Performance of a Multitone CDMA System with Interference Canceller in a Multipath Fading Channel

*Seung-Keun Park, **Byeong-Gwon Kang and *Hee-Chang Chung

Abstract

In this paper, we analyze the effects of interference canceller on the performance of multitone DS/CDMA system proposed by Vandendorpe[5]. There are various kinds of interference canceller suggested by different researchers including parallel and successive cancellers and we adopt a canceller used by Yoon et al.[9] which is a kind of parallel canceller. We consider three kinds of interferences, that is, multipath interference(MPI), interchannel interference(ICI) and multiple access interference(MAI). The ICI is the interference between multitones. The equations for variances are derived for the inteferences and thermal noise used for signal to noise ratio calculation. We also consider RAKE reception over multipath channel which is modeled as lowpass equivalent linear filter and three stage interference canceller used for performance improvement. We show the performance results for number of canceller stage, diversity order and number of users and draw some conclusions that interference canceller is effective in multitone DS/CDMA system and the performance is further improved with the higher order of diversity and larger number of PN chips.

I. Introduction

Multicarrier code division multiple access (CDMA) systems have gained considerable interests in recent years in the area of wireless personal and multimedia communications [1]. It is mainly because we can get the higher rate data transmission, bandwidth efficiency and interference reduction from the combination of multicarrier technique and CDMA scheme. Four kinds of combination types of code division and multicarrier scheme have been proposed since 1993, such as multicarrier(MC) CDMA, multicarrier (MC) DS-CDMA, multitone(MT) CDMA and narrowband multicarrier (NMC) DS-CDMA.

These four schemes were proposed by different persons, that is, MC-CDMA by N. Yee, J-P. Linnartz and G. Fettweis[2], K. Fazel and L. Papke[3], MC DS-CDMA by V. Dasilva and E. S. Sousa[4], MT CDMA by L.Vandendorpe[5], and NMC DS-CDMA by S. Kondo and L. B. Milstein[6]. The system explanations of the first three schemes were presented by R. Prasad and S. Hara[1] with performance comparison, and we now describe the key characteristics of each of the four schemes.

MC CDMA transmitter spreads the original data stream over different subcarriers using a given spreading code in the frequency domain. Each PN code chip modulates different subcarrier. Thus we can see the PN sequence aligned in the frequency domain and each subcarrier spectrum bandwidth is the same with that of original data. MC DS CDMA transmitter spreads the serial-to-parallel converted data streams using a given spreading code in the time domain so that the spectrum of each subcarrier can satisfy the orthogonality condition with the minimum frequency separation. This scheme can lower the transmission data rates in each subcarrier so that a relatively larger chip time makes it easier to synchronize the spreading sequences. MT CDMA transmitter also spreads the serial-to-parallel converted data streams using a given spreading code in the time domain so that the spectrum of each subcarrier before spreading operations can satisfy the orthogonality condition with the minimum frequency separation. However, the resulting spectrum of each subcarrier no longer satisfies the orthogonality condition because of longer spreading codes with relatively higher chip rates according to the number of subcarriers and severe overlapping between subcarrier spectra. In the NMC DS CDMA system, the available frequency spectrum is divided into M equi-width frequency bands, and each frequency band is used to transmit a narrowband DS waveform without overlapping between each band spectrum. In this paper, we consider only the MT CDMA system.

Meanwhile, the capacity of a cellular CDMA system is mainly limited by a variety of interferences. And so, many papers have been proposed and investigated various interference cancellation schemes involving successive cancellation [7][8][9], and parallel cancellation [10]. We derive and analyze the performance of a BPSK modulated multitone DS/CDMA communication system with interference cancellation scheme proposed by Y. C. Yoon et al.[9] in

* Wireless standard Section PEC ETRI
** Department of Electrical and Electronic Engineering Soonchunhyang University

Manuscript Received : October 8, 1998.

this paper. Various interference terms are considered including multipath interfernce(MPI) term, interchannel interference (ICI) term due to overlapping between carrier spectra, multiple access interference(MAI) term and background Gaussian noise. Vandendorpe[5] presented performances of a QPSK modulated MT CDMA system, however, the results are not correct. That is, the author insisted that the performance was improved as the number of carriers increased, however, the performances were almost same for various carrier numbers, for example carrier numbers M = 4 and 32 with fixed bandwidth[11]. We can also see the similar results in [6]. And another technical schemes are needed for performance improvement in the MT CDMA system, for example, convolutional coding and interference cancellation. Also the paper submitted by Y. C. Yoon et al.[9] have some faults in deriving the interference variances and performance results which show too good performance to be processed in a multipath fading channel.

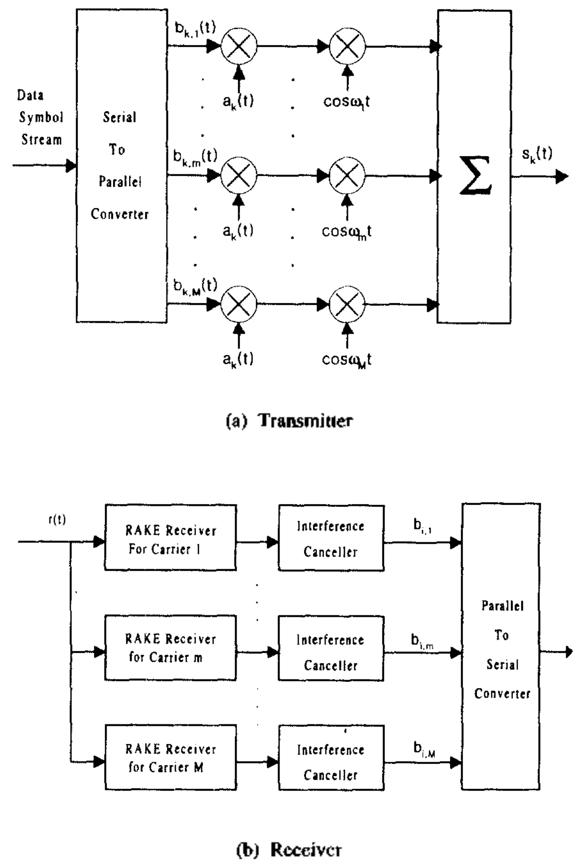
In this paper, we adopt the cancellation technique of [9] to the BPSK multitone CDMA system and show performance improvement by canceling the interferences. Also we accurately derive the performance equations and numerical results. The next section describes the system and channel model, especially the receiver model with correlator outputs. Section III gives the analysis of the canceller performance in the MT CDMA receiver system. Section IV gives numerical results from theoretical calculation. And the conclusions are drawn in Section V.

