Efficient Performance Enhancement Scheme for Adaptive Antenna Arrays in a Rayleigh Fading and Multicell Environments

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Abstract

In this paper, an efficient performance enhancement scheme for an adaptive antenna array under the flat and the frequency-selective Rayleigh fadings is proposed. The proposed signal enhancement scheme is the modified linear signal estimator which combines the rank N approximation by reducing noise eigenvalues(RANE) and Toeplitz matrix approximation(TMA) methods into the linear signal estimator. The proposed performance enhancement scheme is performed by not only reducing the noise component from the signal-plus-noise subspace using RANE but also having the theoretical property of noise-free signal using TMA. Consequently, the key idea of the proposed performance enhancement scheme is to greatly enhance the performance of an adaptive antenna array by removing all undesired noise effects from the post-correlation received signal. The proposed performance enhancement scheme applies at the Wiener maximal ratio combining(MRC) method which has been widely used as the conventional adaptive antenna array. It is shown through several simulation results that the performance of an adaptive antenna array using the proposed signal enhancement scheme is much superior to that of a system using the conventional method under several environments, i.e., a flat Rayleigh fading, a fast frequency-selective Rayleigh fading, a perfect/imperfect power control, a single cell, and a multicell.

Key words : Adaptive Antenna Array, Array Signal Processing, DS/CDMA.

Ⅰ. Introduction

In recent years, a great deal of research has been carried out on wideband DS/CDMA based on mobile communication system for IMT-2000/UMTS(International Mobile Telecommunications 2000/Universal Mobile Telecommunications System)[1],[2]. It is designed to provide a wide range of services to mobile and stationary users in a variety of application areas and operating environment. In such a system, all users transmit data symbol to the base station(the reverse link) over the same radio frequency(RF) channel. Thereby, each user's transmission causes the multiple access interference (MAI) with all other users. MAI significantly leads to degradation of the reverse link performance. One prominent technology to reduce the MAI component efficiently is to employ an adaptive antenna array at the base station[3],[4]. An adaptive antenna array provides diversity effect to suppress spatial interference sources such as a multipath fading. The degradation of code orthogonality due to multipath fading also has a direct impact on DS/CDMA system performance. In urban cells with fully adaptive beamforming, [5] shows the improvement of the overall orthogonality factor.

The most important aspect at an adaptive antenna array is to choose an algorithm to operate at the system. Algorithm researches for calculating a weight vector of an adaptive antenna array have been developed from viewpoint of not only an adaptive digital filter analysis in time domain such as least mean squares(LMS), recursive least squares(RLS), and direct matrix inversion(DMI)[6]~[10] but also a spectral analysis in the space domain based on eigenspace-based techniques [11]~[13]. Generally, by combining the received signal and the above appropriate algorithm in an adaptive antenna array, it can considerably reduce the MAI component of cochannel users at the base station. In this paper, we used the Wiener maximal ratio combining(MRC) method that not only has a good performance compared to the above adaptive digital filter analysis algorithms but also requires relatively small quantity of computation compared to the above spectral analysis algorithms[6],[12],[13].

The performance enhancement scheme for an adaptive antenna array in DS/CDMA system is first proposed in the literature [14]. In [14], this method only applied at the geometrically based single bounce elliptical(GBSB) channel in a microcellular environment. Specifically, the forward/backward averaging method as the performance enhancement scheme was basically used to reduce the correlation relation of the received signal. However, when this method is used at the frequency-selective

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Rayleigh fadings, we found through several simulations that the more the number of users a cell, the worse the performance of a adaptive antenna array[14].

In this paper, we introduce the efficient performance enhancement scheme to improve the performance of an adaptive antenna array under the flat and the frequency-selective Rayleigh fadings. The proposed signal enhancement scheme is the modified linear signal estimator with the rank N approximation by reducing noise eigenvalues(RANE) and Toeplitz matrix approximation(TMA) methods. The proposed performance enhancement scheme is performed by not only reducing the noise effect from the signal-plus-noise subspace using RANE but also having the theoretical property of noise-free signal using TMA. The linear signal estimator using in this paper was proposed by Ephraim and Van Trees in 1995[15],[16] to improve the performance of speech communication systems. The key idea is to decompose the vector space of the noisy signal into a "signal subspace" and an orthogonal "noise subspace" under the assumption that the additive noise is white. Also, we analyzed the modeling of the other-cell interference to assess the performance of the proposed signal enhancement scheme in a multicell environment including the inter-cell interference as well as the intra-cell interference. Most authors have analyzed the performance of an adaptive antenna array within a single cell[6]~[8],[12]~[14].

The organization of this paper is as follows. Section II presents the system model. Section III presents the proposed performance enhancement scheme for an adaptive antenna array that is the most important point in this paper. In section IV, an error performance analysis in the flat and the frequency-selective Rayleigh fadings is developed. Also, we presented the modeling of other-cell interference to apply the proposed performance enhancement scheme in the multicell. Simulation results are provided in section V, and conclusions in section VI.

Ⅱ. System Model

In order to simulate a simple system, we consider a single cell with DS/CDMA system employing BPSK. Fig. 1 shows how the receiver processes L paths from a generic user.

After the received signal is converted to the baseband signal, we can write the $K\times 1$ complex baseband signal vector $\mathbf{x}(t)$ of an array with K antenna elements at the base station as

$$\mathbf{x}(t)=\sum_{l=1}^{L}\sum_{i=1}^{K_u}\varphi_i\alpha_{l,i}(t)S_{l,i}e^{-\phi_{l,i}(t)}\mathbf{a}(\theta_{l,i})+\mathbf{n}(t) \tag{1}$$

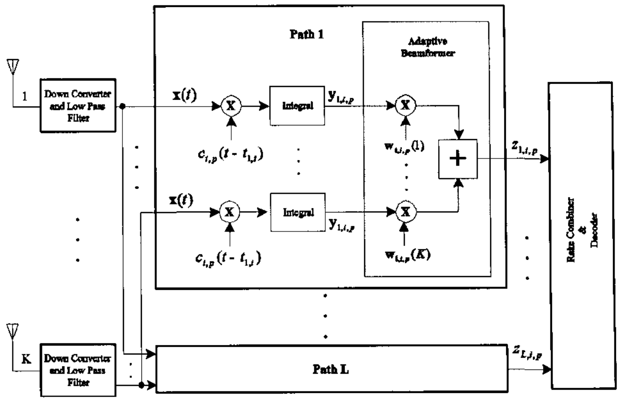


Fig. 1. Two-dimensional(temporal-spatial) rake receiver for DS/CDMA system($t_{1,i}=\tau_{1,i}$).

