PAPER

Joint Antenna Diversity and Frequency-Domain Equalization for Multi-Rate MC-CDMA

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SUMMARY For the reception of MC-CDMA signals in a frequency-selective fading channel, frequency-domain equalization is necessary before despreading. In this paper, joint antenna diversity combining and one-tap frequency-domain equalization is considered (simply referred to as the joint antenna diversity & equalization, in this paper). A receiver structure for joint antenna diversity & equalization is presented and the unified weights based on minimum mean square error (MMSE) criterion are found in the presence of multi-users with different spreading factors and transmit powers. For comparison, antenna diversity combining after despreading using MMSE combining (MMSEC) is also considered. The achievable bit error rate (BER) performances with joint antenna diversity & equalization and with antenna diversity after MMSEC despreading in a frequency-selective Rayleigh fading channel are evaluated by computer simulations and compared.

key words: *MC-CDMA, MMSEC, frequency-domain equalization, antenna diversity, frequency-selective channel*

1. Introduction

In next generation wireless communications systems, very high-speed data transmissions will be required under severe fading environments [1]. Recently, the combination of multi-carrier (MC) and code division multiple access (CDMA), called MC-CDMA, has been attracting much attention and is under extensive study [2]–[5]. In MC-CDMA, each user's data-modulated symbol to be transmitted is spread over a number of subcarriers using an orthogonal spreading sequence defined in the frequency-domain. In wireless multimedia communications, different users may communicate with different data rates and different communication qualities. This can be achieved by assigning different spreading factors and transmit powers to different users in MC-CDMA (called multi-rate MC-CDMA in this paper), as in multi-code direct sequence CDMA (DS-CDMA) [13].

In mobile radio communications, the transmitted signal is scattered by many obstacles located between a transmitter and a receiver, thereby creating a multipath fading channel whose transfer function is not anymore constant over the signal bandwidth. Such a propagation channel is called a frequency-selective fading channel [6]. The received signal suffers from distortion and thus, orthogonality property among different users' signals is partially lost, thereby producing large multi-user interference (MUI). Hence, some frequency-domain equalization technique is necessary. Frequency-domain equalization and despreading called orthogonal restoration combining (ORC) [3], [5] perfectly equalizes the frequency-selective fading channel to restore the orthogonality, but, in turn, produces noise enhancement. The noise enhancement can be significantly reduced by using frequency-domain equalization and despreading called minimum mean square error combining (MMSEC) at the cost of a slight loss of orthogonality [7], [8].

In MC-CDMA with a large spreading factor (or low data rate transmission), the frequency diversity effect can be achieved in a frequency-selective fading channel. However, decreasing the spreading factor (or increasing the data rate) reduces the frequency diversity effect. Antenna diversity reception can be used to improve the transmission performance. In MC-CDMA, antenna diversity combining can be performed in conjunction with frequency-domain equalization. Recently, antenna diversity combining before and after frequency-domain despreading were investigated and the achievable performance was evaluated by computer simulation in [9]. In [9], antenna diversity combining and one-tap frequency-domain equalization for despreading are treated separately. However, antenna diversity combining weight and frequency-domain equalization weight could be unified to simplify the receiver structure. Furthermore, in a multi-rate scenario, different users are transmitting their data at different rates with different quality-of-service requirements, i.e., different spreading factors and transmit powers are assigned to users. To the best of authors' knowledge, there is no literature dealing with unified weights for jointly performing antenna diversity combining and frequency-domain equalization for despreading (simply referred to as joint antenna diversity & equalization, in this paper) and furthermore, the multi-rate scenario has not been considered so far. This is the motivation for this paper. The objective of this paper is two folds: (a) to present a receiver structure for joint antenna diversity & equalization and (b) to present the unified weights based on the MMSE criterion for multi-rate MC-CDMA. For

Manuscript received November 2, 2002.
Manuscript revised May 27, 2003.
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comparison, antenna diversity combining after MMSEC despreading is also considered and maximal ratio combining (MRC) weights for antenna diversity are found. The achievable BER performances with joint antenna diversity & equalization and antenna diversity after MMSEC despreading in a frequency-selective Rayleigh fading channel are evaluated by computer simulations and compared.

The remainder of this paper is organized as follows. Section 2 presents the multi-rate MC-CDMA transmission system model. The unified weights for joint antenna diversity & equalization are presented in Sect. 3. Then, in Sect. 4, the MRC weights for antenna diversity combining after MMSEC despreading are presented. Section 5 presents the performance evaluation of the two antenna diversity combining schemes by computer simulations. Section 6 offers some conclusions and future work.