II. System and Channel Model

In this communication system, we assume that there are K mobile stations asynchronously transmitting over their individual multipath fading channel to a single base station receiver. This section describes the system's transmitter, channel and receiver models.

2.1 Transmitter Model

We consider reverse link of a CDMA cellular system with K active users. The models of transmitter and receiver of a MT DS CDMA system are shown in Fig. 1. and the transmitter model is similar to that of Vandendorpe [5]. The input data symbol stream at the rate of N_t/T_b is split into Nt parallel streams of symbol duration T_b by serial-to-parallel converter. The *m*th symbol stream modulates a tone with a frequency fm. The carriers are orthogonal on the symbol duration and hence are given by $f_m = f_0 + m/T_b$ where T_b is the symbol duration on that carrier. We can consider $f_0 = 0$ for convenience. The transmitted signal for user k can be described by the sum of multiple subcarriers.



(a) Transmitter

(b) Receiver

Figure 1. Model of multitone CDMA communication system.

$$s_k = \sqrt{2P_k}\, a_k(t-T_k) \sum_{m=1}^{M} b_{k,m}(t-T_k)\cos(2\pi f_m(t-T_k)+\Psi_k) \quad (1)$$

where

$$b_{k,m}(t) = \sum_{n=-\infty}^{\infty} b_n^{(k)} P_{T_b}(t-nT_b) \quad (2)$$

$$a_k(t) = \sum_{n=-\infty}^{\infty} a_n^{(k)} P_{T_c}(t-nT_c). \quad (3)$$

P_k and $a_k(t)$ are the transmitted power and the PN sequence, respectively, and $b_{k,m}(t)$ is the data symbol on the *m*th carrier for the kth user. $P_T(t)$ is the unit rectangular pulse defined as $P_T(t) = 1$ for $0 \le t < T$ and $P_T(t) = 0$ otherwise. M is the total number of carriers and f_m means the frequency related with *m*th carrier. The data symbols, $b_n^{(k)}$, are assumed to be independent and identically distributed (i.i.d) sequences with $\Pr[b_n^{(k)} = -1] = \Pr[b_n^{(k)} = 1] = 1/2$. T_c is the chip duration such that $T_c = 1/W$, where W is the channel bandwidth. The pseudo noise sequence, $a_n^{(k)}$, has a period of N such that $T_b = NT_c$. For user **k**, T_k is the transmitter time delay and Ψ_k is the phase offset relative to that of desired user.

2.2 Channel Model

The link for user k can be modeled as a linear filter with complex valued lowpass equivalent impulse response such as

$$h_k(t) = \sum_{l=1}^{L_k} \beta_{k,l}\,\delta(t - t_{k,l})\, e^{j\phi_{k,l}} \tag{4}$$

where subscripts k, l refers to path l for user k. Equation (4) can be characterized by the number of paths L_k, path gains $\beta_{k,l}$, time delays $t_{k,l}$ and phases $\phi_{k,l}$. We assume that the path gains, $\beta_{k,l}$, are independently and identically Rayleigh distributed according to

$$f_\beta(x) = (2x/\rho_o)\exp(-x^2/\rho_o) \tag{5}$$

where $E[\beta^2] = \rho_o$. Each path phases, $\phi_{k,l}$, has a uniform distribution over the interval $[0, 2\pi)$. Furthermore, it is assumed that the sets $\{t_{k,l}\}$, $\{\beta_{k,l}\}$ and $\{\phi_{k,l}\}$ are mutually independent and that the members within each set are i.i.d. random processes for all k and l. As the symbol duration T_b increases with the number of tones, the delay range remains constant given by T_b/M.

2.3 Receiver Model

The received signal at the front end of a base station is the sum of all the signals arriving from each user or each mobile station over multipath channel with the thermal noise such as following equation

$$r(t) = \sum_{k=1}^{K} \int_{-\infty}^{\infty} h_k(\tau) s_k(t-\tau)\, d\tau + n(t) \tag{6}$$

where n(t) is a zero-mean white Gaussian process with a two-sided power spectral density No/2. From (1) and (4), this can be written as

$$\begin{aligned} r(t) = &\sum_{k=1}^{K} \sum_{m=1}^{M} \sum_{l=1}^{L_k} \sqrt{2P_k}\, \beta_{k,l} b_{k,m}(t-\tau_{k,l}) a_k(t-\tau_{k,l}) \\ &\cos(2\pi f_m t - \theta_{k,l}) + n(t) \end{aligned} \tag{7}$$

where the time delay $\tau_{k,l} = t_{k,l} + T_k$ and the phase offset $\theta_{k,l} = (2\pi f_m \tau_{k,l} - \Psi_k - \phi_{k,l}) \bmod 2\pi$. We assume that $\tau_{k,l}$ and $\phi_{k,l}$ are uniformly distributed over $[0, T_b/M]$ and $[0, 2\pi)$, respectively, and there are no gain and phase differences between subcarriers for any user's specified path.