$$S_{l,i}=\left\{b_i\left(\frac{t-\tau_{l,i}}{T_b}\right)c_{i,t}\left(t-\tau_{l,i}\right)+c_{i,p}\left(t-\tau_{l,i}\right)\right\} \tag{2}$$

where $\alpha_{l,i}(t)$, $\tau_{l,i}$ and $\phi_{l,i}$ are the channel gain, the path delay, and the time-varying phase shift for the 1st multipath signal of the ith user. L is the number of multipath signal, K_u is the number of users, φ_i is the amplitude of the ith user, and $b_i(\cdot)$ is the bit of duration T_b. $c_{l,t}(t)$ is the unique PN code of the user's traffic signal as $c_i(\cdot)\,W_k^1$ and $c_{i,p}(t)$ is the unique PN code of user's pilot signal as $c_i(\cdot)\,W_k^0$. W_k^0 and W_k^1 are the 0th and 1st dimensional Walsh codes, respectively. T_b/T_c is the processing gain of the system where T_c is the chip period. $\mathbf{a}(\theta_{l,i})$ is the array response vector for the 1st multipath signal of the ith user as

$$\mathbf{a}(\theta_{l,i})=\left[1\quad e^{j\frac{2\pi d}{\lambda}\sin(\theta_{l,i})},\ \ldots,\ e^{j\frac{2\pi(K-1)d}{\lambda}\sin(\theta_{l,i})}\right]^T \tag{3}$$

where $[\cdot]^T$ denotes the transpose operation, d is the distance between antenna elements, K is the number of antenna elements, λ is the wavelength of the transmitted signal, and $\theta_{l,i}$ is the direction-of-arrival(DOA) for the 1st multipath signal of ith user with respect to the base station. The noise component $\mathbf{n}(t)$ is an additive white Gaussian noise(AWGN) with zero mean

$$E\{\mathbf{n}(t_1)\mathbf{n}^H(t_2)\}=\sigma_n^2\mathbf{I}_K\cdot\delta(t_1-t_2) \tag{4}$$

where σ_n^2 is the noise variance per an antenna element and $(\cdot)^H$ denotes the Hermitian transpose operation. $\mathbf{I}_k$ denotes the $K\times K$ identity matrix. The amplitude φ_i is modeled as a random variable with log-normal distribution, i.e., $10^{x/10}$ where $x\sim N(0,\ \sigma_n^2)$ and σ_n^2 is the standard deviation of the power control error(PCE) in dB[17],[18]. In a system with the perfect power control, all φ_i's are equal.

The pre-correlation signal $\mathbf{x}(t)$ is the composite component of the traffic signal and the pilot signal. Some DS/CDMA systems have used the pilot signal to efficiently calculate a weight vector of an adaptive antenna array in the reverse link[6],[7],[13],[14]. Thus we used the pilot signal to simply estimate a weight vector of an adaptive antenna array in this paper.

The post-correlation signal $\mathbf{y}_{1,i,p}$ for the 1st multipath component of the ith user's pilot signal after despreading can be written as

$$\mathbf{y}_{1,i,p}(t)=\frac{1}{\sqrt{T_b}}\int_{\tau_{1,i}}^{T_b+\tau_{1,i}}\mathbf{x}(t)c_{i,p}^{H}\left(t-\tau_{1,i}\right)dt \tag{5}$$

$$=\varphi_i\alpha_{1,i}(t)\sqrt{T_b}e^{-j\phi_{1,i}(t)}\mathbf{a}(\theta_{1,i})+\mathbf{i}_{1,i,p}(t)+\mathbf{m}_{1,i,p}(t)+\mathbf{n}_{1,i,p}(t) \tag{6}$$

The component $\mathbf{i}_{1,i,p}(t)$ is the self interference(SI) due to only one user's multipath interference as

$$\mathbf{i}_{1,i,p}(t)=\sum_{l=2}^{L}\varphi_i\alpha_{l,i}(t)I_{l,i}e^{-j\phi_{l,i}(t)}\mathbf{a}(\theta_{l,i}) \tag{7}$$

$$I_{l,i}=\frac{1}{\sqrt{T_b}}\int_{\tau_{1,i}}^{T_b+\tau_{1,i}}S_{l,i}c_{i,p}^{H}\left(t-\tau_{1,i}\right)dt \tag{8}$$

The component $\mathbf{m}_{1,i,p}(t)$ is the K_u-1 multiple access interference(MAI) and their L multipath interference as

$$\mathbf{m}_{1,i,p}(t)=\sum_{l=1}^{L}\sum_{k=1,k\neq i}^{K_u}\varphi_k\alpha_{l,k}(t)I_{l,k}e^{-j\phi_{l,k}(t)}\mathbf{a}(\theta_{l,k}) \tag{9}$$

$$I_{l,k}=\frac{1}{\sqrt{T_b}}\int_{\tau_{1,i}}^{T_b+\tau_{1,i}}S_{l,k}c_{i,p}^{H}\left(t-\tau_{1,i}\right)dt \tag{10}$$

The component $\mathbf{n}_{1,i,p}(t)$ is the undesired component due to thermal noise as

$$\mathbf{n}_{1,i,p}(t)=\frac{1}{\sqrt{T_b}}\int_{\tau_{1,i}}^{T_b+\tau_{1,i}}\mathbf{n}(t)c_{i,p}^{H}\left(t-\tau_{1,i}\right)dt \tag{11}$$

We can easily show that the $K\times K$ complex array covariance matrix $\mathbf{R}_{xx,1,i}$ of the pre-correlation signal $\mathbf{x}(t)$ for the 1st multipath signal of the ith user is given by

$$\mathbf{R}_{xx,1,i}=\varphi_i^2\alpha_{l,i}^2(t)\mathbf{a}(\theta_{1,i})\mathbf{a}^H(\theta_{1,i})+\mathbf{Q}_{xx,1,i} \tag{12}$$

$$\mathbf{Q}_{xx,1,i}=\underbrace{\sum_{l=1}^{L}\varphi_i^2\alpha_{l,i}^2(t)\mathbf{a}(\theta_{1,i})\mathbf{a}^H(\theta_{1,i})}_{\text{SI}}+\underbrace{\sum_{l=1}^{L}\sum_{k=1,k\neq i}^{K_u}\varphi_k^2\alpha_{l,k}^2(t)\mathbf{a}(\theta_{1,k})\mathbf{a}^H(\theta_{1,k})}_{\text{MAI}}+\sigma_n^2\mathbf{I}_K \tag{13}$$

The matrix $\mathbf{Q}_{xx,1,i}$ is the complex array covariance matrix of total undesired signals except for the 1st multipath signal of the ith user. We can also obtain the post-processing complex array covariance matrix for the 1st multipath signal of the ith user's pilot as

$$\mathbf{R}_{yy,1,i,p}=2\varphi_i\alpha_{l,i}(t)\mathbf{a}(\theta_{1,i})\mathbf{a}^H(\theta_{1,i})+\underbrace{\mathbf{Q}_{yy,1,i,p}}_{\text{SI+MAI+NOISE}} \tag{14}$$

In this paper, we used the Wiener MRC method [6],[12],[13] as the simple and efficient conventional algorithm to calculate the weight vector of an adaptive antenna array for the 1st multipath signal of the ith user's pilot signal. Generally, the weight vector by the Wiener MRC method is given by

$$\mathbf{w}_{l,i,p}=\mathbf{R}_{xx}^{-1}\left(\frac{1}{M}\sum_{l=1}^{M}\mathbf{y}_{1,i,p}(l)\right) \tag{15}$$

where M is the number of snapshots. For a maximal ratio combining in the two-dimensional Rake receiver, the complex decision of the ith user's pilot signal is given by

$$\mathbf{z}_{i,p}(t)=\sum_{l=1}^{L}\mathbf{w}_{l,i,p}^{H}\mathbf{y}_{l,i,p}(t) \tag{16}$$

Ⅲ. The Proposed Signal Enhancement Scheme

In this section, we propose the efficient performance enhancement scheme using the modified linear signal estimator for an adaptive antenna array. The modified linear signal estimator is to combine the rank N approximation by reducing noise eigenvalues(RANE) and Toeplitz matrix approximation(TMA) methods into the linear signal estimator. The main objective of the proposed performance enhancement scheme is to maximize the noise component reduction while minimizing the distortion of the desired signal.

