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비동기 DS/CDMA 시스템을 위한 역상관 다중사용자 검출기

(A Robust Decorrelating Multiuser Detector for Asynchronous DS/CDMA Communication Systems)

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요 약

본 논문에서는 비동기 DS/CDMA 시스템을 위한 2단 구조의 심볼 단위 역상관 다중사용자 검출기가 제안된다. 기존의 정합필터를 이용하는 검출기의 출력이 한 심볼 지연되어, 선형 역상관 검출기에서 역상관 기저의 선택을 위해 사용되며, 또한 작은 오프셋 값의 칩 타이밍 지연을 고려하여, 단일 상관기 뱅크 대신 Early-Late 상관기 뱅크를 사용하였는데, 이는 간섭 사용자의 정확한 시간 지연을 몰라도 되며 디지털 구현이 용이한 장점을 갖는다. 제안된 검출기의 성능을 검증하기 위하여, 분석된 BER과 모의 실험 결과가 비교된다. 제안된 검출기의 성능 분석은 시간-제한된 신호에 대해 유도되었지만, 모의 실험은 시간-제한된 신호와 대역-제한된 신호에 대해 수행되었다.

Abstract

This paper presents an asynchronous DS/CDMA multiuser detector, which is a two stage, symbol-by-symbol scheme consisting of conventional detectors followed by linear decorrelating detectors. The conventional detector first makes temporal decisions and the detected symbols are delayed by one symbol period to be used for the selection of decorrelating bases in the subsequent decorrelating detection stage. It also employs a bank of early-late correlators in place of a bank of single correlators taking the small offset of chip timing asynchronism into account. The proposed detector requires only the coarse knowledge of relative time delays of interfering users and is suitable for digital implementation. To verify the detector performance, the analytical BER performance will be given and compared with the simulation results for BPSK DS/CDMA signals in AWGN channel. While the performance of the proposed detector will be analyzed for time-limited signal, the simulation is carried out for both the time-limited and band-limited signals. As can be seen in the simulation results, the proposed scheme shows good results.

I. Introduction

In reverse link of a DS/CDMA communication system, the conventional receiver suffers from severe performance degradation as the signal power of interfering users become large. The performance degradation caused by the multiple access interference can be alleviated by power control or by using multiuser detectors which is recently received considerable attention. The optimum multiuser receiver has been first proposed by Verdu in [1] and achieves minimum error probability in asynchronous CDMA system. However, its computational

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complexity is exponential in the number of users involved and the precise knowledge of signal powers and time of arrivals of all the users are required. Hence, sub-optimal receivers with reasonable computational complexity have been considered to ensure the practical implementation. Recently, various sub-optimal receivers with moderate complexity are proposed for synchronous and asynchronous CDMA systems including linear and non-linear approaches. A well-known non-linear multiuser detector is the multi-stage detector[2][3][10], which shows nearly optimal performance when the signal powers of interfering users are strong. However, it still requires the precise knowledge of signal powers of all the users. The decorrelating detector [4][5][6] is one of the linear classes and is optimum when the signal powers of interfering users are not known. Its implementation in synchronous environment is quite clear in the sense that the decorrelation is implemented by using only the known spreading sequences. In asynchronous environment, however, the cross terms in the matched filter outputs depend not only on the relative time delays but also on the adjacent information bits of the interfering users. Thus, the asynchronous decorrelating detector must be a sequence detector and also requires the precise knowledge of time delays [5]. Adaptive MMSE detector [7][8] is another approach to multiuser detection and shows good results for DS/CDMA using only short PN sequences. However, it is not suitable for the systems with long PN sequences. More recently, successive cancellation scheme [9] has been proposed. It detects the data symbols of the stronger users and then subtract the signal due to those users successively in the order of signal powers. However, since the reconstruction of interfering signals is involved, the detection requires extremely accurate estimation of channel parameters.