1) Correlator Outputs with Initial Data Estimates

In this section, we derive expressions for the correlator outputs with initial data estimates for each of the K users. Many symbolic definitions are similar to that of [9], however, the notations are more complex due to the usage of the concept of multitone modulation. If we obtain acquisitions of the spreading sequence of each user i and each path l, the receiver coherently demodulates and despreads the received signal. It is assumed that the correlators are aimed to recover the signal on the first path of the jth subcarrier for user i. At the instant of $t = nT_b + t_{i,1}$, the correlator output for the user i's jth subcarrier can be written as

$$z_{jn}^{(i)} = \int_{(n-1)T+\tau_{i,1}}^{nT_b+\tau_{i,1}} r(t)\, a(t-\tau_{i,1}) \cos(\omega_j - \hat{\theta}_{i,1})\, dt \tag{8}$$

where $\hat{\theta}_{k,l}$ represents the phase estimate of $\theta_{k,l}$ and ω_j is $2\pi f_j$. For compact expression of the correlator output equations, we define some symbolic definitions in advance before giving the full expressions of each output term. For $1 \le i \le K$, $1 \le k \le K$ and $1 \le l \le L_k$, the multiplications of path gain and phase difference term are defined as

$$G_c(i,k,l) = \beta_{k,l}\cos(\theta_{k,l} - \hat{\theta}_{i,1}) \tag{9}$$

$$G_s(i,k,l) = \beta_{k,l}\sin(\theta_{k,l} - \hat{\theta}_{i,1}). \tag{10}$$

Furthermore, we define four partial cross correlation functions such that

$$\begin{aligned} P_{k,i}^{m}(\tau) = &\int_{(n-1)T_b+\tau_{i,1}}^{\tau} a_k(t-\tau_{k,l}+T_b)\, a_i(t-\tau_{i,1}) \\ &\cos[(\omega_m-\omega_j)t]\, dt \end{aligned} \tag{11}$$

$$\begin{aligned} \hat{P}_{k,i}^{m}(\tau) = &\int_{\tau}^{nT_b+\tau_{i,1}} a_k(t-\tau_{k,l})\, a_i(t-\tau_{i,1}) \\ &\cos[(\omega_m-\omega_j)t]\, dt \end{aligned} \tag{12}$$

$$\begin{aligned} Q_{k,i}^{m}(\tau) = &\int_{(n-1)T_b+\tau_{i,1}}^{\tau} a_k(t-\tau_{k,l}+T_b)\, a_i(t-\tau_{i,1}) \\ &\sin[(\omega_m-\omega_j)t]\, dt \end{aligned} \tag{13}$$

$$\begin{aligned} \hat{Q}_{k,i}^{m}(\tau) = &\int_{\tau}^{nT_b+\tau_{i,1}} a_k(t-\tau_{k,l})\, a_i(t-\tau_{i,1}) \\ &\sin[(\omega_m-\omega_j)t]\, dt \end{aligned} \tag{14}$$

where τ is any arbitrary value between $(n-1)T_b + \tau_{i,1}$ and $nT_b + \tau_{i,1}$. With considerations of (9) and (10), the full cross correlation between the signal of user k's lth path normalized with respect to its received signal power and user i's spreading signal is defined as

$$f_c(i,k,l,n,y) = G_c(i,k,l)[b_{n-1}^{(k)} P_{k,i}^{y}(\tau) + b_n^{(k)} \hat{P}_{k,i}^{y}(\tau)] \tag{15}$$

$$f_s(i,k,l,n,y) = G_s(i,k,l)[b_{n-1}^{(k)} Q_{k,i}^{y}(\tau) + b_n^{(k)} \hat{Q}_{k,i}^{y}(\tau)]. \tag{16}$$

Using the above definitions, we can write the correlator output as five parts such that

$$D_n^{(i)} = \sqrt{P_i/2}\ b_n^{(i)} T_b G_c(i,i,1) \tag{17}$$

$$I_n^{(i)} = \sum_{l=2}^{L_i} \sqrt{P_i/2}\ f_c(i,i,l,n,j) \tag{18}$$

$$J_n^{(i)} = \sum_{\substack{m=1 \\ m\neq j}}^{M} \sum_{l=1}^{L_i} \sqrt{P_i/2}[f_c(i,i,l,n,m) + f_s(i,i,l,n,m)] \tag{19}$$

$$M_n^{(i)} = \sum_{\substack{k=1 \\ k\neq i}}^{K} \sum_{\substack{m=1 \\ m\neq j}}^{M} \sum_{l=1}^{L_i} \sqrt{P_k/2}[f_c(i,k,l,n,m) + f_s(i,k,l,n,m)] \tag{20}$$

$$\eta_n^{(i)} \int_{(n-1)T_b+\tau_{i,1}}^{nT_b+\tau_{i,1}} n(t)\, a_i(t-\tau_{i,1}) \cos(\omega_j t - \hat{\theta}_{i,1})\, dt \tag{21}$$

$D_n^{(i)}$ represents the desired signal of the ith user, $I_n^{(i)}$ the output of its multipath interference(MPI), $J_n^{(i)}$ the output of the interchannel interference(ICI), $M_n^{(i)}$ the output of the multiple access interference(MAI), and $\eta_n^{(i)}$ the output of the noise signal. The next step is obtaining initial data estimates for usage in interference cancellation.

