논문 00-25-3B-9 한국통신학회논문지 '00-3 Vol.25 No.3B

Performance Improvement of Multiuser Detection using Antenna Array in CDMA Base Station

Jong Gil Nam*, Won Cheol Lee** *Regular Members*

ABSTRACT

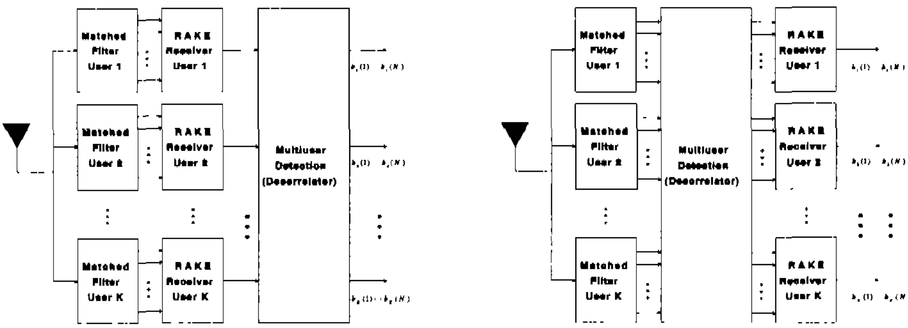
This paper analyses the performance of joint receiving structure consisting of the decorrelating multiuser detection and beamformer-RAKE receiver for DS-CDMA communications systems. In asynchronous transmission, as the number of simultaneous users increase, the capacity of CDMA system becomes severely reduced due to the nonideal orthogonality between user-assigned PN sequences and improper power control. Accordingly, the CDMA receiving system becomes vulnerable to the multiple access interferences and the near-far problem under multipath fading channel environment. To withstand these undesired performance degradations, this paper proposes the new type of multiuser detection which has a form of the hybrid structure of concatenating beamformer-RAKE receiver and decorrelating multiuser detection. The beamformer-RAKE receiver performs temporal and spatial diversity combining with alleviating fading effect and suppressing undesired interferences, and the multiuser detection plays a role of making the receiver robust to the near-far problem. Regarding the individual merit on the usage of either multiuser detection or beamformer-RAKE receiver, the hybrid one is expected to produce the enhanced performance in multipath fading CDMA channel. However major drawback of using decorrelating multiuser detection for practical deployment is arised from its computational complexity, which is exponentially increased as more number of users and transmitted symbols involve. To diminish the computational complexity, this paper exploits an efficient block Toeplitz inversion technique using matrix Levinson polynomial will be introduced. And this paper provides the mathematical analysis to show the efficiency of the proposed joint structure under the multipath propagation environment. And results of a series of exhaustive computer simulations are presented in order to demonstrate the overall performance of the proposed hybrid structure in multipath fading CDMA channel.

I. Introduction

Due to many unique features of code division multiple-access (CDMA) systems, various wireless communication systems such as digital cellular, PCS and WLL are currently adopting the direct sequence type of spreading spectrum technique [1]. Considering the rapid increase of subscribers, the capacity of CDMA communication systems is required to be expanded. However it is well described that the system capacity very limited by surrounding channel environments such as inconsistent fading effects, various cochannel interferences and near-far problem which are inevitably experienced in CDMA wireless cahnnel.

The multipath propagation is generally characterized by its own delay and incident angle spread, and in order to fortify the wanted signal power, various diversity reception techniques are introduced so as to improve detected symbol quality and increase the system capacity ultimately. Towards this, the spatio and temporal combiner adopting the antenna array have been developed, and its superior performance has been verified throughout various articles [2,3]. Here to emphasize the specified user's signal, the spatial

* 숭실대학교 정보통신공학과(wlee@saint.soongsil.ac.kr)
** 숭실대학교 정보통신공학과 통신신호처리연구실(jonggilnam@hanmail.net)
논문번호 : 99072-0225, 접수일자 : 1999년 2월 25일
※ 본 연구는 정보통신부 지원에 의해 수행되었습니다.



(a) RAKE-MUD (b) MUD-RAKE

Fig. 1 Conventional multiuser detection structures in multipath fading channel.

combiner like beamformer performs the coherent combining of spatially spreaded multipath components over the spatial domain. Then it is followed by a bank of matched filters which performs despreading with respect to delayed versions of the specific user-assigned PN sequence. In order to aggregate multipath components in temporal domain, the RAKE receiver can be utilized at final stage.

In multipath fading channel, the near-far problem has been issued as one of major devastating factors in CDMA communications systems. And the system capacity is critically dependent on how well this deteriorate effect could be suppressed in the presence of large number of simultaneous users. Many feasible solutions [4-6,15-16] have been developed to combat the near-far problem with employing the enhanced power control for a practical deployment. But under the fast fading environment over the asynchronous transmission, it is skeptical to achieve the notable performance improvement by the usage of power control only. According to many works [4-8,10-11], the multiuser detection has been strongly considered to be deployed on the CDMA receiving systems due to its superior performance to the near-far problem. So far along the usage of the multiuser detection in the multipath fading channel, RAKE receiver would be accompanied with so as to combine timely spreaded multipath components.

To make a further progress, the multiuser detection using optimum decorrelator based on a prior knowledge of the highly structured cross-correlation properties between delayed versions of signature waveforms associated to every users. Eventhough the optimum decorrelating detector has a series of out standing features which lead the capability increase without a prior knowledge of distribution of received energy, but its practical deployment has been precluded due to its enormous amount of computations. Furthermore, in the multipath propagation situation, the amount of computation gets even severe because each multipath component associated to a specific user is considered as other users signal. To lessen the computational burden on the decorrelating detector, various suboptimum schemes are proposed such as the multistage decorrelator, the successive interference cancellation and the LTI decorrelator [4-5,15].

Fig.1 shows two types of the hybrid structure including the multiuser detection and the RAKE receiver, and its attributes are well explained in [6]. Among the available multiuser detection schemes, the decorrelating detector motivated by the work of Lupas and Verdue[7,8] has been known as an optimum detector which accomplishes a perfect cancellation of interferences, and provides a resistance to the near-far problem upon a prior knowledge of the highly structured crosscorrelation between signature waveforms assigned to the simultaneous users. To consolidate the gain improvement under the multipath fading channel, the multiuser decorrelating detector accompanied by the diversity combining technique could be applicable to enhance the quality of the wanted users signal. Throughout these overall decorrelating process, the capacity increase of the CDMA communications systems could be experienced.

To achieve the full decorrelation between all the multipathed user signals, reasonably speaking, decorrelator should be placed in front of the RAKE receiver [6,10,11] as in Fig.1(b). This paper will exploit the joint multiuser detection structure shown in Fig.1(b) together with involving beamformer in front as depicted in Fig.2. Consequently the receiving process provides the robustness to the multipath fading as well as the near-far problem. Furthermore in order to conduct the decorrelating process at more reduced complexity for practical use, this paper introdues the matrix Levinson polynomials [12] to accomplish the inversion via linear filtering whose coefficients are updated based on the knowledge of crosscorrelation matrices. Observing the inherent structure of crosscorrelation matrix, it could be possible to complete the decorrelating process by making use of only a few number of matrix coefficients which are associated with the matrix Levinson polynomials.

This paper organized as follows: Section II introduces the joint structure consisting of beamformer-RAKE receiver and optimum decorrelating detector with providing its comparable features. Besides these, in depth mathematical derivations relevant to the signal modeling, the two-dimensional combining process and the optimum decorrelating multiuser detection will be provided. Section III proposes the efficient decorrelating process by introducing a set of matrix Levinson polynomials which is utilized to complete the inversion of the crosscorrelation matrix via linear filtering. In section IV, computer simulations are conducted to verify the performance of the proposed joint structure together with applying the efficient decorrelating process. Finally, in section V conclusions are drawn from the work presented in this paper.

