \infty$).

Fig. 7. BER performance versus interference bandwidth to system bandwidth ratio p ($K = 50$, $w = 20$ dB, $E_b/N_0 \to \infty$).

For the multiple-carrier systems, $\overline{\gamma}_m|_{K_h, q}$ are no longer identical for all m. For the MC DS system, the two middle subchannels are overlaid by the interference. The SNRs of the outputs from those two subchannels are identical and given by

$$r_1 = \left\{ 2K_h L/(3N) + 4\left[w\hat{\sigma}_J^2/N + (E_b/N_0)^{-1} \right] \right\}^{-1} \quad (29)$$

while the SNRs of the outputs from the other two subchannels free of interference are equal to r_0. Since the MC DS/SFH system uses one affected and one clean subchannel at any instant, and employs two antennas, it also has two affected diversity branches and two clean ones. Therefore, $P_e|_{K_h, q}$ is identical for the two systems and given by (B10)

$$P_e|_{K_h, q} = A_{10}^2 \mu_0^2 (3 - 2\mu_0) + 2A_{10}^2 A_{01} \mu_0$$
$$+ 2A_{01}^2 A_{10} \mu_1 + A_{01}^2 \mu_1^2 (3 - 2\mu_1) \quad (30)$$

where $A_{ij} = r_j/(r_j - r_i)$, $\mu_i = (1 - \sqrt{r_i/(1 + r_i)})/2$, and r_0 and r_1 are given by (25) and (29), respectively.

In Fig. 6, the BER performance is plotted versus the interference power to signal power ratio w, while the interference bandwidth $B_J = B_s$. It is shown that the BERs of the pure DS and hybrid DS/SFH systems rises drastically with the interference power. Therefore, they are not suitable candidates for CDMA overlay. The MC systems have certain advantages because they can make use of their clean diversity branches and keep the BER at a lower level. Their performance as w approaches infinity is no worse than that only the unaffected diversity branches are
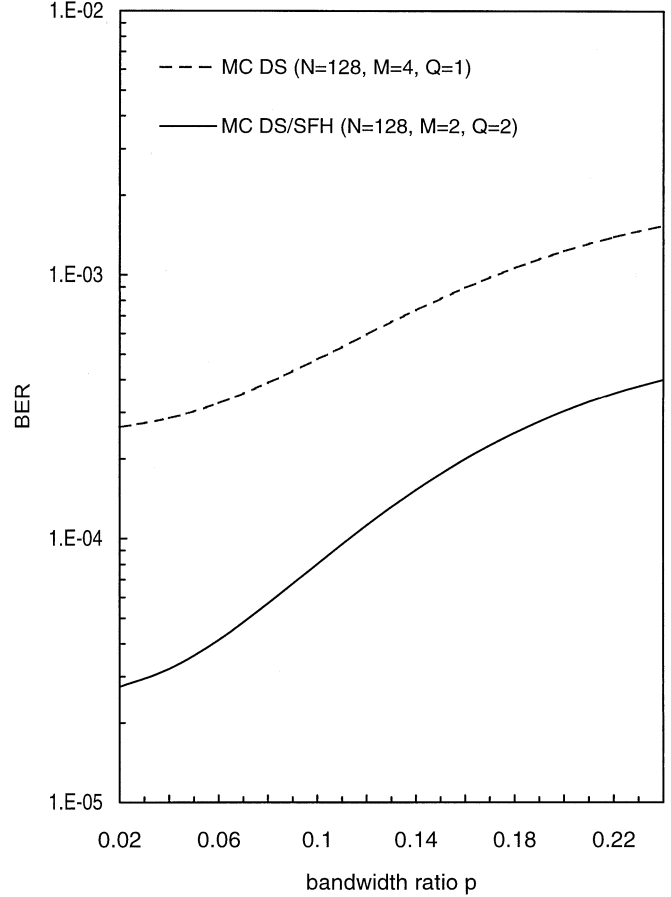
used. It is also seen that the MC DS/SFH system outperforms the MC DS system for any interference power. From (30), one can see that given K_h, the two systems have the same BER. However, the MC DS/SFH system has a larger Q that results in a smaller mean value of K_h and, therefore, a higher average SNR per diversity branch. In general, the MC DS/SFH system can suppress the narrow-band interference as effectively as the MC DS system while undergoing lower MAI.