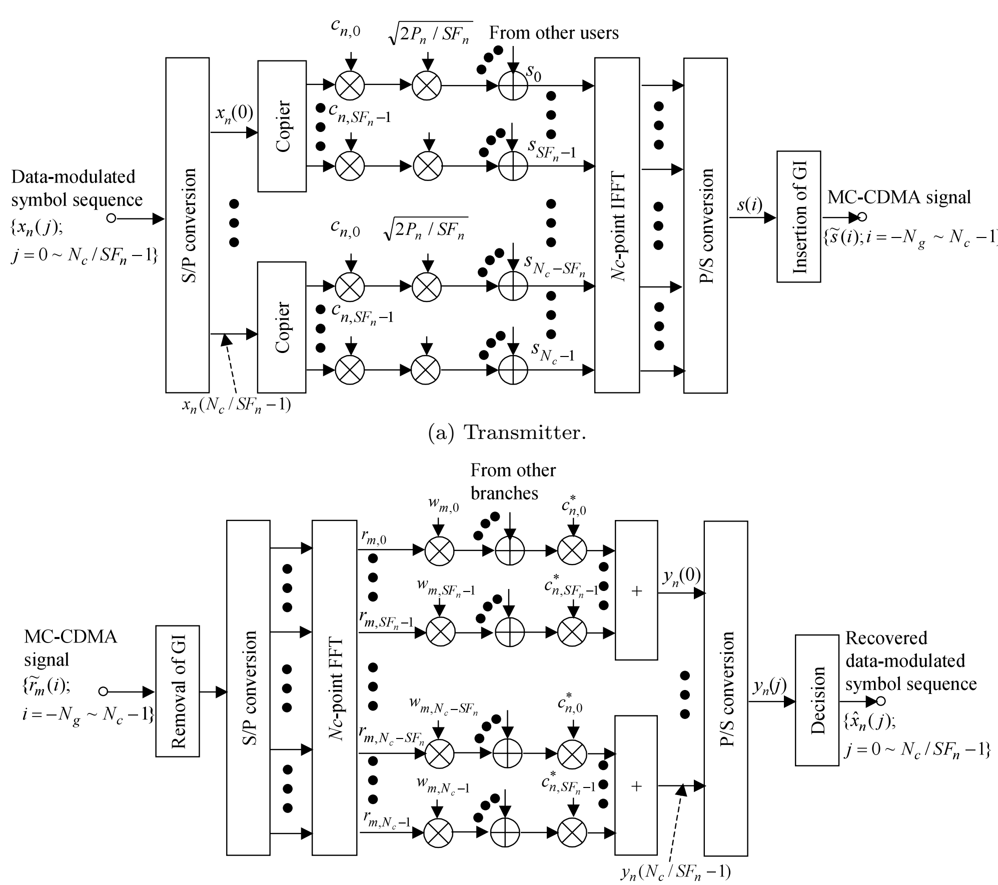
2. Multi-Rate MC-CDMA Signal Transmission Model

MC-CDMA transmitter and receiver structures for joint antenna diversity & equalization are illustrated in Fig. 1. We consider MC-CDMA having N_c subcarriers with a carrier spacing of $1/T_s$ to transmit N users' data in parallel. Each user's data is spread using the frequency-domain orthogonal spreading code with different spreading factor. Without loss of generality, we consider the time interval of one signaling period, i.e., $0 \leq t < T$ with $T = T_s + T_g$, where T_s and T_g are respectively the effective symbol length and the guard interval (GI). Throughout the paper, discrete-time representation of the MC-CDMA signal is used.

The MC-CDMA transmitter structure is illustrated in Fig. 1(a). Let $x_n(j)$ and SF_n be the jth data-modulated symbol with $|x_n(j)|$=1 and the spreading factor of the nth user, respectively. During $0 \leq t < T$, $\{x_n(j);\ n = 0 \sim N-1,\ j = 0 \sim N_c/SF_n - 1\}$ are transmitted for N users, where N_c/SF_n is an integer. The orthogonal spreading codes, $\{c_{n,j};\ n = 0 \sim N-1,\ j = 0 \sim SF_n - 1\}$ with $|c_{n,j}| = 1$, chosen from orthogonal variable spreading factor (OVSF) code tree [10], [11], are considered. The OVSF codes satisfy

$$\frac{1}{SF_n}\sum_{k=0}^{SF_n-1} c_{n,k}c^*_{m,k \bmod SF_m} = \begin{cases} 1 & \text{if } m = n \\ 0 & \text{otherwise} \end{cases} \quad (1)$$

for $n = 0 \sim N-1$, where * denotes complex conjugation. As an example, the OVSF code tree for



(a) Transmitter.

(b) nth user receiver using joint antenna diversity & equalization.

Fig. 1 MC-CDMA transmitter and receiver structures.