In this paper, we will derive a decorrelating detection scheme for asynchronous DS/CDMA system, which is a symbol-by-symbol scheme and whose performance is insensitive to the small offset of interfering users time delays. The decorrelating detection scheme developed here is a two-stage detector, which, however, is different from that in [3]. First, it uses the conventional detector to temporally make decisions. Then the detected symbols are delayed for the selection of decorrelating bases in the subsequent decorrelating detection for the user currently under consideration. In addition, it employs a bank of early-late correlator pairs in place of a bank of single correlators, similar to that in [8], by which the decorrelating detector performance is insensitive to the small variation of interfering users time delays relative to that of the user under consideration and the proposed detector requires only the coarse knowledge of relative time delays of interfering users, which makes it suitable for digital implementation. Although the computational burden is heavier than that of conventional detector, the fast development of DSP technique can make the realization possible in near future.

Ⅱ. Signal Model and System Description

Let us consider a base band BPSK DS/CDMA signal. The chip matched filter output can be expressed as

$$r(t)=\sum_{i}\sum_{k=1}^{K}\sum_{j=0}^{N-1}A_k b_k(i)s_{k,j}g_T(t-(iN+j)T_c-\tau_k)+n(t) \tag{1}$$

with K = The number of users
N= Processing gain
T_c= Chip duration
A_k = Amplitude of user k ; assumed to be constant.
τ_k = Time delay of user k
$b_k(i)$ = ith BPSK data of user k
S_{kj} = jth normalized spreading code of user k

$g_T(t)$ = Combined chip shaping and matched filter impulse response of receiver and transmitter, respectively.

$n(t)$ = Gaussian noise with zero-mean and variance σ^2 of which the spectral envelop is $T_c\sqrt{G(f)}$, where $G(f)$ is Fourier transform of $g(t)$. In this paper, however, it will be assumed to be white.

Assume that user 1 is under consideration and the timing is synchronized to user 1 (i.e. $\tau_1=0$). Then, the time delay τ_k of user k can be expressed as $\tau_k = d_kT_c + \delta_k$, where d_k is an integer satisfying $0\le d_k < N$ and δ_k is a real value satisfying $0\le \delta_k < T_c$. In the derivation of the proposed demodulator, we will also assume the followings

1. d_k's are known for all k, while δ_k's are not.
2. The received signal amplitude A_k's are constant for all k.
3. The $g_T(t)$ is strictly time-limited in the interval $[-T_c, T_c]$ and satisfying

$$g_T(t)=\begin{cases}1 & t=0\\ <1 & |t|<T_c\\ 0 & |t|\ge T_c\end{cases} \tag{2}$$

With these assumptions, the sampled version of received signal for ith symbol period of user 1 can be represented in vector form as

$$\begin{aligned}\mathbf{r}(i) = A_1b_1(i)\mathbf{f}_1^0+\sum_{k=2}^{K}A_k[g_T(T_c-\delta_k)b_k(i)\mathbf{f}_k^{d_k+1}+g_T(\delta_k)b_k(i)\mathbf{f}_k^{d_k}\\ +g_T(T_c-\delta_k)b_k(i-1)\mathbf{g}_k^{d_k+1}+g_T(\delta_k)b_k(i-1)\mathbf{g}_k^{d_k}]+\mathbf{n}(i)\end{aligned} \tag{3}$$

where,

$$\mathbf{f}_k^m=(0,\ldots,0,s_{k,0},s_{k,1},\ldots,s_{k,N-m-1})^T \tag{4a}$$

$$\mathbf{g}_k^m=(s_{k,N-m},\ldots,s_{k,N-2},s_{k,N-1},0,..,0)^T \tag{4b}$$

$\mathbf{n}(i)$ is Gaussian noise vector with zero-mean and covariance $\sigma^2\mathbf{I}$, $\mathbf{I}$ is K × K identity matrix. Now, let us define the followings

$$p_k\equiv b_k(i)b_k(i-1)\in\{1,-1\} \tag{5}$$

$$\bar{\mathbf{s}}_k^{d_k}(p_k)=\mathbf{g}_k^{d_k}+p_k\mathbf{f}_k^{d_k} \tag{6}$$