Fig. 7 illustrates the effects of the interference bandwidth on the performance of the MC DS/SFH and the MC DS systems ($K = 50$, $w = 20$ dB, $E_b/N_0 \to \infty$). It is seen that the performance of both systems degrades as the interference bandwidth increases, but the MC DS/SFH system keeps its advantages. The BERs of both systems for different system loads (K) are given in Fig. 8 ($w = 20$ dB, $B_J = B_s$, $E_b/N_0 \to \infty$), which shows that the advantages of the MC DS/SFH system become more noticeable with heavy system loads.

Finally, we consider the cases that the narrow-band interference is not located in the center of the system bandwidth. The performance of the MC DS/SFH and the MC DS systems with respect to the location of the interference central frequency f_J ($B_J = B_s$, $w = 50$ dB) is shown in Fig. 9. It is seen that both systems perform best when the interference is limited within one subchannel ($f_J = f_s$, $s = 1, \ldots, 4$) because more clean subchannels are available, especially for the MC DS system. In those situations, the MC DS system has three unaffected di-
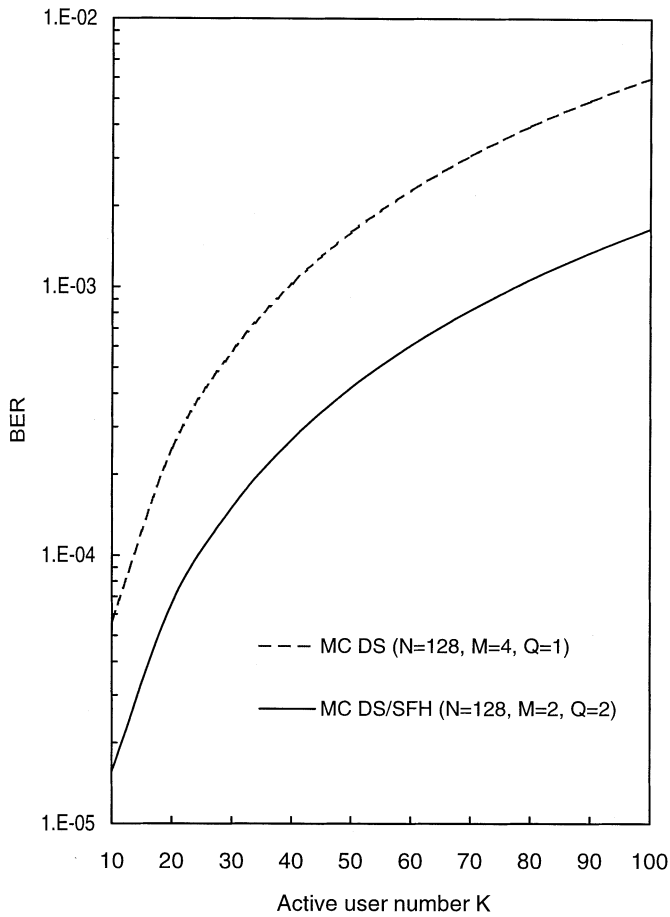
Fig. 8. BER performance versus active user number K ($w = 20$ dB, $B_J = B_s$, $E_b/N_0 \to \infty$).

Fig. 9. BER with respect to interference central frequency f_J ($w = 50$ dB, $B_J = B_s$, $E_b/N_0 \to \infty$).

versity branches while the MC DS/SFH system has only two in the worst case. That explains why the performance of the MC DS system is better when active user number is small (e.g., $K = 10$). However, when user number is large so that MAI is comparable to the narrow-band interference, the MC DS/SFH still performs better.

V. CONCLUDING REMARKS

The MC DS/SFH-CDMA systems are investigated in a multipath Rayleigh-fading channel, with or without narrow-band interference. The performance is compared with that of the MC DS system, along with the hybrid DS/SFH system and the pure DS system. The results show that in a multipath channel, it is advisable to divide the total bandwidth into subchannels with a mutual separation of the channel coherence bandwidth, and use frequency and antenna diversity instead of multipath diversity. Though MC DS system can suppress the narrow-band interference effectively via frequency diversity, it does not help to reduce the MAI. The MC DS/SFH system, with an appropriate combination of MC and SFH techniques, can provide sufficient frequency diversity and reduce MAI simultaneously, and as a result, achieves better performance.