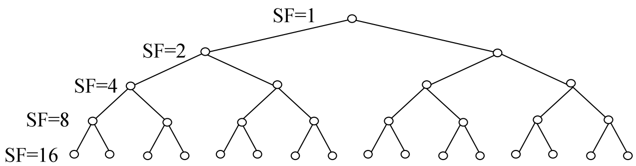


Fig. 2 OVSF code tree of orthogonal codes of $SF_n = 1 \sim 16$.

$SF_n = 1 \sim 16$ is illustrated in Fig. 2.

The composite modulated symbol s_k to be transmitted on the kth subcarrier may be expressed using the equivalent baseband representation as

$$s_k = \sum_{n=0}^{N-1} \sqrt{\frac{2P_n}{SF_n}} c_{n,k \bmod SF_n} x_n \left(\left\lfloor \frac{k}{SF_n} \right\rfloor \right) \tag{2}$$

for $k = 0 \sim N_c - 1$, where P_n represents the transmit power of the nth user and $\lfloor a \rfloor$ denotes the largest integer smaller than or equal to a. The N_c-point inverse fast Fourier transform (IFFT) is applied to the sequence $\{s_k;\ k{=}0{\sim}\ N_c - 1\}$ to generate the MC-CDMA signal $\{s(i);\ i = 0 \sim N_c - 1\}$ in time-domain:

$$s(i) = \frac{1}{N_c} \sum_{k=0}^{N_c-1} s_k \exp\left(j2\pi i \frac{k}{N_c} \right). \tag{3}$$

After insertion of the N_g-sample GI, the resultant MC-CDMA signal $\{\tilde{s}(i); i = -N_g \sim N_c - 1\}$ is transmitted over a propagation channel, where

$$\tilde{s}(i) = s(i \bmod N_c). \tag{4}$$

Since the IFFT sampling period is $\Delta T = T_s/N_c$, we have $T_g = N_g \Delta T$ and $T = T_s + T_g = T_s(1{+}N_g/N_c)$.

A ΔT-spaced time delay model for the propagation channel is assumed. M-branch antenna diversity reception is considered. Assuming L independent propagation paths with distinct time delays $\{\tau_l\}$, the impulse response $h_m(\tau)$ of the multipath channel experienced by the mth antenna, $m = 0 \sim M-1$, may be expressed as

$$h_m(\tau) = \sum_{l=0}^{L-1} h_m^{(l)} \delta(\tau - \tau_l) \tag{5}$$

with $\sum_{l=0}^{L-1} E[|h_m^{(l)}|^2] = 1$, where $\delta(t)$ is the delta function and $E[.]$ denotes ensemble average operation. Time dependency of the channel has been dropped for simplicity. The time delays $\{\tau_l\}$ are assumed to be integer multiple of the FFT sampling period.

The MC-CDMA receiver structure using joint antenna diversity & equalization is illustrated in Fig. 1(b). Ideal sampling timing is assumed. The MC-CDMA signal received on the mth antenna is sampled at the rate of $\Delta T^{-1} = N_c/T_s$ to obtain $\{\tilde{r}_m(i);\ i = -N_g \sim N_c{-}1\}$, which is expressed as

$$\tilde{r}_m(i) = \sum_{l=0}^{L-1} h_m^{(l)} \tilde{s}(i - \tau_l/\Delta T) + \eta_m(i), \tag{6}$$

where $\eta_m(i)$ represents the additive white Gaussian noise (AWGN) process with the single sided power spectrum density N_0. The N_g-sample GI is removed and the N_c-point FFT is applied to decompose the received MC-CDMA signal into the N_c-subcarrier components $\{r_{m,k};\ k = 0 \sim N_c - 1\}$:

$$r_{m,k} = \sum_{i=0}^{N_c-1} \tilde{r}_m(i) \exp\left(-j2\pi k \frac{i}{N_c} \right) \tag{7}$$

for $m = 0 \sim M - 1$. Denoting the channel gain at the kth subcarrier on the mth antenna by $\xi_{m,k}$, the kth subcarrier component $r_{m,k}$ received on the mth antenna is represented as

$$r_{m,k} = \xi_{m,k} \sum_{n=0}^{N-1} \sqrt{\frac{2P_n}{SF_n}} c_{n,k \bmod SF_n} x_n \left(\left\lfloor \frac{k}{SF_n} \right\rfloor \right) + \eta_{m,k}, \tag{8}$$

where $\{\xi_{m,k};\ k = 0 \sim N_c - 1\}$ and $\{\eta_{m,k};\ k = 0 \sim N_c - 1\}$ are respectively the Fourier transforms of the channel impulse response $h_m(\tau)$ and the AWGN process $\eta_m(i)$. They are given by

$$\begin{cases} \xi_{m,k} = \sum_{l=0}^{L-1} h_m^{(l)} \exp\left(-j2\pi k \frac{\tau_l/\Delta T}{N_c} \right) \\ \eta_{m,k} = \sum_{i=0}^{N_c-1} \eta_m(i) \exp\left(-j2\pi k \frac{i}{N_c} \right) \end{cases}. \tag{9}$$

$\{\xi_{m,k};\ m = 0 \sim M - 1\}$ on the kth subcarrier are the independent and identically distributed (iid) complex random variables with zero-mean and unit variance and $\{\eta_{m,k};\ m = 0 \sim M - 1\}$ are the iid complex Gaussian variables with zero mean and variance $2N_0/T_s$.