Then, (3) reads

$$\begin{aligned}\mathbf{r}(i)=A_1b_1(i)\mathbf{s}_1^0+\sum_{k=2}^{K}A_kb_k(i)[g_T(\delta_k)\bar{\mathbf{s}}_k^{d_k}(p_k)\\ +g_T(T_c-\delta_k)\bar{\mathbf{s}}_k^{d_k+1}(p_k)]+\mathbf{n}(i)\end{aligned} \tag{7}$$

The inner product of $\mathbf{r}(i)$ and $\mathbf{s}_1$ yields

$$\begin{aligned}y_1(i)=A_1b_1(i)+\sum_{k=2}^{K}A_kb_k(i)[g_T(\delta_k)\mathbf{s}_1^T\bar{\mathbf{s}}_k^{d_k}(p_k)\\ +g_T(T_c-\delta_k)\mathbf{s}_1^T\bar{\mathbf{s}}_k^{d_k+1}(p_k)]+n_1'(i)\end{aligned} \tag{8}$$

where, $S^T_K S_k$was set to 1 for all k and $n'_1(i) = S^T_1\,\mathbf{n}(i)$. The second term in the right hand side of the above equation is interfering cross terms to be rejected. To decouple the cross terms corresponding to the value of P_k, the vector $\bar{S}_k^{d_k}(P_k)$ and $\bar{S}_k^{d_k+1}(P_k)$ can be used for decorrelating bases for user k. Thus, we must first decide P_k, i.e. whether the information bit of kth interfering user is changed or not in the symbol period of user 1. The conventional detector is used for the decision of it. And then, the linear decorrelating detector finally detects the data symbol in symbol-by-symbol scheme.

Regarding $\bar{S}_k^{d_k}(P_k)$ and $\bar{S}_k^{d_k+1}(P_k)$ as two independent interfering signals, let us define the extended spreading code set and the corresponding cross correlation matrix as

$$\bar{\mathbf{S}}(\mathbf{p})=(\mathbf{s}_1^0,\bar{\mathbf{s}}_1^1(p_1),\ \bar{\mathbf{s}}_2^{d_2}(p_2),\bar{\mathbf{s}}_2^{d_2+1}(p_2),\ldots,\ \bar{\mathbf{s}}_K^{d_K}(p_K),\ \bar{\mathbf{s}}_K^{d_K+1}(p_K)) \tag{9}$$

and

$$\bar{\mathbf{R}}(\mathbf{p})=\bar{\mathbf{S}}^T(\mathbf{p})\bar{\mathbf{S}}(\mathbf{p}) \tag{10}$$

where $\bar{S}(p)$ is N×2K, $\bar{R}(p)$ is 2K×2K matrix and $\mathbf{p}=(p_1, .. , P_k)$. The corresponding correlator output vector can be expressed as

$$\bar{\mathbf{y}}(i)=\bar{\mathbf{S}}^T(\mathbf{p})\mathbf{r}(i)=\bar{\mathbf{R}}(\mathbf{p})\overline{\mathbf{A}\mathbf{b}}(i)+\bar{\mathbf{n}}(i) \tag{11}$$

where,

$$\begin{aligned}\bar{\mathbf{A}}=diag(A_1,0,\ g_T(\delta_2)A_2,g_T(T_c-\delta_2)A_2,..\\ ..,g_T(\delta_K)A_K,g_T(T_c-\delta_K)A_K)\end{aligned}$$

$\overline{\mathbf{b}}(i) = (\ b_1(i),\ b_1(i),\ldots,b_K(i),b_K(i)\ )$

$\overline{\mathbf{n}}(i)$=The noise vector with its covariance $\sigma^2\,\overline{\mathbf{R}}(\mathrm{p})$.

It should be noted that $\overline{\mathrm{R}}(\mathrm{p})$ does not depend o δ_K. The (11) can be regarded as matched filter outputs of a synchronous DS/CDMA system with extended user set and can be decoupled in symbol-by symbol scheme by using the inverse cross correlation matrix $\overline{\mathrm{R}}^{-1}(\hat{\mathrm{p}})$..
The proposed detector structure is shown in Fig. 1 Passing through the early-late correlator bank, i first decides whether the information bit of interfering users is changed or not in the symbo period of the user under consideration (user 1) Then, the set of decorrelating bases $\overline{\mathrm{S}}(\hat{\mathrm{p}})$ based on the detected value $\hat{\mathrm{P}}_k$ and appropriate early-late correlator outputs are selected and passed to the inverse cross correlation matrix filter $\overline{\mathrm{R}}^{-1}(\hat{\mathrm{p}})$.