Note that the performance comparison of the MC DS/SFH with conventional MC CDMA is made under the condition of the same diversity order. Using multiple antennas most likely
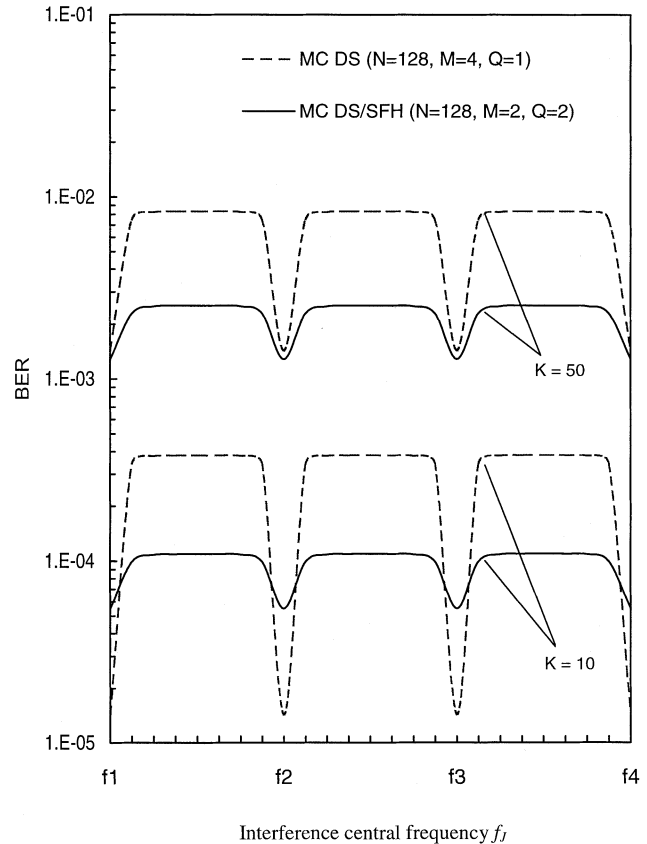
in base stations to obtain diversity in the MC DS/SFH-CDMA system may seem to be unfair to other systems. However, the MC DS-CDMA system employs multiple carriers to achieve frequency diversity. In terms of the number of correlators and gain estimators required in the receiver, the proposed system is no more complicated than the others. We admit that antenna diversity may cost more in implementation, but it provides better performance.

APPENDIX A
THE DERIVATION OF $\sigma_J^2(s_m)$

Substituting $J^{(s_m, v)}(t)$ in (11) by (6a) with $s = s_m$ and neglecting the high-frequency terms, we have

$$J_{ml}^{(v)} = \int_{\lambda T}^{(\lambda+1)T} \hat{J}^{(s_m, v)}[t - (L - 1 - l)T_c]a_i(t)\,dt \quad \text{(A1)}$$

where

$$\hat{J}^{(s_m, v)}(t) = \Delta(s_m)\,\mathrm{Re}\left\{ \left[J_I^{(v)}(t) + jJ_Q^{(v)}(t) \right] \right.$$
$$\left. \cdot \exp\left[\pm j2\pi(B_s/2 - B_J/4)t - j\Omega_{i,l}^{(s_m, v)} \right] \right\} \quad \text{(A2)}$$

where

$$\Omega_{i,l}^{(s_m, v)} = \Phi_{i,l}^{(s_m, v)} + 2\pi[mB_H \pm (B_s/2 - B_J/4)](L - 1 - l)T_c.$$

The autocorrelation function of $\hat{J}^{(s_m, v)}(t)$ is given by (7) with $f_s = 0$, i.e.,

$$\hat{R}_J(\tau)$$
$$= \Delta(s_m)P_J/2\,\mathrm{sinc}(\pi B_J\tau/2)\cos[2\pi(B_s/2 - B_J/4)\tau]. \quad \text{(A3)}$$