3-1 The Modified Linear Signal Estimator Using the RANE and TMA Methods

For presentation simplicity, we can rewrite (5) as follows

$$\mathbf{y}_{1,i,p}(t)=\mathbf{d}_{1,i,p}(t)+\mathbf{i}_{1,i,p}(t)+\mathbf{m}_{1,i,p}(t)+\mathbf{n}_{1,i,p}(t)$$

$$=\mathbf{d}_{1,i,p}(t)+\mathbf{c}_{1,i,p}(t)+\mathbf{n}_{1,i,p}(t) \tag{17}$$

where

$$\mathbf{d}_{1,i,p}(t)=\varphi_i\alpha_{l,i}(t)\sqrt{T_b}e^{-j\phi_{1,i}(t)}\mathbf{a}(\theta_{1,i}) \tag{18}$$

$$\mathbf{c}_{1,i,p}(t)=\mathbf{i}_{1,i,p}(t)+\mathbf{m}_{1,i,p}(t) \tag{19}$$

The interference signal vector $\mathbf{c}_{1,i,p}(t)$ is modeled as a complex Gaussian random vector with zero mean and covariance matrix $\mathbf{c}_{1,i,p}=E\{\mathbf{c}_{1,i,p}(t)\,\mathbf{c}^{H}_{1,i,p}(t)\}$. When the number of cochannel users is large, $\mathbf{c}_{1,i,p}(t)$ can be modeled as spatially white Gaussian noise[19],[20]. In this case

$$\mathbf{C}_{1,i,p} = \sigma_I^2 \mathbf{I}_K \tag{20}$$

where σ_I^2 is given by [19]. The covariance matrix $\mathbf{Q}_{yy,1,i,p}$ of the total interference-plus-noise vector in (14) is then given by

$$\mathbf{Q}_{yy,1,i,p} = \sigma^2 \mathbf{I}_K \tag{21}$$

where $\sigma^2 = \sigma_I^2 + \sigma_n^2$.

The model (17) simply can be rewritten as

$$\mathbf{y}_{1,i,p}(t) = \mathbf{d}_{1,i,p}(t) + \mathbf{v}_{1,i,p}(t) \tag{22}$$

where $\mathbf{v}_{1,i,p}(t)$ is the total interference-plus-noise component given by

$$\mathbf{v}_{1,i,p}(t) = \mathbf{c}_{1,i,p}(t) + \mathbf{n}_{1,i,p}(t) \tag{23}$$

Let $\hat{\mathbf{y}}_{l,i,p}(t) = \mathbf{B}\mathbf{y}_{l,i,p}(t)$ be a linear signal estimator of the desired signal $\mathbf{d}_{1,i,p}(t)$ where B is a $K \times K$ matrix. The residual signal obtained this estimation is given by

$$\begin{aligned} \mathbf{r}(t) &= \hat{\mathbf{y}}_{1,i,p}(t) - \mathbf{d}_{1,i,p}(t) \\ &= (\mathbf{B} - \mathbf{I}_K)\mathbf{d}_{1,i,p}(t) + \mathbf{B}\mathbf{w}_{1,i,p}(t) = \mathbf{r}_d(t) + \mathbf{r}_w(t) \end{aligned} \tag{24}$$

where $\mathbf{r}_d(t) \equiv (\mathbf{B} - \mathbf{I}_K)\mathbf{d}_{1,i,p}(t)$ and $\mathbf{r}_w(t) \equiv \mathbf{B}\mathbf{w}_{1,i,p}(t)$. Let

$$\bar{\varepsilon}_d^2 \equiv tr\left[E\{\mathbf{r}_d(t)\mathbf{r}_d^H(t)\}\right] = tr\left[(\mathbf{B} - \mathbf{I}_K)\mathbf{R}_{dd}(\mathbf{B} - \mathbf{I}_K)^H\right] \tag{25}$$

be the energy of the signal vector $\mathbf{r}_d(t)$. $R_{dd} = E\{\mathbf{d}_{l,i,p}(t)\, \mathbf{d}^H_{1,i,p}(t)\}$ Similarly, let

$$\bar{\varepsilon}_w^2 \equiv tr\left[E\{\mathbf{r}_w(t)\mathbf{r}_w^H(t)\}\right] = \sigma_w^2 tr\left[\mathbf{B}\mathbf{B}^H\right] \tag{26}$$

denote the energy of the total interference-plus-noise vector $\mathbf{r}_w(t)$.

The optimal estimator minimizes $\bar{\varepsilon}_d^2$ while constraining $\bar{\varepsilon}_w^2$.

$$\min\ \bar{\varepsilon}_d^2 \text{ subject to } g(w) = \bar{\varepsilon}_w^2 - \alpha\sigma_w^2 \le 0 \tag{27}$$

where $0 \le \alpha \le 1$[16]. Lagrangian method is a classical method for solving a constrained minimization problem [21]. The Lagrangian of a constrained minimization problem is defined to be the scalar-valued function

$$L(\mathbf{B},\mu) = \bar{\varepsilon}_d^2 + \mu\left(\bar{\varepsilon}_w^2 - \alpha\sigma_w^2\right) \tag{28}$$

and

$$\mu\left(\bar{\varepsilon}_w^2 - \alpha\sigma_w^2\right) = 0 \quad \text{for } \mu \ge 0 \tag{29}$$

where μ is the Lagrange multiplier. The stationary feasible point **B** of the Lagrangian is the potential solution of the constrained optimization problem.

$$\nabla_B L(\mathbf{B},\mu) = 0 \tag{30}$$

where $\mu \ge 0$. Therefore, we can obtain

$$\nabla_B L(\mathbf{B},\mu) = 2\mathbf{B}\mathbf{R}_{dd} - 2\mathbf{R}_{dd} + 2\mathbf{B}\mu\sigma_w^2\mathbf{I}_K = 0 \tag{31}$$

$$\mathbf{B} = \mathbf{R}_{dd}\left(\mathbf{R}_{dd} + \mu_1\sigma_w^2\mathbf{I}_K\right)^{-1} \tag{32}$$

μ_1 is the Lagrange multiplier in connection with adjustable input noise level $\mu_1\sigma_w^2$ after despreading. In reality, because the channel vector $\mathbf{a}(\theta_{1,i})$ in (18) is not estimated perfectly, the desired signal covariance matrix $\mathbf{R}_{dd}$ can't be applied at (32). Although $\mathbf{R}_{dd}$ is used in (32), the estimation of $\mathbf{R}_{dd}$ from $\mathbf{R}_{xx}$ has the much computational complexity burden. Hence, $\mathbf{R}_{dd}$ can be replaced by $\mathbf{R}_{xx}$ to output the effect of the optimal estimator maximally.