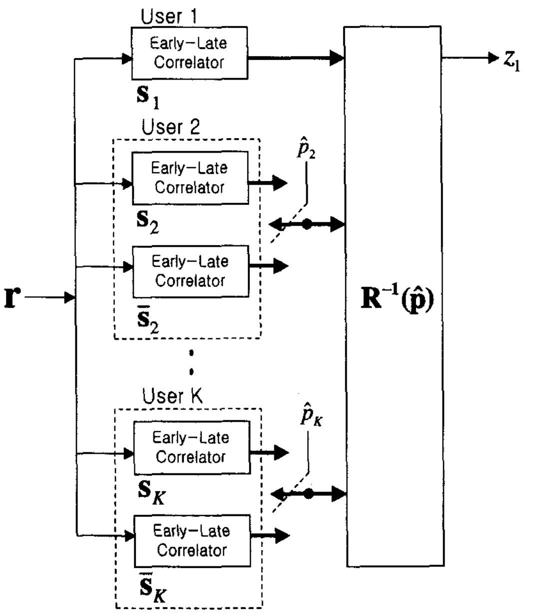


그림 1. 확장된 확산 코드집합을 갖는 역상관 검출기
Fig. 1. Decorrelating detector with extended spreading code set.

In case of using the detected values $\hat{P}_k$, the matrix filter output vector can be represented as

$$\begin{aligned}\overline{\mathbf{z}}(i) &= \overline{\mathbf{R}}^{-1}(\hat{\mathbf{p}})\overline{\mathbf{S}}^T(\hat{\mathbf{p}})\mathbf{r}(i)\\ &= \overline{\mathbf{R}}^{-1}(\hat{\mathbf{p}})\overline{\mathbf{R}}(\hat{\mathbf{p}},\mathbf{p})\overline{\mathbf{A}\mathbf{b}}(i) + \overline{\mathbf{R}}^{-1}(\hat{\mathbf{p}})\overline{\mathbf{n}}(i) \qquad (12)\end{aligned}$$

where, $\overline{\mathbf{R}}(\hat{\mathbf{p}},\mathbf{p}) = \overline{\mathbf{S}}^T(\hat{\mathbf{p}})\overline{\mathbf{S}}(\mathbf{p})$.

To facilitate the evaluation of BER performance, let us define $\mathrm{W}(\hat{\mathrm{p}},\mathrm{p})$ such that

$$\overline{\mathbf{R}}(\hat{\mathbf{p}},\mathbf{p}) = \overline{\mathbf{R}}(\hat{\mathbf{p}}) + \mathbf{W}(\hat{\mathbf{p}},\mathbf{p}) \qquad (13)$$

Then the (13) can be represented as

$$\overline{\mathbf{z}}(i) = \overline{\mathbf{A}\mathbf{b}}(i) + \overline{\mathbf{R}}^{-1}(\hat{\mathbf{p}})\mathbf{W}(\hat{\mathbf{p}},\mathbf{p})\overline{\mathbf{A}\mathbf{b}}(i) + \overline{\mathbf{R}}^{-1}(\hat{\mathbf{p}})\overline{\mathbf{n}}(i) \quad (14)$$

The last term in the right hand side of (14) is gaussian noise with zero-mean and covariance $\sigma^2\overline{\mathrm{R}}^{-1}(\mathrm{p})$. The second term is the unresolved interference due to the detection error of p_k. Let us now pay attention to $\mathrm{W}(\hat{\mathrm{p}},\mathrm{p})$. From (13), we have