II. Concatenated Multiuser Detection with Decorrelating Detector and Beamformer-RAKE Receiver

Suppose that single path signal is impinged on single antenna in asynchronous CDMA communication system, each users signal is arrived at the different time instant with having different attenuation gain. So the received signal is composed with the timely spreaded signals with having different incident angles. If it is matched filtered on the basis of the specific users signature waveform, the resulting sample value involves the wanted signal component as well as the superposed interferences arised from undesired users signals together with additive noises. To alleviate the performance degradation caused by these deteriorate interferences, optimum decorrelation process accompanied with certain diversity combining technique could be adopted to suppress these undesired interference over the temporal and spatial domain.

2.1 Multiple Lineer Constrained Beamformer

Considering slowly-varying fading channel in which symbol interval is smaller than coherence time of the channel. Suppose that the received signal at an array with P antennas is considered as the superposition of K-users signals coming through L distinct paths individually. Here the received signal vector

at antenna array can be expressed as

$$\mathbf{r}(t)=\sum_{n=0}^{N}\sum_{k=1}^{K}b_k(n)\sum_{l=0}^{L}\sqrt{w_{k,l}(n)}\, s_k(t-(n-1)T-\tau_{k,l})\, c_{k,l}(n)\ \mathbf{a}(\theta_{k,l})(n)+\mathbf{n}(t) \tag{1}$$

where N is the number of the data bits transmitted, $b_k(n)\in\{-1,+1\}$ is the nth transmitted bit of the kth user, $w_{k,l}$ is the signal power of lth path of kth user, $s_k(t)$ is time-limited signature waveform with support [0 T]. Also $c_{k,l}$ is the complex fading coefficient for lth path of kth user and $\tau_{k,l}$ is distinct relative delay of lth path of kth user. In (1), $\mathbf{n}(t)$ is the additive white Gaussian noise, and $\mathbf{a}(\theta_{k,l})$ is the array response to the lth path of the kth user whose pth component has the form of $e^{-j\frac{2\pi d}{\lambda}(p-1)\sin\theta_{k,l}}$.

To perform the spatial combining, the beamformer network is utilized so as to make an appropriate beam pattern whose main beams are placed in the directions of the multipath components associated to the specific user [9], whereas producing nulls to the multiple access interferences. But due to the lack of number of antennas it is hard to generate exact nulls to all the directions of interferences. This paper introduces a beamforming scheme satisfying the given multiple linear constraints, it is confined to produce an appropriate a series of main beams only which are placed at the directions of multipath components for a certain user. A prior knowledge to update the weight vector for the above beamformer is the incident angle profile corresponding to the specific user. In this paper the incident angle profile is presumed to be known onward. Thus, the beamformer weight vector for each user can be calculated via minimizing the array output power subject to satisfying a pre-determined set of linear constraints. Here the cost function for updating the optimum weight vector given by

$$\begin{aligned}\min \mathbf{J} = &\ \mathbf{w}_k^H \boldsymbol{\Phi}_{rr} \mathbf{w}_k \\ &+\lambda_{k,l}(\mathbf{w}_k^H \mathbf{a}(\theta_{k,l})-1)+\cdots \\ &+\lambda_{k,L}(\mathbf{w}_k^H \mathbf{a}(\theta_{k,L})-1)\end{aligned} \tag{2}$$

where $\boldsymbol{\Phi}_{rr}=E[\mathbf{r}(t)\ \mathbf{r}^H(t)]$ is the pre-correlation matrix obtained from antenna array outputs. The above cost function is comprised with the array output power and a set of linear equations together with multiplicative Lagrange multipliers. Suppose the channel is time-invariant, after some derivations, the optimum weight vector can be expressed as [9,14]

$$\mathbf{w}_{opt,k}= \boldsymbol{\Phi}_{rr}^{-1}\mathbf{D}_k(\mathbf{D}_k^H \boldsymbol{\Phi}_{rr}^{-1}\mathbf{D}_k)^{-1}\mathbf{1}_L \tag{3}$$

where $\mathbf{1}_L$ is $L\times 1$ vector whose components are all the unity, and $\mathbf{D}_k$ is given by

$$\mathbf{D}_k=[\mathbf{a}_k^{(1)},\ \mathbf{a}_k^{(2)},\cdots,\ \mathbf{a}_k^{(P)}]^T \tag{4}$$

$$\mathbf{a}_k^{(P)}=[a^{(P)}(\theta_{k,1}),a^{(P)}(\theta_{k,2}),\cdots,a^{(P)}(\theta_{k,L})]^T \tag{5}$$

where $a^{(p)}(\theta_{k,l})$ is the pth antenna response to the direction of $\theta_{k,l}$, $a^{(p)}(\theta_{k,l})=e^{-j\frac{2\pi d}{\lambda}(p-1)\sin\theta_{k,l}}$

Considering the situation of slowly time-varying channel, i.e., regarding the movement of transmitting platform, the optimum weight vector should be altered at least over each symbol duration. Thus the optimum weight vector turns out to be a function of symbol index n, then it has the form of the following ;

$$\begin{aligned}\mathbf{w}_{opt,k}(n)= &\ \boldsymbol{\Phi}_{rr}^{-1}(n)\ \mathbf{D}_k(n) \\ &(\mathbf{D}_k^H(n)\ \boldsymbol{\Phi}_{rr}^{-1}(n)\ \mathbf{D}_k(n))^{-1}\mathbf{1}_L\end{aligned} \tag{6}$$

where $\boldsymbol{\Phi}_{rr}(n)$ is the estimate of the array output covariance matrix, and the collection of array response at the nth symbol can be expressed as

$$\mathbf{D}_k(n)=[\mathbf{a}_k^{(1)}(n),\ \mathbf{a}_k^{(2)}(n),\cdots,\ \mathbf{a}_k^{(P)}(n)]^T \tag{7}$$

Major contribution of the above constrained beamformer on the increase of system capacity attributes to its capability of increasing powers of the specific multipath components associated to the desired users signal whereas diminishing those of either the multiple access interferences or the background noise. On the usage of the optimum weight vector as in (6), the beamformer constitutes its main beams to the directions of every incomming multipath signals of the k th user and shows relatively low spatial gains to those of interferences. Moreover referring to the cost function as in (2), since the total receiving power is minimized while the given linear constraints are fulfilled, the nulls are produced provided that the strong interference impinges to the array. As a result, the overall SINR (Signal to Interference and Noise Ratio) could be enhanced at the receiving array.

2.2 Signal Modeling after Matched Filtering

Here relative time delays due to the asynchronous transmission are presumed to be less than one chip duration and those due to the multipath propagation are less than one symbol duration. Considering multipathed signals associated to each user as other users signals, inherent overlapped portion between adjacent symbols can be observed attributed to the asynchronous transmission. Thus the matched filter output sample of kth user at the nth symbol time instant is determined by adjacent of (n-1)th as well as (n+1)th symbol corresponding to its own multipath components and those of other users [6]. Let $(1\times L)$ vector $\mathbf{y}_k(n)$ denote the matched filter output vector relevant to kth user, which has the form of the collection of p samples in column fashion. Here each $1\times L$ row vector consists of output samples to the kth matched filter bank. Thus, the output vector can be expressed as

$$\mathbf{y}_k(n) = \mathbf{w}^H_{opt}(n)\ \mathbf{X}_k(n) \tag{8}$$

where $\mathbf{A}^H$ denotes conjugate transpose of matrix $\mathbf{A}$, and $\mathbf{X}_k(n)$ is expressed by

$$\mathbf{X}_k(n) = \mathbf{DWCbR}_k + \mathbf{N} \tag{9}$$

In (9) each matrix components are given by

$$\mathbf{D} = [\ \mathbf{D}(n-1),\ \mathbf{D}(n),\ \mathbf{D}(n+1)] \tag{10}$$

$$\mathbf{W} = diag(\ \mathbf{W}(n-1),\ \mathbf{W}(n),\ \mathbf{W}(n+1)) \tag{11}$$

$$\mathbf{C} = diag(\ \mathbf{c}(n-1),\ \mathbf{c}(n),\ \mathbf{c}(n+1)) \tag{12}$$

$$\mathbf{b} = diag(\ \mathbf{b}(n-1),\ \mathbf{b}(n),\ \mathbf{b}(n+1)) \tag{13}$$

$$\mathbf{R}_k = [\ \mathbf{R}_k(-1),\ \mathbf{R}_k(0),\ \mathbf{R}_k(1)]^T \tag{14}$$