Therefore, the variance of $J_{ml}^{(v)}$, conditioned on s_m, is given by

$$
\begin{aligned}
\sigma_J^2(s_m) &= \int_{\lambda T}^{(\lambda+1)T} \int_{\lambda T}^{(\lambda+1)T} E\left\{ \hat{J}^{(s_m, v)}[t_1 - (L-1-l)T_c] \right. \\
&\quad \left. \cdot \hat{J}^{(s_m, v)}[t_2 - (L-1-l)T_c] \right\} \\
&\quad \cdot E[a_i(t_1)a_i(t_2)]\, dt_1\, dt_2 \\
&= \int_{\lambda T}^{(\lambda+1)T} \int_{\lambda T}^{(\lambda+1)T} \hat{R}_J(t_1 - t_2)R_i(t_1 - t_2)\, dt_1\, dt_2 \\
&= \int_{-T}^{T} \hat{R}_J(\tau)R_i(\tau)(T - |\tau|)\, d\tau \qquad (A4)
\end{aligned}
$$

where $R_i(\tau)$ is the autocorrelation function of $a_i(t)$ and triangular in $(-T_c, T_c)$. Substituting $\hat{R}_J(\tau)$ in (A4) by (A3), one obtains

$$
\begin{aligned}
\sigma_J^2(s_m) &= \Delta(s_m)P_J \int_0^{T_c} \sin c(\pi B_J \tau/2) \\
&\quad \cdot \cos[2\pi(B_s/2 - B_J/4)\tau](1 - \tau/T_c)(T - \tau)\, d\tau \\
&= \Delta(s_m)\frac{P_J T^2}{N} \hat{\sigma}_J^2 \qquad (A5)
\end{aligned}
$$

where

$$
\hat{\sigma}_J^2 = \int_0^1 \sin c(\pi p S\tau) \cos[2\pi(1 - pS/2)\tau](1 - \tau)(1 - \tau/N)\, d\tau \qquad (A6)
$$

where p is given by (4) and $S = MQ$.

APPENDIX B
THE DERIVATION OF $P_e|_{K_h, q}$

Since $\beta_{il}^{(s_m, v)}$ is Rayleigh distributed, the SNR $\gamma_{ml}^{(v)}|_{\beta_{il}^{(s_m, v)}, K_h, q}$ in (17) has a chi-square distribution, conditioned on K_h and q. Its characteristic function is given by

$$
\psi_{ml}^{(v)}(ju) = \frac{1}{1 - ju\overline{\gamma}_m|_{K_h, q}} \qquad (B1)
$$

where $\overline{\gamma}_m|_{K_h, q}$ is the mean of $\gamma_{ml}^{(v)}|_{\beta_{il}^{(s_m, v)}, K_h, q}$ over $\beta_{il}^{(s_m, v)}$ and given by

$$
\begin{aligned}
\overline{\gamma}_m|_{K_h, q} &= E_\beta \left[\gamma_{ml}^{(v)}\Big|_{\beta_{il}^{(s_m, v)}, K_h, q} \right] \\
&= \left\{ 2K_h L/(3N) + D\left[\Delta(q + mQ)w\hat{\sigma}_J^2/N \right. \right. \\
&\quad \left. \left. + (E_b/N_0)^{-1} \right] \right\}^{-1}. \qquad (B2)
\end{aligned}
$$

Because $\{\beta_{il}^{(s_m, v)}\}$ are mutually independent, the characteristic function of $\gamma_b|_{\{\beta_{il}^{(s_m, v)}\}, K_h, q}$ is simply the product of all $\psi_{ml}^{(v)}(ju)$, i.e.,

$$
\begin{aligned}
\psi_b(ju) &= \prod_{v=1}^{V} \prod_{m=0}^{M-1} \prod_{l=0}^{L-1} \psi_{ml}^{(v)}(ju) \\
&= \prod_{m=0}^{M-1} \frac{1}{\left(1 - ju\overline{\gamma}_m|_{K_h, q}\right)^{VL}}. \qquad (B3)
\end{aligned}
$$

The pdf of γ_b, $f_\gamma(\gamma_b|K_h, q)$, conditioned on K_h and q, can be obtained by inverse Fourier transform of $\psi_b(ju)$, and $P_e|_{(K_h, q)}$ is given by

$$
P_e|_{K_h, q} = \int_0^\infty P_e(\gamma_b)f_\gamma(\gamma_b|K_h, q)\, d\gamma_b. \qquad (B4)
$$

When all diversity branches have the same average SNR equal to r, i.e., $\psi_b(ju) = 1/(1 - jur)^D$, where D is the diversity order, $P_e|_{K_h, q}$ has a closed form solution [10]

$$
P_e|_{K_h, q} = P(D, r) = \mu^D \sum_{i=0}^{D-1} \binom{D-1+i}{i}(1 - \mu)^i \quad (B5)
$$

where $\mu = [1 - \sqrt{r/(1+r)}]/2$.