$$\begin{aligned}\mathbf{W}(\hat{\mathbf{p}},\mathbf{p}) &= \overline{\mathbf{S}}^T(\hat{\mathbf{p}})[\overline{\mathbf{S}}(\hat{\mathbf{p}}) - \overline{\mathbf{S}}(\mathbf{p})]\\ &= \overline{\mathbf{S}}^T(\hat{\mathbf{p}})[(\mathbf{G}+\mathbf{F}\hat{\mathbf{P}}) - (\mathbf{G}+\mathbf{F}\mathbf{P})]\\ &= \overline{\mathbf{S}}^T(\hat{\mathbf{p}})\mathbf{F}[\hat{\mathbf{P}} - \mathbf{P}] \qquad (15)\end{aligned}$$

where, $\mathbf{F} = (\mathbf{f}_1^0,\mathbf{f}_1^1,\mathbf{f}_2^{d_2},\mathbf{f}_2^{d_2+1},\ldots,\mathbf{f}_K^{d_K},\mathbf{f}_K^{d_K+1})$,

$\mathbf{G} = (\mathbf{g}_1^0,\mathbf{g}_1^1,\mathbf{g}_2^{d_2},\mathbf{g}_2^{d_2+1},\ldots,\mathbf{g}_K^{d_K},\mathbf{g}_K^{d_K+1})$,

$\mathrm{P} = diag(0, p_1, p_2, p_2,\ldots, p_K, p_K)$.

Note that $p_k, \hat{p}_k \in \{1,-1\}$ for all k and $\overline{\mathbf{S}}^T(\hat{\mathbf{p}})\mathbf{F}$ is the partial cross correlation matrix of PN sequences for given $\hat{\mathbf{p}}$.

The first component of $\overline{z}(i)$, which is the decision statistic of the user 1, reads

$$z_1(i) = A_1 b_1(i) + \sum_{k=2}^{K}(\hat{p}_k - p_k)I_k(\hat{\mathbf{p}},b_k) + \overline{n}_1'(i) \qquad (16)$$

where

$I_k(\hat{\mathbf{p}},b_{\mathrm{k}}) = A_k\lfloor g_T(\delta_k)q_{1,2k}(\hat{\mathbf{p}}) + g_T(T_c - \delta_k)q_{1,2k+1}(\hat{\mathbf{p}})\rfloor b_k$,

$q_{m,n}(\hat{\mathbf{p}})$ is (m,n)th component of matrix $\mathbf{Q}(\hat{\mathbf{p}}) = \mathbf{R}^{-1}(\hat{\mathbf{p}})\overline{\mathbf{S}}^T(\hat{\mathbf{p}})\mathbf{F}$, and the gaussian noise $\overline{n}_1'(i)$ has zero-mean and approximate variance $\sigma^2\overline{\mathbf{R}}_{11}^{-1}(\hat{\mathbf{p}})$.

Ⅲ. Performance Analysis

From (16), the BER of user 1 is given by

$$P_{e1} = E_{\mathbf{b}}\left\lfloor E_{\mathbf{p},\bar{\mathbf{p}}}\left\lfloor P(n_1' > A_1 - \sum_{k=2}^{K}(\hat{p}_k - p_k) I_k(\hat{\mathbf{p}}, b_k)\right\rfloor\right\rfloor$$
$$= 2^{-2(K-1)} \sum_{b_i, p_i, i\neq 1} \sum_{\hat{p}_j, j\neq 1} Q\left(\sqrt{\frac{A_1^2 - \sum_{k=2}^{K}(\hat{p}_k - p_k) I_k(\hat{\mathbf{p}}, b_k)}{\sigma^2 \hat{\mathbf{R}}_{11}^{-1}(\hat{\mathbf{p}})}}\right) P(\hat{\mathbf{p}} \mid \mathbf{p}) \tag{17}$$

where, $Q(x) = (1/\sqrt{2\pi})\int_x^{\infty} e^{-y^2/2} dy$. E_b and $E_{p,\bar{p}}$ denote expectations over the ensemble of independent, uniformly distributed **b** and $(\mathrm{p}, \bar{\mathrm{p}})$, respectively. Since the conventional detector temporally decides each users data symbol independently, the probability $\mathrm{P}(\hat{\mathrm{p}} \mid \mathrm{p})$ can be written

$$P(\hat{\mathbf{p}} \mid \mathbf{p}) = \prod_{k=2}^{K} P(\hat{p}_k \mid p_k) \tag{18}$$

where, ,

$$P(\hat{p}_k \mid p_k) = \begin{cases} P_e(p_k) & \text{if } \hat{p}_k \neq p_k \\ 1 - P_e(p_k) & \text{otherwise} \end{cases},$$
$$P_e(p_k) = P(\hat{p}_k \neq p_k)$$