The followings are detailed descriptions of (10)-(14)

$$\mathbf{D}(n) = [\ \mathbf{D}_1(n),\ \mathbf{D}_2(n), \cdots,\ \mathbf{D}_k(n)] \tag{15}$$

$$\mathbf{W}(n) = diag(\sqrt{w_{1,1}(n)}, \sqrt{w_{1,2}(n)}, \cdots, \sqrt{w_{1,L}(n)}, \sqrt{w_{2,1}(n)}, \cdots, \sqrt{w_{K,L}(n)}) \tag{16}$$

$$\mathbf{c}(n) = diag(c_{1,1}(n), c_{1,2}(n), \cdots, c_{1,L}(n), c_{2,1}(n), \cdots, c_{K,L}(n)) \tag{17}$$

$$\mathbf{b}(n) = diag(b_1(n)\ 1^T_L, b_2(n)\ 1^T_L, \cdots, b_K(n)\ 1^T_l) \tag{18}$$

where $\mathbf{N}$ is the $P\times L$ noise matrix whose component has white Gaussian property in statistical point of view. In slowly time-vary channel, matrices $\mathbf{D}$, $\mathbf{W}$ and $\mathbf{C}$ are almost constant matrices over the a few symbol duration. Here it is noticeable fact that the crosscorrelation matrix $\mathbf{R}_k$ consists of three $KL\times L$ matrices which is independent of the symbol index provided that the characteristic of the fading channel is stationary over each block of symbols observed with the help of the sliding window. And its component matrix is determined from correlating between replicas of signature waveforms of all users which are retrieved based on the knowledge of the arrival time delay instants. And each component matrix in (14) can be expressed as, for i = -1,0,1,

$$\mathbf{R}_k(i) = [\ \mathbf{R}^1_k(i),\ \mathbf{R}^2_k(i), \cdots,\ \mathbf{R}^L_k(i)] \tag{19}$$

$$\mathbf{R}^l_k(i) = [\ \mathbf{R}^l_{k,1}(i),\ \mathbf{R}^l_{k,2}(i), \cdots,\ \mathbf{R}^l_{k,K}(i)]^T \tag{20}$$

$$\mathbf{R}^l_{k,m}(i) = [R^{l,1}_{k,m}(i), R^{l,2}_{k,m}(i), \cdots, R^{l,L}_{k,m}(i)] \tag{21}$$

In (21), $1\times L$ vector $\mathbf{R}^l_{k,m}(i)$ is composed with $\{R^{l,j}_{k,m}(i)\}^L_{l,j=1}$ given by

$$R^{l,j}_{k,m}(i) = \frac{1}{T}\int_{\tau_{k,1}}^{T+\tau_{k,l}} s_{k,l}(t-\tau_{k,l})\, s_{m,j}(t-iT-\tau_{m,j})dt \tag{22}$$

As mentioned previously, the above is the crosscorrelation between signature waveform relevant to the jth multipath of the mth user with

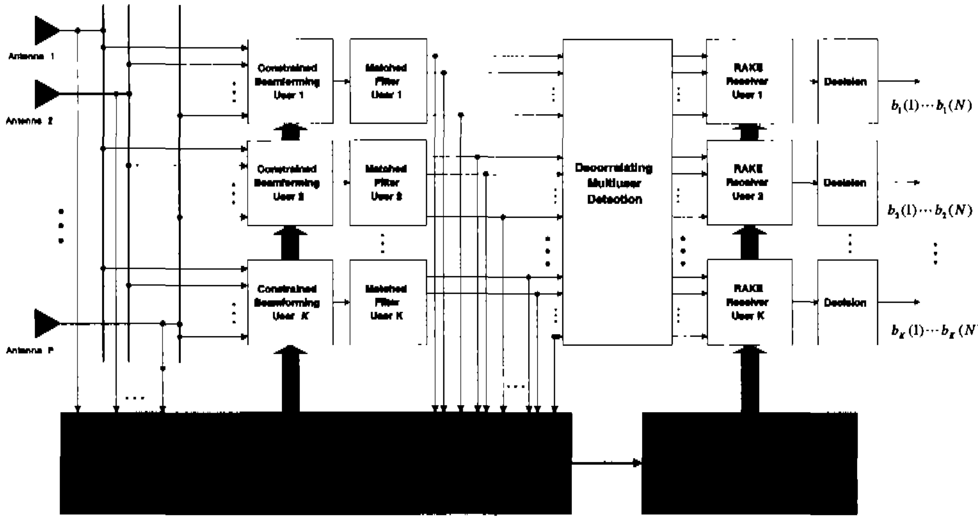


Fig. 2 The proposed structure of multiuser detection with combined beamformer-RAKE and decorrelating detector for asynchronous CDMA systems.

respect to the lth multipathed signature waveform of the kth user. Here it can be easily proved that the crosscorrelation term $R_{k,m}^{l,j}(i)$, i = -1,0,1, is equivalent to the unity only when k and l are equivalent to m and j respectively at the same time.

To make further progress, substituting (9) into (8) gives rise to

$$\mathbf{y}_k(n) = \mathbf{p}_k(n)\ \mathbf{WCbR}_k + \mathbf{n} \qquad (23)$$

where, referring to (6),

$$\begin{aligned}\mathbf{p}_k(n) &= \mathbf{w}_{opt,k}^{H}(n)\ \mathbf{D} \\ &= \mathbf{1}_L^T(\ \mathbf{D}_k^*(n)\ \Phi_{rr}^{-1}(n)\ \mathbf{D}_k(n))^{-1} \\ &\quad \mathbf{D}_k^*(n)\ \Phi_{rr}^{-1}(n)\ \mathbf{D} \\ &= [\ \mathbf{p}_{n,n-1}^{(k)} \vdots\ \mathbf{p}_{n,n}^{(k)} \vdots\ \mathbf{p}_{n,n+1}^{(k)}]\end{aligned} \qquad (24)$$

The overall KL matched filter outputs corresponding to all the users at nth symbol can be expressed in vector form as follows;

$$\mathbf{y}(n) = \mathbf{R} \boxtimes \mathbf{PWCB} + \mathbf{N} \qquad (25)$$

where $\mathbf{y}(n) = [\ \mathbf{y}_1^T(n),\ \mathbf{y}_2^T(n), \cdots,\ \mathbf{y}_K^T(n)]^T$ and $\boxtimes$ denotes the elementwise product, and

$$\mathbf{R} = [\ \mathbf{R}(-1),\ \mathbf{R}(0),\ \mathbf{R}(1)] \qquad (26)$$

$$\mathbf{P} = [\ \mathbf{P}_1^T(n),\ \mathbf{P}_2^T(n), \cdots,\ \mathbf{P}_k^T(n)]^T \qquad (27)$$

$$\mathbf{B} = [\ \mathbf{B}(n-1),\ \mathbf{B}(n),\ \mathbf{B}(n+1)]^T \qquad (28)$$

$\mathbf{R}(i)$, i=-1,0,1 is $KL \times KL$ matrix which consists of $\mathbf{R}_k(i)$'s given by

$$\mathbf{R}(i) = [\ \mathbf{R}_1^T(i),\ \mathbf{R}_2^T(i), \cdots,\ \mathbf{R}_k^T(i)]^T \qquad (29)$$

In (26)-(27)

$$\mathbf{P}_k(n) = [\ \mathbf{p}_k^T(n),\ \mathbf{p}_k^T(n)\ , \cdots,\ \mathbf{p}_k^T(n)]^T \qquad (30)$$

$$\mathbf{B}(n) = [b_1(n)\ \mathbf{1}_L^T, \cdots, b_k(n)\ \mathbf{1}_L^T]^T \qquad (31)$$

where $\mathbf{P}_k(n)$ in (30) is $L \times 3KL$ matrix

2.3 Optimum Decorrelation Detector for Multiuser Detection

Eventhough the spatial diversity combining process is performed in appropriate, if the power control does not work properly and/or the wanted user locates at the same direction of a certain other user, the detection process becomes deteriorated. To mitigate the reduction of system capacity caused by faulty power control and ambiguity of incient angles, this paper introduces the integrated system consisting of the beamformer and the multiuser detection together as in Fig.2.

