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Multiple Access Performance of Parallel Combinatory Spread Spectrum Communication Systems in Nonfading and Rayleigh Fading Channels

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SUMMARY This paper describes the multiple access performance of parallel combinatory spread spectrum (PC/SS) communication systems in nonfading and Rayleigh fading multipath channels. The PC/SS systems can provide the high-speed data transmission capability by transmitting multiple pseudo-noise sequences out of a pre-assigned sequence set. The performance is evaluated in terms of average bit error rate (BER) by numerical computation. In nonfading white gaussian channel, the PC/SS systems are superior to conventional direct sequence spread spectrum (DS/SS) systems under the identical spreading factor condition. In Rayleigh fading channel, the performance of the PC/SS system without diversity is poorer than that of the DS/SS system. By including the explicit and implicit diversity, the performance of the PC/SS system becomes better than that of conventional DS/SS systems. A longer spreading sequence is assignable to a PC/SS system having the spreading factor equal to that in the conventional DS/SS system. Hence, the error control coding is easily. It is found that the PC/SS systems including diversity and Reed-Solomon coding improves the multiple access performance.
key words: *communication theory, radio communication, code division multiple access, Rayleigh fading*

1. Introduction

Recently, spread spectrum (SS) or code-division multiple access (CDMA) systems receive growing attention as a promising way to efficient use of radio frequency. This is especially true in cellular mobile communication systems, indoor wireless communications and consumer communications [1]-[10]. SS techniques offer desirable properties in anti-interference and anti-interception, and these are additional advantages to apply the SS systems in those.

In the future wireless mobile communication systems for personal multimedia systems desirable are two critical factors: high-speed data transmission capability and efficient use of radio frequency. The radio frequency is a limited source, and so there is no guarantee to keep sufficient frequency bandwidth being left for conventional direct sequence spread spectrum (DS/SS) scheme. To this problem, DS/SS systems using orthogonal codes or M-ary/SS systems [7]-[10] have been studied as a possible solution. It provides excellent performance, when many PN sequences are assignable. However, the required length of assigned PN sequences increases as the powers of the number of data bits grows. Thus little performance improvement is achieved if a spreading factor is low. Accordingly, we need to develop an alternative SS scheme with high data rate transmission capability.

For this purpose, *parallel combinatory spread spectrum* (PC/SS) communication systems have been proposed [11]. In the PC/SS system, multiple pseudo-noise (PN) sequences are simultaneously transmitted out of a pre-assigned spreading sequence set. The PN sequences for transmission depend on the state of a set of data bits. Direct sequence spread spectrum system using orthogonal codes or M-ary/SS system [7]-[10] is just as a subset of the PC/SS system. We have studied the PC/SS systems mainly on single user case [11]-[14].

This paper investigates the basic multiple access performance of the PC/SS communication systems in nonfading and Rayleigh fading multipath channels. Particularly, comprehensive treatment of fading and multipath effects in mobile radio channel are considered in channel environment.

In a fading channel, it is known that the diversity technique and the error-control coding are very effective to reduce the performance degradation. Hence, the performance improvement by these techniques is studied in this paper.

In the next section, a transmitter and receiver model of the PC/SS systems are described. In Sect. 3, we analyze the interference of the PC/SS systems in multiple access environment, and calculate the average bit error rate (BER) performance. The analysis covers the performance in nonfading additive white gaussian noise (AWGN) channel and in Rayleigh fading multipath channel. The BER performance of the PC/SS systems combining the error control coding are discussed in Sect. 4. We will conclude our results in Sect. 5.

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2. System Description

In conventional DS/SS systems, a single PN sequence is assigned to each user, and data is transmitted by modulating the assigned PN sequence. In PC/SS systems, a set of M orthogonal PN sequences is assigned to each user. Data is transmitted by modulating R PN sequences that are specified by the R-out-of-M combination among M PN sequences.

Let the system consist of Q users. A particular set of M orthogonal PN sequences is assigned to a user.

$$\boldsymbol{a}^{(q)}=[a_1^{(q)}(t),\ a_2^{(q)}(t),\ a_3^{(q)}(t),\ \cdots,\ a_M^{(q)}(t)] \tag{1}$$

where $a_i^{(q)}(t)$ are a PN sequences for the qth user, and are expressed by

$$a_i^{(q)}(t)=\sum_{j=0}^{N-1} a_{i,j}^{(q)}\, P_{T_c}(t-jT_c),\ a_{i,j}^{(q)}\in\{+1,\ -1\} \tag{2}$$

where $a_{i,j}^{(q)}$ is the jth element of ith spreading sequence for the qth user. N is the length of the assigned sequences. $P_{T_c}(t)$ represents the chip waveform. If it is a rectangular pulse,

$$P_{T_c}(t)=\begin{cases}1 & 0\leqq t<T_c\\ 0 & t<0,\ t\geqq T_c\end{cases} \tag{3}$$