The detection error probability of p_k can be given by using the bit error probability of conventional detector $P_{e,conv}(b_k)$ as follows

$$P_e(p_k) = P(p_k \neq \hat{p}_k) = 2P_{e,conv}(b_k)[1 - P_{e,conv}(b_k)] \tag{19}$$

$P_{e,conv}(b_k)$ is given by

$$P_{e,conv}(b_k) = 2^{-2(K-1)} \sum_{b_j(0), b_j(-1), j\neq 1} Q\left([A_1 - \sum_{k=2}^{K} I_k] \Big/ \sigma\right) \tag{20}$$

where,

$$I_k = g_T(\delta_k) A_k [\mathbf{s}_1^T \mathbf{f}_k^{d_k} b(0) + \mathbf{s}_1^T \mathbf{g}_k^{d_k} b(-1)]$$
$$+ g_T(T_c - \delta_k) A_k [\mathbf{s}_1^T \mathbf{f}_k^{d_k+1} b(0) + \mathbf{s}_1^T \mathbf{g}_k^{d_k+1} b(-1)].$$

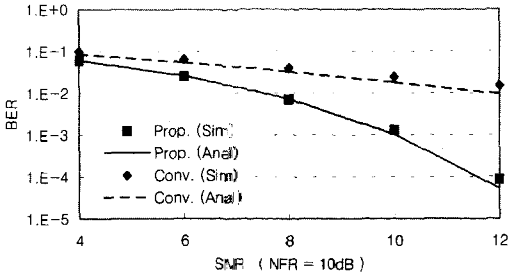
(20) will be used for the performance evaluation of the conventional detector in the next section.

Ⅳ. Simulations and Numerical Results

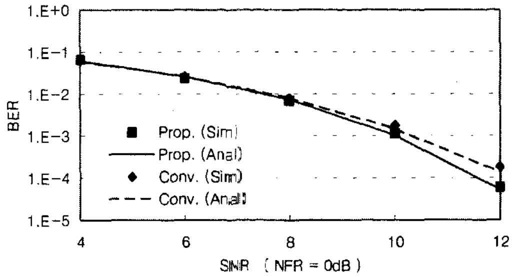
First, we will evaluate the BER of the proposed detector comparing with that of the conventional detector. Although the proposed scheme was derived for time-limited signal, the simulation is performed for band-limited signal filtered by raised cosine filter with roll-off of 0.5. The processing gain and the number of users were set to N = 63 and K = 4, respectively, and the spreading sequences were given by randomly generated binary codes. The results are shown in Fig. 2-a), b) and c), each of which were simulated with NFR_k = 10, 0 and -10dB for all *k*, respectively, where $NFR_k(\mathrm{dB}) = 10\log_{10}(A_k^2/A_1^2)$. The analytical values for the conventional and the proposed detector have been computed from (20) and (17), respectively. It is shown in the figures that the performance of the proposed detector is insensitive to near-far ratio. And, even for the band-limited signal, the simulated bit error probability is fairly close to the analytical values from (17). Indeed, in the derivation of (17), we assumed that the PN pulse is strictly time-limited in the interval $[-T_c, T_c]$.

Fig. 3 shows the BER versus NFR plots from which we can see the effect of band limitation to the detector performance. In the figure, the simulated performances for time-limited signal and band-limited signals with the roll-off factor 1.0 and 0.5 are compared with the analytical values of (17). In this simulation, the processing gain and the number of users were set to N = 15 and K = 3, respectively. And the signal to background noise power ratio was fixed to 10dB. For the time-limited signal, the simulated values are very close to the analytical values. For band limited signals, the performance of the proposed detector is degraded as the NFR grows. It is due to the side lobes of the band-limited chip pulse that was not taken into account in our analysis.

