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A Combined Antenna Arrays and Reverse-Link Synchronous DS-CDMA System over Frequency-Selective Fading Channels with Power Control Error

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An improved antenna array (AA) has been introduced, in which reverse-link synchronous transmission technique (RLSTT) is incorporated to effectively make better an estimation of covariance matrices at a beamformer-RAKE receiver. While RLSTT is effective in the first finger at the RAKE receiver in order to reject multiple-access interference (MAI), the beamformer estimates the desired user's complex weights, enhancing its signal and reducing cochannel interference (CCI) from the other directions. In this work, it is attempted to provide a comprehensive analysis of user capacity which reflects several important factors such as the shape of multipath intensity profile (MIP), the number of antennas, and power control error (PCE). Theoretical analysis, confirmed by the simulations, demonstrates that the orthogonality provided by employing RLSTT along with AA may make the DS-CDMA system insensitive to the PCE even with fewer numbers of antennas.

Keywords and phrases: antenna arrays, reverse-link synchronous DS-CDMA, frequency-selective fading channel, power control error.

1. INTRODUCTION

DS-CDMA systems exhibit a user capacity limit in the sense that there exist a maximum number of users that can simultaneously communicate over multipath fading channels and maintain a specified level of performance per user. This limitation is caused by cochannel interference (CCI) which includes both multiple-access interference (MAI) between the multiusers, and intersymbol interference (ISI) which arises from the existence of different transmission paths. A promising approach to increase the system capacity is the use of spatial processing with an antenna array (AA) at base station (BS) [1, 2, 3, 4, 5, 6]. Generally, the AA system consists of spatially distributed antennas and a beamformer which generates a weight vector to combine the array output. Several algorithms have been proposed in the spatial signal processing to design the weights in the beamformer. For example, a new space-time processing framework for the beamforming with AA in DS-CDMA has been proposed in [2], where a code-filtering approach was used in each receiving antenna in order to estimate the optimum weights in the beamformer.

For a terrestrial mobile system, reverse-link synchronous transmission technique (RLSTT) has been proposed to reduce interchannel interference over a reverse link [7]. In the RLSTT, the synchronous transmission in the reverse link can be achieved by adaptively controlling the transmission time in each mobile station (MS). In a similar way to the closed-loop power control technique, the BS computes the time difference between the reference time generated in the BS and the arrival time of the dominant signal transmitted from each MS, and then transmits timing control bits, which order MSs to "advance" or "delay" their transmission times. The considered DS-CDMA system uses orthogonal reverse-link spreading sequences and the timing control algorithm that allows the main paths to be synchronized.

In this paper, an improved AA has been introduced, in which RLSTT is incorporated to effectively make better an estimation of covariance matrices at a Beamformer-RAKE receiver. While RLSTT is effective in the first finger at the RAKE receiver in order to reject MAI, the beamformer estimates the desired user's complex weights, enhancing its signal and reducing CCI from the other directions. In this work, it is attempted to provide a comprehensive analysis of user capacity which reflects several important factors such as the shape of multipath intensity profile (MIP), the number of antennas, and power control error (PCE). Of particular interest are the trade-offs encountered among parameters such as the number of receiving antennas and PCE. The paper is organized as follows. In Section 2, channel and system models are described. The AA system with RLSTT is introduced and its theoretical analysis is derived to investigate the trade-offs among the system parameters in Section 3. Section 4 shows numerical results mainly focusing on the system capacity. Finally, a concluding remark is given in Section 5.

2. CHANNEL AND SYSTEM MODEL

We consider a BPSK-modulated DS-CDMA system over a multipath fading channel. Assuming K active users ($k = 1, 2, \ldots, K$), the low-pass equivalent signal transmitted by user k is presented as

$$s^{(k)}(t) = \sqrt{2P_k}\, b^{(k)}(t) g^{(k)}(t) a(t) \cos\left[\omega_c t + \phi^{(k)}\right], \quad (1)$$

where $a(t)$ is a pseudonoise (PN) randomization sequence which is common to all the channels in a cell to maintain the CDMA orthogonality, $g^{(k)}(t)$ is an orthogonal channelization sequence, and $b^{(k)}(t)$ is user k's data waveform. In (1), P_k is the average transmitted power of the kth user, ω_c is the common carrier frequency, and $\phi^{(k)}$ is the phase angle of the kth modulator to be uniformly distributed in $[0, 2\pi)$. The orthogonal chip duration T_g and the PN chip interval T_c is related to data bit interval T through processing gain

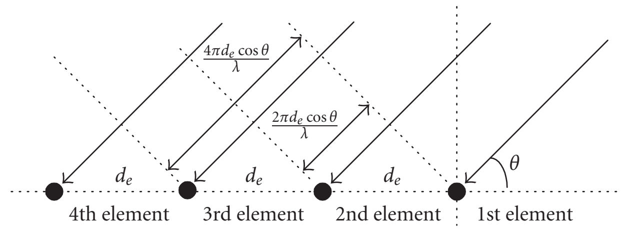


FIGURE 1: Antenna array model geometry.