$$\mathbf{R}^N = \begin{bmatrix} \mathbf{R}(0) & \mathbf{R}(1) & 0 & 0 & 0 & \cdots & 0 \\ \mathbf{R}(-1) & \mathbf{R}(0) & \mathbf{R}(1) & 0 & 0 & \cdots & 0 \\ 0 & \mathbf{R}(-1) & \mathbf{R}(0) & \mathbf{R}(1) & 0 & \cdots & 0 \\ \cdots & \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \\ 0 & 0 & \cdots & 0 & \mathbf{R}(-1) & \mathbf{R}(0) & \mathbf{R}(1) \\ 0 & 0 & \cdots & 0 & 0 & \mathbf{R}(-1) & \mathbf{R}(0) \end{bmatrix} \quad (33)$$

$$\mathbf{P}^N = \begin{bmatrix} \mathbf{P}_{1,1} & \mathbf{P}_{1,2} & 0 & 0 & 0 & & \cdots & 0 \\ \mathbf{P}_{2,1} & \mathbf{P}_{2,2} & \mathbf{P}_{2,3} & 0 & 0 & & \cdots & 0 \\ 0 & \mathbf{P}_{3,2} & \mathbf{P}_{3,3} & \mathbf{P}_{3,4} & 0 & & \cdots & 0 \\ \cdots & & & & \cdots & & \cdots & \\ 0 & 0 & & & \cdots & 0 & \mathbf{P}_{N,N-1} & \mathbf{P}_{N,N} \end{bmatrix} \quad (34)$$

As shown in Fig.2, the multiuser detection placed right after the beamformer is enforced only on the refined signal resulted from the coherent spatial combining of multipath components. Here the spatial combining is performed by superimposing every multipath components corresponding to a specific user. The major role of beamformer is confined to produce relatively high gain beams on the directions to the incomming multipath components whereas the undesired interferences are relatively deemphasized.

Regarding the increase of the complexity as more number of data symbols are transmitted, the sliding window technique can be utilized [16]. Assume that the N number of data symbols are processed at each time, the stored $N\times 1$ sample vector resulted from the two dimensional combining process can be expressed as

$$\begin{aligned} \mathbf{y} &= [\ \mathbf{y}^T(1),\ \mathbf{y}^T(2), \cdots,\ \mathbf{y}^T(N)]^T \\ &= \mathbf{R}^N \boxtimes\ \mathbf{P}^N\ \mathbf{W}^N C^N\ \mathbf{B}^N + \mathbf{N} \end{aligned} \quad (32)$$

where each matrix involved in the above formulation is represented as follows;

$$\mathbf{W}^N = diag(\ \mathbf{W}(1),\ \mathbf{W}(2), \cdots,\ \mathbf{W}(N)) \quad (35)$$

$$\mathbf{C}^N = diag(\ c(1),\ c(2), \cdots,\ c(N)) \quad (36)$$

$$\mathbf{B}^N = [\ \mathbf{B}^T(1),\ \mathbf{B}^T(2), \cdots,\ \mathbf{B}^T(N)]^T \quad (37)$$

In (34), for $n = 1 \rightarrow N$ and $i = -1,0,1$

$$\mathbf{P}_{n,n-i} = [\ \mathbf{P}_{n,n-i}^{(1)^T},\ \mathbf{P}_{n,n-i}^{(2)^T}, \cdots,\ \mathbf{P}_{n,n-i}^{(K)^T}]^T \quad (38)$$

$$\mathbf{P}_{n,n-i}^{(k)} = [\ \mathbf{p}_{n,n-i}^{(k)^T},\ \mathbf{p}_{n,n-i}^{(k)^T}, \cdots,\ \mathbf{p}_{n,n-i}^{(k)^T}]^T \quad (39)$$

$$\mathbf{p}_{n,n-i}^{(k)} = \mathbf{w}_{opt,k}^H(n)\ \mathbf{D}(n-i) \quad (40)$$

In (34), $\mathbf{R}^N$ is the highly structured crosscorrelation matrix which is obtained from correlating in between a series of timely spreaded replicas of simultaneous users' signature waveforms. Here the prototypes of delayed users' signature waveforms are precedently generated at the base station based on the knowledge of time delays corresponding to incoming multipath signals to corresponding every users signals. In the proposed algorithm, the decorrelation process is viewed as multiplying the inverse of precedently calculated crosscorrelation matrix $\mathbf{R}^N$ on the sample vector as in (26). Thus multiplying the inverse on the both sides gives rise to

$$\mathbf{z} = \mathbf{R}^{N^{-1}}\ \mathbf{y} \quad (41)$$

where I is an NKL dimensional identity matrix. As shown in (41), the usage of the optimum decorrelator with applying the exact inversion of the crosscorrelation matrix gives rises to the result of (42). Here it is noteworthy that in the resulting form it could not see the effect of the constrained beamformer, and the last expression is the same as the result from using the single sensor. But the detection capabilities between with and without using the constrained beamformer become quite comparable because of the difference of background noise power. In the case of using the constrained beamformer, the noise power level is getting diminished as the more number of sensors are used so that the array output SNR enhanced.

Here the crosscorrelation matrix $\mathbf{R}^N$ has a form of the band-type block Toeplitz matrix of size $NKL \times NKL$. Afterward the decorrelation process, the first term on the right side in (41) does not include the crosscorrelation matrix, it can be explained that inter-correlations between every multipathed signature waveforms are eliminated. In other words, the multiple access interferences arised from cross coupling between overlapped adjacent symbols between every multipath component are intentionally removed. Another considerable fact from the resulting formulation in (41) is that it turns out to be possible to detect every user's symbol without knowing their relevant attenuation gain factors. Throughout the two dimensional diversity combining and decorrelation process, the residual term as in (41) is still remaining afterward, but this value could be dismissible attributed to the beamforming process as the signal quality refining step. But still due to the enormous amount of computations, the optimum decorrelator seems quite far from the practical application. In next section, by exploiting the intrinsic structure(Block Toeplitz matrix) of the crosscorrelation matrix, more efficient decorrelating technique in some suboptimal manner will be introduced.

2.4 Coherent Multipath Combining using RAKE Receiver

To achieve the further gain improvement along the receiving process, the joint structure utilizes the maximal ratio combiner such as RAKE receiver could be utilized to combine existing resolvable multipath components. In order to process the coherent multipath combining, it is necessary to estimate channel parameter which are known as fading coefficients. Provided that these values are perfectly recovered, multiplying estimated channel parameter to output vector as in (41), and then combining all multipath signals for specific user in (41).

$$\begin{aligned}\hat{\mathbf{B}} &= \mathbf{E}\mathbf{C}^{N^*}\mathbf{z} \\ &= \mathbf{E}\mathbf{W}^N\mathbf{C}^N\mathbf{C}^{N^*}\mathbf{B}^N + \tilde{\mathbf{N}}\end{aligned} \quad (42)$$

$$= \mathbf{E}\mathbf{W}^N\boldsymbol{\Omega}^N\mathbf{B}^N + \tilde{\mathbf{N}} \quad (43)$$

where

$$\boldsymbol{\Omega}^N = diag[\,\boldsymbol{\Omega}(1),\ \boldsymbol{\Omega}(2), \cdots,\ \boldsymbol{\Omega}(N)] \quad (44)$$

$$\begin{aligned}\boldsymbol{\Omega}(n) = diag[&|a_{1,1}|^2, |a_{1,2}|^2, \cdots \\ &, |a_{1,L}|^2, |a_{2,1}|^2, \cdots, |a_{K,L}|^2]\end{aligned} \quad (45)$$

$$\mathbf{E} = \begin{bmatrix} \mathbf{1}_L^T & \mathbf{0}_L & \mathbf{0}_L & \cdots & \mathbf{0}_L \\ \mathbf{0}_L & \mathbf{1}_L^T & \mathbf{0}_L & \cdots & \mathbf{0}_L \\ \vdots & \vdots & \ddots & \cdots & \vdots \\ \mathbf{0}_L & \mathbf{0}_L & \mathbf{0}_L & \cdots & \mathbf{1}_L^T \end{bmatrix} \quad (46)$$

where $\mathbf{0}_L$ is $1 \times L$ vector whose components are zero and $\mathbf{E}$ is $NK \times NKL$ matrix.

