논문 99-24-4B-24 한국통신학회논문지 '99-4 Vol.24 No.4B

Performance Enhancement of CDMA Receiving System on Base Station Using a Spatio-temporal Combiner

Yong Song*, Won Cheol Lee** *Regular Members*

ABSTRACT

Considering the wireless CDMA communication system over the multipath fading channel, the spatio-temporal processing at receiver seems quite beneficial because it could alleviate the fading effect and suppress various interferences over space and time domains simultaneously. In this paper, in order to discriminate multipath components corresponding to the desired user while maximizing SINR(Signal to Interferences plus Noise Ratio), the spatial filtering technique using the adaptive beamformer is employed which is followed by the rake receiver so as to coherently combine timely spreaded user's desired signal constructively. Towards this, this paper discusses the efficient techniques for obtaining the beamformer weights in recursive manner together with estimating the fading coefficients including gains and relevant phases. To show the superiority of the proposed composite processing for the spatio-temporal combiner, the BER and the SINR curves will be shown.

I. 서 론

In recent years, due to the rapid increase of subscribers in the CDMA mobile communication arena, increased capacities of the mobile celluar system are demanded. Due to the wave propagation phenomenon, the wireless CDMA communication system critically suffers from multipath fading while imposing spatial diversity as well as time diversity. Since the bandwidth of a channel is much larger than the coherence bandwidth in the CDMA communication system, the wireless channel turns out to be frequency selective. By making use of the intrinsic correlation property of PN sequences[1] these multipaths could be resolvable, and those can be coherently combined to increase a particular user's signal power. Specially in the current CDMA system, in order to combine these multipath signals constructively, the rake receiver plays a major role of combining[2, 3].

In the CDMA communication system, since each channel is shared with numerous users in the area of the same cell, from the aspect of discriminating the specific user's signal, the signals transmitted by other users could be considered as co-channel interference. Besides, due to the usage of the nonideal orthogonal PN code allocated to each user, after despreading the received signal with the particular user's PN sequence, residual cross terms remain, which result in the self and the multiple access interference. Here it is worthwhile to mention that these interferences create serious detection errors and induce the reduction of the capacity in the CDMA communication system[4]-[6].

In this paper, the author focuses on the enhanced CDMA receiving system deploying an adaptive beamformer followed by rake receiver in concatenated fashion. Thus a series of main beams are produced for spatially discriminating multipath components associated with the designated user and temporally combined constructively.

This paper consists of following sections. Section II introduces the modeling of the received

* 선린대학 정보통신과(ysong@sunlin.ac.kr)

** 숭실대학교 정보통신공학과(wlee@saint.soongsil.ac.kr)

논문번호 : 98162-0413, 접수일자 : 1998년 4월 13일

※ This research work was supported by the Korea Research Foundation under contract no. 971-0917-106-2

baseband signal, in vector form along the antenna array together with their statistical aspects. Section III proposes the method of computing the optimum weight vector subject to satisfying the maximum SINR. In section IV, the recursive way of updating the beamformer weight vector under the slowly time-varying fading channel will be proposed. Section V discusses the estimation technique for the fading gains and phases in recursive manner.

II. Received CDMA Signal Modeling under Multipath Fading Channel

Considering the cellular CDMA communication channel in a macro-cell environment, the characteristics of this wireless channel are generally described as frequency selective and slow fading. Suppose the linear M-sensor array is mounted on the base station and multiple users are sending their own signals asynchronously. Here it is presumed that the angle spread raised from local scatteres is neglectible such that incident angles are distinctive due to remote scatteres. While K active users are residing inside of the cell, the received signals along the uniform linear array can be modeled in vector form, i.e.,

$$\mathbf{x}(t)=\sum_{k=1}^{K}\sum_{p=1}^{L_k}\alpha_{k,p}e^{j\phi_{k,p}}s_k(t-\tau_{k,p})\mathbf{a}(\theta_{k,p})+\mathbf{n}(t) \tag{1}$$

where k^{th} user's signal $s_k(t)$ has a form of PN modulated information-bearing symbols by k^{th} user's PN sequence $c_k(t)$ and $\mathbf{a}(\theta_{k,p})$ is the array response vector which is equivalent to the direction vector. And in (1), $[\tau_{k,l}]_{l=1}^{L_k}$ are time delays associated with k^{th} user. The multiplicative complex fading coefficient $\alpha_{k,p}e^{j\phi_{k,p}}$ corresponding to p^{th} multipath component of k^{th} user consists of the Rayleigh distributed gain $\alpha_{k,p}$ and uniformly distributed phase $\phi_{k,p}$.

Without a loss of generality, this paper considers the 1^{st} user as a wanted user and the time delays $\tau_{1,l}$, $l=1\cdots L_1$, are presumed to be a priori known. Then it is well-described in [7] that the covariance matrix of the sensor output signal vector has the form of

$$\mathbf{R}_{xx}=\alpha_{1,l}^2\mathbf{a}(\theta_{1,l})\mathbf{a}^*(\theta_{1,l})+\sum_{p=1,p\neq l}^{L}\alpha_{1,p}^2\mathbf{a}(\theta_{1,p})\mathbf{a}^*(\theta_{1,p})$$
$$+\sum_{k=1}^{K}\sum_{p=1}^{L}\alpha_{k,p}^2\mathbf{a}(\theta_{k,p})\mathbf{a}^*(\theta_{k,p})+\sigma_n^2\mathbf{I}. \tag{2}$$

To discriminate multipath components associated with the wanted user in temporal domain, the bank of parallel correlator depicted in Fig.1 can be utilized. Here each tap delay is identical to the length of a single chip duration. For the sake of convenience, this paper assumes that the number of multipaths associated with each user are the same, i.e., $L\equiv L_1=L_2=\cdots=L_k$.