$N = T/T_c$. We assume, for simplicity, that T_g equals T_c. The complex lowpass impulse response of the vector channel associated with the kth user may be written as [3]

$$\mathbf{h}_k(\tau) = \sum_{l=0}^{L^{(k)}-1} \beta_l^{(k)} \exp\left(j\varphi_l^{(k)}\right) \mathbf{V}\left(\theta_l^{(k)}\right) \delta\left[\tau - \tau_l^{(k)}\right], \quad (2)$$

where $\beta_l^{(k)}$ is the Rayleigh fading strength, $\varphi_l^{(k)}$ is its phase shift, and $\tau_l^{(k)}$ is the propagation delay. The kth user's lth path array response vector is expressed as

$$\mathbf{V}\left(\theta_l^{(k)}\right) = \left[1 \ \exp\left(\frac{-j2\pi d \cos\theta_l^{(k)}}{\lambda}\right) \cdots \exp\left(\frac{-j2(M-1)\pi d \cos\theta_l^{(k)}}{\lambda}\right)\right]^T. \quad (3)$$

Throughout this paper, we consider that the array geometry, which is the parameter of the antenna aperture gain, is a uniform linear array (ULA) of M identical sensors in Figure 1. All signals from MS arrive at the BS AA with mean angle of arrival (AOA) $\theta_l^{(k)}$ which is uniformly distributed in $[0, \pi)$. Assuming Rayleigh fading, the probability density function (pdf) of signal strength associated with the kth user's lth propagation path, $l = 0, 1, \ldots, L^{(k)} - 1$, is presented as

$$p\left(\beta_l^{(k)}\right) = \frac{2\beta_l^{(k)}}{\Omega_l^{(k)}} \exp\left(-\frac{\left(\beta_l^{(k)}\right)^2}{\Omega_l^{(k)}}\right), \quad (4)$$

where $\Omega_l^{(k)}$ is the second moment of $\beta_l^{(k)}$ with $\sum_{l=0}^{\infty} \Omega_l = 1$, and we assume it is related to the second moment of the initial path strength $\Omega_0^{(k)}$ for exponentially decaying MIP as

$$\Omega_l^{(k)} = \Omega_0^{(k)} \exp(-l\delta), \quad \text{for } 0 < l \le L^{(k)} - 1,\ \delta \ge 0, \quad (5)$$

where δ reflects the rate at which the decay of average path strength as a function of path delay occurs. Note that a more realistic profile model may be the exponential MIP.

The receiver is a coherent RAKE receiver with AA, where the number of fingers L_r is a variable less than or equal to $L^{(k)}$ which is the number of resolvable propagation paths associated with the kth user. Perfect estimates of the channel parameters are assumed. The complex received signal is expressed as

$$\mathbf{r}(t) = \sqrt{2P}\sum_{k=1}^{K}\sqrt{\lambda_k}\sum_{l=0}^{L^{(k)}-1}\beta_l^{(k)}\mathbf{V}\left(\theta_l^{(k)}\right)b^{(k)}\left(t-\tau_l^{(k)}\right) \times g^{(k)}\left(t-\tau_l^{(k)}\right)a\left(t-\tau_l^{(k)}\right)\cos\left[\omega_c t+\psi_l^{(k)}\right]+\mathbf{n}(t), \tag{6}$$

where P is the average received power and $\psi_l^{(k)}$ is the phase of the lth path associated to the kth carrier. λ_k corresponds to the PCE of the kth user which is a random variable due to imperfect power control [8]. We consider λ_k to be log-normally distributed with standard deviation σ_{λ_k} dB. In other words, $\lambda_k = 10^{(x/10)}$, where the variable x follows a normal distribution. $\mathbf{n}(t)$ is an $M\times 1$ spatially and temporally white Gaussian noise vector with a zero mean and covariance which is given by $E\{\mathbf{n}(t)\mathbf{n}^H(t)\} = \sigma_n^2\mathbf{I}$, where $\mathbf{I}$ is the $M\times M$ identity matrix, σ_n^2 is the antenna noise variance with $\eta_0/2$, and the superscript H denotes the Hermitian-transpose operator. When the received signal is matched to the reference user's code, the lth multipath matched filter output for the interest user ($k=1$) can be expressed as

$$\begin{aligned}\mathbf{y}_l^{(1)} &= \int_{\tau_l^{(1)}}^{\tau_l^{(1)}+T}\mathbf{r}(t)\cdot g^{(1)}\left(t-\tau_l^{(1)}\right)a\left(t-\tau_l^{(1)}\right)\cos\left[\omega_c t+\psi_l^{(1)}\right]dt \\ &= \mathbf{S}_l^{(1)}+\mathbf{I}_{l,\mathrm{mai}}^{(1)}+\mathbf{I}_{l,\mathrm{si}}^{(1)}+\mathbf{I}_{l,\mathrm{ni}}^{(1)}.\end{aligned} \tag{7}$$