When the SNRs of different branches are not identical (in the presence of narrow-band interference), $\psi_b(ju)$ can be expressed by a partial fraction expansion, i.e.,

$$
\psi_b(ju) = \frac{1}{\prod_i (1 - jur_i)^{D_i}} = \sum_i \sum_{j=1}^{D_i} \frac{B_{i,j}}{(1 - jur_i)^j} \quad (B6)
$$

where $\sum_i D_i = D$ and

$$
B_{i,j} = \frac{(-r_i)^{(j-D_i)}}{(D_i - j)!} \left\{ \frac{d^{(D_i - j)}}{dx^{(D_i - j)}} \left[(1 - r_i x)^{D_i} \psi_b(x) \right] \right\}\Bigg|_{x=1/r_i} \quad (B7)
$$

where $(d^k/dx^k)[f(x)]$ represents the kth derivative of $f(x)$. Since the inverse Fourier transform and integration are linear, the corresponding probability of bit error can be written as

$$
P_e|_{K_h, q} = \sum_i \sum_{d=1}^{D_i} B_{i,d}P(d, r_i). \qquad (B8)
$$

For example, when $D = 4$, two branches have SNR r_0 and the other two branches have SNR r_1 (the case we meet in the numerical section)

$$
\psi_b(ju) = \frac{1}{(1 - jur_0)^2(1 - jur_1)^2} \qquad (B9)
$$

one has

$$
\begin{aligned}
P_e|_{K_h, q} &= A_{10}^2 P(2, r_0) + 2A_{10}^2 A_{01}P(1, r_0) \\
&\quad + 2A_{01}^2 A_{10}P(1, r_1) + A_{01}^2 P(2, r_1) \\
&= A_{10}^2 \mu_0^2(3 - 2\mu_0) + 2A_{10}^2 A_{01}\mu_0 \\
&\quad + 2A_{01}^2 A_{10}\mu_1 + A_{01}^2 \mu_1^2(3 - 2\mu_1) \quad (B10)
\end{aligned}
$$

where $A_{ij} = r_j/(r_j - r_i)$ $(i \neq j, ri \neq rj)$ and $\mu_i = [1 - \sqrt{r_i/(1+r_i)}]/2$.

APPENDIX C
THE MONTE CARLO SIMULATION MODEL

For rectangular chips and $0 \leq nT_c \leq \tau \leq (n+1)T_c \leq T$, as shown in [11], the partial cross-correlation functions $R_{ki}(\tau)$ and $\hat{R}_{ki}(\tau)$ are given by

$$
\begin{aligned}
R_{ki}(\tau) &= C_{ki}(n - N)T_c + [C_{ki}(n+1-N) - C_{ki}(n-N)] \\
&\quad \cdot (\tau - nT_c) \\
\hat{R}_{ki}(\tau) &= C_{ki}(n)T_c + [C_{ki}(n+1) - C_{ki}(n)](\tau - nT_c) \quad (C1)
\end{aligned}
$$

where the discrete aperiodic cross-correlation term $C_{ki}(\cdot)$ is defined as

$$
C_{ki}(n) = \begin{cases} \sum\limits_{j=0}^{N-1-n} a_k^{(j)} a_i^{(j+n)}, & 0 \leq n \leq N-1 \\[2ex] \sum\limits_{j=0}^{N-1+n} a_k^{(j-n)} a_i^{(j)}, & -(N-1) \leq n \leq 0 \\[2ex] 0, & \text{else.} \end{cases} \quad (C2)
$$