After the temporal combining, decision can be made on the samples. Then retrieved nth data for kth user becomes as follows

$$\hat{b}_k(n) = \operatorname{sgn}(\ \hat{\mathbf{B}}_{(n-1)K+k}), \quad n = 1 \rightarrow N \quad (47)$$

Ⅲ. Proposed Suboptimum Decorrelating Detector using Matrix Levinson Polynomials

Eventhough the proposed optimum decorrelating detector devastates the near-far problem, its complexity is still burdensome whose required computational complexity is approximately upto $O((NKL)^3) + O((NKL)^2)$. As mentioned in previous section, the structure of the crosscorrelation matrix is characterized as band-type block Toeplitz such that its inversion process can be alternatively accomplished by matrix linear filtering where involving coefficients are constructed by a series of matrix Levinson polynomials. Moreover as shown in (33), only two distinct sub-cross-correlation matrices, i.e., $\mathbf{R}(0)$ and

R(1), compose whole cross correlation matrix R^N. Accordingly it is noticeable fact that only a few matrix coefficients of matrix Levinson polynomials could be required for realizing suboptimal decorrelating linear filter. This section focuses on an efficient and fast decorrelating process via matrix linear filtering at the low computational complexity rather than direct matrix inversion. Towards this, this section introduces new inversion technique using matrix Levinson polynomials, as a result, the amount of computations needed for decorrelating process turns out to be diminished spectacularly.

3.1 Block Toeplitz Inversion using Matrix Levinson Polynomials

It is well-known that scalar Levinson polynomial is characterized as an orthogonal polynomial which has the capability of interpolating the given number of scalar autocorrelation type of coefficients. Thus, in many application field such as speech processing, system identification or inverse scattering problem the Levinson polynomial has been celebrately used for modeling a certain statistical process or estimating unknown parameters [17-22]. In fact, a stochastic signal of AR(N)-type can be modeled exactly by the Levinson polynomial of order N. Moreover the inversion of a positive definite matrix of size $N+1$ having hermitian Toeplitz property can be obtained by using Levinson polynomial of order N [12].

As a extension of scalar Levinson polynomial, its matrix version has been employed for multichannel system identification [23] or modeling multivariate statistical signal whose autocorrelations are in matrix form. Actually comparing to scalar Levinson polynomial, its matrix version is composed with matrix coefficients. Those can be updated from given matrix autocorrelations of size $KL\times KL$, i.e., $\{R(n)\}_{n=0}^{N}$, are provided in recursive manner. Specially it is of interest that if there exists a hermitian positive definite block Toeplitz matrix, then its inversion can be expressed in terms of coefficients of four types of matrix Levinson polynomials, i.e., $A_n(z)$, $B_n(z)$, $C_n(z)$ and $D_n(z)$. And those can be generated by the following recursions;

$$E_k^H A_k(z) = A_{k-1}(z) - z S_k \tilde{C}_{k-1}(z) \tag{48}$$

$$E_k^H A_k(z) = A_{k-1}(z) - z S_k \tilde{C}_{k-1}(z) \tag{49}$$

$$C_k(z) F_k = C_{k-1}(z) - z \tilde{A}_{k-1}(z) S_k \tag{50}$$

$$D_k(z) F_k = D_{k-1}(z) - z \tilde{B}_{k-1}(z) S_k \tag{51}$$

The above recursions starts with the following initial matrices

$$\begin{aligned} A_0(z) &= C_0(z) = R(0)^{-1/2} \\ B_0(z) &= D_0(z) = R(0)^{1/2}/2 \end{aligned} \tag{52}$$

Here S_k in (48)-(51), $k=1\to N$ are a sequence of free matrix parameters that are strictly bounded by unity, i.e.,

$$I - S_k S_k^H > 0, \ k=1\to N \tag{53}$$

and provided S_k's are chosen to be [24]

$$\begin{aligned} S_k &= \Big\{ A_{k-1}(z) \big(\sum_{i=0}^{k} R(i) z^i \big) \Big\}_k C_{k-1}(0) \\ &= A_{k-1}(z) \Big\{ \big(\sum_{i=0}^{k} R(i) z^i \big) C_{k-1}(z) \Big\}_k \end{aligned} \tag{54}$$

Furthermore the matrices E_k and F_k in (48)-(51) satisfy the matrix factorizations

$$\begin{aligned} E_k^H E_k &= I - S_k S_k^H \\ F_k^H F_k &= I - S_k^H S_k \end{aligned} \tag{55}$$

For uniqueness, these matrix factor E_k and F_k may be chosen to be lower triangular with positive diagonal elements. Further,

$$\begin{aligned} \tilde{A}_k(z) &= z^k A_{k^*}(z) \\ &= z^k A_k^H(1/z^H), \ k\geq 1 \end{aligned} \tag{56}$$

represents the matrix polynomial reciprocal to $A_k(z)$. Let T_N be a block Toeplitz correlation matrix of size $(N+1)K\times(N+1)K$ given by

$$\mathbf{T}_N=\begin{bmatrix} \mathbf{R}(0) & \mathbf{R}(1) & \cdots & \mathbf{R}(N) \\ \mathbf{R}(-1) & \mathbf{R}(0) & \cdots & \mathbf{R}(N-1) \\ \vdots & \vdots & \ddots & \cdots \\ \mathbf{R}(-N) & \mathbf{R}(-N+1) & \cdots & \mathbf{R}(0) \end{bmatrix} \quad (57)$$

whose elements are matrix correlations. Then its inversion can be expressed in terms of either matrix coefficients of { $\mathbf{A}_N(z)$, $\mathbf{C}_N(z)$,} or those of { $\mathbf{C}_N(z)$, $\mathbf{D}_N(z)$,} shown as

$$\begin{aligned}\mathbf{T}_N^{-1} &= \mathbf{M}_0^{(N)}\mathbf{M}_0^{(N)^H}-\mathbf{N}_N^{(N)}\mathbf{N}_N^{(N)^H} \\ &= \mathbf{N}_0^{(N)^H}\mathbf{N}_0^{(N)}-\mathbf{M}_N^{(N)^H}\mathbf{M}_N^{(N)}\end{aligned} \quad (58)$$

where

$$\mathbf{M}_0^{(N)}\cong\begin{bmatrix} \mathbf{A}_0^H & 0 & 0 & \cdots & 0 \\ \mathbf{A}_1^H & \mathbf{A}_0^H & 0 & \cdots & 0 \\ \mathbf{A}_2^H & \mathbf{A}_1^H & \mathbf{A}_0^H & \cdots & 0 \\ \vdots & \vdots & \vdots & \cdots & 0 \\ \mathbf{A}_N^H & \mathbf{A}_{N-1}^H & \mathbf{A}_{N-2}^H & \cdots & \mathbf{A}_0^H \end{bmatrix}$$

$$\mathbf{N}_0^{(N)}\cong\begin{bmatrix} \mathbf{C}_0^H & 0 & \cdots & 0 \\ \mathbf{C}_1^H & \mathbf{C}_0^H & \cdots & 0 \\ \vdots & \vdots & \cdots & 0 \\ \mathbf{C}_N^H & \mathbf{C}_{N-1}^H & \cdots & \mathbf{C}_0^H \end{bmatrix} \quad (59)$$

$$\mathbf{M}_N^{(N)}\cong\begin{bmatrix} 0 & 0 & \cdots & 0 & 0 \\ \mathbf{A}_N & 0 & \cdots & 0 & 0 \\ \vdots & \vdots & \cdots & \vdots & \vdots \\ \mathbf{A}_1 & \mathbf{A}_2 & \cdots & \mathbf{A}_N & 0 \end{bmatrix}$$

$$\mathbf{N}_N^{(N)}\cong\begin{bmatrix} 0 & 0 & \cdots & 0 & 0 \\ \mathbf{C}_N & 0 & \cdots & 0 & 0 \\ \vdots & \vdots & \cdots & \vdots & \vdots \\ \mathbf{C}_1 & \mathbf{C}_2 & \cdots & \mathbf{C}_N & 0 \end{bmatrix} \quad (60)$$