3. Joint Antenna Diversity and Equalization before Despreading

On each subcarrier, the M received signal samples are weighted by $\{w_{m,k};\ m = 0 \sim M - 1\}$ and combined to obtain

$$y_k = \sum_{m=0}^{M-1} w_{m,k} r_{m,k}. \tag{10}$$

The instantaneous combining error is defined as

$$\varepsilon_k = s_k - y_k = s_k - \boldsymbol{w}_k^T \boldsymbol{r}_k, \tag{11}$$

where $\boldsymbol{w}_k = (w_{0,k}, w_{1,k}, \cdots, w_{M-1,k})^T$ and $\boldsymbol{r}_k = (r_{0,k}, r_{1,k}, \cdots, r_{M-1,k})^T$ with the superscript T denoting the transposition. From [14], the set of weights that minimizes the mean square error (MSE) $E[|\varepsilon_k|^2]$ can be expressed in matrix form as

$$\boldsymbol{w}_k^* = \boldsymbol{R}^{-1} \boldsymbol{p}, \tag{12}$$

where $\boldsymbol{R}^{-1}$ is the inverse of the M-by-M correlation matrix $\boldsymbol{R} = E[\boldsymbol{r}_k \boldsymbol{r}_k^H]$ and $\boldsymbol{p} = E[\boldsymbol{r}_k s_k^*]$ represents the M-by-1 cross-correlation vector with the superscript H being the Hermitian transposition and $E[.]$ being the ensemble average operation over the N users' data-modulated symbols $\{x_n(j);\ n = 0 \sim N-1\}$ and noise samples $\{\eta_{m,k};\ m = 0 \sim M-1\}$ for the given set of $\{\xi_{m,k};\ m = 0 \sim M-1\}$. It is assumed that $\{x_n(j);\ n = 0 \sim N-1\}$ are zero-mean and independent random variables. The simple-to-use closed-form solution for $\boldsymbol{w}_k = (w_{0,k}, w_{1,k}, \cdots, w_{M-1,k})^T$ can be obtained, after some manipulation, as

$$w_{m,k} = \frac{\xi_{m,k}^*}{\sum_{m=0}^{M-1} |\xi_{m,k}|^2 + \left(\sum_{n=0}^{N-1} \frac{\Gamma_{n,eff}}{SF_n}\right)^{-1}}, \tag{13}$$

where

$$\Gamma_{n,eff} = \frac{P_n T_s}{N_0} = \frac{\Gamma_n}{1 + T_g/T_s} \tag{14}$$

is the *effective* average received symbol energy-to-AWGN power spectrum density ratio E_s/N_0 and Γ_n represents the average received E_s/N_0 for the nth user. Letting M=1 in Eq. (13) gives the well-known MMSEC weight given in [7].

To obtain the decision variable $y_n(j)$ for the jth data-modulated symbol of the nth user, despreading is performed by multiplying SF_n soft samples $\{y_k;\ k = jSF_n \sim (j+1)SF_n - 1\}$ by orthogonal spreading code $\{c_{n,k \bmod SF_n};\ k = jSF_n \sim (j+1)SF_n - 1\}$ and summing:

$$y_n(j) = \sum_{k=jSF_n}^{(j+1)SF_n-1} y_k c_{n,k \bmod SF_n}^*. \tag{15}$$

Using $\{y_n(j);\ j = 0 \sim N_c/SF_n - 1\}$, the transmitted data-modulated symbol sequence is recovered, which is denoted by $\{\hat{x}_n(j);\ j = 0 \sim N_c/SF_n - 1\}$.

Observing Eq. (13), the unified weight for joint antenna diversity & equalization can be represented as the product of two weights:

$$w_{m,k} = w_{m,k}^{(1)} w_k^{(2)}, \tag{16}$$

where

$$\begin{cases} w_{m,k}^{(1)} = \xi_{m,k}^* \\ w_k^{(2)} = \dfrac{1}{\sum_{m=0}^{M-1} |\xi_{m,k}|^2 + \left(\sum_{n=0}^{N-1} \frac{\Gamma_{n,eff}}{SF_n}\right)^{-1}} \end{cases}. \tag{17}$$