$$\mathbf{B} = \mathbf{R}_{xx}\left(\mathbf{R}_{xx} + \mu_2\sigma_w^2\mathbf{I}_K\right)^{-1} \tag{33}$$

where μ_2 is the Lagrange multiplier in connection with adjustable input noise level $\mu_2\sigma_w^2$ before despreading. Let us denote the eigenvalues of the array covariance matrix $\mathbf{R}_{xx}$ by λ_i and the corresponding eigenvectors by $\hat{e}_i$. Applying the eigendecomposition method to (33), **B** can be rewritten as

$$\mathbf{B} = \sum_{m=1}^{K} g_m \hat{e}_m \hat{e}_m^H \tag{34}$$

where g_m denotes the mth diagonal element of the eigenvalue of $\mathbf{R}_{xx}$ as

$$g_m \equiv \frac{\hat{\lambda}_m}{\hat{\lambda}_m + \mu_2\sigma_w^2} \tag{35}$$

Hence, the linear signal estimator is given by

$$\hat{\mathbf{y}}_{1,i,p}(t) = \left[\sum_{m=1}^{K} g_m \hat{e}_m \hat{e}_m^H\right]\mathbf{y}_{1,i,p}(t) \tag{36}$$

Next, we want to decrease the undesired effect(noise eigenvalues) from the signal-plus-noise subspace of $\mathbf{R}_{xx}$ in (34). The noise power component is estimated by the eigendecomposition or the singular value decomposition (SVD) method. Namely, the undesired effect of the noise eigenvectors in **B** is decreased by applying the RANE method. Generally, the smallest signal eigenvalues of $\mathbf{R}_{xx}$ are almost equal to the (K − N) noise eigenvalues. We therefore assume that $\mathbf{R}_{xx}$ possesses the theoretical rank N.

[Theorem 1] : (Rank N Approximation by reducing Noise Eigenvalues: RANE) Let $\mathbf{R}_{xx}$ denote given matrix contained in a set of Hermitian complex $K\times K$ matrices and $\widetilde{\mathbf{R}}_{xx}$ be the described matrix with the (K − N) smallest eigenvalues where N<K. The (K − N) smallest eigenvalues of $\mathbf{R}_{xx}$ are almost equal to 0. The optimal matrix $\widetilde{\mathbf{R}}_{xx}$ then is given by the following expression so that $\| \mathbf{R}_{xx} - \widetilde{\mathbf{R}}_{xx} \|_F$ is minimized

$$\tilde{\mathbf{R}}_{xx} = \sum_{k=1}^{N} \hat{\lambda}_k \hat{e}_k \hat{e}_k^H - \hat{\lambda} \sum_{k=N+1}^{K} \hat{e}_k \hat{e}_k^H \tag{37}$$

in which

$$\hat{\lambda} = \frac{1}{K-N} \sum_{k=N+1}^{K} \hat{\lambda}_k \tag{38}$$

and λ_k are nonnegative singular values which are ordered in the monotonically nonincreasing fashion $\lambda_k \geq \lambda_{k+1}$, $\hat{e}_k$ is a eigenvector associated with λ_k, and $\| \cdot \|_F$ denotes the Frobenius norm, i.e., for any matrix S, $\| S \|_F = trace(SS^H)$.

[Proof] : See [22].

Applying (37) to (33), the linear signal estimator applying the RANE method is given by

$$\tilde{\mathbf{B}} = \tilde{\mathbf{R}}_{xx} \left(\tilde{\mathbf{R}}_{xx} + \mu_2 \sigma_w^2 \mathbf{I}_K \right)^{-1} \tag{39}$$

Generally, $\widetilde{\mathbf{B}}$ doesn't have the theoretical properties of noise-free signal in spite of using the RANE method. To complement this problem, we must find a matrix that $\widetilde{\mathbf{B}}$ fully approximates at a $K\times K$ matrix belonging to a class of linear structured matrices[22]. One of matrices having the desired theoretical properties is a Hermitian Toeplitz matrix that has a special form of a Hermitian persymmetric matrix[22],[23].

[Theorem 2] : (Toeplitz Matrix Approximation: TMA). Let $\widetilde{\mathbf{B}}$ denote a given matrix contained in $C^{K\times K}$ and $\tilde{\tilde{\mathbf{B}}}$ be contained in the set of Hermitian-Toeplitz matrices with a special form of Hermitian persymmetric matrix. The $K\times K$ Hermitian-Toeplitz matrix which minimizes $\| \mathbf{B} - \tilde{\tilde{\mathbf{B}}} \|_F$ is then expressed as

$$\tilde{\tilde{b}}(1,m) = \frac{1}{K-m} \sum_{l=1}^{K-m} \tilde{b}(l+m,l), \quad 0 \leq m \leq K-1 \tag{40}$$

in which $\tilde{b}(l, n)$ denotes the component in the lth row and the nth column of $\widetilde{\mathbf{B}}$ and $\tilde{\tilde{b}}(l, m)$ denotes the component in the 1st row and the mth column of $\tilde{\tilde{\mathbf{B}}}$ having the Hermitian Toeplitz structure.

[Proof] : See [23].

It follows that Toeplitz structured matrix approximation is made by taking the average of the components along all subdiagonals. Therefore $\tilde{\tilde{\mathbf{B}}}$ can easily obtain all rows and columns using (40) and Toeplitz function in MATLAB. Finally, the post-correlation signal $\bar{\hat{\mathbf{y}}}_{1,i,p}(t)$ with the proposed signal enhancement scheme is given by

$$\bar{\hat{\mathbf{y}}}_{1,i,p}(t) = \tilde{\tilde{\mathbf{B}}} \mathbf{y}_{1,i,p}(t) \tag{41}$$

At this time, the weight vector for the 1st multipath signal of the ith user's pilot signal in an adaptive antenna array is also given by

$$\bar{\mathbf{w}}_{1,i,p} = \mathbf{R}_{xx,l,i}^{-1} \left[\frac{1}{M} \sum_{l=1}^{M} \bar{\hat{\mathbf{y}}}_{1,i,p}(l) \right] \tag{42}$$

Ⅳ. Performance Analysis

The main objective of this section is to present a closed form expression for the probability density function(pdf) of signal-to-interference-plus-noise ratio (SINR) for MRC when the desired and interfering signals fade independently with Rayleigh statistics. Also, we deal with the modeling of the other-cell interference to access the performance of an adaptive antenna array using the proposed signal enhancement scheme in the multicell environment.