3.2 Proposed Suboptimum Decorrelating Detector using Matrix Levinson Polynomials

As discussed in last subsection, any hermitian block Toeplitz matrix satisfying positive definite property can be inversed by producing matrix Levinson polynomial. To apply this inversion technique on the decorrelating detector, highly structured crosscorrelation matrix should calculated a priori based on the knowledge of time delays of first incoming multipath components associated to all the users. Refering to the structure of overall crosscorrelation matrix $\mathbf{R}^N$ as in (33), it is definitely hermitian block Toeplitz and positive definite. Moreover only two distinct crosscorrelation terms, i.e., $\mathbf{R}(0)$ and $\mathbf{R}(1)$, are involved in constructing overall crosscorrelation matrix. This implies that only a few number of coefficients of matrix Levinson polynomials are good enough to represent inverse of the crosscorrelation matrix. For example, the number of effective coefficients is determined as three, the inverse matrix of $\mathbf{R}^N$ can be approximately expressed in terms matrix coefficients associated to the first and third kind matrix Levinson polynomials, i.e., $\mathbf{A}_3(z)$ and $\mathbf{C}_3(z)$, of order 3 as the following

$$\mathbf{R}^{1N^{-1}}\cong\mathbf{M}_0^{(3)}\mathbf{M}_0^{(3)^H}-\mathbf{N}_N^{(3)}\mathbf{N}_N^{(3)^H} \quad (61)$$

where

$$\mathbf{N}_N^{(3)}=\begin{bmatrix} 0 & 0 & 0 & \cdots & 0 \\ 0 & 0 & 0 & \cdots & 0 \\ 0 & 0 & 0 & \cdots & 0 \\ \vdots & \vdots & \vdots & \cdots & \vdots \\ \mathbf{C}_2^{(3)^H} & 0 & 0 & \cdots & 0 \\ \mathbf{C}_1^{(3)^H} & \mathbf{C}_2^{(3)^H} & 0 & \cdots & 0 \end{bmatrix} \quad (63)$$

and with the help of (62) and (63)

$$\begin{aligned} \mathbf{m}_0 &= \mathbf{A}_0^{(3)^H}\mathbf{A}_0^{(3)} \\ \mathbf{m}_1 &= \mathbf{A}_0^{(3)^H}\mathbf{A}_1^{(3)} \\ \mathbf{m}_2 &= \mathbf{A}_0^{(3)^H}\mathbf{A}_2^{(3)} \\ \mathbf{m}_3 &= \mathbf{A}_1^{(3)^H}\mathbf{A}_2^{(3)}+\mathbf{A}_0^{(3)^H}\mathbf{A}_1^{(3)} \\ \mathbf{m}_4 &= \mathbf{A}_1^{(3)^H}\mathbf{A}_1^{(3)}+\mathbf{A}_0^{(3)^H}\mathbf{A}_0^{(3)} \\ \mathbf{m}_5 &= \mathbf{A}_2^{(3)^H}\mathbf{A}_2^{(3)}+\mathbf{A}_1^{(3)^H}\mathbf{A}_1^{(3)} \\ &\quad +\mathbf{A}_0^{(3)^H}\mathbf{A}_0^{(3)} \\ \mathbf{n}_0 &= \mathbf{C}_2^{(3)^H}\mathbf{C}_2^{(3)} \\ \mathbf{n}_1 &= \mathbf{C}_2^{(3)^H}\mathbf{C}_1^{(3)} \\ \mathbf{n}_2 &= \mathbf{C}_1^{(3)^H}\mathbf{C}_1^{(3)}+\mathbf{C}_2^{(3)^H}\mathbf{C}_2^{(3)} \end{aligned} \quad (64)$$

$$\mathbf{M}_0^{(3)}=\begin{bmatrix} \mathbf{A}_0^{(3)^H} & 0 & 0 & \cdots & 0 & 0 & 0 \\ \mathbf{A}_1^{(3)^H} & \mathbf{A}_0^{(3)^H} & 0 & \cdots & 0 & 0 & 0 \\ \mathbf{A}_2^{(3)^H} & \mathbf{A}_1^{(3)^H} & \mathbf{A}_0^{(3)^H} & \cdots & 0 & 0 & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots & \vdots & \vdots \\ 0 & \cdots & 0 & \cdots & \mathbf{A}_2^{(3)^H} & \mathbf{A}_1^{(3)^H} & \mathbf{A}_0^{(3)^H} \end{bmatrix} \quad (62)$$

$$\hat{\mathbf{z}}(1) = \mathbf{m}_0 \mathbf{y}(1) + \mathbf{m}_1 \mathbf{y}(2) + \mathbf{m}_2 \mathbf{y}(3) \quad (65)$$

$$\hat{\mathbf{z}}(2) = \mathbf{m}_1^H \mathbf{y}(2) + \mathbf{m}_4 \mathbf{y}(3) + \mathbf{m}_3 \mathbf{y}(4) + \mathbf{m}_2 \mathbf{y}(5) \quad (66)$$

$$\hat{\mathbf{z}}(n) = \mathbf{m}_2^H \mathbf{y}(n-1) + \mathbf{m}_3^H \mathbf{y}(n-1) + \mathbf{m}_5 \mathbf{y}(n) + \mathbf{m}_3 \mathbf{y}(n+1) + \mathbf{m}_2 \mathbf{y}(n+2) \quad n = 3 \rightarrow N-2 \quad (67)$$

$$\hat{\mathbf{z}}(N-1) = \mathbf{m}_2^H \mathbf{y}(N-3) + \mathbf{m}_3^H \mathbf{y}(N-2) + (\mathbf{m}_5 - \mathbf{n}_0)\mathbf{y}(N-1) + (\mathbf{m}_3 - \mathbf{n}_1)\mathbf{y}(N) \quad (68)$$

$$\hat{\mathbf{z}}(N) = \mathbf{m}_2^H \mathbf{y}(N-2) + (\mathbf{m}_3^H - \mathbf{n}_1^H)\mathbf{y}(N-1) + (\mathbf{m}_5 - \mathbf{n}_2)\mathbf{y}(N) \quad (69)$$

Therefore the decorrelating detector can be accomplished by linear filtering whose coefficients are updated from (64). Along linear decorrelating filter accompanied by the sliding window technique, the output samples after decorrelating detector are as follows

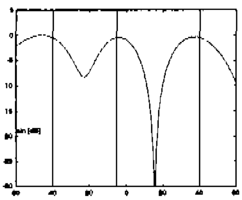
where $\mathbf{y}(n)$ is described in (25).

The computational complexity of the proposed suboptimum decorrelating detection scheme is only $O(2(KL)^3 + (14N+2)(KL)^2)$ when three matrix Levinson coefficients are in effect. In case that two effective coefficients are used, the complexity becomes more reduced into $O(4/3(KL)^3 + (7N+2)(KL)^2)$. For both cases, the computational complexity is really comparable to that of conventional optimum decorrelating detector i.e., $O((NKL)^3) + O((NKL)^2)$. And another advantage of using the proposed scheme is that the computational complexity becomes independent of the number of symbols to be detected.