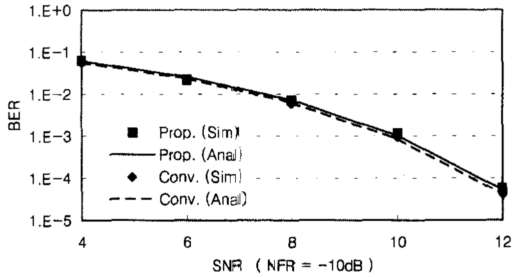
Finally, Fig. 4 shows the BER versus the number of users at power-controlled condition in which the received powers of all the users are equal. The BER performance of the proposed detector appears to be better than that of conventional detector even in the perfectly power controlled condition.



a) NFR$_k$ = 10dB for all k

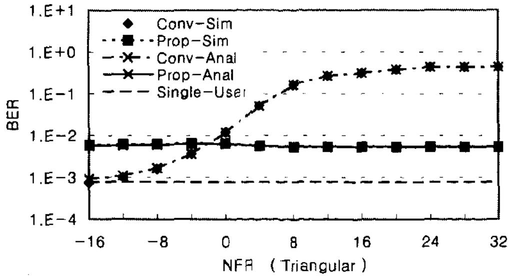


b) NFR$_k$ = 0dB for all k

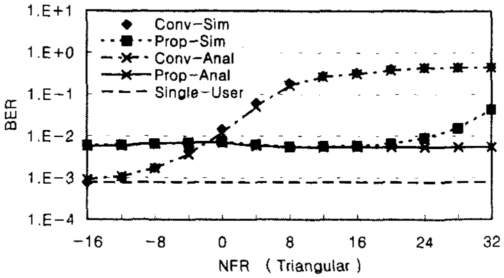


c) NFR$_k$ = -10dB for all

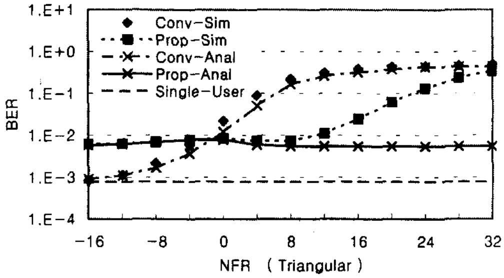
그림 2. SNR에 따른 BER성능 비교. K = 4, N = 63
Fig. 2. BER versus SNR plot. K = 4, N = 63.



a) Time-limited



b) Roll-off : 1.0



c) Roll-off : 0.5

그림 3. NFR에 따른 BER성능 비교
Fig. 3. Comparison of BER as NFR varies.

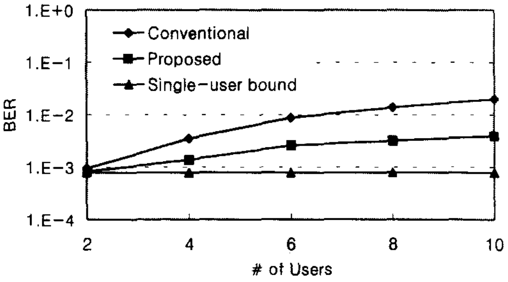


그림 4. 사용자수에 따른 BER 성능, SNR = 10dB, N = 15
Fig. 4. BER versus the number of users at perfect power control, SNR = 10dB, N = 15.

V. Conclusion

In this paper, a decorrelating detection scheme has been derived for asynchronous DS/CDMA system. While decorrelating detector has an advantage that it does not need to know received powers of interfering users, it must be a sequence detector in application to asynchronous CDMA system. Thus we developed the detector structure as a two-stage detector which makes it implemented in symbol-by symbol scheme. And by use of a bank of early-late correlator pairs in place of a bank of single correlators, we make the detector performance insensitive to the small variation of interfering users time delays relative to that of the user under consideration, which makes it suitable for digital implementation. The proposed detector shows strict near-far resistance for time-limited signal when the signal powers of interfering users are strong. And, even for the band-limited signal, the proposed scheme appeared to be superior to the conventional detector. However, when the interfering signal powers are weak, the conventional detector is superior to the proposed scheme. Thus, the second, decorrelation stage should be applied selectively only to the users whose signal power is weak.

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