Once the code sequences are obtained, $R_{ki}(\tau)$ and $\hat{R}_{ki}(\tau)$ are determined by τ. Note that $I_{kl'}^{(m, v)}$ in (14) is a zero-mean Gaussian random variable if τ_k is given, which follows from the fact that

$$\beta_{kl'}^{(s_m, v)} \cos[\Phi_{kl'}^{(s_m, v)} - \Phi_{il}^{(s_m, v)}]$$

is a zero-mean Gaussian variable with variance ρ, therefore,

$$b_k^{(\lambda-1)} \beta_{kl'}^{(s_m, v)} \cos[\Phi_{kl'}^{(s_m, v)} - \Phi_{il}^{(s_m, v)}]$$

and

$$b_k^{(\lambda)} \beta_{kl'}^{(s_m, v)} \cos[\Phi_{kl'}^{(s_m, v)} - \Phi_{il}^{(s_m, v)}]$$

are also zero-mean Gaussian variables with the same variance. The conditional variance of $I_{kl'}^{(m, v)}$ is given by

$$\sigma_{kl'}^2 = 2\rho P \left[R_{ki}^2 \left(\Gamma_{l'l}^k \right) + \hat{R}_{ki}^2 \left(\Gamma_{l'l}^k \right) \right]. \tag{C3}$$

It is also obvious that $\{I_{kl'}^{(m, v)}\}$ are all uncorrelated since

$$E\{\cos[\Phi_{k_1 l_1}^{(s_m, v)} - \Phi_{k_2 l_2}^{(s_m, v)}]\} = 0, \quad \text{for } k_1 \neq k_2 \text{ or } l_1 \neq l_2.$$

Therefore, conditioned on $b_i^{(\lambda)}$, $\beta_{il}^{(s_m, v)}$, $\{\tau_k\}$, K_h, and q, $\xi_{ml}^{(v)}(\lambda)$ is a Gaussian random variable. The SNR of $\xi_{ml}^{(v)}(\lambda)$ is defined as

$$\gamma_{ml}^{(v)} \Big|_{\beta_{il}^{(s_m, v)}, \{\tau_k\}, K_h, q}$$

$$= \frac{E^2 \left[\xi_{ml}^{(v)}(\lambda) \Big| \beta_{il}^{(s_m, v)} \right]}{2 \operatorname{Var} \left[\xi_{ml}^{(v)}(\lambda) \Big| \{\tau_k\}, K_h, q \right]}$$

$$= \frac{E_c T \left(\beta_{il}^{(s_m, v)} \right)^2}{2\rho}$$

$$\cdot \left[\sum_{k=1}^{K_h} \sum_{l'=0}^{L-1} \sigma_{kl'}^2 + \sigma_J^2(q + mQ) + N_0 T \right]^{-1}. \tag{C4}$$

The instantaneous SNR of $\xi(\lambda)$, conditioned on $\{\beta_{il}^{(s_m, v)}\}$, $\{\tau_k\}$, K_h, and q is given by

$$\gamma_b \Big|_{\beta_{il}^{(s_m, v)}, \{\tau_k\}, K_h, q}$$

$$= \sum_{v=1}^{V} \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} \gamma_{ml}^{(v)} \Big|_{\beta_{il}^{(s_m, v)}, \{\tau_k\}, K_h, q}$$

$$= \sum_{v=1}^{V} \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} \frac{\left(\beta_{il}^{(s_m v)} \right)^2}{2\rho}$$

$$\cdot \left\{ T^{-2} \sum_{k=1}^{K_h} \sum_{l'=1}^{L-1} \left[R_{ki}^2(\Gamma_{l'l}^k) + \hat{R}_{ki}^2 \left(\Gamma_{l'l}^k \right) \right] \right.$$

$$\left. + D \left[\Delta(q + mQ) w \hat{\sigma}_J^2 / N + (E_b/N_0)^{-1} \right] \right\}^{-1}. \tag{C5}$$

To obtain the average BER P_e, the conditional BER is averaged over the statistics of γ_b, that is

$$P_e = \int_0^\infty P_e(\gamma_b) f_\gamma(\gamma_b) \, d\gamma_b \tag{C6}$$

where $f_\gamma(\gamma_b)$ represents the pdf of γ_b. It is difficult to get the analytical results, therefore Monte Carlo integration over $\{\beta_{il}^{(s_m, v)}\}$, $\{\tau_k\}$, K_h, and q is performed to get P_e.

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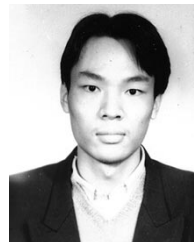
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