When a reference signal is not available, a common criterion for optimizing the weight vectors and this criterion is to maximize the signal-to-interference plus noise ratio (SINR). In (7), $\mathbf{u}_l^{(1)} = \mathbf{I}_{l,\mathrm{si}}^{(1)}+\mathbf{I}_{l,\mathrm{mai}}^{(1)}+\mathbf{I}_{l,\mathrm{ni}}^{(1)}$ is a total interference plus noise for the lth path of interest user. By solving the following problem, we can obtain the optimal weights to maximize the SINR [9]:

$$\mathbf{W}_{l(\mathrm{opt})}^{(1)} = \max_{\mathbf{W}\neq\mathbf{0}}\frac{\mathbf{W}_l^{(1)^H}\mathbf{R}_{yy}\mathbf{W}_l^{(1)}}{\mathbf{W}_l^{(1)^H}\mathbf{R}_{uu}\mathbf{W}_l^{(1)}}, \tag{8}$$

where $\mathbf{R}_{yy}$ and $\mathbf{R}_{uu}$ are the second-order correlation matrices of the received signal subspace and the interference plus noise subspace, respectively. Here, $\mathbf{R}_{uu}$ can be estimated by the code-filtering approach in [2], which is presented as

$$\mathbf{R}_{uu} = \frac{N}{N-1}\left(\mathbf{R}_{rr}-\frac{1}{N}\mathbf{R}_{yy}\right), \tag{9}$$

where $\mathbf{R}_{rr}$ means the covariance matrix of the received signal prior to RAKE. The solution corresponds to the largest eigenvalue ($\lambda_{\max}$) of the generalized eigenvalue problem in the matrix pair ($\mathbf{R}_{yy}, \mathbf{R}_{uu}$). Therefore, we can obtain the maximum SINR when the weight vector $\mathbf{W}_{l(\mathrm{opt})}^{(1)}$ equals the principal eigenvector of the matrix pair, which is presented as

$$\mathbf{R}_{yy}\cdot\mathbf{W}_{l(\mathrm{opt})}^{(1)} = \lambda_{\max}\cdot\mathbf{R}_{uu}\cdot\mathbf{W}_{l(\mathrm{opt})}^{(1)}. \tag{10}$$

From (7) and (8), the corresponding beamformer output for the lth path of interest user is

$$\begin{aligned}\hat{z}_l^{(1)} &= \mathbf{W}_l^{(1)^H}\cdot\mathbf{y}_l^{(1)} \\ &= \hat{S}_l^{(1)}+\hat{I}_{l,\mathrm{mai}}^{(1)}+\hat{I}_{l,\mathrm{si}}^{(1)}+\hat{I}_{l,\mathrm{ni}}^{(1)},\end{aligned} \tag{11}$$

where

$$\begin{aligned}\hat{S}_l^{(1)} &= \sqrt{P\lambda_1/2}\beta_l^{(1)}C_{ll}^{(1,1)}b_0^{(1)}T, \\ \hat{I}_{l,\mathrm{mai}}^{(1)} &= \sqrt{P/2}\sum_{k=2}^{K}\sqrt{\lambda_k}\sum_{j=0}^{L^{(k)}-1}\beta_j^{(k)}C_{lj}^{(l,k)} \\ &\quad\times\left\{b_{-1}^{(k)}RW_{k1}\left[\tau_{lj}^{(k)}\right]+b_0^{(k)}\widehat{RW}_{k1}\left[\tau_{lj}^{(k)}\right]\right\}\cos\left[\Psi_{lj}^{(k)}\right], \\ \hat{I}_{l,\mathrm{si}}^{(1)} &= \sqrt{P\lambda_1/2}\sum_{\substack{j=0\\ j\neq l}}^{L^{(1)}-1}\beta_j^{(1)}C_{lj}^{(1,1)}\left\{b_{-1}^{(1)}RW_{11}\left[\tau_{lj}^{(1)}\right]\right. \\ &\quad\left.+b_0^{(1)}\widehat{RW}_{11}\left[\tau_{lj}^{(1)}\right]\right\}\cos\left[\Psi_{lj}^{(1)}\right], \\ \hat{I}_{l,\mathrm{ni}}^{(1)} &= \int_{\tau_l^{(1)}}^{\tau_l^{(1)}+T}\mathbf{W}_l^{(1)^H}\cdot\mathbf{n}(t)g^{(1)}\left(t-\tau_l^{(1)}\right) \\ &\quad\times a\left(t-\tau_l^{(1)}\right)\cos\left[\omega_c t+\psi_l^{(1)}\right]dt,\end{aligned} \tag{12}$$

with $b_0^{(1)}$ being the information bit to be detected, $b_{-1}^{(1)}$ the preceding bit, $\tau_{lj}^{(k)} = \tau_j^{(k)}-\tau_l^{(1)}$, and $\psi_{lj}^{(k)} = \psi_j^{(k)}-\psi_l^{(1)}$. $\mathbf{W}_l^{(1)} = [w_{l,1}^{(1)}w_{l,2}^{(1)}\cdots w_{l,M}^{(1)}]^T$ is the $M\times 1$ weight vector for the lth path of the first user. $C_{lj}^{(1,k)} = \mathbf{W}_l^{(1)^H}\cdot\mathbf{V}(\theta_j^{(k)})$ represents the spatial correlation between the array response vector of the kth user at the jth multipath and the weight vector of the interest user at the lth path. RW and $\widehat{RW}$ are Walsh-PN continuous partial cross-correlation functions defined by $RW_{k1}(\tau) = \int_0^{\tau} g^{(k)}(t-\tau)a(t-\tau)\cdot g^{(1)}(t)a(t)dt$ and $\widehat{RW}_{k1}(\tau) = \int_{\tau}^{T} g^{(k)}(t-\tau)a(t-\tau)g^{(1)}(t)a(t)dt$. From (11), we can obtain the Rake receiver output from MRC combining $\hat{z}^{(1)} = \sum_{l=0}^{L_r}\beta_l^{(1)}\cdot\hat{z}_l^{(1)}$ and see that the outputs of the lth branch, $l = 0, 1, \ldots, L_r-1$, consist of four terms. The first term represents the desired signal component to be detected. The second term represents the MAI from $(K-1)$ other simultaneous users in the system. The third term is the self-interference (SI) for the reference user. Finally, the last term is AWGN.