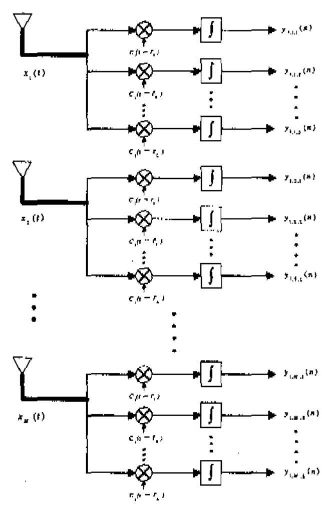


Fig. 1. Structure of a bank of parallel correlators using the spatio-temporal combiner

After the M parallel correlators, M sets of output vector $\mathbf{y}_{1,m}$, $m=1\rightarrow M$, are produced as

$$\mathbf{y}_{1,m}=[y_{1,m,1}(n)\ \ y_{1,m,2}(n)\ \ \cdots\ \ y_{1,m,L}(n)] \tag{3}$$

where the l^{th} component is the correlation output between sensor output and the delayed version of PN code assigned to first user, i.e.,

$$y_{1,m,l}(n) = \frac{1}{\sqrt{T_b}} \int_{(n-1)T_b+\tau_{1,l}}^{nT_b+\tau_{1,l}} x_m(t) c_1(t-\tau_{1,l}) dt$$
$$= \sqrt{T_b} b_1(n) a_m(\theta_{1,l}) + i_{m,1,l}(n) + n_{m,1,l}(n) \quad (4)(5)$$

where $a_m(\theta_{1,l})$ is the m^{th} sensor response, and

$$i_{m,1,l}(n) = i_{self,m,1,l}(n) + i_{MAI,m,1,l}(n). \quad (6)$$

In (6), $i_{self,m,1,l}(n)$ is denoted as the residual self interference (SI) experienced at m^{th} sensor and $i_{MAI,m,1,l}(n)$ is the multiple access interference (MAI) and $n_{m,1,l}(n)$ is the noise component. Here it can be stated that both $i_{self,m,1,l}(n)$ and $i_{MAI,m,1,l}(n)$ have the form of i.i.d with zero mean and variance $2T_C$[8].

After the parallel correlator, outputs are rearranged into L groups of discriminated multipath components denoted by

$$\mathbf{z}_{1,l}(n) = [y_{1,1,l}(n) \;\; y_{1,2,l}(n) \;\; \cdots \;\; y_{1,M,l}(n)],\; l = 1 \to L. \quad (7)$$

Here the auto-correlation matrix of $\mathbf{z}_{1,l}(n)$ is represented by

$$\mathbf{R}_{zz,1,l} \equiv E\{\mathbf{z}_{1,l}(n)\mathbf{z}^*_{1,l}(n)\} = G\beta\alpha^2_{1,l}\mathbf{a}(\theta_{1,l})\mathbf{a}^*(\theta_{1,l}) + \mathbf{R}_{uu,1,l} \quad (8)$$

where β is a modulation index, i.e., $\beta=1$ for BPSK and $\beta=2$ for QPSK, and $\mathbf{R}_{uu,1,l}$ corresponds to the covariance matrix relevant to the undesired interference and noise component. Thus, (2) can be simply represented into

$$\mathbf{R}_{xx} = \beta\alpha^2_{1,l}\mathbf{a}(\theta_{1,l})\mathbf{a}^*(\theta_{1,l}) + \mathbf{R}_{uu,1,l}. \quad (9)$$

Furthermore with the help of (9), (8) can be rewritten as

$$\mathbf{R}_{zz,1,l} = (G-1)\beta\alpha^2_{1,l}\mathbf{a}(\theta_{1,l})\mathbf{a}^*(\theta_{1,l}) + \mathbf{R}_{xx}. \quad (10)$$

Provided that the large number of multipaths impinge on the array, the interference plus noise terms behave like an white gaussian random process with its variance $\hat{\sigma}^2_n$ [8]. As a result, the covariance matrix in (8) can be simply approximated into

$$\mathbf{R}_{zz,1,l} \approx G\beta\alpha^2_{1,l}\mathbf{a}(\theta_{1,l})\mathbf{a}^*(\theta_{1,l}) + \hat{\sigma}^2_n\mathbf{I}. \quad (11)$$