2) Outputs of RAKE Receiver

In this paper, we consider a RAKE receiver which coherently combines the signals over each path and if we define the correlator output $z_{jn}^{(i)}$ of equation (8) as $z_{jn,d}^{(i)}$, the decision statistic of the RAKE receiver output using maximal ratio combining diversity can be written as

$$Z_n^{(i)} = \sum_{d=1}^{D} \beta_{i,d}\, z_{jn,d}^{(i)} \tag{22}$$

where d is a parameter for diversity order and $\beta_{i,d}$ is the estimated path gain at the receiver. We assume the fading and noise on each path are mutually and statistically independent and that the average power per path is identical for all paths. The bit error rates after maximal ratio combining diversity can be derived from the following equation[12]

$$P_e = [(1-\mu)/2]^D \sum_{j=0}^{D-1} \binom{D-1+j}{j} [(1+\mu)/2]^j \tag{23}$$

where D is the order of diversity and

$$\mu = \sqrt{\gamma_{ca}/(1+\gamma_{ca})}\ . \tag{24}$$

The γ_{ca} is the average signal-to-noise ratio(SNR) per path.

3) Outputs of Single Stage Interference Canceller

In this subsection, we derive the correlator outputs subtracted the interference terms using one stage interference canceller. For doing this, we must derive the replica of the full cross correlation expressed as

$$\begin{aligned} \hat{f}_c(i,k,l,x,y,1) &= \hat{G}_c(i,k,l)\,[\, b_{x-1}^{(k)}(0)\, P_{k,l}^{y}(\tau) \\ &\quad + b_x^{(k)}(0)\, \hat{P}_{k,l}^{y}(\tau)] \end{aligned} \tag{25a}$$

$$\begin{aligned} \hat{f}_s(i,k,l,x,y,1) &= \hat{G}_s(i,k,l)\,[\, b_{x-1}^{(k)}(0)\, Q_{k,l}^{y}(\tau) \\ &\quad + b_x^{(k)}(0)\, \hat{Q}_{k,l}^{y}(\tau)]. \end{aligned} \tag{25b}$$

And number 1 in the parenthesis means the stage number of interference canceller and y represents the carrier number to be correlated. By subtracting the replicas of the interference terms from the decision statistic of (22), we can get the results one stage cancelled for user i's xth data bit given by

$$Z_x^{(i)}(1) = Z_x^{(i)} - I_x^{(i)}(1) - J_x^{(i)}(1) - M_x^{(i)}(1). \tag{26}$$

We can express the subtracted interference terms using the definition of (25a) and (25b) such as

$$I_x^{(i)}(1) = \sum_{l=2}^{L_i} \sqrt{P_i/2}\ \hat{f}_c(i,i,l,x,j,1) \tag{27}$$

$$J_x^{(i)}(1) = \sum_{\substack{m=1 \\ m\neq j}}^{M} \sum_{l=1}^{L_i} \sqrt{P_i/2}[\hat{f}_c(i,i,l,x,m,1) + \hat{f}_s(i,i,l,x,m,1)] \tag{28}$$

$$M_x^{(i)}(1) = \sum_{\substack{k=1 \\ k\neq i}}^{K} \sum_{\substack{m=1 \\ m\neq j}}^{M} \sum_{l=1}^{L_i} \sqrt{P_k/2}[\hat{f}_c(i,k,l,x,m,1) + \hat{f}_s(i,k,l,x,m,1)]\ . \tag{29}$$

By substituting (22), (27), (28) and (29) into (26), we obtain new correlator outputs with one stage interference cancellation given by

$$Z_x^{(i)}(1) = D_x^{(i)} + \tilde{I}_x^{(i)}(1) + \tilde{J}_x^{(i)}(1) + \tilde{M}_x^{(i)}(1) + \eta_x^{(i)} \tag{30}$$

where

$$\tilde{I}_x^{(i)}(1) = \sum_{l=2}^{L_i} \sqrt{P_i/2}\ \tilde{f}_c(i,i,l,x,j,1) \tag{31}$$

$$\tilde{J}_x^{(i)}(1) = \sum_{\substack{m=1 \\ m\neq j}}^{M} \sum_{l=1}^{L_i} \sqrt{P_i/2}[\tilde{f}_c(i,i,l,x,m,1) + \tilde{f}_s(i,i,l,x,m,1)] \tag{32}$$

$$\tilde{M}_x^{(i)}(1) = \sum^{K} \sum^{M} \sum^{L_i} \sqrt{P_k/2}[\tilde{f}_c(i,k,l,x,m,1) + \tilde{f}_s(i,k,l,x,m,1)]\ . \tag{33}$$

We define the subtracted version of full cross correlation such as

$$\tilde{f}_c(i,k,l,x,m,1) = f_c(i,k,l,x,m) - \hat{f}_c(i,k,l,x,m,1) \tag{34a}$$

$$\hat{f}_s(i,k,l,x,m,1) = f_s(i,k,l,x,m) - \hat{f}_s(i,k,l,x,m,1). \quad (34b)$$

Rewriting (34a) and (34b) more detailed with enough accurate channel and phase estimates as

$$\hat{f}_c(i,k,l,x,m,1) = G_c(i,k,l)[\hat{b}_{x-1}^{(k)}(0)P_{k,i}^{m}(\tau) + \hat{b}_x^{(k)}(0)\hat{P}_{k,i}^{m}(\tau)] \quad (35a)$$

$$\hat{f}_s(i,k,l,x,m,1) = G_s(i,k,l)[\hat{b}_{x-1}^{(k)}(0)Q_{k,i}^{m}(\tau) + \hat{b}_x^{(k)}(0)\hat{Q}_{k,i}^{m}(\tau)] \quad (35b)$$

where

$$\hat{b}_x^{(k)}(0) = b_x^{(k)} - b_x^{k}(0). \quad (36)$$

The symbol bit for user i's xth bit is decided based on hard decision at the single stage interference canceller output given by

$$b_x^{(i)}(1) = sgn[Z_x^{(i)}(1)] \quad (37)$$

4) Outputs of Multiple Stage Interference Canceller

After one stage cancelling of interferences by (26) or (30), we can get a more accurate data decisions, $b_x^{(k)}(1)$ for $1 \le k \le K$. For multiple stage interference canceller, this decision result should be an input to a second stage canceller for generating even more accurate replicas of the interference terms and these terms are removed from the original decision statistic by the same manner of (26). In the general case of V-stage multiple interference canceller, above steps are repeated in each stage. The correlator output of the vth stage canceller($1 \le \nu \le V$) for the xth bit of the ith user is

$$Z_x^{(i)}(\nu) = D_x^{(i)} + \tilde{I}_x^{(i)}(\nu) + \tilde{J}_x^{(i)}(\nu) + \tilde{M}_x^{(i)}(\nu) + \eta_x^{(i)} \quad (38)$$