3. PERFORMANCE OF AA WITH RLSTT IN RAYLEIGH FADING CHANNEL WITH PCE

In our analysis, the evaluation is carried out for the case in which the arrival time of paths is modeled as synchronous in the first branch (i.e., for main paths) but as asynchronous in the rest of the branches (i.e., for multipaths). With the well-known Gaussian approximation, we model the MAI terms in the first branch and the other branches as a Gaussian process with variances equal to the MAI variances for $l=0$ and for $l\geq 1$, respectively. Extending the derived results in [7], the variance of MAI for $l = 0$, conditioned on the values of $\beta_l^{(1)}$ and λ_k, is

$$\overline{\sigma}^2_{\text{mai},0} = \frac{E_b T(2N-3)}{12N(N-1)} \left\{\beta_0^{(1)}\right\}^2 \sum_{k=2}^{K} \lambda_k \sum_{j=1}^{L^{(k)}-1} \Omega_j^{(k)} \zeta_{0j}^{(1,k)^2}. \quad (13)$$

Similarly, the variance of MAI for $l \geq 1$ is

$$\overline{\sigma}^2_{\text{mai},l} = \frac{E_b T(N-1)}{6N^2} \left\{\beta_l^{(1)}\right\}^2 \sum_{k=2}^{K} \lambda_k \sum_{j=0}^{L^{(k)}-1} \Omega_j^{(k)} \zeta_{lj}^{(1,k)^2}, \quad (14)$$

where $E_b = PT$ is the signal energy per bit, and $\zeta_{lj}^{(1,k)^2} = E\lfloor \{C_{lj}^{(1,k)}\}^2 \rfloor$ is the second-order characterization of the spatial correlation between the array response vector of the kth user at the jth multipath and the weight vector of interest user at the lth path, of which more detailed derivation is described in the appendix. The conditional variance of $\overline{\sigma}^2_{\text{si},l}$ is approximated by [10]:

$$\overline{\sigma}^2_{\text{si},l} \approx \frac{E_b \lambda_1 T}{4N} \left\{\beta_l^{(1)}\right\}^2 \sum_{\substack{j=0 \\ j\neq l}}^{L^{(1)}-1} \Omega_j^{(1)} \zeta_{lj}^{(1,1)^2}. \quad (15)$$

The variance of the AWGN term, conditioned on the value of $\beta_l^{(1)}$, is calculated as

$$\overline{\sigma}^2_{\text{ni},l} = \frac{T\eta_0 \zeta_{ll}^{(1,1)^2}}{4M} \cdot \left\{\beta_l^{(1)}\right\}^2. \quad (16)$$

Therefore, the output of the receiver is a Gaussian random process with mean

$$U_s = \sqrt{\frac{E_b \lambda_1 T}{2}} \sum_{l=0}^{L_r-1} \left\{\beta_l^{(1)}\right\}^2 \zeta_{ll}^{(1,1)} \quad (17)$$

and the total variance equal to the sum of the variance of all the interference and noise terms. From (13), (14), (15), and (16), we have

$$\begin{aligned}
\overline{\sigma}^2_T &= \overline{\sigma}^2_{\text{mai},0} + \sum_{l=1}^{L_r-1} \overline{\sigma}^2_{\text{mai},l} + \sum_{l=0}^{L_r-1} \left(\overline{\sigma}^2_{\text{si},l} + \overline{\sigma}^2_{\text{ni},l}\right) \\
&= E_b T \Omega_0 \\
&\quad \times \Bigg\{ \frac{(2N-3)\{q(L_r,\delta)-1\}\lambda_I \zeta_0^2 \cdot \left\{\beta_0^{(1)}\right\}^2}{12N(N-1)} \\
&\quad + \frac{(N-1)q(L_r,\delta)\lambda_I \zeta^2 \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2}{6N^2} \\
&\quad + \frac{\lambda_1\{q(L_r,\delta)-1\}\left(\zeta_0^2 \cdot \left\{\beta_0^{(1)}\right\}^2 + \zeta^2 \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2\right)}{4N} \\
&\quad + \frac{\eta_0\left(\zeta_0'^2 \cdot \left\{\beta_0^{(1)}\right\}^2 + \zeta'^2 \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2\right)}{4ME_b\Omega_0} \Bigg\}.
\end{aligned} \quad (18)$$