$w_{m,k}^{(1)}$ is identical with the maximal ratio combining (MRC) weight [12] for the mth antenna and $w_k^{(2)}$ is the weight common to all antennas. In antenna diversity before MMSEC despreading of [9], MRC is carried out first and then, MMSEC despreading is performed; the MMSEC weight for the kth subcarrier is given by $w_k^{(c)} = |h_k^{(c)}|/(D^{(c)}|h_k^{(c)}|^2 + N^{(c)})$, where $h_k^{(c)}$ is the equivalent channel gain, $D^{(c)}$ the total power-to-each user power ratio and $N^{(c)}$ the noise power, all after antenna diversity combining based on MRC. In [9], $h_k^{(c)}$, $D^{(c)}$ and $N^{(c)}$ are estimated using known pilot. Assuming ideal estimation and $\Gamma_n = \Gamma$ for all N users (i.e., $P_n = P$), it can be shown that $h_k^{(c)} = 2(P/SF)\sum_{m=0}^{M-1} |\xi_{m,k}|^2$, $D^{(c)} = N$, and $N^{(c)} = 2(N_0/T_s)\sum_{m=0}^{M-1} |\xi_{m,k}|^2$. Therefore, the MMSEC weight after MRC antenna diversity turns out to be $w_k^{(c)} = 1\Big/\left(\sum_{m=0}^{M-1} |\xi_{m,k}|^2 + \left(\frac{N}{SF}\Gamma\right)^{-1}\right)$ and equals $w_k^{(2)}$ of Eq. (17). As a consequence, MRC antenna diversity before MMSEC despreading of [9] is equivalent to the joint antenna diversity & equalization. This shows that antenna diversity and frequency-domain equalization can be unified rather than implementing them separately.

4. Antenna Diversity Combining after MMSEC Despreading

The receiver structure using antenna diversity combining after MMSEC despreading (called post-MMSEC antenna diversity in this paper) is illustrated in Fig. 3. The received signal sample on the kth subcarrier of the mth antenna is given by Eq. (8). First, despreading using MMSEC is performed on each antenna and then, antenna diversity combining is applied. The MMSEC weight is obtained by letting M=1 in Eq. (13) as

$$w_{m,k} = \frac{\xi_{m,k}^*}{|\xi_{m,k}|^2 + \left(\sum_{n=0}^{N-1} \frac{\Gamma_{n,eff}}{SF_n}\right)^{-1}}. \tag{18}$$

Despreading using MMSEC gives

$$\begin{aligned} y_{m,n}(j) &= \sum_{k=jSF_n}^{j(SF_n+1)-1} w_{m,k} r_{m,k} c_{n,k \bmod SF_n}^* \\ &= \sqrt{\frac{2P_n}{SF_n}} x_n(j) \left(\sum_{k=jSF}^{(j+1)SF_n-1} w_{m,k}\xi_{m,k}\right) \\ &\quad + \sum_{\substack{l=0\\ \neq n}}^{N-1} \sqrt{\frac{2P_l}{SF_l}} x_l\left(\left\lfloor \frac{k}{SF_l} \right\rfloor\right) \\ &\quad \times \left(\sum_{k=jSF_n}^{(j+1)SF_n-1} w_{m,k}\xi_{m,k} c_{l,k \bmod SF_l} c_{n,k \bmod SF_n}^*\right) \\ &\quad + \sum_{k=jSF_n}^{(j+1)SF_n-1} w_{m,k}\eta_{m,k} c_{n,k \bmod SF_n}^*. \end{aligned} \tag{19}$$

The first term is the desired signal component, the sec-

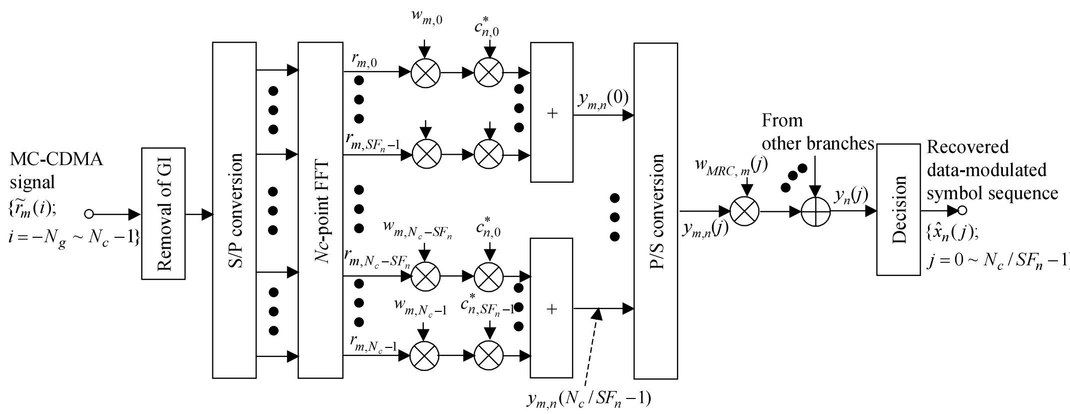


Fig. 3 MC-CDMA receiver structure using antenna diversity combining after MMSEC despreading.