The MPI, ICI and MAI output terms at the vth stage are respectively given by

$$\tilde{I}_x^{(i)}(\nu) = \sum_{l=2}^{L_i} \sqrt{P_i/2}\, \tilde{f}_c(i,i,l,x,j,\nu) \quad (39)$$

$$\tilde{J}_x^{(i)}(\nu) = \sum_{\substack{m=1\\m\ne j}}^{M}\sum_{l=1}^{L}\sqrt{P_i/2}[\tilde{f}_c(i,i,l,x,m,\nu) + \tilde{f}_s(i,i,l,x,m,\nu)] \quad (40)$$

$$\tilde{M}_x^{(i)}(\nu) = \sum^{K}\sum^{M}\sum^{L_k}\sqrt{P_k/2}[\tilde{f}_c(i,k,l,x,m,\nu) + \tilde{f}_s(i,k,l,x,m,\nu)] \quad (41)$$

The generated version of the full cross correlation at the νth stage canceller with enough accurate gain and phase estimates becomes

$$\hat{f}_c(i,k,l,x,m,\nu) = G_c(i,k,l)[\hat{b}_{x-1}^{(k)}(\nu-1)P_{k,i}^{m}(\tau) + \hat{b}_x^{(k)}(\nu-1)\hat{P}_{k,i}^{m}(\tau)] \quad (42a)$$

$$\hat{f}_s(i,k,l,x,m,\nu) = G_s(i,k,l)[\hat{b}_{x-1}^{(k)}(\nu-1)Q_{k,i}^{m}(\tau) + \hat{b}_x^{(k)}(\nu-1)\hat{Q}_{k,i}^{m}(\tau)] \quad (42b)$$

where

$$\hat{b}_x^{(k)}(\nu) = b_x^{(k)} - b_x^{k}(\nu). \quad (43)$$

As in the previous subsection, the hard decision of the vth stage is performed and we finally get to the last stage bit decision $b_x^{(i)}(V)$ by going through each stage step by step in order of $\nu = 1, 2, \ldots, V$.

$$b_x^{(i)}(\nu) = \mathrm{sgn}[Z_x^{(i)}(\nu)]. \quad (44)$$

III. Performance Analysis

In this section we derive the variances of signal and interference terms for the three cases discussed in the previous section. We can derive bit error probabilities using the signal-to-noise ratio.

3.1 Performance with Initial Data Estimates

From (17)-(21), the variances of each term can be calculated. First, the variance of desired signal term is given by

$$E[(D_n^{(i)})^2] = \frac{P_i}{2}\, T_b^2 E[G_c^2(i,i,1)] \quad (45)$$

where

$$E[G_c^2(i,i,1)] = E[\beta_{i,1}^2]E[\cos^2 \wedge\theta_{i,1}] = \frac{1}{2}\, E[\beta_{i,1}^2][1 + \exp(-2\sigma^2_{\wedge\theta_{i,1}})]. \quad (46)$$

This result is from the fact that $E[\cos bx] = \exp(-b^2\sigma_x^2/2)$ for a constant b and a zero-mean Gaussian random variable x. The variances of the MPI, ICI and MAI can be written as, respectively,

$$Var[I_n^{(i)}] = \frac{P_i}{2}\,(L_i - 1)E[f_c^2(i,i,l,n,j)] \quad (47)$$

$$Var[J_n^{(i)}] = \frac{P_i}{2}(L-1)\sum^{M}\{E[f_c^2(i,i,l,n,m)] + E[f_s^2(i,i,l,n,m)]\} \quad (48)$$

$$Var[M_n^{(i)}(1)] = \frac{P_k}{2}\,(K-1)L_k\sum_{m=1}^{M}\{E[f_c^2(i,k,l,n,m)] + E[f_s^2(i,k,l,n,m)]\} \quad (49)$$

where

$$E[f_c^2(i,i,l,n,j)] = \frac{1}{2}\, E[\beta_{i,l}^2]\, \frac{2T_b^2}{3N} \tag{50}$$

$$E[f_c^2(i,k,l,n,m)] = E[G_c^2(i,k,l)] \{E[(P_{k,i}^m(\tau))^2] + E[(\hat{P}_{k,i}^m(\tau))^2]\} \tag{51}$$

$$E[f_s^2(i,k,l,n,m)] = E[G_s^2(i,k,l)] \{E[(Q_{k,i}^m(\tau))^2] + E[(\hat{Q}_{k,i}^m(\tau))^2]\}. \tag{52}$$