4-1 Frequency-Selective Rayleigh Fading

The SINR at the output of an adaptive antenna array is defined as:

$$\text{SINR} = 10\log_{10} \frac{\mathbf{w}\mathbf{R}_{dd}\mathbf{w}^H}{\mathbf{w}\mathbf{Q}_{xx}\mathbf{w}^H} \tag{43}$$

which is a function of the weights, the correlation matrix of the desired signal($\mathbf{R}_{dd}$), and the additive white Gaussian noise plus cochannel interference(CCI)($\mathbf{Q}_{xx}$). We suppose that there are L resolvable paths, the SINR can be derived as

$$\gamma_b = \sum_{l=1}^{L} \gamma_l \tag{44}$$

where $E_b = P_b \cdot T_b$ and $N_o = \sigma^2 \cdot 4T_b$ and γ_b is the instantaneous SINR at the output of the adaptive antenna array for the lst path in multipath fading. Since the fading on L multipath signals is mutually independent, γ_l is statistically independent. Since γ_b has a gamma distribution, with scaling parameter $\bar{\gamma}_l$ and shaping parameter L, pdf of γ_b is given by [24]

$$f_{\gamma_b}(\gamma_b)=\frac{\gamma_b^{L-1}}{(\bar{\gamma}_b K)^L (L-1)!}e^{-\frac{\gamma_b}{(\bar{\gamma}_l K)}} \tag{45}$$

where $\bar{\gamma}_1$ is the average SINR per multipath signal per antenna for user. As MAI and SI have Gaussian distribution, the error probability as a function of the γ_b is given by

$$P_b(\gamma_b)=\frac{1}{2}\mathrm{erfc}\left(\sqrt{\gamma_b}\right) \tag{46}$$

We can easily show that the error probability under the perfect power control is given by

$$\begin{aligned} P_e &= \int_0^\infty P_b(\gamma_b) f_{\gamma_b}(\gamma_b)d\gamma_b \\ &= \int_0^\infty \frac{\gamma_b^{L-1}}{(\bar{\gamma}_b K)^L (L-1)!}e^{-\frac{\gamma_b}{(\bar{\gamma}_l K)}}\frac{1}{\sqrt{\pi}}\int_{\sqrt{\gamma_b}}^\infty e^{-t^2}dt d\gamma_b \\ &= (1-\hbar_1)^L \sum_{l=0}^{L-1}(1+\hbar_1)^l \frac{(L-1+l)!(2L-1)!!}{l!(2L-1)!2^{L+1}} \\ &= \left(\frac{1-\hbar_1}{2}\right)^L \binom{2L-1}{L} \sum_{l=0}^{L-1}\left(\frac{1+\hbar_1}{2}\right)^l \end{aligned} \tag{47}$$

where $\hbar_1=\sqrt{\frac{\bar{\gamma}_1 K}{1+\bar{\gamma}_1 K}}$. If we assume that each user has an identically distributed power control error(PCE), i.e., $E\{\varphi_i\}=E\{\varphi\}$, the error probability under the imperfect power control is given by

$$P_e=\left(\frac{1-\hbar_1}{2}\right)^L \binom{2L-1}{L} \exp\left[\frac{(\hbar_2\sigma_m)^2}{2}\right] \sum_{l=0}^{L-1}\left(\frac{1+\hbar_1}{2}\right)^l \tag{48}$$

where $\hbar_2=(\ln 10)/10$ and σ_m is the standard deviation of the power control error in dB.

4-2 Flat Rayleigh Fading

In many circumstances, the channels vary slowly relative to the symbol rate. This channel model is referred to as a flat Rayleigh fading channel. For a flat Rayleigh channel(i.e., $L=1$) under the perfect power control, we then have

$$P_e=\frac{1}{2}-\frac{1}{2}\hbar_1=\frac{1}{2}\left(1-\sqrt{\frac{\bar{\gamma}_l K}{1+\bar{\gamma}_l K}}\right) \tag{49}$$

Under the imperfect power control, the error probability is given by

$$P_e=\frac{\exp\left[\frac{(\hbar_2\sigma_m)^2}{2}\right]}{2}\left(1-\sqrt{\frac{\bar{\gamma}_l K}{1+\bar{\gamma}_l K}}\right) \tag{50}$$

4-3 Modeling of the Other-Cell Interference

In DS/CDMA systems, each base station receives not only the interference from mobiles in the home cell(intra-cell interference) but also the interference from mobiles located in adjacent cells(inter-cell or other-cell interference). CCI is caused by the other-cell interference as well as the intra-cell interference in the multicell environment. There have been several researches regarding the characteristics of the other-cell interference in DS/CDMA systems[18],[25]~[27]. The amount of other-cell interference is determined by the path loss, power control scheme, handoff algorithm, and the number of users per a cell. Due to the number of those influencing factors and their random nature, exact characterization of the other-cell interference is difficult. Mean and variance of the other-cell interference varies dynamically as time passes.

For analytical convenience, hexagonal cells are approximated by circular cells with radius R. $x_{m,n}$ represents the distance between the nth mobile in the mth cell and the base station of the home cell, which is given by [25]

$$x_{m,n}=\sqrt{\psi_m^2+r_{m,n}^2+2\psi_m r_{m,n}\cos\theta_{m,n}} \tag{51}$$

where Ψ_m is the distance between the mth base station of surrounding cells and the home base station, $\gamma_{m,n}$ represents the distance between the nth mobile and its base station in the mth cell, $0\le\gamma_{m,n}\le R$. If a power control scheme with equal received power at the base station is deployed, the interference caused by the nth mobile in the mth cell would be given by

$$I_{out}=S_p\left(\frac{r_{m,n}}{x_{m,n}}\right)^\xi \tag{52}$$

where S_p is the power received at the base station in the case of the perfect power control and ξ is the path loss exponent. We assume that mobiles are evenly distributed over the circle. The probability distribution function of a mobile in the circle will be given by

$$f(r,\theta)=\begin{cases}\frac{1}{\pi R^2} & \text{for} \quad 0\le r_{m,n}\le R \quad \text{and} \quad 0\le\theta\le 2\pi \\ 0 & \text{otherwise}\end{cases} \tag{53}$$

The average other-cell interference from a mobile in the mth cell will be given by

$$\begin{aligned} E\{I_{out}\} &= \iint I_{out}(r,\theta)f(r,\theta)rd\theta dr \\ &= \int_0^{2\pi}\int_0^R S_p\left(\frac{r_{m,n}}{\sqrt{\psi_m^2+r_{m,n}^2+2\psi_m r_{m,n}\cos\theta_{m,n}}}\right)^\xi \frac{r_{m,n}}{\pi R^2}drd\theta \end{aligned} \tag{54}$$

Table 1. The relative other-cell interference factor f_{out} as function of the path loss law exponent considering different numbers of tiers without shadowing effect.

Relative other-cell interference factor				
Number of tiers	$\xi=2$	$\xi=3$	$\xi=4$	$\xi=5$
1	0.2277	0.1521	0.1216	0.1105
2	0.2471	0.1638	0.1248	0.1104
3	0.2858	0.1670	0.1253	0.1115
4	0.2890	0.1683	0.1255	0.1115
5	0.2900	0.1689	0.1256	0.1115

Assuming the wave propagation with pass loss proportional to the 4th power of distance, the integration can be evaluated analytically[25], and is given by

$$E\{I_{out}\}=4\cdot S_p\left[\tilde{\psi}^2\ln\left(\frac{\tilde{\psi}^2}{\tilde{\psi}^2-1}\right)-\frac{4\tilde{\psi}^4-6\tilde{\psi}^2+1}{2\left(\tilde{\psi}^2-1\right)^2}\right] \tag{55}$$

where $\tilde{\Psi}=\frac{\Psi_m}{R}$. The relative other-cell interference factor(f_{out}) is defined as the ratio of the average interference received from the other cells(E$\{I_{out}\}$) and the interference generated by users in the home cell(I_h). Therefore, f_{out} can be written as

$$f_{out}\equiv\frac{K_u\cdot E\{I_{out}\}}{I_h}=\frac{K_u\cdot E\{I_{out}\}}{(K_u-1)S_p} \tag{56}$$

Table 1 shows the relative other-cell interference factor(f_{out}) as function of the pass loss law exponent(ξ) considering several number of cell tiers around the home cell. It is clear that for larger values of ξ, the influence of the other cell tiers is decreasing with increasing tier number. We can therefore know from Table 1 that the value of f_{out} depends on the value of ξ and on the number of cell tiers. The values of the Table 1 is less than the values reported in [26], [27] that was obtained by simulation assuming log-normal shadowing. The values of our simulation was found analytically for the situation without shadowing.