At the output of the receiver, SNR may be written in a more compact form as γ_s:

$$\begin{aligned}
\gamma_s = \Bigg\{ & \frac{(2N-3)\{q(L_r,\delta)-1\}\lambda_I}{6N(N-1)} \cdot \frac{\zeta_0^2 \cdot \left\{\beta_0^{(1)}\right\}^2}{\zeta_0' \cdot \left\{\beta_0^{(1)}\right\}^2 + \zeta' \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2} \\
& + \frac{(N-1)q(L_r,\delta)\lambda_I}{3N^2} \cdot \frac{\zeta^2 \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2}{\zeta_0' \cdot \left\{\beta_0^{(1)}\right\}^2 + \zeta' \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2} \\
& + \frac{\{q(L_r,\delta)-1\}\lambda_1}{2N} \cdot \frac{\zeta_0^2 \cdot \left\{\beta_0^{(1)}\right\}^2 + \zeta^2 \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2}{\zeta_0' \cdot \left\{\beta_0^{(1)}\right\}^2 + \zeta' \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2} \\
& + \frac{\eta_0}{2M\Omega_0 E_b} \cdot \frac{\zeta_0'^2 \cdot \left\{\beta_0^{(1)}\right\}^2 + \zeta'^2 \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2}{\zeta_0' \cdot \left\{\beta_0^{(1)}\right\}^2 + \zeta' \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2} \Bigg\}^{-1} \\
& \times \frac{\lambda_1\left(\zeta_0' \cdot \left\{\beta_0^{(1)}\right\}^2 + \zeta' \cdot \sum_{l=1}^{L_r-1}\left\{\beta_l^{(1)}\right\}^2\right)}{\Omega_0},
\end{aligned} \quad (19)$$

where $q(L_r,\delta) = \sum_{l=0}^{L_r-1}\exp(-l\delta) = 1 - \exp(-L_r\delta)/1 - \exp(-\delta)$, $\lambda_I = \sum_{k=2}^{K}\lambda_k$, and $\Omega_0^{(k)} = \Omega_0$. $\zeta_{lj}^{(k,m)^2} = \zeta_0^2$ when $k \neq m$ or $l \neq j$ for $l = 0$, $\zeta_{lj}^{(k,m)^2} = \zeta^2$ when $k \neq m$ or $l \neq j$ for $l > 0$, $\zeta_{lj}^{(k,m)^2} = \zeta_0'^2$ when $k = m$ and $l = j$ for $l = 0$, and $\zeta_{lj}^{(k,m)^2} = \zeta'^2$ when $k = m$ and $l = j$ for $l > 0$. In [11], the pdf of $\lambda_I = \sum_{k=2}^{K}\lambda_k$ for $K-1$ users is an approximately lognormal distribution, with the following logarithmic mean and variance, which is presented as

$$p(\lambda_I) = \frac{1}{\sqrt{2\pi}\sigma_{\lambda_I}\lambda_I} \exp\left[-\frac{(\ln\lambda_I - m_{\lambda_I})^2}{2\sigma_{\lambda_I}^2}\right], \quad (20)$$

where

$$\begin{aligned}
\sigma_I^2 &= \ln\left(\frac{1}{K-1}\exp\left(\sigma_{\lambda_I}^2\right) + \frac{K-2}{K-1}\right), \\
m_I &= \ln(K-1) + m + \frac{\sigma_{\lambda_I}^2}{2} \\
&\quad - \frac{1}{2}\ln\left(\frac{K-2}{K-1} + \frac{1}{K-1}\exp\left(\sigma_{\lambda_I}^2\right)\right).
\end{aligned} \quad (21)$$

This method is valid for a logarithmic standard deviation σ_λ less than 4 dB. To evaluate the average bit error probability, $P_e^l(\lambda_1,\lambda_I)$, conditioning on the values of λ_1 and λ_I follows as

$$\begin{aligned}
&P_e^l(\lambda_1,\lambda_I) \\
&= \int_0^\infty\int_0^\infty Q(\sqrt{\gamma_s}) \sum_{k=1}^{L_r-1}\frac{\pi_k}{\Omega_k}\exp(-x/\Omega_k) \cdot \frac{1}{\Omega_0}\exp(-y/\Omega_0)\,dx\,dy,
\end{aligned} \quad (22)$$

where $\pi_k = \Pi_{i=1,i\neq k}^{L_r-1}(x_k/(x_k - x_i)) = \Pi_{i=1,i\neq k}^{L_r-1}(\Omega_k/(\Omega_k - \Omega_i))$, $Q(x) = (1/\sqrt{2\pi})\int_x^\infty \exp(-u^2/2)du$. The average bit error