In the transmitter, an input data stream with bit duration T_d are converted to K streams in parallel. The bit duration of a parallel branch is $T(=KT_d=NT_c)$. A part of K branches, K-R branches, are used to specify the R transmitting PN sequences among the M orthogonal PN sequences. The procedure of choosing R sequences for transmission is as follows. At first, the data of K-R bits in the same time-slot are encoded into a constant weight code (CWC) of length M and weight R by a suitable encoding method: e.g. Shalkwijk's constant weight coding method [18]. This is referred to as $(M,\ R)$ constant weight code. The element of a CWC corresponds to the on-off sign for M PN sequences in Eq. (1). Since an $(M,\ R)$ CWC has R nonzero elements, R PN sequences for simultaneously transmission can be determined by choosing PN sequences that correspond to the nonzero element. The set of transmitting R PN sequences for the qth user is expressed by

$$\boldsymbol{v}^{(q)}=(v_1^{(q)}(t),\ v_2^{(q)}(t),\ v_3^{(q)}(t),\ \cdots,\ v_R^{(q)}(t)). \tag{4}$$

$$v_i^{(q)}(t)=\sum_{j=0}^{N-1} v_{i,j}^{(q)}\, P_{T_c}(t-jT_c) \tag{5}$$

The other data of R bits in the same time-slot defines the state of transmitting PN sequences. It is written by

$$\boldsymbol{b}_q=(b_{q1},\ b_{q2},\ b_{q3},\ \cdots,\ b_{qR}). \tag{6}$$

$$b_{qi}\in\{+1,\ -1\}$$

The state word of the transmitting sequences, the carrier and the transmitting PN sequence set are altogether multiplied to form a transmitting signal.

$$s_q(t)=\sum_{i=1}^{R} A_q b_{qi} v_i^{(q)}(t)\cos\omega_c t \tag{7}$$

A_q is the amplitude of ith transmitting PN sequence of the qth user. Every transmitting PN sequence is assumed to have the same power. ω_c is the carrier frequency.

In the receiver, the received signal $r(t)$ passes through M matched filters (or correlators). A matched filter detects an assigned spreading sequence that coincides with the counterpart in a transmitter. The filter output is described as

$$\boldsymbol{Z}=(Z_1,\ Z_2,\ Z_3,\ \cdots,\ Z_M) \tag{8}$$

$$Z_i=\int_0^T r(t)\, a_i^{(q)}(t)\cos\omega_c t dt \tag{9}$$

In the decision block, according to the descending order of the absolute value of Z_i, R elements are decoded to '1,' and the others are decoded to '0.' Noncoherent detection [15] is assumed in this part. This procedure forms an $(M,\ R)$ CWC which estimates R PN sequences in transmission. A part of the data along with the $(K$-$R)$ branches in the same time-slot is obtained by decoding the CWC estimate.

From the matched filter outputs in (8) and the $(M,\ R)$ CWC, the other part of R-bit data is obtained by demodulating the state of those R transmitted PN sequences. Coherent detection is assumed. Finally, the receiver output with bit duration T_d is obtained through parallel to serial conversion.

3. Performance Analysis

In this section, we analyze the multiple access performance of the PC/SS systems in terms of the average bit error rate (BER). We discuss the BER performance in two types of channels. One is a nonfading AWGN channel, and the other is a Rayleigh fading multipath channel.

3.1 BER in Nonfading AWGN Channel

In AWGN channel, the received signal $r(t)$ consists of the signals transmitted by Q users with different PN sequence sets and noise. It is expressed by

$$r(t)=\sum_{q=1}^{Q} s_q(t-\tau_q)+\eta(t) \tag{10}$$

where $\eta(t)$ is the AWGN component of which power spectral density is $N_0/2$.

In multiple access environment, the output signal of a matched filter consists of a desired signal, interference and thermal noise. Let us consider the receiver of the first user. The matched filter output Z_i in (9) is expressed as follows.

$$Z_i = D_i + I_s(i) + I_c(i) + \eta_i \tag{11}$$

$$D_i = \frac{1}{2} A_{1i} T b_{1i} \tag{12}$$

$$I_s(i) = \frac{1}{2} A_1 \sum_{\substack{j=1 \\ j \neq i}}^{R} [b_{1j}^{(-1)} R_{i1,j1}(0) + b_{1j}^{(0)} \hat{R}_{i1,j1}(0)] = 0 \tag{13}$$

$$I_c(i) = \frac{1}{2} \sum_{q=2}^{Q} A_q \cos\phi_q \sum_{j=1}^{R} [b_{qj}^{(-1)} R_{i1,jq}(\tau_q) + b_{qj}^{(0)} \hat{R}_{i1,jq}(\tau_q)] \tag{14}$$

$$\eta_i = \int_0^T \eta(t)\, a_i^{(1)}(t) \cos\omega_c t dt \tag{15}$$

D_i represents the desired component. $I_s(i)$ is the *self-interference* from other transmitted sequences of the desired user, and $I_c(i)$ is the multiple access interference (MAI) from undesired users. η_i is the AWGN term of the matched filter output. Since orthogonal PN sequences are assigned for a user, the *self-interference* $I_s(i)$ is zero in this case.

$b_{qj}^{(-1)}$ and $b_{qj}^{(0)}$ are the previous and present states of jth transmitted PN sequence for qth user, respectively. They are $+1$ or -1 with equal probability. τ_q and ϕ_q are the difference in arrival time and phase between the first user and the qth user, respectively. It is clear that τ_1 and ϕ_1 are 0. $R_{iq,jq}(\tau_{lq})$ and $\hat{R}_{iq,jq}(\tau_{lq})$ are the partial crosscorrelation functions defined in [16]:

$$R_{iq,jq}(\tau) = \int_0^{\tau} v_{jq}(t-\tau)\, a_{iq}(t)\, dt \tag{16}$$

$$\hat{R}_{iq,jq}(\tau) = \int_{\tau}^{T} v_{jq}(t-\tau)\, a_{iq}(t)\, dt \tag{17}$$