ond the MUI component, and the third the noise component due to AWGN. We treat the sum of MUI and noise due to AWGN as an equivalent zero-mean Gaussian noise $z_{m,n}(j)$, hence Eq. (19) simplifies to

$$y_{m,n}(j) = \sqrt{\frac{2P_n}{SF_n}} \left(\sum_{k=jSF_n}^{(j+1)SF_n-1} w_{m,k}\xi_{m,k} \right) x_n(j) + z_{m,n}(j). \quad (20)$$

The variance σ_m^2 of $z_{m,n}(j)$ is given by

$$\sigma_m^2 = \frac{2N_0}{T_s} \sum_{k=jSF_n}^{(j+1)SF_n-1} |w_{m,k}|^2 \cdot \left(1 + |\xi_{m,k}|^2 \sum_{\substack{l=0 \\ \neq n}}^{N-1} \frac{\Gamma_{l,eff}}{SF_l} \right). \quad (21)$$

Antenna diversity combining based on MRC after MMSEC despreading is performed as follows:

$$y_n(j) = \sum_{m=0}^{M-1} w_{MRC,m}(j) y_{m,n}(j), \quad (22)$$

where $w_{MRC,m}(j)$ represents the MRC weight given, from Eqs. (20) and (21), as

$$w_{MRC,m}(j) = \frac{\displaystyle\sum_{k=jSF_n}^{(j+1)SF_n-1} w_{m,k}\xi_{m,k}}{\displaystyle\sum_{k=jSF_n}^{(j+1)SF_n-1} |w_{m,k}|^2 \left(1 + |\xi_{m,k}|^2 \sum_{\substack{l=0 \\ \neq n}}^{N-1} \frac{\Gamma_{l,eff}}{SF_l} \right)}. \quad (23)$$

We also consider two other combining methods, approximate MRC and equal gain combining (EGC) [12]. The former is obtained by assuming that the noise variance σ_m^2 is equally likely for all antennas, $m = 0 \sim M-1$, in the case of a severe frequency-selective fading channel. The weights are given by

$$\begin{cases} w_{approx.MRC,m}(j) = \displaystyle\sum_{k=jSF_n}^{(j+1)SF_n-1} w_{m,k}\xi_{m,k} \\ w_{EGC,m}(j) = 1 \end{cases}. \quad (24)$$

5. Computer Simulation

Table 1 shows the simulation condition. We assume MC-CDMA using N_c=256 subcarriers with a carrier spacing of $1/T_s$ and a GI of $T_g = T_s/8$ (i.e., N_g=32) and ideal coherent quaternary phase shift keying (QPSK) data-modulation/demodulation. IFFT and FFT sampling period ΔT is $\Delta T = T_s/256$. We assume two classes of users, i.e., the spreading factors of SF_0=256 for one class and SF_1=128 for the other. As a propagation channel, a simple 2-path (L=2) Rayleigh fading model with equal average power is assumed. In the simulation, the time delay difference of $\Delta\tau$ ($= \tau_1 - \tau_0$) $= 4\Delta T$ is assumed. (Since it can be implied from [15, Fig. 9] that the achievable BER performance is insensitive to the time delay difference $\Delta\tau$ if $\Delta\tau/\Delta T > N_c/SF$ but less than the GI, the same BER performance can be obtained for $32 \geq \Delta\tau/\Delta T \geq 2(1)$ when $SF = 128$ (256).)

In Sects. 2 and 3, we considered the orthogonal spreading codes only. But, in practical applications, $\{s_k\}$ is further multiplied by a long scramble code sequence in order to make the MUI noise-like. In the simulation, an m-sequence with a repetition period of 4095 chips is used as the scramble code sequence in addition to the orthogonal spreading codes.

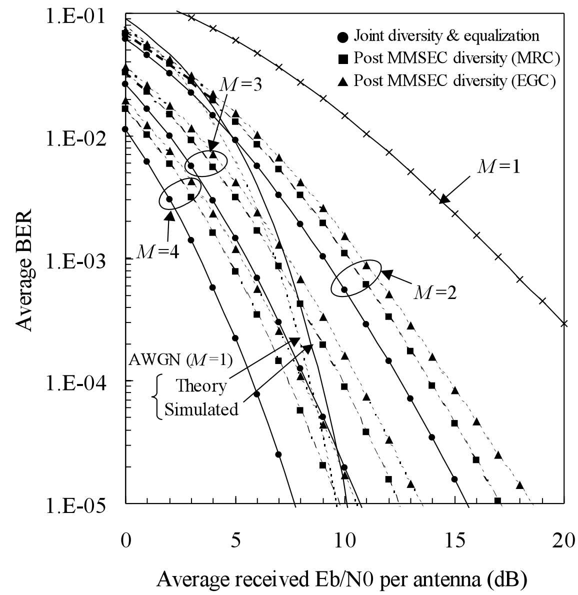
First, we consider the case that all users are using the same spreading factor $SF = 256$. We measured the average BER performances achievable using joint antenna diversity & equalization and using post-MMSEC antenna diversity using MRC, approximate MRC, and

Table 1 Simulation condition.