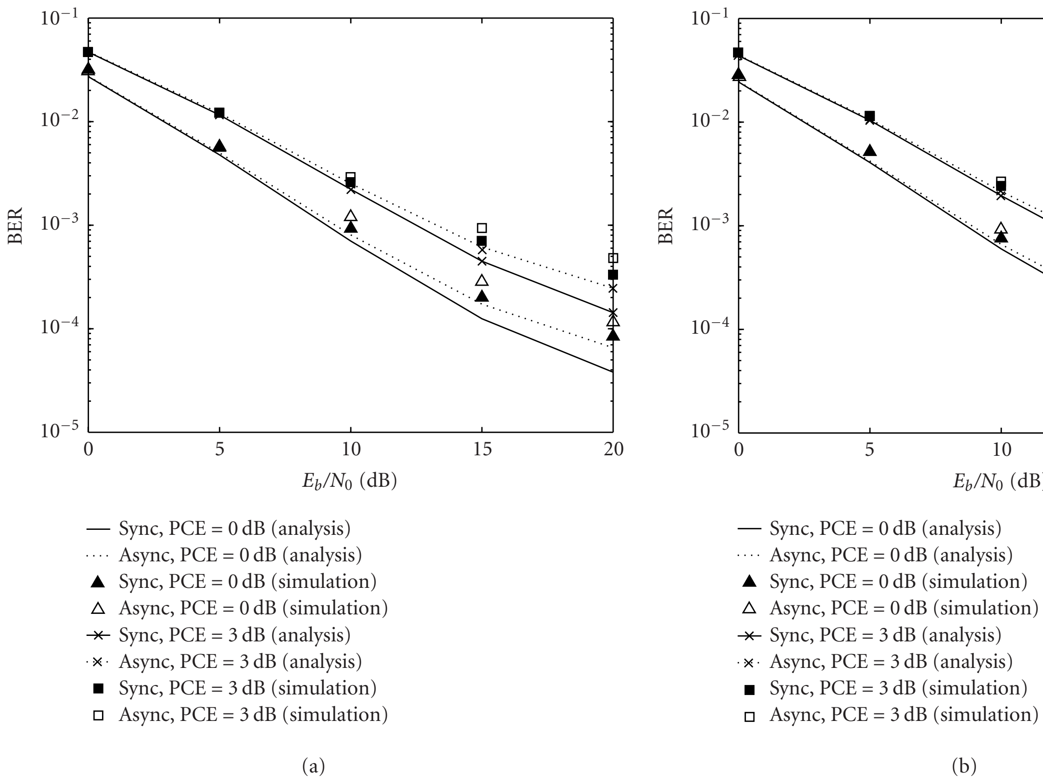


Figure 2: Analytical results versus simulation results. (Number of users = 12, $M = 4$, $L_r = L^{(k)} = 2$, PCE = 0 and 3 dB.) (a) $\delta = 1.0$, (b) $\delta = 0.2$.

probability P_e is calculated as

$$P_e = \frac{1}{\sqrt{\pi}} \int_{-\infty}^{\infty} \frac{1}{\sqrt{\pi}} \int_{-\infty}^{\infty} P_e^l \big(\exp\big(\sqrt{2}\sigma_{\lambda_1} z_1 + m_{\lambda_1}\big), \exp\big(\sqrt{2}\sigma_{\lambda_l} z_l + m_{\lambda_l}\big)\big) \times \exp\big[-z_1^2\big] dz_1 \exp\big[-z_l^2\big] dz_l, \tag{23}$$

where $z_1 = (\ln \lambda_1 - m_{\lambda_1})/\sqrt{2}\sigma_{\lambda_1}$ and $z_l = (\ln \lambda_l - m_{\lambda_l})/\sqrt{2}\sigma_{\lambda_l}$. This integration can be easily obtained by using the Hermite polynomial approach, which requires only summation and no integration [12]:

$$P_e = \frac{1}{\pi} \sum_{l=1}^{h} w_l \sum_{n=1}^{h} w_n P_e^l \big(\exp\big(\sqrt{2}\sigma_{\lambda_1} x_n + m_{\lambda_1}\big), \exp\big(\sqrt{2}\sigma_{\lambda_l} x_l + m_{\lambda_l}\big)\big). \tag{24}$$

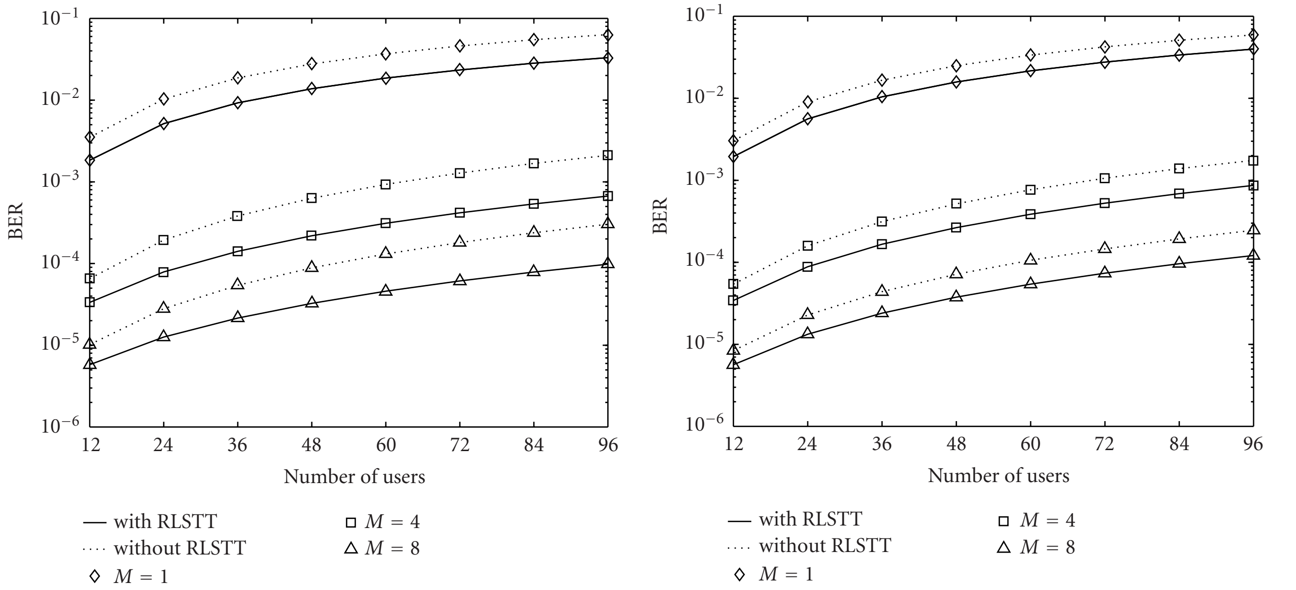
4. NUMERICAL RESULTS

In this section, we have investigated the user capacity of AA system both with RLSTT and without RLSTT, considering several important factors such as the shape of MIP, the number of antennas, and the PCE. In all evaluations, processing gain is assumed to be 128, and the number of paths and taps in RAKE is assumed to be the same for all users and denoted by two. The decaying factor is considered as 1.0 or 0.2 for the exponential MIP. The sensor spacing is half of the carrier wavelength.