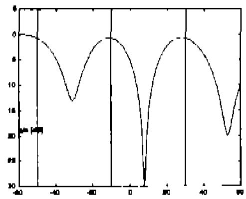
Ⅳ. Computer Simulations

In this section, the performance of the proposed multiuser detection scheme presented in this paper is investigated under the multipath fading channel from the aspect of BER behavior. Here the structure of proposed joint multiuser detection as shown in Fig.2 consists of two types of diversity combiners such as the RAKE and beamformer and additionally the decorrelating detector in concatenated fashion. For comparison's sake, the performances resulting from using either the conventional structure involving RAKE receiver prior to the decorrelating detector or the proposed joint structure adopting the suboptimum decorrelating linear filter are also analyzed. In order to demonstrate the overall performance, the BER curves are depicted which underlay performance degradations in the presence of multiple access interferences and unequalized power ratios between simultaneous users transmitted signals. Besides, to show the feasibility of deploying the proposed structure for a commercial use, the efficiency of computational complexity is also demonstrated by providing required amount of computations for processing proposed decorrelating detection scheme.

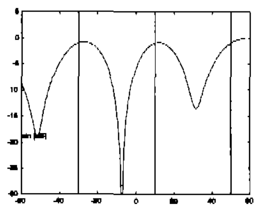
To conduct the series of computer simulations, background scenario is assumed as follows; Gold sequence of length 31 is assigned as the signature waveform to the three users with designated PN offsets, the size of sliding widow is affixed to the length of 100 bits. And BPSK is used for modulation in all cases. To produce the effect of improper power control, the fractional powers corresponding to existing users are presumed to be 0.1, 0.5 and 1 respectively, Thus the power ratios are 10dB and 5dB with respect to the third user. Moreover three distinct paths are taken to be considered as resolvable multipath components for each user whose arrival angle profile of three incoming multipaths is $[-40°, -5°, 40°]$ $[-30°, 10°, 50°]$ $[-50°, -10°, 30°]$ in the order of first, second and third user. To simulate the slowly time-varying multipath fading channel, the fading coefficients are assumed to be fixed at each frame.



(a) first user



(b) second user



(c) third user

Fig. 3 Beam pattern of constrained beamformer with antenna arrays (P=4)
(x label : angel, y label : array gain)

Fig.3 shows the beampatterns derived from the linear multiple constrained beamformer whose main beams are placed exactly at the multipath incoming directions relevant to each user. But due to the lack of antennas nulls to the cochannel interferences could not be provided. Thus, as shown in Fig.3 it is impossible to null out these interferences property. Since the number of distinct directions of self-interferences are less than that of antennas, making nulls to the incomming directions of the self interferences could be possible, the details are introduced in [14].

Fig.4 depicts the averaged BER(Bit Error Rate) behaviors deliberated from three types of multiuser detections as increasing the signal-to-noise ratio; the first name as the conventional makes use of neither any diversity reception technique nor the decorrelating, the second method employs utilizes only RAKE receiver after the decorrelating detector and the third type has a form of the proposed multiuser detection structure involving two dimensional combiner together with optimum decorrelating detector. In third case, the number of antennas is presumed to be 4 or 8. Here it can be noticed that the proposed multiuser detection scheme outperforms to either the conventional or the method using the combined RAKE receiver and the optimum decorrelating detector only. Fig.4 also shows the performance gain as increasing the number of antennas.

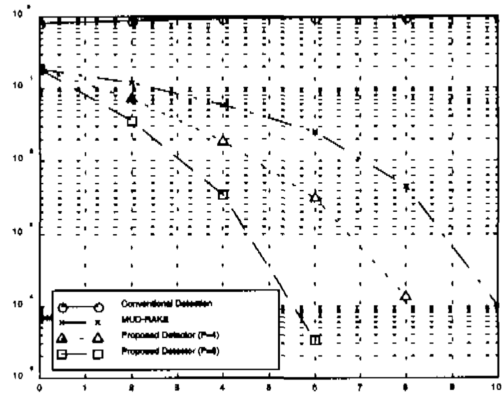


Fig 4 Comparison of three distinct approaches; conventional detection by employing matched filter only, multiuser detection with RAKE receiver and the proposed detector, NFR(Near-far-ratio) is 10dB, and the number of antenna array P=4 and P=8.
(x label : Eb/No, y label : BER)

To apply the proposed suboptimum decorrelating process presented in this paper, the matrix Levinson polynomial of type I and type III, i.e., $\mathbf{A}_N(z)$ and $\mathbf{C}_N(z)$ of order 10 generated on the basis of the highly structured crosscorrelation matrices. Here the reason to choose the order of matrix Levinson polynomial as 10 instead of 99 ($=N-1$), i.e., N is the actual number of symbols to be detected and L is a designated number of multipath components, is straightforward. Eventhough the order of matrix Levinson polynomials increases, their coefficients are not so much variant above the order of three. Here on the recursive calculation of a series of matrix Levinson polynomials, as a prior knowledge, two $KL\times KL$ crosscorrelation matrices together with 8 zero matrices are used. Among those first two crosscorrelation matrices are updated from correlating between every user's prototype signature waveforms embedded on the first incomming multipath signal on the basis of its time delays. Observing the calculated matrix coefficients of type I and type III matrix Levinson polynomials of order 10, it can easily notice that only two or three coefficients are dominant. From this reason, in this paper, either three or two effective matrix coefficients are involved in constructing linear decorrelating filter.

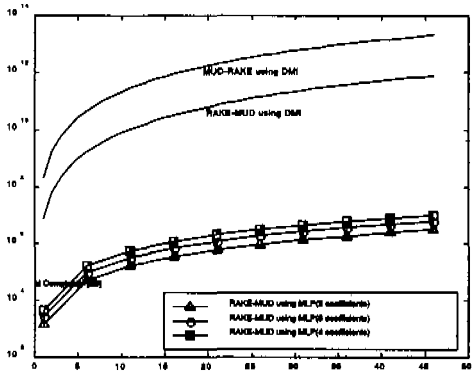


Fig. 5 A comparision of computational complexity (x label : # of users, ylabel : complexity)

The above Fig.5 shows the comparision of computational complexity between several decorrelation processes as number of users increases. The top line indicates the amount of computation induced from using an optimum decorrelating detector whose structure shown in Fig.1(b). The second from the top shows that in the case of using the structure of the proposed optimum decorrelating detector. The rest are related to that of the proposed suboptimum decorrelating detector using the effective number of matrix coefficient associated with corresponding matrix Levinson polynomials. So it can conclude that the computational complexity is stringently decreased provided that the proposed suboptimum decorrelating detector is utilized.

Finally, to demonstrate the performance of the proposed suboptimum decorrelating detector, BER curves are shown in Fig.6. For comparison's sake, the resulting BER curves from using optimum decorrelating detector and proposed its suboptimum version on the basis of the proposed multiuser detection process are overlaid. As shown in Fig.6, provided that the three effective Levinson matrix coefficients are used, the resulting BER curves obtained from individual schemes are hard to be distinguished regardless of producing considerable difference on their computational complexities. The above results tell us that the proposed suboptimum decorrelating detector for the multiuser detection could be quite applicable to the practical implementation.

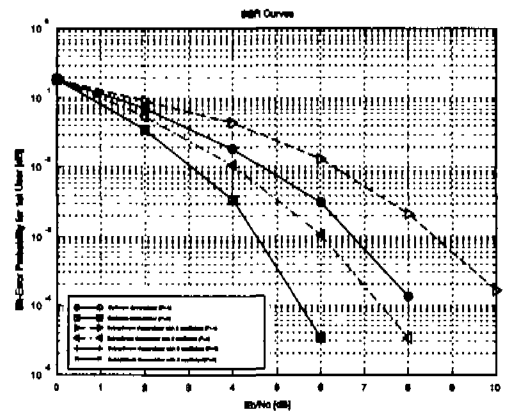


Fig. 6 BER curves of optimum and suboptimum decorrelating detector based on the proposed multiuser detection structure

V. Conclusions

This paper analyzed the performance of jointly concatenated decorrelating multiuser detection and beamformer-RAKE receiver in multipath fading CDMA channel. In order to achieve the performance gain together with the computational efficiency, the beamformer and RAKE receiver has been intentionally employed which plays a role of combining diversity branches both in temporal and spatial domain independently in conjunction with the decorrelating multiuser detection. In this paper, we claimed the new joint structure in which the beamforming process is accomplished precedently, the intermediate decorrelation detection is performed and the RAKE receiver is utilized them at the end of the overall receiving process. While deriving the proposed hybrid structure, the amount of computational complexity for the optimum decorrelating process is dependent on the number of multipath components, Simultaneous users and symbols to be detected. Here the major role of the beamformer is emphasizing only the true multipath components whereas relatively deemphasizing the unwanted interferences over the spatial domain. Towards this, the multiple linear constrained beamformer has been utilized to retrieve the refined multipath signals associated to the specific user.