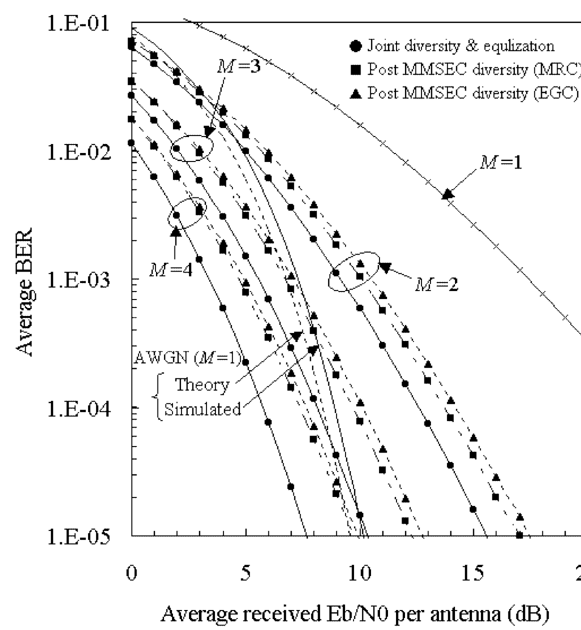
MC-CDMA	No. of subcarriers	N_c=256
	Effective symbol length	T_s=256ΔT
	Guard interval	T_g=32ΔT (T_g/T_s=1/8)
	Spreading factor	SF=256 and 128
	Data-modulation	QPSK
Rayleigh fading channel	No. of paths	L=2
	Time delay difference	Δτ=4ΔT

EGC with M as a parameter for $SF = 256$. The number of users is (a) 100% of SF and (b) 50% of SF. The results are plotted as a function of average received E_b/N_0 (=$0.5E_s/N_0$) in Fig. 4. For comparison, the theoretical BER performance curve for the AWGN channel (no fading) is plotted together with the simulated one (note that the performance difference between the theoretical and simulated BER curves for the AWGN channel is due to the power loss of 0.5 dB owing to the GI insertion). Since post-MMSEC antenna diversity using MRC and that using approximate MRC are found to give identical BER performances, only the results for MRC are plotted. It can be seen that the joint antenna diversity & equalization can achieve better BER performance than the post-MMSEC antenna diversity schemes for all values of M. The gain of the joint antenna diversity & equalization in the required E_b/N_0 value for the average BER=10^{-4} over the post-MMSEC antenna diversity using MRC is about 1.4 dB, 1.6 dB and 1.7 dB when M=2, 3 and 4, respectively, for the case of $N = SF$ (see Fig. 4(a)). Performance superiority of the joint antenna diversity & equalization against the post-MMSEC antenna diversity can also be seen when $N = 0.5SF$ (see Fig. 4(b)). For each diversity scheme, the BER performance when $N = 0.5SF$ is slightly better than when $N = SF$. This is due to less MUI.

It can be seen from computer simulation results presented in [9, Fig. 7] that EGC antenna diversity after MMSEC despreading achieves the best BER performance and is better than MRC antenna diversity before MMSEC despreading (which is equivalent to our joint antenna diversity & equalization as discussed in Sect. 3). Furthermore, in [9], EGC antenna diversity after MMSEC despreading is seen to be better than MRC antenna diversity after MMSEC despreading (see [9, Fig. 6]). These are opposite to our result of the ideal channel estimation case, i.e, the joint antenna diversity & equalization achieves the best BER performance. In computer simulations of [9], practical channel estimation using pilot is carried out for computing the MRC and MMSEC weights. This suggests that the achievable BER performance with antenna diversity combining and frequency-domain equalization is sensitive to channel estimation errors (this is also suggested in [15]) and improving channel estimation is important for joint antenna diversity & equalization.



(a) $SF = 256$ and $N = 256$



(b) $SF = 256$ and $N = 128$

Fig. 4 Comparison of average BER performances with joint antenna diversity & equalization and post-MMSEC antenna diversity. Single class of users is assumed. (a) $(SF, N) = (256, 256)$, (b) $(SF, N) = (256, 128)$.