Figure 2 shows uncoded BER performance as a function of E_b/N_0, when the number of users is twelve and the number of antennas is four in the exponential MIP. Two decay factors are considered, and both perfect power control (PCE = 0 dB) and imperfect power control (PCE = 3 dB) are assumed. The results confirm that the analytical results are well matched to the simulation results. It is noted that using RLSTT together with AA may enhance the performance, since RLSTT tends to make better the estimation of covariance matrices for beamformer-RAKE receiver.

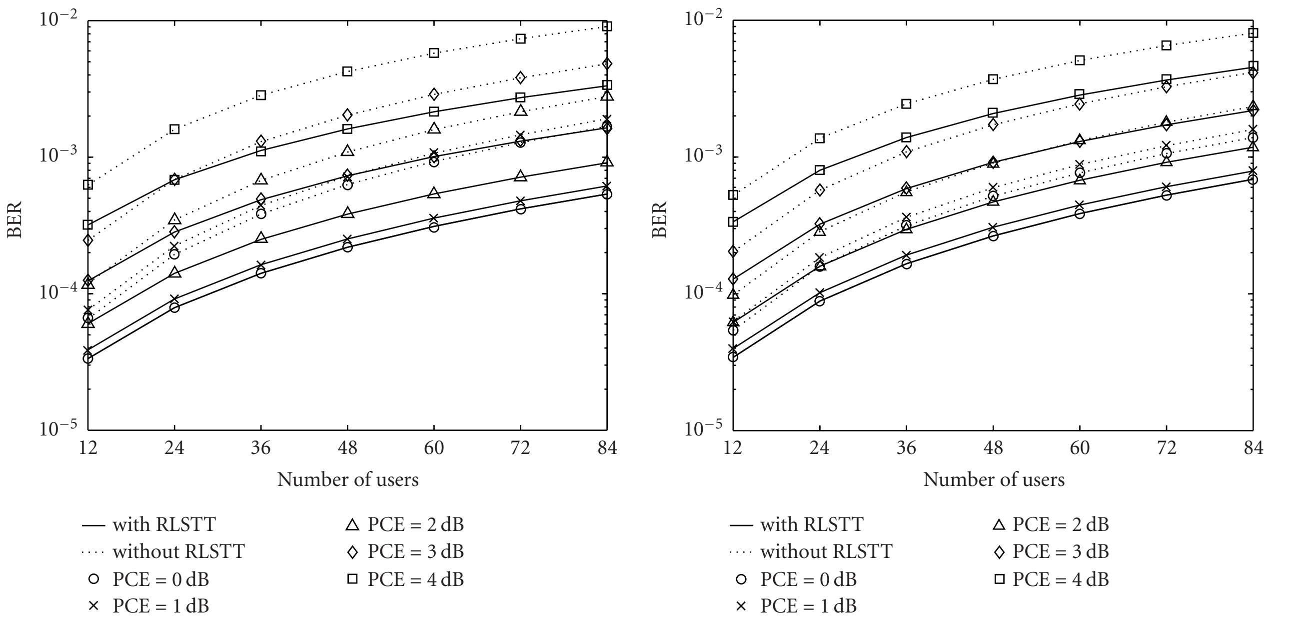
The BER curves are plotted as functions of the number of users in Figure 3 when $E_b/N_0 = 20$ dB and power control is perfect. The number of antennas is chosen among one, four, or eight. It is shown that AA with RLSTT demonstrates significant performance gain when the number of users increases, even though the performance improvement decreases when the number of antenna increases. For example, in the case of four antennas, while AA without RLSTT supports 60 users, AA with RLSTT supports more than 96 users at a BER of 10^{-3}, showing an enhancement of 50%.

Figure 4 shows the BER system performance as a function of the number of users, when $M = 4$, $E_b/N_0 = 20$ dB,



(a) (b)

Figure 3: BER versus number of users in AA with RLSTT and AA without RLSTT ($E_b/N_0 = 20$ dB, $M = 1, 4,$ and 8, $L_r = L^{(k)} = 2$, PCE = 0 dB). (a) $\delta = 1.0$, (b) $\delta = 0.2$.



(a) (b)

Figure 4: BER versus number of users in AA with RLSTT and AA without RLSTT ($E_b/N_0 = 20$ dB, $M = 4$, $L_r = L^{(k)} = 2$, PCE = $0, 1, 2, 3,$ and 4 dB). (a) $\delta = 1.0$, (b) $\delta = 0.2$.

and power control is imperfect. The curves are parameterized by different PCE values such as PCE $= 0, 1, 2, 3,$ and 4[dB], and show that RLSTT makes DS-CDMA system with AA insensitive to the PCE and thus increases the achievable overall system capacity. At BER $= 5 \times 10^{-4}$, AA with RLSTT when PCE = 2 dB can support even greater number of users, about 35% more than AA without RLSTT when power control is perfect (PCE = 0 dB), even though its capacity when PCE = 2 dB is degraded about 28% in comparison to the perfect power control.