Since the PN sequence used here is random, the interference term is approximated by a zero mean gaussian random variable [16], [17]. The variance of the MAI is expressed by

$$\begin{aligned} \mathrm{Var}\{I_c(i)\} &= E\{I_c(i)^2\} \\ &= \frac{1}{4} E\left\{\sum_{q=2}^{0} (A_q \cos\phi_q)^2\right\} \sum_{j=1}^{R} E\{[b_{qj}^{(-1)} \\ &\quad R_{iq,jq}(\tau_q) + b_{qj}^{(0)} \hat{R}_{iq,jq}(\tau_q)]^2\} \end{aligned} \tag{18}$$

The variance of the crosscorrelation terms in the second factor in the above expression are obtained by a similar way for the conventional DS/SS scheme [16], [17]. That is

$$E\{[b_{qj}^{(-1)} R_{iq,jq}(\tau_q) + b_{qj}^{(0)} \hat{R}_{iq,jq}(\tau_q)]^2\} = \frac{2T^2}{3N}. \tag{19}$$

It is assumed that the received power of all interfering users is equal to that of a desired user ($A_1^2 = A_q^2 = A^2$). Then, the variance of MAI is expressed as follows:

$$\mathrm{Var}\{I_c(i)\} = \frac{(Q-1) R A^2 T^2}{12N} \tag{20}$$

The variance of the AWGN component is $N_0 T/4$. Thus, the total variance of Z_i is

$$\mathrm{Var}(Z_i) = \frac{(Q-1) R A^2 T^2}{12N} + \frac{N_0 T}{4} \tag{21}$$

Using (12) and (21), the average SNR at the matched filter output is expressed by

$$\gamma_0 = \frac{1}{2} \frac{D_i^2}{\mathrm{Var}(Z_i)} = \left[\frac{2R(Q-1)}{3N} + \frac{1}{\gamma_n}\right]^{-1} \tag{22}$$

$\gamma_n = A^2 T / 2N_0$ stands for the SNR in the absence of interference. Strictly speaking, γ_0 and γ_n represents a half of the SNR at the matched filter output. We will call these simply SNR. γ_0 has the relation with the SNR per information bit γ_b as follows [11]:

$$\gamma_0 = \frac{K}{R}\gamma_b = \frac{R + \lfloor \log_2({}_M C_R) \rfloor}{R}\gamma_b \tag{23}$$

where $\lfloor x \rfloor$ stands for the largest integer not exceeding x.

The average BER is calculated as follows (See Appendix.)

$$\begin{aligned} P_{eg} &= \frac{1}{2K} \sum_{x=1}^{R} [(R-x)\,\mathrm{erfc}(\sqrt{\gamma_0}) \\ &\quad + \{x - (R-x)\,\mathrm{erfc}(\sqrt{\gamma_0})\} P_{cc1}^{R-x} (1-P_{cc1})^x] \\ &\quad + \frac{K-R}{2K}(1 - P_{cc1}^R) \end{aligned} \tag{24}$$

$$P_{cc1} = \sum_{n=1}^{M-R} (-1)^n \binom{M-R}{n} \frac{1}{n+1} \exp\left(-\frac{n\gamma_0}{n+1}\right) \tag{25}$$

3.2 BER in Rayleigh Fading Multipath Channel

In the case of multipath fading channel, the channel is assumed to be discrete and time-invariant. The channel impulse response of a user is given by

$$h_q(\tau) = \sum_{l=1}^{L} \beta_{lq} \delta(\tau - \tau_{lq}) \exp(j\phi_{lq}). \tag{26}$$

where β_{lq}, τ_{lq}, ϕ_{lq} are the path gain, delay and phase of lth path in qth user, respectively. It is assumed that ϕ_{lq} is uniformly distributed over $[0, 2\pi]$. β_{lq} has the Rayleigh distribution. L stands for the number of resolvable paths that is given by

$$L = \left\lfloor \frac{T_m}{T_c} \right\rfloor + 1. \tag{27}$$

where T_m stands for the multipath delay spread. To simplify the analysis, we assume that every user has the same number of multipaths and all paths are mutually independent.

The received signal consists of multipath faded signals and noise. The received signal is expressed by

$$r(t) = \sum_{q=1}^{Q} \sum_{l=1}^{L} \beta_{lq} \exp(j\phi_{lq})\, s_q(t - \tau_{lq}) + \eta(t) \tag{28}$$

In the multipath environment, the matched filter output suffers from the *self-interference*, MAI and noise. Let us consider the first user. In this case, the terms of D_i, $I_s(i)$ and $I_c(i)$ in (11) are expressed by

$$D_i=\frac{1}{2}\beta_{11}A_1Tb_{1i} \tag{29}$$

$$I_s(i)=\frac{1}{2}A_1\sum_{l=2}^{L}\sum_{j=1}^{R}\beta_{l1}\cos\phi_{l1}[b_{1j}^{(-1)}R_{i1,j1}(\tau_{l1}) + b_{1j}^{(0)}\hat{R}_{i1,j1}(\tau_{l1})] \tag{30}$$

$$I_c(i)=\frac{1}{2}\sum_{q=2}^{Q}A_q\sum_{l=1}^{L}\sum_{j=1}^{R}\beta_{lq}\cos\phi_{lq}[b_{qj}^{(-1)}R_{i1,jq}(\tau_{lq}) + b_{qj}^{(0)}\hat{R}_{i1,jq}(\tau_{lq})] \tag{31}$$

τ_{lq} and ϕ_{lq} stands for the difference of arrival time and phase between the first path of the first user and the lth path of the qth user, respectively. It is clear that τ_{11} and ϕ_{11} are 0. It is also assumed that $+1$ or -1 has the equal probability.