Furthermore the cross-correlation matrix between $\mathbf{z}_{1,l}$ and $\mathbf{z}_{1,q}$ corresponding to the l^{th} and the q^{th} multipaths, can be expressed as

$$\mathbf{C}_{zz,1,l,q} \equiv E\{\mathbf{z}_{1,l}\mathbf{z}^*_{1,q}\}$$
$$= G\beta\alpha_{1,l}\alpha_{1,q}e^{j(\phi_{1,l}-\phi_{1,q})}\mathbf{a}(\theta_{1,l})\mathbf{a}^*(\theta_{1,q}) + \mathbf{C}_{uu,1,l,q}, \quad (12),\ (13)$$

where

$$\mathbf{C}_{uu,1,l,q} \equiv E\{(\mathbf{I}_{self,1,l} + \mathbf{I}_{MAI,1,l} + \mathbf{n}_{1,l})(\mathbf{I}_{self,1,q} + \mathbf{I}_{MAI,1,q} + \mathbf{n}_{1,q})^*\} \quad (14)$$

Here, due to the same reason as in (11), (13) can be also simplified into the following[9]

$$\mathbf{C}_{zz,1,l,q} \approx G\beta\alpha_{1,l}\alpha_{1,q}e^{j(\phi_{1,l}-\phi_{1,q})}\mathbf{a}(\theta_{1,l})\mathbf{a}^*(\theta_{1,q}). \quad (15)$$

Considering the above relationship it can state that the relative phase difference between l^{th} multipath and q^{th} multipath, i.e., $\phi_{1,l}-\phi_{1,q}$, could be obtained from (15) based on the observed cross correlation matrix and the set of directional vectors.

Ⅲ. Beamforming subject to maximizing SINR

Although L number of multipath components are discriminated by employing M banks of parallel correlators in temporal domain, due to the usage of the non-ideal orthogonal PN codes, the self and the multiple access interference are still remaining. In order to avoid the performance degradation raised from the various interferences, the receiver structure could be reconfigured as in Fig. 2 so as to achieve the high SINR(Signal to Interference plus Noise Ratio) for robust detection.

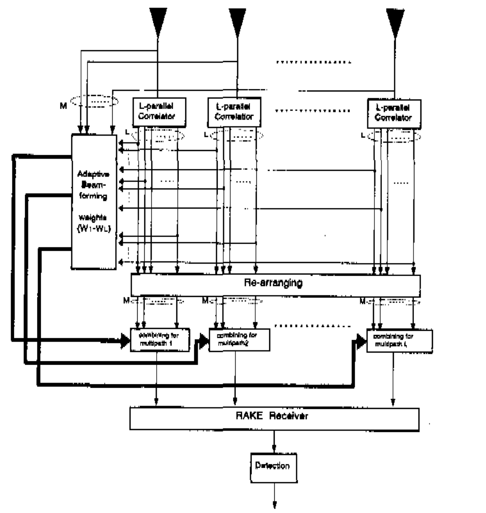


Fig. 2. A basic structure of CDMA receiver using the spatio-temporal combiner

According to[7], the optimal beamformer weights enforcing the weights to satisfy the maximum SINR have the form of

$$\mathbf{w}_{opt,1,l} \equiv \mathbf{R}_{uu,1,l}^{-1}\mathbf{a}(\theta_{1,l}). \tag{16}$$

And the resulting maximized SINR can be expressed as

$$\begin{aligned} SINR_{\max} &= G\beta\alpha_{1,l}^2 \frac{\mathbf{w}_{opt,1,l}^*\mathbf{a}(\theta_{1,l})\mathbf{a}^*(\theta_{1,l})\mathbf{w}_{opt,1,l}}{\mathbf{w}_{opt,1,l}^*\mathbf{R}_{uu,1,l}\mathbf{w}_{opt,1,l}} \\ &= G\beta\alpha_{1,l}^2\mathbf{a}^*(\theta_{1,l})\mathbf{R}_{uu,1,l}^{-1}\mathbf{a}(\theta_{1,l}). \end{aligned} \tag{17}$$

To simplify for updating the weight vector, this paper will introduce the alternative approach without involving the inversion computation corresponding to the intermediate correlation matrix $\mathbf{R}_{uu}$. Provided that the array output correlation matrix $\mathbf{R}_{xx}$ is nonsingular and Hermitian matrix, this assumption is valid in practical situation, $\mathbf{R}_{xx}$ can be factorized into two matrices as $\mathrm{R}_{xx}^{1/2*}$ and $\mathrm{R}_{xx}^{1/2*}$. Multiplying $\mathrm{R}_{xx}^{-1/2*}$ and $\mathrm{R}_{xx}^{-1/2}$ on both sides of (10) from the left and right side respectively gives rise to

$$\mathbf{R}_{xx}^{-\frac{1}{2}*}\mathbf{R}_{zz,1,l}\mathbf{R}_{xx}^{-\frac{1}{2}} = (G-1)\beta\alpha_{1,l}^2\mathbf{R}_{xx}^{-\frac{1}{2}*}\mathbf{a}(\theta_{1,l})\mathbf{a}^*(\theta_{1,l})\mathbf{R}_{xx}^{-\frac{1}{2}} + \mathbf{I}. \tag{18}$$