These variances can be calculated from the fact that $E[G_c^2(i,k,l)] = E[G_s^2(i,k,l)] = E[\beta_{k,l}^2]/2$ and

$$\sum_{\substack{m=1\\m\neq j}}^{M} \{E[P_{i,i}^m(\tau))^2] + E[(\hat{P}_{i,i}^m(\tau))^2] + E[(Q_{i,i}^{m(\tau))^2}] + E[(\hat{Q}_{i,i}^m(\tau))^2]\}$$
$$= \sum_{\substack{m=1\\m\neq j}}^{M} \frac{T_b^2}{\pi^2(m-j)^2} \left\{1 - \frac{N}{2\pi(m-j)} \sin\frac{2\pi(m-j)}{N}\right\} \tag{53}$$

$$\sum_{m=1}^{M} \{E[P_{k,i}^m(\tau))^2] + E[(\hat{P}_{k,i}^m(\tau))^2] + E[(Q_{k,i}^{m(\tau))^2}] + E[(\hat{Q}_{k,i}^m(\tau))^2]\}$$
$$= \left[\frac{2T_b^2}{3N} + \sum_{\substack{m=1\\m\neq j}}^{M} \frac{T_b^2 N}{\pi^2(m-j)^2} \left\{1 - \frac{N}{2\pi(m-j)} \sin\frac{2\pi(m-j)}{N}\right\}\right] \tag{54}$$

Finally, the mean and variance of the noise term are $E[\eta_n^{(i)}] = 0$, and

$$Var[\eta_n^{(i)}] = N_o T_b/4 \tag{55}$$

From (45), (47), (48), (49) and (55), we can obtain the averaged SNR per path at the RAKE receiver output for the ith user's jth carrier defined as

$$\gamma_{ca}^{(i)} = \frac{E[(D_n^{(i)})^2]}{Var[I_n^{(i)}] + Var[J_n^{(i)}] + Var[M_n^{(i)}] + Var[\eta_n^{(i)}]} \tag{56}$$

3.2 Performance with Single Stage Interference Canceller

We present the performance analysis of the single stage canceller with accurate channel estimates in this subsection. From (31)-(33), we can obtain the variances of MPI, ICI and MAI leaving the expectation of desired signal squared and noise variances unchanged. The variance of MPI term can be written as

$$Var[\tilde{I}_x^{(i)}(1)] = \frac{P_i}{2}(L_i - 1)\, E[\tilde{f}_c^2(i,i,l,x,j,1)] \tag{57}$$

and the variances of ICI and MAI terms can be written as

$$Var[\tilde{J}_x^{(i)}(1)] = \frac{P_i}{2}(L_i - 1) \sum_{\substack{m=1\\m\neq j}}^{M} \{E[\tilde{f}_c^2(i,i,l,x,m,1)] + E[\tilde{f}_s^2(i,i,l,x,m,1)]\} \tag{58}$$

$$Var[\tilde{M}_x^{(i)}(1)] = \frac{P_i}{2}(K-1)L_k \sum_{m=1}^{M} \{E[\tilde{f}_c^2(i,k,l,x,m,1)] + E[\tilde{f}_s^2(i,k,l,x,m,1)]\} \tag{59}$$

where

$$E[\tilde{f}_c^2(i,k,l,x,m,1)] = E[(\tilde{b}_x^{(k)}(0))^2]\, E[G_c^2(i,k,l)] \{E[P_{k,i}^m(\tau))^2] + E[(\hat{P}_{k,i}^m(\tau))^2]\} \tag{60}$$

$$E[\tilde{f}_s^2(i,k,l,x,m,1)] = E[(\tilde{b}_x^{(k)}(0))^2]\, E[G_s^2(i,k,l)] \{E[Q_{k,i}^m(\tau))^2] + E[(\hat{Q}_{k,i}^m(\tau))^2]\} \tag{61}$$

The term of data symbol difference between the original bit $b_x^{(k)}$ and hard decision bit $b_x^{(k)}(0)$ is defined in (36) and its variance is determined from the conditional probability of initial data decisions

$$\Pr[b_x^{(k)}(0) = b_x^{(k)} \mid b_x^{(k)}] = 1 - p_e^{(k)}(0) \tag{62}$$

$$\Pr[b_x^{(k)}(0) = -b_x^{(k)} \mid b_x^{(k)}] = p_e^{(k)}(0). \tag{63}$$

By the definition of (36), we obtain

$$\Pr[\tilde{b}_x^{(k)}(0) = 0 \mid b_x^{(k)}] = 1 - p_e^{(k)}(0) \tag{64}$$

$$\Pr[\tilde{b}_x^{(k)}(0) = 2b_x^{(k)} \mid b_x^{(k)}] = p_e^{(k)}(0) \tag{65}$$

and the variance of $\tilde{b}_x^{(k)}(0)$ is given by

$$E[(\tilde{b}_x^{(k)}(0))^2] = 4p_e^{(k)}(0). \tag{66}$$

From (45), (57), (58), (59) and (55), we can get the averaged SNR per path out of the one stage canceller such as

$$\gamma_{ca}^{(i)} = \frac{E[(D_n^{(i)})^2]}{Var[\tilde{I}_n^{(i)}(1)] + Var[\tilde{J}_n^{(i)}(1)] + Var[\tilde{M}_n^{(i)}(1)] + Var[\eta_n^{(i)}]}. \tag{67}$$

3.3 Performance with Multiple Stage Interference Canceller

We derive the interference variances of the multiple stage canceller in this section using the similar method of the previous section. For V stage canceller, the decision statistic and bit error rate are resulted in each stage, and all the bit error rates of the previous stages are needed to obtain the final stage's bit error rate $p_e^{(i)}(V)$. That is, the calculation of bit error probability is repeated in each stage. As in the previous section, we can get the three interference terms, MPI, ICI and MAI, respectively, given by

$$Var[\tilde{I}_x^{(i)}(\nu)] = \frac{P_i}{2}(L_i - 1)\, E[\tilde{f}_i^2(i, i, l, x, j, \nu)] \qquad (68)$$

$$Var[\tilde{J}_x^{(i)}(\nu)] = \frac{P_i}{2} L_i \sum_{\substack{m=1 \\ m \neq j}}^{M} \{E[\tilde{f}_c^2(i, i, l, x, m, \nu)] + E[\tilde{f}_s^2(i, i, l, x, m, \nu)]\} \qquad (69)$$