The *self-interference* consists of the $(L-1)$ interfering paths from the delayed PN sequences themselves and $(R-1)(L-1)$ interfering paths from the other transmitted PN sequences in a desired user. Since an orthogonal PN sequence set is assigned for every user, the *self-interference* of the first path is zero. Thus, there are $R(L-1)$ interfering paths as the *self-interference* to a desired user. When τ_{lq} is close to T, the auto-correlation function for two bits could overlap with each other. However, this cannot happen, unless T_m, delay spread, approaches to T, that is just K times of the input data duration T_d. On the other hand, if τ_{lq} is small, the auto-correlation function for the same bit might overlap. Also however, as long as τ_{lq} is smaller than chip duration T_c, path separation is impossible in principle, and thus this effect is supposed to be accounted in β_{l1} in Eq. (30). Hence, for rough performance estimation, the *self-interference* can be calculated in the gaussian approximation.

Since a PC/SS system transmits R PN sequences simultaneously to one user, any matched filter output at a specific user can be interfered by RL PN sequences as MAI from every undesired user. R MAI components from a user have the same path gain and phase difference.

To simplify the interference analysis, the gaussian approximation is adopted to calculate the self and multiple access interferences as in Ref. [3], [17]. The variance of the *self-interference* and MAI terms are

$$E[I_s(i)^2]=\frac{A_1^2}{4}E\left[\sum_{l=2}^{L}\beta_{l1}^2\cos^2\phi_{l1}\sum_{j=1}^{R}\{b_{1j}^{(-1)}R_{i1,j1}(\tau_{l1})+b_{1j}^{(0)}\hat{R}_{i1,j1}(\tau_{l1})\}^2\right] \tag{32}$$

$$E\{I_c(i)^2\}=\sum_{q=2}^{Q}\frac{A_q^2}{4}E\left[\sum_{l=1}^{L}\beta_{lq}^2\cos^2\phi_{lq}\sum_{j=1}^{R}\{b_{qj}^{(-1)}R_{i1,jq}(\tau_{lq})+b_{qj}^{(0)}\hat{R}_{i1,jq}(\tau_{lq})\}^2\right] \tag{33}$$

Since the gaussian approximation is applied to the *self-interference* computation, it leads to the simplified elimination of τ_{lq}. Using the calculation of crosscorrelation function in (19), (32) and (33) are written as

$$E[I_s(i)^2]=\frac{RA_1^2T^2}{12N}E\left\{\sum_{l=2}^{L}\beta_{l1}^2\right\} \tag{34}$$

$$E\{I_c(i)^2\}=\frac{RT^2}{12N}E\left\{\sum_{q=2}^{Q}A_q^2\sum_{l=1}^{L}\beta_{lq}^2\right\} \tag{35}$$

In this paper, it is assumed that the average received power of every interfering users are equal to that of a desired user, and the average received power is equal in all paths. Then applying the analysis of [3] and [17],

$$E[I_s(i)^2]=\frac{R(L-1)T^2}{12N}E(\beta^2) \tag{36}$$

$$E[I_c(i)^2]=\frac{RL(Q-1)T^2}{12N}E(\beta^2) \tag{37}$$

Accordingly, the total variance of noise and interference term is expressed by

$$\mathrm{Var}(Z_i)=\frac{R(L-1)A^2T^2}{12N}E(\beta^2)+\frac{(Q-1)RLA^2T^2}{12N}E(\beta^2)+\frac{N_0T}{4}. \tag{38}$$

The average SNR per matched filter output of the transmitted sequence is expressed by

$$\bar{\gamma}_0=\frac{E(D_i^2)}{\mathrm{Var}(Z_i)}=\left(\frac{2R(LQ-1)}{3N}+\frac{1}{\bar{\gamma}_n}\right)^{-1} \tag{39}$$

In a multipath channel, the error rate performance of the SS system can be improved by using the explicit and implicit diversity. The implicit diversity effect is achieved by using RAKE receivers. In this paper, the received signal on every branch is assumed to be equally distributed and mutually independent. It is also assumed that the maximal ratio combining is applied. The probability density function of the received SNR is given by [15]

$$p(\gamma)=\frac{\gamma^{DP-1}}{(DP-1)!\,\bar{\gamma}_c^{DP}}\exp\left(-\frac{\gamma}{\bar{\gamma}_c}\right) \tag{40}$$

where D and P is the order of explicit and implicit diversity, respectively. $\bar{\gamma}_c$ is the average SNR per diversity branch. In this paper, we assume that the average path gain is equal for all diversity branches. This is related to the average SNR per transmitted PN sequence;

$$\bar{\gamma}_0=P\bar{\gamma}_c. \tag{41}$$

The BER in (24) is the function of SNR per transmitted PN sequence γ_0. From (24) and (40), the BER of the PC/SS systems in the multipath fading channel is expressed by

$$P_{eb}=\int_0^{\infty} P_{eg}(\gamma_0)\,p(\gamma_0)\,d\gamma_0. \tag{42}$$

3. 3 Numerical Results and Discussion

The multiple access performance of the PC/SS systems is evaluated through BER performance and is compared with that of conventional DS/SS systems.

Figure 1 displays the BER curves against the number of simultaneous users without error control coding or diversity in nonfading AWGN channel and flat (frequency-nonselective) Rayleigh fading channel. Here we have two cases of the PC/SS systems as follows:

(a) $M=16$, $R=3$, $K=12$, $N=256$
(b) $M=48$, $R=2$, $K=12$, $N=256$

The spreading factor of the PC/SS system without error control coding is expressed by N/K as implied in Sect. 2. In Fig. 1, the spreading factor is set at 21.33. For comparison, the BER of conventional DS/SS systems with $N=21$ is also plotted.