By letting

$$\mathbf{B} \equiv \mathbf{R}_{xx}^{-\frac{1}{2}*}\mathbf{R}_{zz,1,l}\mathbf{R}_{xx}^{-\frac{1}{2}} \quad \text{and} \quad \mathbf{b} \equiv \mathbf{R}_{xx}^{-\frac{1}{2}*}\mathbf{a}(\theta_{1,l}) \tag{19}$$

then (18) can be rewritten as

$$\mathbf{Bb} = \left((G-1)\beta\alpha_{1,l}^2\|\mathbf{b}\|^2 + 1\right)\mathbf{b}, \tag{20}$$

where $\|\cdot\|$ denotes the euclidean norm. Using the eigenvector corresponding to the maximum eigenv- alue of **B**, (20) also can be rewritten equivalently as

$$\mathbf{Bv}_{\max} = \lambda_{\max}\mathbf{v}_{\max}, \tag{21}$$

where $\mathbf{v}_{\max} \equiv \tilde{c}\,\mathbf{b}$ is the eigenvector corresponding to the maximum eigenvalue $\lambda_{\max}$ and $\tilde{c}$ is a complex scalar. The directional vector for l^{th} multipath component of the first user can be obtained by multiplying the correlation matrix $\mathbf{R}_{xx}$ on the computed eigenvector $\mathbf{v}_{\max}$ together with proper normalization step, i.e.,

$$\mathbf{a}(\theta_{1,l}) = \frac{\mathbf{R}_{xx}^{\frac{1}{2}*}\mathbf{v}_{\max}}{\tilde{c}}. \tag{22}$$

Moreover, comparing (20) with (21) gives rise to the following equality, i.e.,

$$\lambda_{\max} = (G-1)\beta\alpha_{1,l}^2\|\mathbf{b}\|^2 + 1 = (G-1)\beta\alpha_{1,l}^2\mathbf{a}^*(\theta_{1,l})\mathbf{R}_{xx}^{-1}\mathbf{a}(\theta_{1,l}) + 1. \tag{23}$$

From (23) the magnitude of fading gain for l^{th} multipath component of the first user can be computed by

$$\alpha_{1,l} = \sqrt{\frac{\lambda_{\max}-1}{(G-1)\beta\mathbf{a}^*(\theta_{1,l})\mathbf{R}_{xx}^{-1}\mathbf{a}(\theta_{1,l})}} = \sqrt{\frac{\lambda_{\max}-1}{(G-1)\beta\|\mathbf{b}\|^2}}. \tag{24}$$

In order to update the directional vector, it is necessary to calculate the eigenvector $\mathbf{v}_{\max}$. And this can be computed by solving the maximizing the Rayleigh quotient ξ as shown below[10];

$$\max_{\mathbf{v}} \xi = \max_{\mathbf{v}} \frac{\mathbf{v}^*\mathbf{B}\mathbf{v}}{\mathbf{v}^*\mathbf{v}}. \tag{25}$$

In addition to the knowledge of the directional vector $\mathbf{a}(\theta_{1,l})$ as in (22), the inverse of $\mathbf{R}_{uu,1,l}$ should be acquired as in (16) so as to obtain the optimum weight vector. To avoid this intermediate step resulting in the reduction of the computational complexity, this paper proposes another approach which allows to represent the optimal weight vector in terms of the eigenvector $\mathbf{v}_{max}$ and $\mathbf{R}_{xx}$ rather than $\mathbf{R}_{uu,1,l}$. To make a further progress, multiplying $\mathbf{R}_{xx}^{-1/2*}$ and $\mathbf{R}_{xx}^{-1/2}$ on the both sides of (9) gives rise to

$$\mathbf{R}_{xx}^{-\frac{1}{2}*}\mathbf{R}_{xx}\mathbf{R}_{xx}^{-\frac{1}{2}} \equiv \mathbf{I} = \beta\alpha_{1,l}^2\mathbf{R}_{xx}^{-\frac{1}{2}*}\mathbf{a}(\theta_{1,l})\mathbf{a}^*(\theta_{1,l})\mathbf{R}_{xx}^{-\frac{1}{2}} + \mathbf{R}_{xx}^{-\frac{1}{2}*}\mathbf{R}_{uu,1,l}\mathbf{R}_{xx}^{-\frac{1}{2}}. \tag{26}$$

Here, with the help of, $\mathbf{b} = \mathbf{R}_{xx}^{-1/2*}\mathbf{a}(\theta_{1,l})$, letting

$$\mathbf{D} \equiv \mathbf{R}_{xx}^{-\frac{1}{2}*}\mathbf{R}_{uu,1,l}\mathbf{R}_{xx}^{-\frac{1}{2}} = \mathbf{I} - \beta\alpha_{1,l}^2\mathbf{b}\mathbf{b}^* \tag{27}$$

and using the Matrix Inversion Lemma[11], the inversion matrix of $\mathbf{R}_{uu,1,l}$ can be expressed as

$$\mathbf{R}_{uu,1,l}^{-1} = \mathbf{R}_{xx}^{-\frac{1}{2}}\mathbf{D}^{-1}\mathbf{R}_{xx}^{-\frac{1}{2}*} = \mathbf{R}_{xx}^{-\frac{1}{2}}\left\{\mathbf{I} - \frac{\mathbf{b}\mathbf{b}^*}{\mathbf{b}^*\mathbf{b} - \frac{1}{\beta\alpha_{1,l}^2}}\right\}\mathbf{R}_{xx}^{-\frac{1}{2}*}. \tag{28}$$

Thus substituting (28) into (16), the optimum weight vector satisfying the maximum SINR can be represented in terms of $\mathbf{R}_{xx}$, $\alpha_{1,l}$ and $\mathbf{b}$ as follows

$$\mathbf{w}_{opt,1,l} = \left(\frac{1}{1-\beta\|\mathbf{b}\|^2\alpha_{1,l}^2}\right)\mathbf{R}_{xx}^{-\frac{1}{2}}\mathbf{b} = \left(\frac{1}{1-\beta\|\mathbf{b}\|^2\alpha_{1,l}^2}\right)\mathbf{R}_{xx}^{-1}\mathbf{a}(\theta_{1,l}). \tag{29)(30}$$

Consequently, once the directional vector is estimated or the eigenvector $\mathbf{v}_{max}$ is computed, the beamformer weight vector for l^{th} multipath of the first user can be calculated by using (30). Regardless of the two distinct expressions in (16) and (30) for the optimum weight vector, it can be easily shown that corresponding maximum SINRs are the same as (17).