The received signal corresponding to K_u in each of N_{oc} cells under the perfect power control, for the reverse link of the home cell, is given by

$$\mathbf{x}(t)=\underbrace{\sum_{n=1}^{K_u}\sum_{m=1}^{L}\alpha_{m,n}(t)S_{m,n}e^{-j\phi_{m,n}(t)}\mathbf{a}\left(\theta_{m,n}\right)}_{\text{home cell}}+\underbrace{\sum_{n=1}^{K_u\cdot N_{oc}}\sum_{m=1}^{L}f_{out}\alpha_{m,n}(t)S_{m,n}e^{-j\phi_{m,n}(t)}\mathbf{a}\left(\theta_{m,n}\right)+\mathbf{n}(t)}_{\text{other cell}} \tag{57}$$

Table 2. Parameters employed in the simulation.

Parameter	Specification
Carrier Frequency	2 GHz
Link	Reverse Link
Chip Rate	3.84 Mcps
Signal Structure	Pilot & Traffic Signal
Geometry of Array	Uniform Linear Array
Radiation Power of Antenna	Omni-directional
Number of Antenna Elements	2, 4, 6, 8
Processing Gain(PG)	8, 64
Symbol Rate	480 ksps, 60 ksps
Number of Simulations	500

where N_{oc} is the number of other cells. The post-correlation signal vector for the 1st multipath signal of the ith user's pilot signal at the home cell is given by

$$\mathbf{y}_{1,i,p}=\frac{1}{\sqrt{T_b}}\int_{\tau_{1,i}}^{T_b+\tau_{1,i}}\mathbf{x}(t)c_{i,p}^H\left(t-\tau_{1,i}\right)dt$$
$$=\alpha_{1,i}(t)\sqrt{T_b}e^{-j\phi_{1,i}(t)}\mathbf{a}(\theta_{1,i}) \tag{58}$$
$$+\underbrace{\underbrace{\mathbf{i}_{1,i,p}(t)+\mathbf{m}_{1,i,p}(t)}_{\text{home cell}}+\mathbf{M}_{oc}(t)+\mathbf{n}_{1,i,p}(t)}_{\text{Total interference}} \tag{59}$$

where $\mathbf{M}_{oc}(t)$ is the interference component by the other cells

$$\mathbf{M}_{oc}(t)=\sum_{n=1}^{K_u\cdot N_{oc}}\sum_{m=1}^{L}f_{out}\alpha_{m,n}(t)e^{-j\phi_{m,n}(t)}\mathbf{a}\left(\theta_{m,n}\right)m_{oc} \tag{60}$$

$$m_{oc}=\frac{1}{\sqrt{T_b}}\int_{\tau_{m,n}}^{T_b+\tau_{m,n}}S_{m,n}c_{i,p}^H\left(t-\tau_{m,n}\right)dt \tag{61}$$

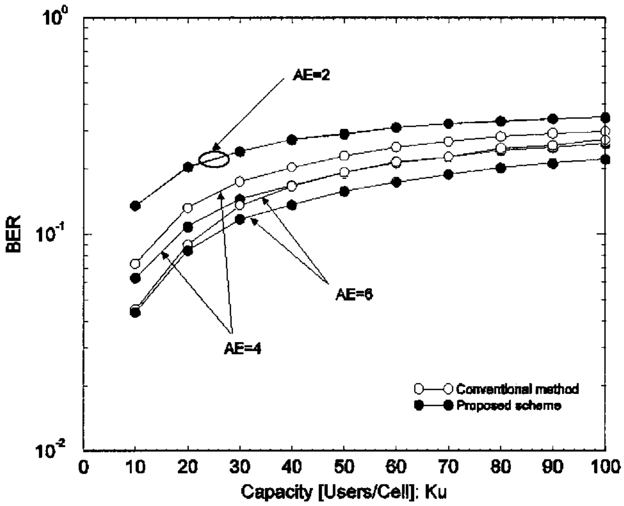
V. Simulation Results

This section deals with the performance evaluation of an adaptive antenna array using the proposed performance enhancement scheme in terms of the bit error rate(BER). In the simulation, we consider the reverse link of asynchronous DS/CDMA system. A cell site is configured with three 120° sectors and each of them is equipped with 0.5λ spaced antenna array. Besides the parameters given in Table 2, simulation conditions have the 4 multipaths per a user and 4 snapshots. The DOA value of the desired user's signal is set to θ=0°. The number of multipath components at the receiver is assumed to be equal to the number of Rake receiver fingers(i.e., an optimum system is assumed). All the multipath components are also assumed to be independent and uncorrelated. $\tau_{l,i}$ and $\theta_{l,i}$, are independent random variables uniformly distributed over [0, T_b] and

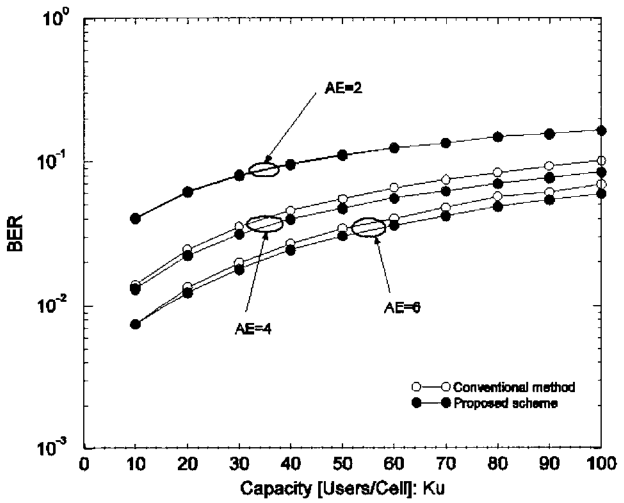
[0, 2π], respectively. $\bar{\lambda}_i$ approximately is [45.04406 8.0108 0.1334 0.0286] and [43.4322 15.8773 4.8696 0.1701 0.0364 0.0047] according to the number of antenna elements($1 \le i \le K$, K=4 and 6) through several simulations, respectively. We assume that N in (37) is $K/2$ because noise eigenvalues are almost close to 0[28]. In particular, we assess the performance of the proposed signal enhancement scheme under realistic slow/fast fading scenarios. In flat Rayleigh fading, the channel varys very slowly relative to the symbol rate. The channel Doppler spread is approximately 55 Hz for a mobile speed of 30 km/h and carrier frequency of 2.0 GHz. In fast frequency-selective Rayleigh fading, the channel Doppler spread also is approximately 222 Hz for a mobile speed of 120 km/h and the fading rate for fast fading is given by $0.125/T_b$.