In nonfading AWGN channel, the BER of the PC/SS systems is superior to that of conventional DS/SS systems. For PC/SS systems, case (b) is better than (a). This is because the SNR per transmitted PN sequence in (b) is larger than that in (a), owing to Eq. (23).

In Rayleigh fading channel, BER difference between two PC/SS systems is very small. Unfortu-

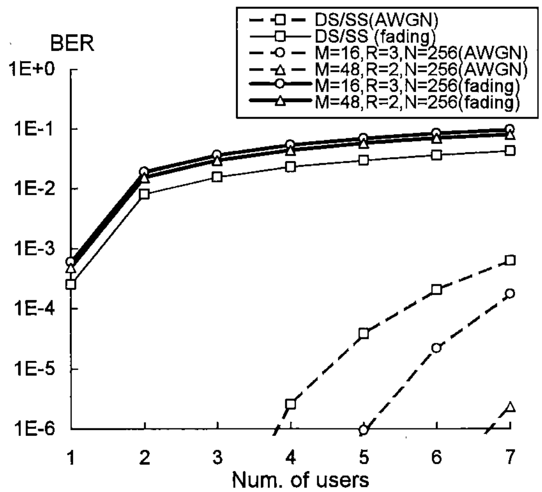

Fig. 1 BER performance of PC/SS systems in AWGN channel and flat (frequency-nonselective) Rayleigh fading channel. $N/K=21.33$, no explicit and implicit diversity, average SNR/bit $=30$ dB.

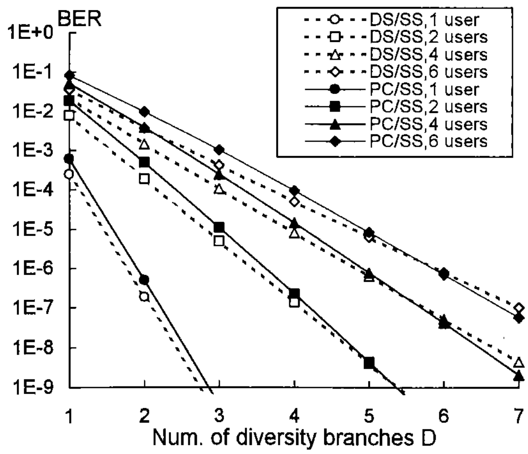

Fig. 2 The effect of explicit diversity to the BER of PC/SS systems in flat Rayleigh fading channel. $L=1$, $P=1$ (no implicit diversity) and average SNR/bit/branch$=30$ dB. PC/SS system ($M=16$, $R=3$, $K=12$, $N=256$) in solid line. DS/SS system ($N=21$) in dotted line.

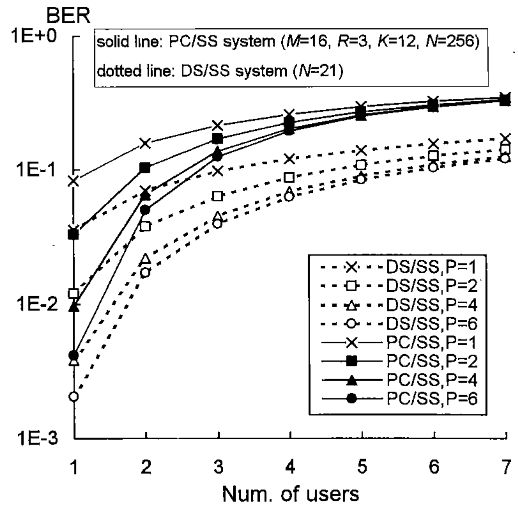

Fig. 3 BER performance of the PC/SS systems for no explicit diversity, P-th order of implicit diversity. $L=6$, and average SNR/bit$=30$ dB.

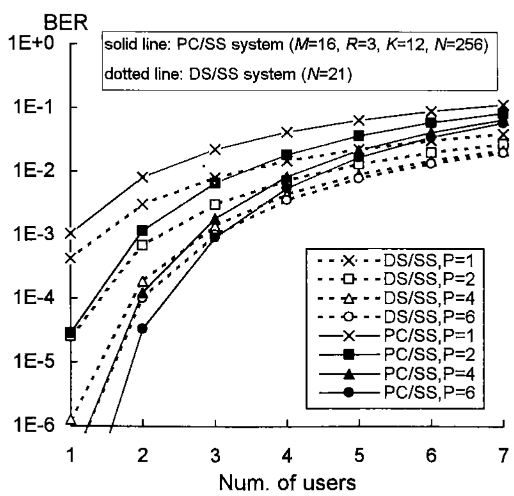

Fig. 4 BER performance of the PC/SS systems for 3-order of explicit diversity, P-th order of implicit diversity. $L=6$, and average SNR/bit$=30$ dB.

nately, the BER of the PC/SS system is slightly worse than conventional DS/SS systems in the fading channel. This is because the performance degradation in low SNR is much larger than that of DS/SS systems. Both the PC/SS systems and the DS/SS systems have much more interfering paths in a multipath fading environment than those in a flat fading environment; $L-1$ times many *self-interference* and L times many multiple access interferences. Especially, explicit or implicit diversity is required for the PC/SS systems to provide better BER performance than the DS/SS systems. (After this we evaluate the performance about case (a) only in the PC/SS systems.)