Assuming that the self and multiple access interference are perfectly uncorrelated, i.e., $\mathbf{R}_{uu,1,l} \approx \hat{\sigma}_n^2\mathbf{I}$ as in (11), the weight vector can be more simply represented into[12]

$$\mathbf{w}_{opt,1,l} \approx \mathbf{a}(\theta_{1,l}). \tag{31}$$

According to (31), it is noteworthy that the eigenvector corresponding to the maximum eigenvalue of $\mathbf{R}_{zz,1,l}$ turns out to be the same as the directional vector $\mathbf{a}(\theta_{1,l})$ after the proper normalization. Furthermore, the weight vector becomes identical to the directional vector itself. And its resulting SINR can be written as

$$SINR = \frac{G\beta\alpha_{1,l}^2}{\hat{\sigma}_n^2}. \tag{32}$$

Ⅳ. Recursive computation of beamformer weight vector under the slowly timevarying channel

Considering the alternation of user's position or surrounding local scatterers, the channel characteristics is continuously varying at different time instant. To track the time-varying directional vector for the wanted user and renew the beamformer weight vector over the slowly varying channel, correlation matrix $\mathbf{R}_{xx}$ and $\mathbf{R}_{zz,1,l}$ should be updated recursively at different time instant. And their recursions are as follows

$$\hat{\mathbf{R}}_{xx}(n) = \mu\hat{\mathbf{R}}_{xx}(n-1) + \mathbf{x}(n)\mathbf{x}^*(n) \tag{33}$$

$$\hat{\mathbf{R}}_{zz,1,l}(m) = \mu\hat{\mathbf{R}}_{zz,1,l}(m-1) + \mathbf{z}_{1,l}(m)\mathbf{z}_{1,l}^*(m) \tag{34}$$

where the $\mathbf{x}(n)$ and $\mathbf{z}_{1,l}(m)$ are the sampled snapshot vector before and after parallel correlator at chip rate and symbol rate respectively, and μ is a forgetting factor whose value can be chosen as a positive constant between 0.8 and 1. The value of μ determines how fasting to adapt the alternation of the channel environment[13].

To estimate the directional vector $\mathbf{a}(\theta_{1,l})$ as in (22), it needs to calculate the inversion matrix of $\mathbf{R}_{xx}^{1/2*}$. The inversion of $\mathbf{R}_{xx}(n)$ at the n^{th} time instant can be factorized and it can be recursively updated as follows[7]

$$\mathbf{R}_{xx}^{-1}(n) \equiv \mathbf{R}_{xx}^{-\frac{1}{2}}(n)\mathbf{R}_{xx}^{-\frac{1}{2}*}(n) \tag{35}$$

and where

$$\mathbf{R}_{xx}^{-\frac{1}{2}}(n) = \mu^{-\frac{1}{2}}\mathbf{R}_{xx}^{-\frac{1}{2}}(n-1)(\mathbf{I}-\gamma\mathbf{f}\mathbf{f}^*), \tag{36}$$

$$\mathbf{f} \equiv \mu^{-\frac{1}{2}}\mathbf{R}_{xx}^{-\frac{1}{2}*}(n-1)\mathbf{X}(n),$$
$$\gamma \equiv \frac{1}{(1+\mathbf{f}^*\mathbf{f}+\sqrt{1+\mathbf{f}^*\mathbf{f}})} \tag{37}$$

With the help of (19) and (36), the matrix $\mathbf{B}(n)$ also can be written as

$$\mathbf{B}(n) = \mathbf{R}_{xx}^{-\frac{1}{2}*}(n)\mathbf{R}_{zz,1,l}(m)\mathbf{R}_{xx}^{-\frac{1}{2}*}(n). \tag{38}$$

Then, the eigenvector corresponding to the maximum eigenvalue of $\mathbf{B}(n)$ gives rise to the estimate of directional vector $\mathbf{a}(\theta_{1,l})$ as in (22). Morever if the eigenvector $\mathbf{v}_{max}(n)$ is normalized, i.e., $\mathbf{v}_{max}^*(n)\ \mathbf{v}_{max}(n) \equiv 1$, and the corresponding maximum eigenvalue is equivalent to the result of the following quadratic form

$$\lambda_{max}(n) \equiv \xi_{max}(n) = \mathbf{v}_{max}^*(n)\mathbf{B}(n)\mathbf{v}_{max(n)}. \tag{39}$$