5-1 Effect of the Number of Antenna Elements

We study the impact of varying the number of antenna elements on the performance of the proposed scheme. Fig. 2 and Fig. 3 show the simulation results compared the proposed scheme(eqn. 42) to the conventional method(eqn. 15) according to the number of antenna elements(2, 4, and 6) in case that the processing gain

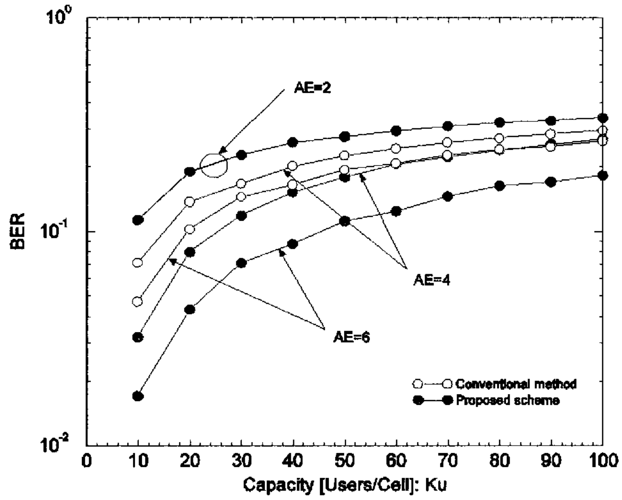


(a) PG=8

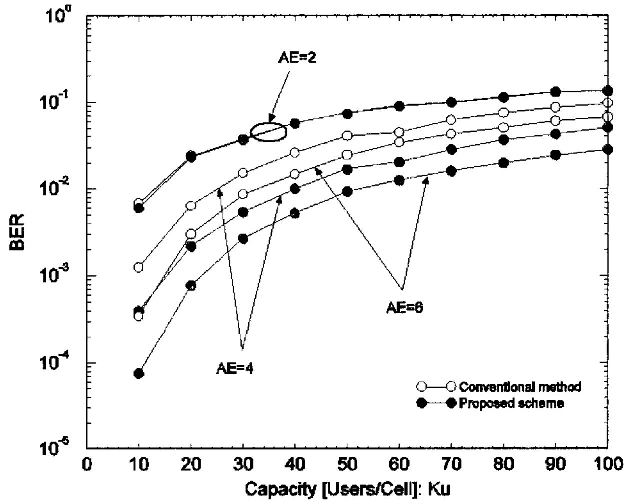


(b) PG=64

Fig. 2. The bit error rate(BER) versus the number of users for comparison between the conventional method and the proposed scheme under the flat Rayleigh fading and the perfect power control environment: PG=Processing Gain, AE=Number of Antenna Elements, SNR=10 dB.



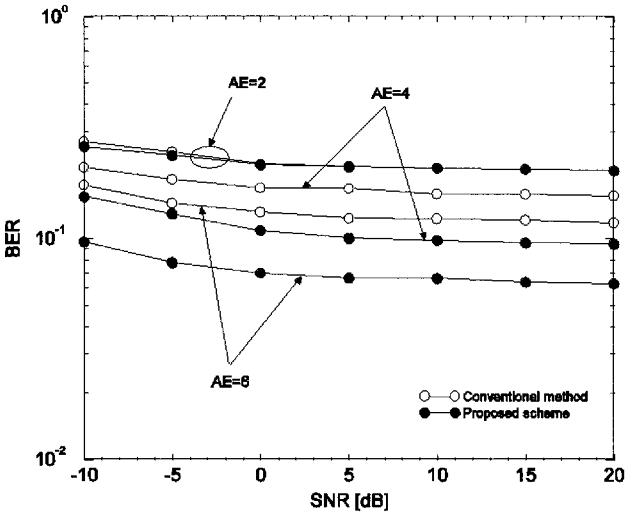
(a) PG=8



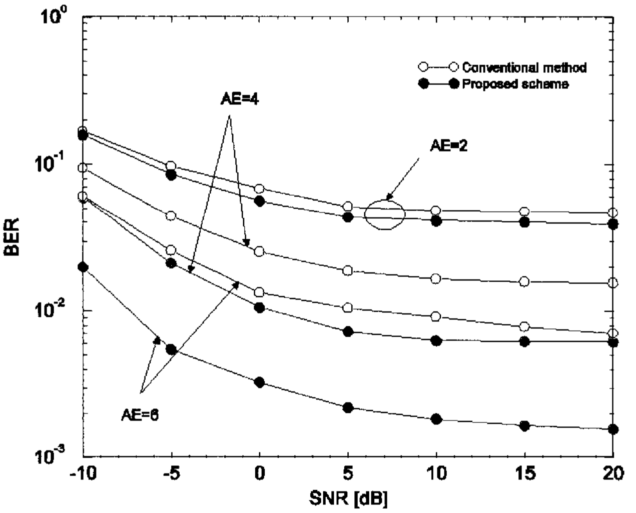
(b) PG=64

Fig. 3. The bit error rate(BER) versus the number of users for comparison between the conventional method and the proposed scheme under the fast frequency-selective Rayleigh fading, the perfect power control, and a single cell environment: PG=Processing Gain, AE=Number of Antenna Elements, SNR=10 dB.

is 8 and 64 under the SNR of 10 dB and the perfect power control environment. The simulation results show that the proposed scheme has a very good BER performance compared to the conventional method. Generally, the more the number of antenna elements, the better the performance of the proposed scheme. In the two-branch space diversity, the conventional method and the proposed scheme have similar performance in all Rayleigh fading channel cases. Especially, when the processing gain is 8, i.e., high data-rate service, the performance of the proposed scheme is much better than that of the conventional method. In case of a BER=10^{-2}



(a) PG=8



(b) PG=64

Fig. 4. The bit error rate(BER) versus the received SNR(dB) for comparison between the conventional method and the proposed scheme under the fast frequency-selective Rayleigh fading, the perfect power control, and a single cell environment: PG=Processing Gain, AE=Number of Antenna Elements, Number of users=20.

under the fast frequency-selective Rayleigh fading and the processing gain of 64, the six-branch space diversity of the proposed scheme provides a capacity of 52 users/cell, while the capacity for the conventional method is only 32 users/cell.

5-2 Effect of the SNR Value

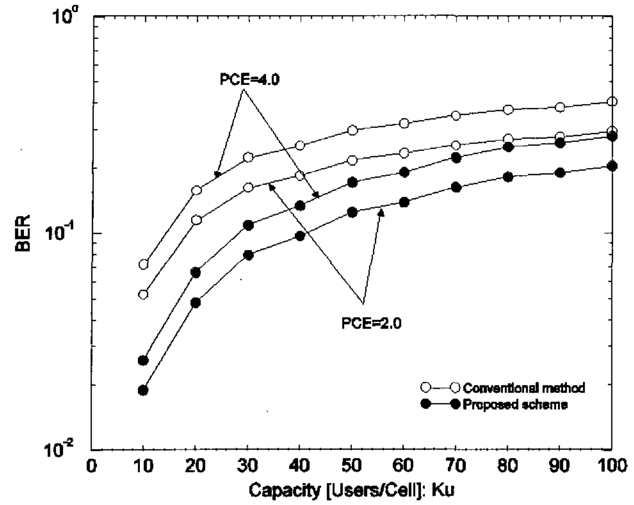
Next, Fig. 4 shows the simulation results in regard of the BER versus the received SNR(dB) for the proposed scheme under 4 antenna elements, 20 users, and the perfect power control environment. Like the previous simulation results, the simulation result is shown that the performance of the proposed scheme is much better than that of the conventional method irrespective of the processing gain. Especially, the more the SNR value and the number of antenna elements increase, the better the performance of the proposed scheme. In the six-branch space diversity and the processing gain of 64, the proposed scheme needs a SNR of -7.5 dB to obtain a BER of 10^{-2}, whereas the conventional method needs a SNR of 5 dB.