Figure 2 shows the effect of the explicit diversity in flat Rayleigh fading channel. As increasing the order of explicit diversity, the PC/SS systems can achieve more BER improvement than that in conventional DS/SS systems. We can find that lower BER can be provided by the PC/SS system with more than 6-order of diversity comparing to the DS/SS system. This is insensitive to the number of simultaneous users.

Figures 3 and 4 display the BER comparison between the PC/SS system and conventional DS/SS system for different orders of implicit diversity in the Rayleigh fading 6-path channel. Figure 3 comes from the case without explicit diversity, and Fig. 4 does from the case with 3 orders of explicit diversity. With no explicit diversity, the PC/SS systems cannot provide superior BER performance to the DS/SS systems. However in the PC/SS systems with 3 orders of explicit diversity, as found in Fig. 4, lower BER performance can be achieved comparing to that of conventional DS/SS system when the orders of implicit diversity exceeds 4. Unlike in the flat fading channel, the diversity effect depend on the number of simultaneous users. For example, in the case of $P=4$, the BER of the PC/SS system is less than that of the DS/SS system under 2-user operation.

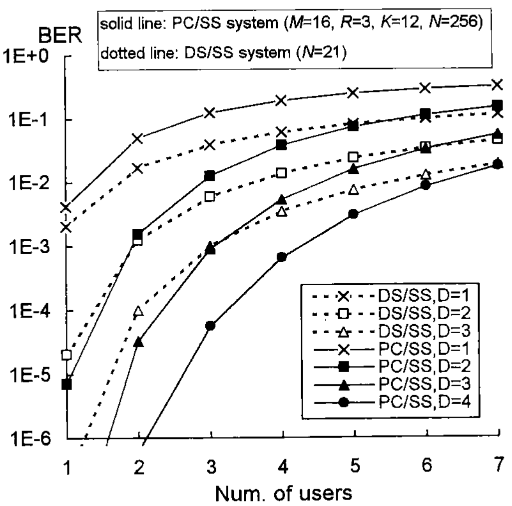

Fig. 5 BER of the PC/SS systems against the order of explicit diversity D. $L=6$, $P=6$ and average SNR/bit=30 dB.

Figure 5 shows the BER comparison for different orders of explicit diversity in multipath channels. The number of multipaths and the order of implicit diversity are both set to 6. Compared to the DS/SS systems, lower BER is achievable by the 3 explicit diversity PC/SS systems under the operation in the presence of three or less users. Especially, at most 1/3 rate of BER is available under the 2-user operation.

The BER reduction by diversity techniques in PC/SS systems is larger than that of conventional DS/SS systems without error-control coding. But the performance advantage over the DS/SS system is not so large in spite of complex receiver configuration involved with diversity. In the next section, we will apply the error control coding to reduce the BER and the complexity in receiver configuration.

4. Performance Improvement by Reed-Solomon Coding

We discuss the effect of error control coding for the multiple access performance of the PC/SS systems. If spreading factor is small, it is difficult to apply error control coding to conventional DS/SS system. Since some spectrum spreading by coding is inevitable, it is impossible to configure sufficient PN sequences with a CDMA system. On the contrary, in PC/SS systems, since the spreading factor is given by $(T_c/T_d)=(N/K)$, longer PN sequences is assignable if the spreading factor is the same with that in DS/SS systems. The error control coding is easily applicable than conventional DS/SS systems.

Since the PC/SS system transmits K bits of data during one PN period, 2^K-ary Reed-Solomon coding brings about a significant improvement in error rate performance in AWGN channel (See Ref. [13]).

Applying bounded distance decoding, the BER is given by

$$P'_{eb}=\frac{1}{2n}\left\{\sum_{j=2m+1}^{n} jW_j \sum_{i=m+1}^{n}\sum_{d=0}^{m}\phi(i,j,d)P_{e1}^{i}P_c^{n-i} + \sum_{j=m+1}^{n}\left({}_nC_i q^i-\sum_{i=m+1}^{n}\sum_{d=0}^{m}\phi(i,j,d)W_j\right) iP_{e1}^{i}P_c^{n-i}\right\} \tag{43}$$

where

$$\phi(i,j,d)=\begin{cases} {}_jC_x\cdot{}_{n-j}C_{d-x} \\ \quad x=(j-i+d)/2\text{: integer \&} \\ \quad \min[2n-i-j,\ i+j]\geqq d|j-i| \\ 0 \quad \text{otherwise} \end{cases} \tag{44}$$

$$P_c=P_{cc1}^{R} \tag{45}$$

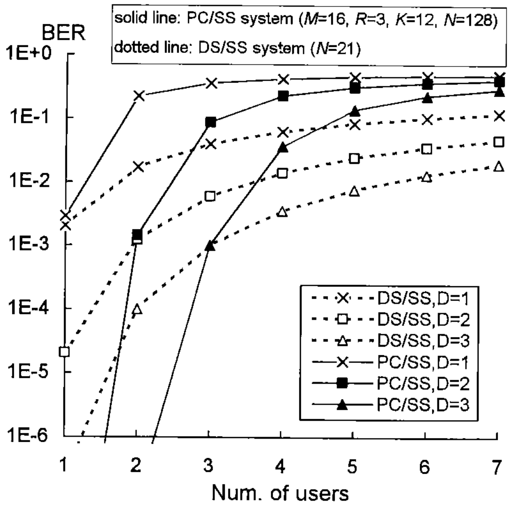

Fig. 6 BER performance of the PC/SS systems with (16, 8)-RS coding. $L=6$, $P=6$ and average SNR/information bit=30 dB.

$$P_{e1} \approx (1-P_{cc1}^{R})/(2^{K}-1) \tag{46}$$

W_j is the number of codes with weight $j.n$ is the codeword length. m means error correction capability of the RS code. The number of information symbols h is given by $n-2m$. This RS code is denoted by (n, h)-RS code.