In order to obtain the eigenvector $\mathbf{v}_{max}(n)$, the recursive algorithms such as CGM (Conjugate Gradient Method)[12] or Power Method [14] can be employed. Once the eigenvector $\mathbf{v}_{max}(n)$ corresponding to the maximum eigenvalue of $\mathbf{B}(n)$ subject to $\mathbf{v}^*(n)\ \mathbf{v}(n) = 1$ is obtained, referring to (22), the estimate of directional vector $\mathbf{a}(\theta_{1,l})$ can be represented as follows

$$\hat{\mathbf{a}}(\theta_{1,l}) = \frac{\mathbf{R}_{xx}^{\frac{1}{2}*}(n)\mathbf{v}_{max}(n)}{c}, \tag{40}$$

where the complex scalar c is the first component of the vector $\mathbf{R}_{xx}^{1/2}(n)\ \mathbf{v}_{max}(n)$. Finally, with the help of (30) together with substituting (40) into (30), the optimum weight vector of the beamformer for the l^{th} path of the first user can be calculated as follows

$$\hat{\mathbf{w}}_{opt,1,l}(n) \equiv \eta\mathbf{R}_{xx}^{-1}(n)\hat{\mathbf{a}}(\theta_{1,l})$$
$$= \frac{\eta}{c}\mathbf{R}_{xx}^{-1}(n)\mathbf{R}_{xx}^{\frac{1}{2}*}(n)\mathbf{v}_{max}(n) = \frac{\eta}{c}\mathbf{R}_{xx}^{-\frac{1}{2}}(n)\mathbf{v}_{max}(n) \tag{41}$$

where $\eta = 1/[(1-\beta\|\mathbf{v}_{max}(n)\|^2\hat{\alpha}_{1,l}^2(n))/\|c\|^2]$. Here the method for achieving the estimate of $\alpha_{1,i}$ will be discussed in the next section.

As mentioned earlier, the calculation of weight vector $\hat{\mathbf{w}}_{opt,1,l}(n)$ can be carried out in recursive manner by maximizing the quadratic form. Towards this by adopting CGM, the recursion form for estimating the eigenvector $\mathbf{v}_{max}(n)$ can be expressed as follows;

$$\mathbf{v}_{max}(n+1) = \mathbf{v}_{max}(n) + t_v(n)\mathbf{u}_v(n), \tag{42}$$

where $t_v(n)$ is the adaptive gain and $\mathbf{u}_v(n)$ is the search-direction vector at n^{th} iteration[15]. As discussed in the last section, the simplified formulations as in (11) and (31) allows the beamformer weight vector being equivalent to the eigenvector associated with the maximum eigenvalue of the cross correlation matrix $\mathbf{R}_{zz,1,l}$ up to scalar multiplication. Therefore the optimum beamformer weight vector can be simply expressed as

$$\mathbf{w}_{opt,1,l} \equiv \mathbf{a}(\theta_{1,l}) = \frac{\mathbf{h}_{max}}{\mathbf{h}_{max,1}}, \tag{43}$$

where $\mathbf{h}_{max}$ is the eigenvector associated with the maximum eigenvalue of the correlation matrix $\mathbf{R}_{zz,1,l}$ and $\mathbf{h}_{max,1}$ is the first component of the vector $\mathbf{h}_{max}$. Similar to (42), the beamformer weight vector can be obtained in a recursive manner such that the optimum weight vector at the $(n+1)^{th}$ iteration becomes as follows

$$\mathbf{w}_{opt,1,l}(n+1)=\frac{\mathbf{h}_{max}(n+1)}{\mathbf{h}_{max,1}(n+1)}=\frac{\mathbf{h}_{max}(n)+t_h(n)\mathbf{u}_h(n)}{\mathbf{h}_{max,1}(n+1)}, \quad (44)$$

where $t_h(n)$ is the adaptive gain and $\mathbf{u}_h(n)$ is the search-direction vector at n^{th} iteration.

V. Recursive channel parameter estimation and rake receiver

Regarding the structure of the CDMA receiver eadvantage of temporal diversity embedded on the CDMA signal, the rake receiver can be adopted for combining temporally spreaded versions of a particular user's signal constructively. However in general the fading coefficients denoted as the channel parameters are unknown, so that the receiver must exploit these values to complete the overall rake receiver.

Over the slowly-varying multipath fading channel the covariance matrices of the pre- and post-parallel correlator output signal vector can be utilized for estimating each multipath attenuation gain factor. To see this, once the directional vector $\mathbf{a}(\theta_{1,l})$ for l^{th} path of the first user is estimated as in (40), the maximum eigenvalue of the matrix $\mathbf{B}(n)$ also can be obtained from the knowledge of the eigenvector $\mathbf{v}_{max}(n)$, which is

$$\lambda_{1,l,\max}(n)=\mathbf{v}^{*}_{\max}(n)\mathbf{B}(n)\mathbf{v}_{\max}(n). \quad (45)$$

With a help of (24), the recursively updated eigenvector $\mathbf{v}_{max}(n)$ as in (42) can be utilized for estimating the magnitude fading gain of the multipath l^{th} component given by

$$\hat{\alpha}_{1,l}(n)\cong\sqrt{\frac{(1-\mu)\|\tilde{C}\|^2(\lambda_{1,l,\max}(n)-1)}{(1-\mu^n)(G-1)\beta\|\mathbf{v}_{max}(n)\|^2}}, \quad l=1\rightarrow L. \quad (46)$$