To achieve more efficient computation along the decorrelating detection process, this paper introduced the matrix Levinson polynomials whose matrix coefficients are recursively updated from a priori determined crosscorrelation matrix elements. Since the highly structured crosscorrelation matrix not only has the form of hermitian block Toeplitz type but most of its matrix elements are equivalent to zeros matrices, its inversion can be calculated from a few number of matrix coefficients corresponding to out least two types of matrix Levinson polynomials.

As a result, the decorrelation detection process can be virtually accomplished by deriving the suboptimum linear filter whose coefficients are constructed from the precedently updated matrix Levinson polynomials. Here it is worthwhile to mention that the computational complexity becomes significantly reduced without giving any major performance degradations. As shown in simulation results, in the case of using only three effective coefficients associated to type I and type III matrix Levinson polynomials, the performance of the proposed suboptimum linear filtering technique is the almost same as that of optimum decorrelating detector.

By conducting computer simulations, the superior performance of the proposed joint structure consisting of the beamformer-RAKE receiver together with the decorrelating multiuser detection has been confirmed from BER point of view. And since the total amount of computation required for processing decorrelating detector becomes greatly reduced, the proposed concatenated structure employing the proposed decorrelting linear filter could be deployable to the current or future mobile communication systems.

References

[1] W. C. Y. Lee, Mobile Communications Engineering, New York : McGraw-Hill Telecommunications second ed.

[2] Ayman F. Naguib Adaptive Antennas for CDMA Wireless Networks Ph.ds thesis, University of Stanford, August. 1996.

[3] Richard B. Ertel, Spatial Filtering with Adaptive Antenna Array Algorithms in DS-CDMA Communications Systems, Masters thesis, Pennsylvania State University, 1996.

[4] Shimon Moshavi, Multi-User Detection for DS-CDMA Communications, IEEE Commun. Mag., vol.8, pp.675--682, October. 1996.

[5] A. Duel-Hallen, J. Holtzman, and Z. Zvonar, Multi-user Detection for CDMA Systems, IEEE Pers. Commun vol.2, no.2, pp.46-58, Apr. 1995.

[6] Markku Juntti, Multiuser Demodulation for DS-CDMA Systems in Fading Channel, Ph.d's thesis, University of Oulu, September. 1997.

[7] R. Lupas and S. Ver여, Near-Far Resistance of Multi-user Detectors in Asynchronous Channels, IEEE Trans. Comm., vol.38, pp.496-508, Apr. 1990.

[8] R. Lupas and S. Ver여, Linear Multi-User Detectors for Synchronous Code-Division Multiple-Access Channels, IEEE Trans. Info. Theory, vol.41, no.4, pp.123-36, July 1995.

[9] Jong Gil Nam , Yong Song and Won Cheol Lee, Combined Beamformer-RAKE Receiver and Multiuser Detection for Asynchronous CDMA Systems, The Conference of The Korean Institute of Communication Sciences, no.2, pp.2324-2327, September, 1998.

[10] Z. Zvonar and D. Brady, Linear Multipath-Decorrelating Receivers for CDMA Frequency-Selective Fading Channel, IEEE Trans. Commun., vol.44, no.6, pp.650-53, June 1996.

[11] Z. Zvonar, Combined Multi-user Detection and Diversity Reception for Wireless CDMA Systems, IEEE Trans. Vehic. Tech., vol.45, no.1, pp.205-11, Feb. 1996

[12] Won Cheol Lee and Jong Gil Nam, Block Toeplitz Matrix Inversion using Levinson Polynomials Submitted to The Journal of The Korean Institute of Communication Sciences.

[13] John Litva and Titus Kwok-Yeung Lo, Digital Beamforming in Wireless Communications, Artech House, Boston

[14] Yong Song , Jong Gil Nam and Won Cheol Lee, A Spatio-Temporal Receiver Slowly-Varying Wireless CDMA Channel, The Conference of The Korean Institute of Communication Sciences, no.2, pp.2319-2322, September, 1998.

[15] P. Patel and J. Holtzman, Analysis of a Simple Successive Interference Cancellation Scheme in a DS/CDMA System, IEEE JASC, vol.12, no.5, pp.796-807, June. 1994.

[16] Wijayasuriya, G. Norton, and J. McGeehan, A Near-Far Resistant Sliding Decorrelating Algorithm for Multi-user Detectors in DS-CDMA Systems, in Proc. Globecom92, Orlando, FL, pp. 1331- 1338, Dec. 1992.

[17] W. Trench, An Algorithm for The Inversion of Finite Toeplitz Matrices, J. Soc. Indust. Appl . Math.,12, pp.515-522, 1964.

[18] B. Friendlander, M. Morf, T. Kailath and L. Ljung, New Inversion Formulas for Matrices Classified in Terms of Their Distance from Toeplitz Matrices, Linear Algebra and its Applicaton, 27, pp.31-60, 1979.

[19] T. Kailath, A. Vieira and M. Morf, Inverses of Toeplitz Operators, Innovations, and Orthogonal Polynomials, SIAM Rev., 20, pp.106-119, 1978.

[20] H. Akaike, Block Toplitz Matrix Inversion, SIAM J. Appl. Math., 24, pp.234-241, 1973. I. S. Iohvidov, Hankel and Toplitz Matrices and Forms: Algebraic Theory, Translated by G.P.A. Thijsse, Boston: Birkhauser, 1982.

[21] I.C. Gohberg and A. A. Semencul, On The Inversion of Finite Toplitz Matrices and Their Continual Analoga, Mat. Issled., vol.7, no.2 pp.201-223 (Russian), 1972.

[22] S. U. Pillai and T. I. Shim, Spectrum Estimation and System Identification, Springer-Verlag, New York, 1993.

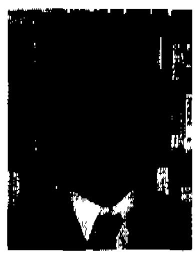
[23] S. U. Pillai and W. C. Lee, Toeplitz Inverses and Levinson Polynomials, Technical report, Department of Electrical Engineering, Polytechnic University, Brooklyn, New York, September 1992.

남 종 길(Jong Gil Nam) 정회원

1998년 : 호서대학교 공과대학 정보통신공학과 졸업 (공학사)
1999년 : 숭실대학교 대학원 정보통신공학과 석사과정

<주관심 분야> CDMA System, Smart Antenna, Multiuser Detection

이 원 철(Won-Cheol Lee) 정회원



1986년 : 서강대학교 이공대학 전자공학과 학사
1988년 : 연세대학교 대학원 전자공학과 석사
1994년 : New York Polytechnic University 박사

1994년 7월~1995년 7월 : postdoctoral Fellow (Polytechnic University)
1994년 1월~1994년 12월 : IEEE Trans. on Signal Processing 논문 심사 위원
1995년 9월~현재 : 숭실대학교 공과대학 정보통신공학과 조교수
1995년 9월~현재 : 연세대학교 신호처리 연구센터 연구원
1995년 9월~현재 : 대한음향학회 편집위원
1998년 1월~현재 : 대한음향학회 이사
1995년 1월~현재 : 한국통신학회 편집위원

<주관심 분야> 디지털 시스템 인지, 이동통신 시스템, 음성 신호 부호화 및 레이다 신호 처리