Figure 6 shows the BER of the (16, 8)-RS coding PC/SS system. To give the same total spreading factor in previous computations, we set the PN length N at 128. The total spreading factor is actually around 21, and the average SNR per information bit is 30 dB, which are both the same as those in the case of Figs. 1 and 2. Suppose a six multipath environment on the PC/SS systems of implicit diversity. As found in Figs. 5 and 6, the BER of the PC/SS system are better than that of conventional DS/SS system when the required BER is less than 10^{-3}. By applying the RS coding under this requirement, much more BER improvement can be achieved in proportion to the number of explicit diversity branches. Let us consider the 2-simultaneous user operation, and let us see the 3-path explicit diversity plots in Fig. 5. BER of the PC/SS system is about 1/3 of a DS/SS system. If RS coding is used, see Fig. 6, BER of the PC/SS system drastically decreases to less than 1/100 of the DS/SS system. In order to clear the level of 10^{-6} BER, 4-path explicit diversity is needed, unless RS coding is used. By contrast, just 3-path explicit diversity suffices to maintain the same BER quality, if (16, 8)-RS coding is applied. RS coding is thus useful to reduce the number of explicit diversity branches in PC/SS receiver.

Figures 7 and 8 show BER performance versus the average SNR per information bit in the case of 2 simultaneous users. As same as in Fig. 6, the number of multiple paths and the orders of implicit diversity

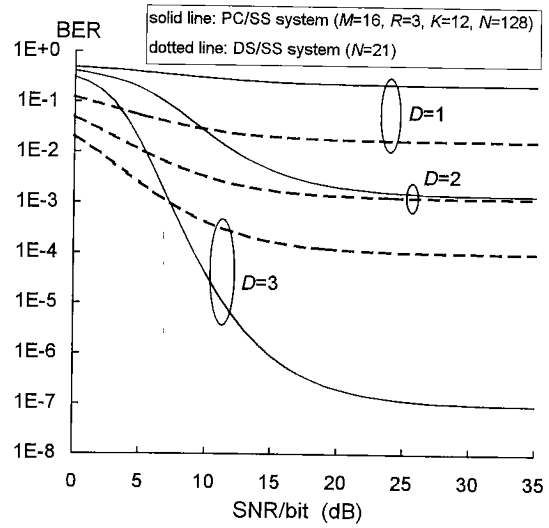

Fig. 7 BER of the PC/SS systems with (16, 8)-RS coding under the condition of 2-user, 6 paths and 6-implicit diversity.

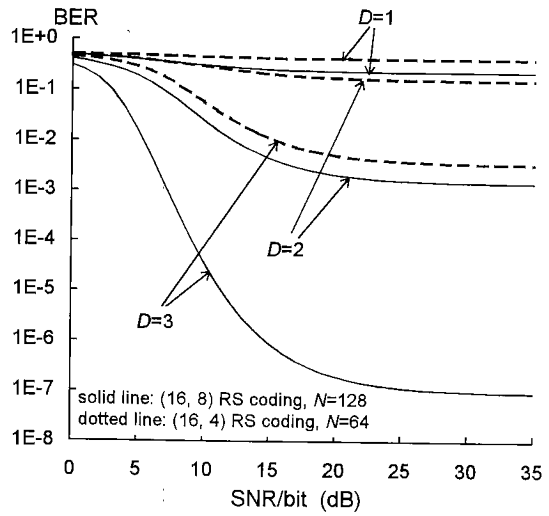

Fig. 8 BER comparison of the PC/SS systems with RS coding for different coding rate. $M=16$, $R=3$, $K=12$, 2 users, 6 paths and 6-implicit diversity.

are both six.

Because of the SNR penalty caused by MAI, the BER of both PC/SS and DS/SS systems approaches to a constant against increase of SNR. In the cases of no explicit diversity and 2-order of explicit diversity, the BER of the RS-coded PC/SS system are still worse than conventional DS/SS systems. As found in the 3-order explicit diversity plots. As found in Fig. 7, the RS-coded PC/SS system gives better performance than conventional DS/SS systems when the average SNR per information bit is more than 8 dB. This crossing point indicates around 10^{-3} BER, which agrees with the results in Figs. 5 and 6. To maintain the level of 10^{-6} BER in the RS-coded PC/SS system under 2-user operation, only 15 dB average SNR per information bit is required.

Figure 8 displays the BER comparison between RS coded PC/SS systems with different coding rates. One is a half rate RS coded system, and the other is a quarter rate RS coded system. In spite of less error correction capability, the half-rate coded system shows much better BER performance than the quarter-rate coded system. The reason is as follows. To keep the equivalent spreading factor condition, a half-length PN sequences must be assigned in quarter-rate coded PC/SS systems comparing to half-rate coded systems. As seen from Eq. (19), this means that the self and multiple access interferences per transmitted PN sequence grow twice. Accordingly, excessively large self and multiple access interferences can produce a large SNR penalty so that it can defeat the error correction capability.

5. Conclusions

The multiple access performance of the PC/SS systems was investigated in nonfading and multipath Rayleigh fading environment.

In nonfading AWGN channel, the PC/SS systems provide the considerable improvement in multiple access performance over the conventional DS/SS systems. The error control coding is not always necessary in this case.