Figure 5 compares the average BER performances using joint antenna diversity & equalization and post-MMSEC antenna diversity with M as a parameter in the presence of the two groups (SF_0=256 and SF_1=128) of users. The numbers of users for SF_0=256 group and SF_1=128 group are N_0=128 and N_1=64, respectively, and $\Gamma_0 = \Gamma_1 = \Gamma$. (Note that the data rate

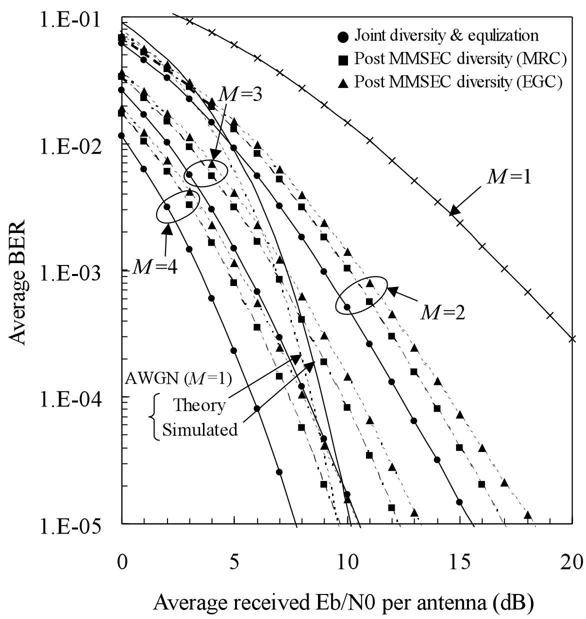


Fig. 5 Average BER performances with joint antenna diversity & equalization and post-MMSEC antenna diversity. Two groups of users are assumed: (SF_0, N_0) = (256, 128) and (SF_1, N_1) = (128, 64). $\Gamma = \Gamma_0 = \Gamma_1$.

of the SF_1=128 group is two times that of the SF_0=256 group and that the average received signal power per subcarrier increases as SF decreases for the given Γ_n since the average received E_b/N_0 per subcarrier is given by Γ_n/SF for the nth user.) Since large spreading factors of SF_0=256 and SF_1=128 are used, sufficient frequency diversity effect is obtained when $\Delta\tau/\Delta T = 4$ [15] and therefore, the two groups achieve almost identical BER performance for the given normalized number of users $\sum_{n=0}^{N-1} N_n/SF_n$. Hence, only the BER results for the SF_0=256 user group are plotted in Fig. 5. However, for the case of SF_1=32 and N_1=16 (i.e., the data rate of SF_1=32 user group is 8 times that of SF_0=256 user group), the SF_1=32 user group may suffer BER performance degradation because the frequency diversity effect reduces. Of course, all users can use the same spreading factor of 256 and apply code-multiplexing to achieve different data rates (i.e., multi-code/multi-rate MC-CDMA). In this case, obviously, all users in two groups can achieve identical BER performance despite of different data rates. However, this is not discussed in this paper. Similar to the case of single class of users, joint antenna diversity & equalization provides better BER performance than the post-MMSEC antenna diversity. It should be noted that the achievable BER performance is almost the same if the normalized number of users is the same if sufficiently large spreading factors are used (compare Fig. 4(a) and Fig. 5).

6. Conclusion

In this paper, the combination of antenna diversity combining and one-tap frequency-domain equalization was considered for the reception of multi-rate MC-CDMA signals in a frequency-selective fading channel. The approach taken in [9] is that antenna diversity combining and frequency-domain equalization are treated separately. However, in this paper, we have shown that antenna diversity combining weight and frequency-domain equalization weight can be unified, leading to joint antenna diversity & frequency-domain equalization. Note that when channel estimation is ideal, the MRC antenna diversity before MMSEC despreading of [9] is equivalent to our joint antenna diversity & equalization.

The achievable BER performances with joint antenna diversity & equalization and post-MMSEC antenna diversity in a frequency-selective Rayleigh fading channel were evaluated by computer simulations. It was found that the joint antenna diversity & equalization can always achieve better BER performance than the post-MMSEC antenna diversity irrespective of the number of diversity antennas; the gain in the required E_b/N_0 value for the average BER=10^{-4} over the post-MMSEC antenna diversity using MRC is about 1.4 dB, 1.6 dB and 1.7 dB when M=2, 3 and 4, respectively. It was also found that when the normalized number of users $\sum_{n=0}^{N-1} N_n/SF_n$ is the same, the same BER performance can be achieved for the different user groups of different spreading factors if sufficiently large spreading factors are used.

In this paper, ideal channel estimation was assumed. When practical channel estimation is used, EGC antenna diversity after MMSEC despreading was found to be the best [9]. However, when channel estimation is ideal, joint antenna diversity & equalization is the best. This suggests that the achievable BER performance with antenna diversity combining and frequency-domain equalization is sensitive to channel estimation errors. Improving channel estimation for the joint antenna diversity & equalization is left as an interesting future study. Another interesting work is a study on combined effect of channel coding and joint antenna diversity & equalization for multi-rate MC-CDMA.

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