$$Var[\tilde{M}_x^{(i)}(\nu)] = \frac{P_k}{2}(K-1) L_k \sum_{m=1}^{M} \{E[\tilde{f}_c^2(i, k, l, x, m, \nu)] + E[\tilde{f}_s^2(i, k, l, x, m, \nu)]\} \qquad (70)$$

where

$$E[\tilde{f}_c^2(i, k, l, x, m, \nu)] = E[(\tilde{b}_x^{(k)}(\nu - 1))^2]\, E[G_c^2(i, k, l)]\, E[(P_{k,i}^m(\tau) + \hat{P}_{k,i}^m(\tau))^2] \qquad (71)$$

$$E[\tilde{f}_s^2(i, k, l, x, m, \nu)] = E[(\tilde{b}_x^{(k)}(\nu - 1))^2]\, E[G_s^2(i, k, l)]\, E[(Q_{k,i}^m(\tau) + \hat{P}_{k,i}^m(\tau))^2] \qquad (72)$$

The term of data symbol difference between the original bit $b_x^{(k)}$ and hard decision bit $b_x^{(k)}(v-1)$ is defined in (43) and the variance of $\tilde{b}_x^{(k)}(\nu - 1)$ is given by

$$E[(\tilde{b}_x^{(k)}(\nu - 1))^2] = 4p_e^{(k)}(\nu - 1). \qquad (73)$$

From (45), (68), (69), (70) and (55), we can derive the SNR out of the one stage canceller and bit error probability

$$\gamma_{ca}^{(i)}(\nu) = \frac{E[(D_n^{(i)})^2]}{Var[I_n^{(i)}(\nu)] + Var[J_n^{(i)}(\nu)] + Var[\tilde{M}_n^{(i)}(\nu)] + Var[\eta_n^{(i)}]} \qquad (74)$$

V. Numerical Results

The effects of different parameters on system performance are investigated. We assumed that the input data symbol duration T_b/N_t and the bandwidth is about N_t/T_b. The symbol duration on each carrier is T_b and the bandwidth on each carrier is also about $1/T_b$. We set all the received powers to be equal with power control, and the average path strength for all users and paths is taken as $E[\beta_{k,l}^2] = 1$ in the theoretical calculations.

Fig. 2 shows the BER versus Eb/No when the number of users K = 5, the number of carriers M = 12, the number of chips N = 128, the number of multipaths L = 4 and the order of diversity D = 2. In this case, the use of interference canceller is essential for communication. As the number of canceller stage increases, the performance is improved rapidly. However, there exists error floor for each stage. Fig. 3 is a plot of BER versus Eb/No when the number of PN chips are increased to N = 256 with the other parameters are same with those of Fig. 2. Without canceller, the degree of improvement is very small. Especially, the case of stage number S = 2 shows the most dramatic improvement. At the Eb/No = 30dB, the bit error rates are about $6*10^{-4}$ in Fig. 2 and $2*10^{-6}$ in Fig. 3. The number of spreading chips can be an important design factor. Fig. 4 presents the effects of diversity on the system performance with K = 5, M = 12, N = 128, L = 4 and the number of canceller stage is fixed to S = 2. The case of D = 2 is also shown in Fig. 2 and we can use it as a comparison reference. In the situation of given parameters, MRC diversity(D = 3) is more effective than interference canceller(S = 3) without consideration of hardware complexity. For example, the required E/No's for BER = 10^{-5} are about 24dB and 14dB in Fig. 2 and Fig. 4, respectively. Fig. 5 is a plot of BER versus number of users with M = 12, N = 256, L = 2, D = 2 and Eb/No = 30 dB. There may be various kinds of graphs according to the number of paths L and the order of diversity D. From

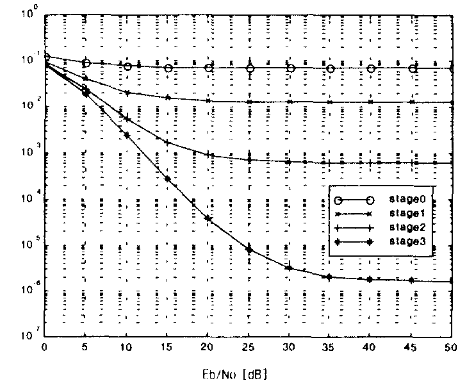


Figure 2. BER versus Eb/No with K = 5, M = 12, N = 128, L = 4, D = 2 for three canceller stages.

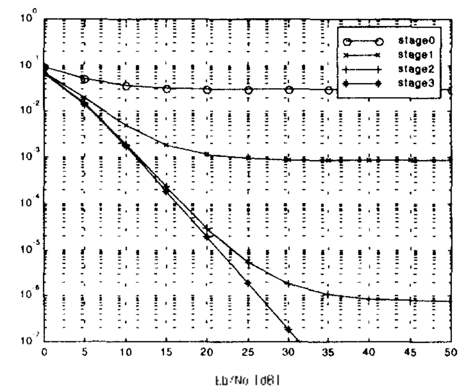


Figure 3. BER versus Eb/No with K = 5, M = 12, N = 256, L = 4, D = 2 for three canceller stages.

the results of the performance evaluation, when there is a limit in the performance improvement by the interference cancellation scheme, the use of higher order of diversity can be recommended.

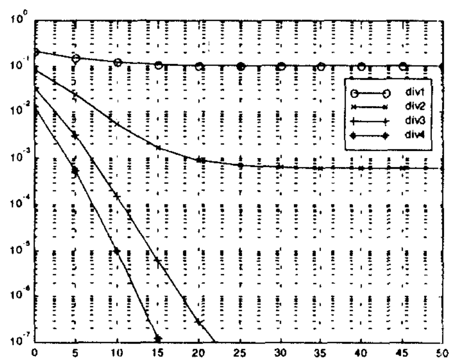


Figure 4. BER versus Eb/No for various diversity orders with K = 5, M = 12, N = 127, L = 4, S = 2.