The multiple access performance of the PC/SS systems is a little bit worse than that of conventional DS/SS systems in Rayleigh fading channel in the absence of diversity. However, the BER performance of the PC/SS system including explicit and implicit diversity technique is better than that of conventional DS/SS systems.

Applying Reed-Solomon coding to the PC/SS system, significant improvement in BER performance is achievable. If available spreading factor is small, it is difficult for conventional DS/SS systems to cope with the multiple access environments. On the contrary, PC/SS systems are expected to implement the CDMA system if error control coding and diversity technique are combined in the implementation.

Simple and efficient system implementation, the performance evaluation in more realistic environment, and more efficient error control are left as further research.

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Appendix: Derivation of the bit error rate in Eqs. (24)-(25)

The BER of the PC/SS systems is given in Ref. [12]. But it is somewhat different from the case in this paper. The BER in [12] gives more pessimistic results than simulated BER performance in low SNR case. In this appendix, we derive the BER while considering the error pattern estimation of transmitted PN codes.

In the mapping method described in Sect. 2, R bits are used to represent the polarity, and K-R bits are used for choosing a set of spreading sequences to be simultaneously transmitted.

To compute the bit error probability of the PC/SS system, we should compute the data error probability depending on both the combination of transmitting sequences and the polarities of transmitting sequences.

First, let us investigate the data error probability depending on the combination of PN sequences. If all of the R absolute values of matched filter outputs for transmitted PN sequences are larger than those of not transmitted sequences, the correct estimation of PN code combination is performed. In this case, it is sufficient to consider the correct decision probability of a transmitted PN sequence P_{cc1}.

P_{cc1} is obtained by applying the analysis in M-ary orthogonal signaling with noncoherent detection [15]. The correct decision probability of a transmitted PN sequence is computed by

$$P_{cc1}=\sum_{n=1}^{M-R}(-1)^{n}\binom{M-R}{n}\frac{1}{n+1}\exp\left(-\frac{n\gamma_0}{n+1}\right) \tag{A·1}$$

γ_0 is the signal to noise ratio of a transmitted PN sequence. Bit error probability of the data depending on the combination of PN sequences is easily computed from (A·1). That is approximately expressed by

$$P_{ebc}\cong(1-P_{cc1}^{R})/2 \tag{A·2}$$

Next, we consider the error probability of the data depending on the $+$ or $-$ state of transmitted PN sequences. If error occurred in the estimation of the transmitted PN sequence set, some signatures of the transmitted sequences will be made errors. The amount of the case that x errors are occurred is written by ${}_{R}C_{x}$. The probability of the case that x errors are occurred is given by

$$P_{cc1}^{R-x}(1-P_{cc1})^{x}. \tag{A·3}$$

Note on a transmitted PN sequence. The number of patterns that the estimation is made error is expressed by ${}_{R-1}C_{x-1}$.

The decision error probability of a $+$ or $-$ state with correct and wrong estimation of transmitted PN sequence are expressed by

$$\begin{cases}\frac{1}{2}\mathrm{erfc}(\sqrt{\gamma_0}) & \text{correct estimation}\\ \frac{1}{2} & \text{wrong estimation}\end{cases} \tag{A·4}$$

Thus, when x estimation errors of transmitted PN sequences are occurred, the bit error probability of $+$ and $-$ state is computed by

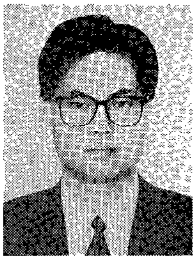
$$\frac{1}{2}\frac{x}{R}P_{cc1}^{R-x}(1-P_{cc1})^{x}+\left\{\frac{1}{2}\mathrm{erfc}(\sqrt{\gamma_0})\right\}\frac{R-x}{R}P_{cc1}^{R-x}(1-P_{cc1})^{x} \tag{A·5}$$

Taking the sum of x, the average bit error probability depending on the state is computed by

$$\begin{aligned}P_{ebs}&=\sum_{x=0}^{R}\left[\frac{1}{2}\frac{x}{R}P_{cc1}^{R-x}(1-P_{cc1})^{x}+\left\{\frac{1}{2}\mathrm{erfc}(\sqrt{\gamma_0})\right\}\frac{R-x}{R}P_{cc1}^{R-x}(1-P_{cc1})^{x}\right]\\&=\frac{1}{2R}\sum_{x=0}^{R}[\{x+(R-x)\mathrm{erfc}(\sqrt{\gamma_0})\}P_{cc1}^{R-x}(1-P_{cc1})^{x}]\end{aligned} \tag{A·6}$$

The average bit error rate can be obtained by taking the weighted sum of error rate in (A·3) and (A·6). That is

$$\begin{aligned}P_{eg}&=\frac{R}{K}P_{ebs}+\frac{K-R}{K}P_{ebc}\\&=\frac{1}{2K}\sum_{x=0}^{R}[\{x+(R-x)\mathrm{erfc}(\sqrt{\gamma_0})\}P_{cc1}^{R-x}(1-P_{cc1})^{x}]+\frac{K-R}{2K}(1-P_{cc1}^{R})\end{aligned} \tag{A·7}$$



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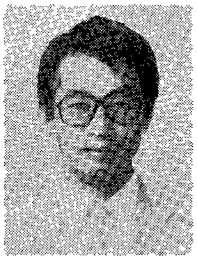

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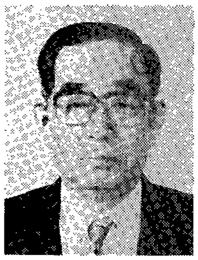

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