Furthermore, the phase terms involving in the fading coefficients should be estimated for constructing the RAKE receiver. Towards this, using (15), the cross correlation matrix between l^{th} and p^{th} parallel correlator output signal vector could gives rise to the relative phase factor. Let $\mathbf{A}_{1,q}\equiv\hat{\mathbf{a}}(\theta_{1,l})\,\hat{\mathbf{a}}(\theta_{1,q})$, then the relative phase factor denoted by $\Delta\phi_{1,q}\equiv\phi_{1,l}-\phi_{1,q}$ can be estimated from the observed cross correlation matrix denoted by $\hat{\mathbf{C}}_{zz,1,l,q}$ and the estimated directional vectors $\hat{\mathbf{a}}(\theta_{1,l})$ and $\hat{\mathbf{a}}(\theta_{1,q})$ as follows

$$\Delta\hat{\phi}_{1,q}=\arg\left(\sum_{i=1}^{M}\sum_{j=1}^{M}\frac{\left[\hat{\mathbf{C}}_{zz,1,l,q}\right]_{i,j}}{\left[\mathbf{A}_{l,q}\right]_{i,j}}\right), \quad l,q=1\rightarrow L. \quad (47)$$

And considering the phase component having uniform distribution over $[-\pi/2,\pi/2]$ with zero mean, the phase term $\phi_{1,i}$ for l^{th} multipath can be estimated as follows

$$\hat{\phi}_{1,l}=\frac{1}{L}\sum_{q=1}^{L}\Delta\hat{\phi}_{l,q}, \quad l=1\rightarrow L. \quad (48)$$

Thus, using (46) and (48) complex fading coefficient for l^{th} multipath of the first user can be estimated such that the rake receiver can be constructed for combining multipath components constructively. However in order to achieve the phase accurately, it is necessary to have enough numbers of multipath components such that the average gives rise to the exact one.

VI. Simulation Results

In order to verify the superior performance of our proposed process for spatio-temporal combiner, computer simulations were conducted under the situation in which multipath rays exist with having their own independent fading coefficients. At the receiving platform, the linearly distributed antenna array was utilized with a half-wavelength distance between two adjacent antennas.

To simulate the wireless multipath fading channel in the CDMA system, the complex fading parameters are provided whose attenuation gain has the Rayleigh distribution and phase has the uniform distribution. Here the fading parameters are presumed to be constant over the transmission. Each user was assigned with distinct cyclic shifted versions of a unique spreading sequence of length 127. And the time delays associated with each received CDMA signals are assumed to be perfectly known prior to spatio-temporal combining process. Another assumption is that the phase component of each fading parameter corresponding to the first incoming multipath for each user is recorvered in accurate. The uplink SNR for all simulations is a priori fixed at -23dB. The arrival angles are uniformly spreaded in the range of $[-60^{\circ} \cdot 60^{\circ}]$. The number of multipath rays over the transmission for each user is taken to be 3. The modulation scheme for transmitting the baseband code modulated sequence was QPSK, i.e., β=2.

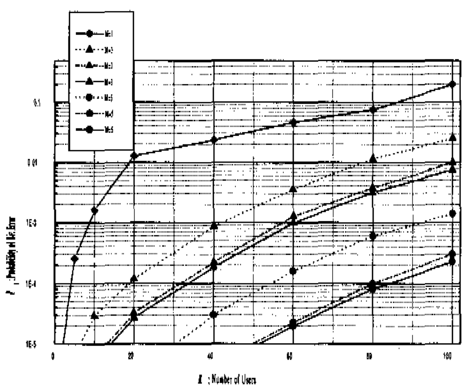


Fig. 3. BER curves for a specified number of antennas M

To investigate the performance quantatively, the BER curves are depicted in Fig.3 as increasing the number of users up to 100. Herein the solid line indicates the BER curve resulted from our proposed scheme. In this case, the optimum weight vector for each beamformer is calculated by (41) which gives rise to the maximum SINR. The dotted line in Fig.3 shows the BER curves in which the optimum weight vector is equivalent to the estimated directional vector. Also the dashed dot line indicates the BER curves resulting from the weight vector by using (16).

To see the behavior of proposed spatio-temporal combiner as increasing the number of antennas, the BER curves are depicted for the specified number of antennas, i.e., M=1, M=3 and M=5. Here it can easily notice via observing the BER curve obtained utilizing (41), our proposed scheme is superior to others. In order to see the performance improvement in the SINR sense, the maximum SINR values curves are shown in Fig.4 as increasing the number of sensors.

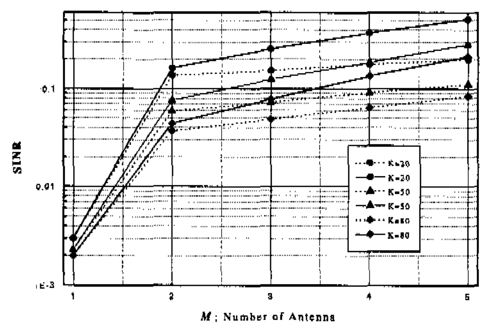


Fig. 4. SINR curves for a specified number of users K

Here it can be observed that the SINR increases as more number of antennas are utilized for the specific number of users.