5-3 Effect of the Power Control Error

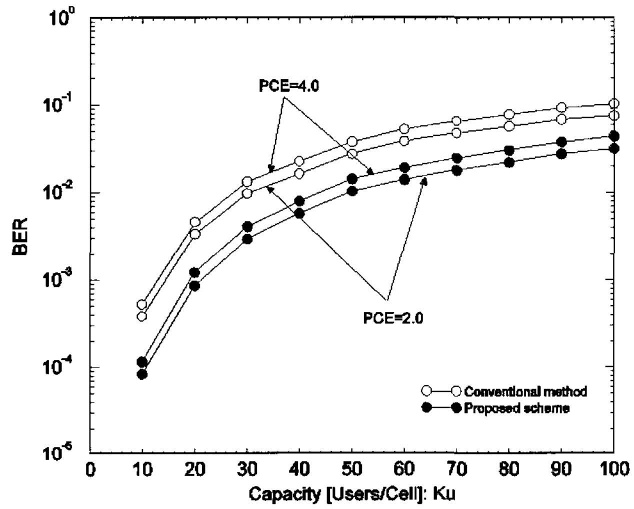
Fig. 5 is plotted as a function of the capacity and the PCE value under 6 antenna elements and the *imperfect* power control environment. We assume that each user has an identically distributed PCE value, i.e., $E\{\varphi^2_i\}= E\{\varphi^2\}$. In the imperfect power control environment, the simulation results are also shown that the performance of the proposed scheme is the superior to that of the conventional method irrespective of both the PCE value and the processing gain. Generally, the BER of the proposed scheme increases according to increasing in the PCE value for the same capacity. For a BER of 10^{-2} under the processing gain of 64, the system capacity of the proposed scheme is approximately 50 and 44 users per cell for PCE=2.0 and 4.0 dB compared to the conventional method, which is approximately 30 and 28 users per cell for PCE=2.0 and 4.0 dB, respectively.

5-4 Effect of the Other-Cell Interference

The previous examples have shown the performance of the proposed scheme in terms of single cell environment. In this example, we study the performance of the proposed scheme under the effect of other-cell interference. We take into account only the first tier of co-channel cells and the fourth-order-of-distance power control, which is the most commonly used form for the reverse link power control. In this simulation, the other-cell interferers are assumed to be uniformly dis-

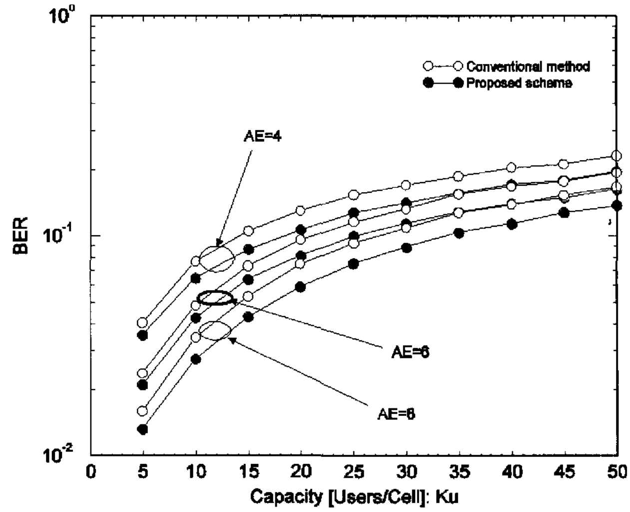


(a) PG=8

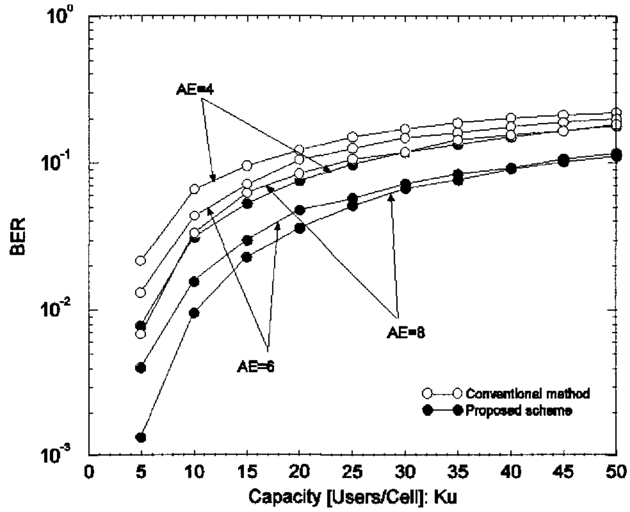


(b) PG=64

Fig. 5. The bit error rate(BER) versus the number of users for comparison between the conventional method and the proposed scheme under the fast frequency-selective Rayleigh fading, the imperfect power control, and a single cell environment: PG=Processing Gain, AE=Number of Antenna Elements=6, SNR=10 dB.



(a) Flat Rayleigh fading



(b) Fast frequency-selective Rayleigh fading

Fig. 6. The bit error rate(BER) versus the number of users for comparison between the conventional method and the proposed scheme under the multicell and the perfect power control environment: PG=Processing Gain=64, AE=Number of Antenna Elements, SNR=10 dB, f_{out}=0.1216.

tributed within their cells and the relative other-cell interference factor(f_{out}) is 0.1216. Fig. 6 shows the bit error probability for different number of users per cell in the first tier environment. The proposed scheme has a good performance compared to the conventional method under all Rayleigh channel environments. Increasing the number of antenna elements in the array cause to increase the system capacity.

Ⅵ. Conclusions

This paper has presented an efficient performance enhancement scheme for improving the performance of the Wiener MRC method-based adaptive antenna array. The proposed performance enhancement scheme is the modified linear signal estimator which combines the rank N approximation by reducing noise eigenvalues (RANE) and Toeplitz matrix approximation(TMA) methods into the linear signal estimator. The main object of the proposed signal enhancement scheme is to improve the performance of a system by reducing all undesired noise effects, i.e., having the theoretical properties of noise-free signal, from the post-correlation received signal. We can know that the performance of an adaptive antenna array using the proposed signal enhancement scheme is much superior to that of a system using the conventional method by several simulation parameters such as the processing gain, the number of antenna elements, the SNR value, and the PCE value under the perfect/imperfect power control, the fast frequency-selective Rayleigh fading, and the flat Rayleigh fading. We also confirm the performance of an adaptive antenna array using the proposed signal enhancement scheme in the multicell environment. Especially, we can know through several simulation results that the proposed signal enhancement scheme has better performance in the fast frequency-selective Rayleigh fading compared to the flat Rayleigh fading.

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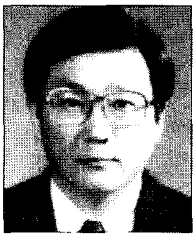

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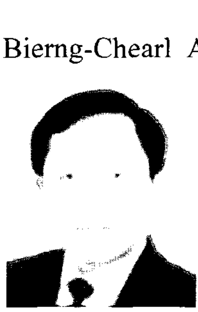

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