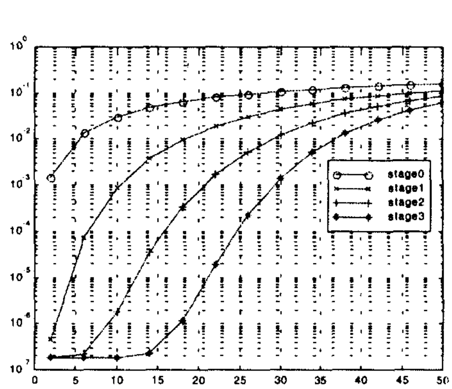


Figure 5. BER versus K with Eb/No = 30dB for three canceller stages. L = 2, D = 2, N = 256, M = 12.

V. Conclusions

In this paper, we analyzed the effects of interference canceller on system performance with consideration of RAKE reception. We use the canceller scheme proposed by [9] in BPSK modulated multitone CDMA system, and derive the variances of MPI, ICI and MAI. We assumed very accurate estimation of path gain and phase. The interference cancellation scheme used in this paper is very effective in performance improvement of MT DS/CDMA system. Also the increase of chip numbers per symbol is the most effective in case of cancellation stage number S = 2. That is, the degree of performance improvement is the largest in this case. And also the order of diversity D = 3 shows a good performance which is available in data communication at Eb/No = 20dB. There may exist a number of calculation results and graphs according to the variations of each parameter and we present four figures. From these results, we can know that the number of chips and the order of diversity are very important parameters in design of multitone CDMA system with interference canceller. The next step of the work in this paper is to evaluate the system performance with consideration of error in path gain and phase estimation.

References

1. R. Prasad and S. Hara, "An overview of multi-carrier CDMA," in *Proc. ISSSTA'96*, pp.107-114, 1996.
2. N. Yee, J-P. Linnartz and G. Fettweis, "Multi-carrier CDMA on indoor wireless radio networks," in *Proc. IEEE PIMRC '93*, Yokohama, Japan, pp.109-113, Sep. 1993.
3. K. Fazel and L. Papke, "On the performance of convolutionally-coded CDMA/OFDM for mobile communication system," in *Proc. IEEE PIMRC '93*, Yokohama, Japan, pp.468-472, Sep. 1993.
4. V. M. Dasilva and E. S. Sousa, "Multicarrier orthogonal CDMA signals for quasi-synchronous communication systems," in *Proc. IEEE ICUPC '93*, Ottawa, Canada, pp.995-999, Oct. 1993.
5. L. Vandendorpe, "Multitone spread spectrum multiple access communications system in a multipath rician fading channel," *IEEE Trans. Vehic. Tech*, vol.44, no.2, pp.327-337, May 1995.
6. S. Kondo and L. B. Milstein, "On the performance of multicarrier DS CDMA systems," *IEEE Trans. Commun.*, vol.44, no.2, pp.238-246, Feb. 1996.
7. A. J. Viterbi, "Very low rate convolutional codes for maximum theoretical performance of spread-spectrum multi-access channels," *IEEE J. Select. Areas Commun.*, vol.8, pp.641-649, May 1990.
8. P. Patel and J. Holtzman, "Analysis of simple successive interference cancellation scheme in DS/CDMA system using correlations," in *Proc.GLOBECOM*, Houston, Tx, 1993.
9. Y. C. Yoon, R. Kohno and H. Imai, "A spread-spectrum multiaccess system with cochannel interference cancellation for multipath fading channels," *IEEE J. Select. Areas Commun.*, vol. 11, no. 7, pp.1067-1075, Sept. 1993.
10. D. Divsalar and M. K. Simon, "Improved CDMA performance using parallel interference cancellation," in *Proc. IEEE MILCOM '94, Ft. Monmouth*, NJ., pp. 911-917, Oct. 1994.
11. B. G. Kang, J. S. Ha and M. J. Kim, "Performance evaluation of up-link channel in al multi-carrier DS/CDMA system," *Proc.ITC-CSCC'96*, Seoul, Korea, July 15-17, pp.432-436, 1996.
12. J. Proakis, *Digital Communications*, Third Ed., McGraw-Hill, 1995.

▲Seung-Keun Park

Seung-Keun Park was born in Seoul, received the B.S. and M.S. in statistics from the University of Korea in 1991 and 1993, respectively. Since 1993 he has been at ETRI, where he is currently a member of engineering staff. His research interests are in the area of RF spectrum, nonlinear estimation, and radio regulation.

▲Byeong-Gwon Kang

Byeong-Gwon Kang was born in Seoul, Korea, on November 18, 1963. He received the B.S. degree in electronics engineering from Yonsei University, Seoul, Korea, in 1982, and the M.S. and Ph. D. degrees from Yonsei University, Seoul, Korea, in 1988, and 1993, respectively. From 1993 to 1997 he was a senior researcher with ETRI in Taejon, Korea. In 1997 he join the faculty of Soonchunhyang University, where he is a professor. His research interests include the W-CDMA and WLL.

▲Hee-Chang Chung

Hee-Chang Chung was born in Kangwon province on June 12, 1954. He received the B.S. degree in electronics engineering from Korea University, Seoul, Korea in 1980, and the M.S. and Ph D. degree from Ajou University, Kyonggi, Korea, in 1989 and 1997, respectively. In 1980 he joined the Resercher of ETRI, taejon, Korea, where he is a principal researcher. His research interests include the radio regulation and spectrum management for GMPCS and Mobile communications.