Ⅶ. Concluding Remarks

This paper shows the performance improvement by employing the spatio-temporal combiner which exploits either the temporal or the spatial diversity over the multipath fading channel encountered on the CDMA communication system. The major contribution in this paper is that the optimum weight vector for a bank of beamformers is calculated in recursive manner only from the correlation matrix of sensor output signals and that of the parallel correlator output signals. Here it is a noticeable fact that the intermediate correlation matrix updated from the interference and noise signals was not involved while obtaining beamformer weights. And each beamformer output is the discriminated multipath signal associated with a wanted user's signal with having maximized SINR. Also this paper proposed the recursive way of estimating channel parameters embedded on the structure of rake receiver. To confirm the superior performance of the proposed spatio-temporal combining process, the BER curves are depicted by conducting the computer simulations. As a result, it is noticed that the error rate turns out to be very low at low SNR. Furthermore the SINR after the spatio-temporal combiner gets improved as the number of antennas as increasing specified number of users.

References

[1] L. F Chang, "Dispersive Fading Effects in CDMA Radio Systems," in *proceedings of First Int. Conf. on Universal Personal Communica- tions*, (Dallas, TX), September 1992.

[2] R. L. Peterson, R. E. Ziemer, and D. E. Borth, *Introduction to Spread Spectrum Communications*, Englewood Cliffs, New Jersey: Prentice- Hall, 1995.

[3] G. L. Stuber, *Principles of Mobile Communication*, KAP. Kluwer Academic Publishers.

[4] A. Naguib and A. Paulraj, and T. Kailath, "Capacity Improvement with Base-Station Antenna Arrays in Cellular CDMA," *IEEE Trans.Veh.Tech., vol.* VT-43(3),pp. 691-698, August 1994.

[5] R. Kohno, H. Imai, M. Hatori, and S. Pasupath, "Combination of an adaptive array antenna and a canceller of interference for direct-sequence spread-spectrum multiple access system," *IEEE Joun. on Sel. Areas of Commun.*, vol. 8, pp. 675-682, May 1990.

[6] J. G. Proakis, *Digital Communications.* McGraw-Hill, second ed., 1989.

[7] A. Naguib, "Adaptive Antennas for CDMA Wireless Networks," Master's thesis, Stanford University, August 1996.

[8] A. Naguib, Arogyaswami Paulraj, "Effects of Multipath and Base-Station Antenna Arrays on Uplink Capacity of cellular CDMA," *GLOB- COM'94*

[9] R. B. Ertel, "Spatial Filtering with Adaptive Antenna Array Algorithms in DS-CDMA Com- munications Systems," *Master's thesis, Pennsyl- vania State University*, 1996

[10] L. Bergamasch, G. Gambolati, G. Din, and M. Putti, "Gradient eigen analysis on nested finite elements," *Advances Engineering Software*, V.27, N.1-2, 10/01/96.

[11] T. Kailath, *Linear System*, Englewood Cliffs, N.J.: Prentice-Hall, 1981.

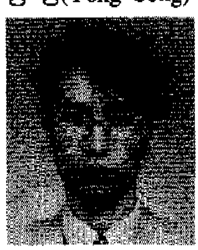
[12] X. Yang, T. K. Sarker and E. Arvas, "A Survey of Conjugate Gradient Algorithems for Solution of Extreme Eigen-Problems of a Symmetric matrix," *IEEE Trans. Acoust. Speech Signal Process.*, vol.37, No.10, October 1989, pp. 1550-1556.

[13] B. Widrow and S. Stern, *Adaptive Signal Processing*. Englewood Cliffs, NJ:Prentice-Hall, 1985.

[14] G. H. Golub and C. F. V. Loan, "Matrix Computations," Baltimore and London: The John Hopkins University Press, Second edition ed., 1989.

[15] S. Choi, and D. Yun, "Design of an Adaptive Antenna Array for Tracking the Source of Maximum Power and Its Application to CDMA Mobile Communications," *IEEE Trans. on Antennas and Propagation*, vol.45, No.9., September 1997.

송 용(Yong Song) 정회원



Yong Song received the B.S. and M.S. degree in the Information and Telecommunication Engineering from Soongsil University, Seoul, in 1996, 1998 respectively.

He has been an full-time

lecturer of Department of Information and Telecommunication, Sunlin University. His major research area includes adaptive signal processing and mobile communication system.

이 원 철(Won Cheol Lee) 정회원

Won Cheol Lee received the B.S. degree in electronic from Seogang University in 1986 and M.S. degree from Yonsei University and Ph.D. degree from Polytechnic University, New York, in 1988, 1994 respectively. He joined Polytechnic University as a Post-Doctoral Fellow from July. 1994 to July. 1995. From Jan. 1994 to Dec. 1994, he was a member of the examining committee of IEEE Trans. on S후미 Processing. He was been an assistant professor of the department of Information and Telecommunication, Soongsil University since Sep. 1995. He has been working in the Research Institute of Yonsei University on signal processing as a researcher since Sep. 1995.

He has been as editor of the Acoustical Society of Korea and executive board since Sep. 1995. and Jan. 1998 respectively. He is currently as editor of the KICS(Korea Institute of Communication Sciences). His major research area includes digital system identification, speech signal coding and mobile communication system, radar signal processing.