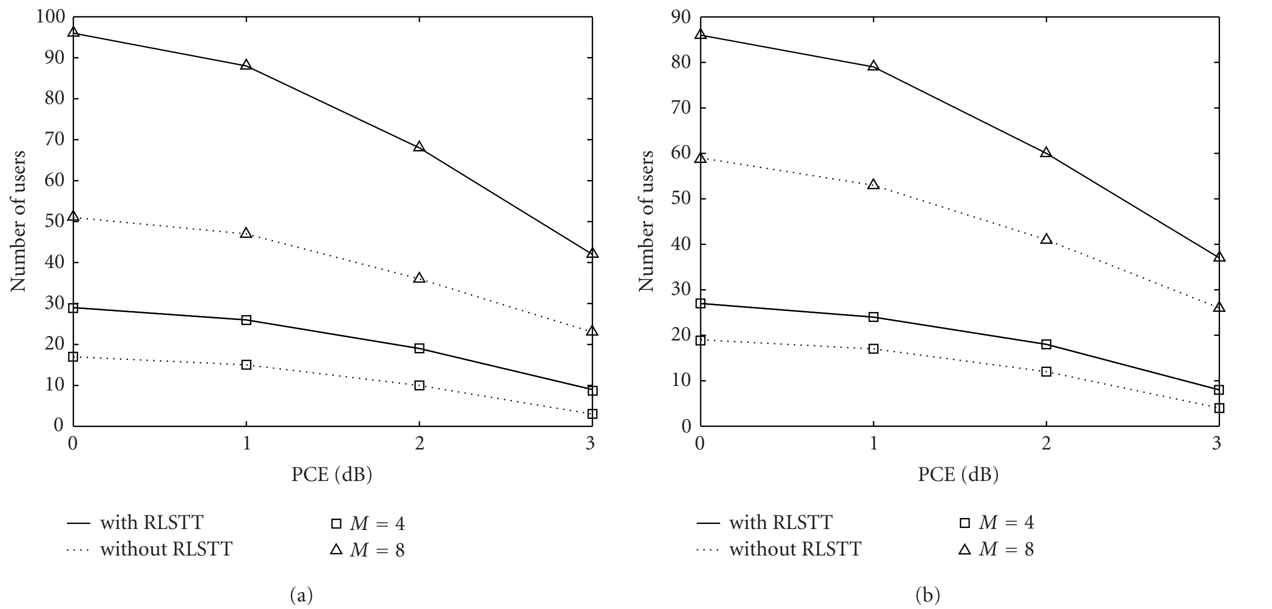


Figure 5: Number of users versus PCE in AA with RLSTT and AA without RLSTT ($E_b/N_0 = 20$ dB, $M = 4$ and 8, $L_r = L^{(k)} = 2$, BER $= 10^{-4}$) (a) $\delta = 1.0$ (b) $\delta = 0.2$.

In Figure 5, the maximum allowable number of users to achieve BER of 10^{-4} is shown as a function of PCE when the number of antenna elements is four or eight. The figure demonstrates while in eight-element AA without RLSTT PCE is required to keep less than 1 dB in order to achieve the user capacity of 50 users, AA with RLSTT may make loose the requirement to 3 dB. The figure can also be used to find the overall system capacity for a given PCE and the number of antenna elements. These results, however, do not take into account effects such as coding and interleaving. Additionally, it is apparent that RLSTT has superior performance and/or reduces the complexity of the system since AA with RLSTT with fewer numbers of antennas can obtain better performance than AA without RLSTT.

5. CONCLUSIONS

In this paper, we presented an improved AA, in which RLSTT is incorporated to effectively make better an estimation of covariance matrices at a beamformer-RAKE receiver, and investigated the user capacity and the performance analysis which reflects several important factors such as the shape of multipath intensity profile (MIP), the number of antennas, and power control error (PCE). The results show that the orthogonality provided by employing RLSTT along with AA may make the DS-CDMA system insensitive to the PCE even with fewer numbers of antennas. Additionally, RLSTT has superior performance and/or reduces the complexity of the system since AA with RLSTT with fewer numbers of antennas can obtain better performance than AA without RLSTT. The consideration of estimation technique such as diagonal loading employed in the proposed system may be an interesting issue for future study.

APPENDIX

SPATIAL CORRELATION STATISTICS

From (10), we can obtain the optimal beamformer weight presented as

$$\mathbf{W}_l^{(k)} = \xi \cdot \mathbf{R}_{uu,l}^{(k)^{-1}} \mathbf{V}\left(\theta_l^{(k)}\right); \tag{A.1}$$

since ξ does not affect the SINR, we can set $\xi = 1$. When the total number of paths is large, a large code length yields $\mathbf{R}_{uu,l}^{(k)} = \sigma_{s,l}^{(k)^2} \cdot \mathbf{I}$ [2]. However, it means that the total undesired signal vector can be modeled as a spatially white Gaussian random vector. Here, $\sigma_{s,l}^{(k)^2}$ is the total interference-plus-noise power. From (7), the total interference-plus-noise for the lth path of the kth user in the matched filter output is shown as

$$\mathbf{u}_l^{(k)} = \mathbf{I}_{l,\mathrm{si}}^{(k)} + \mathbf{I}_{l,\mathrm{mai}}^{(k)} + \mathbf{I}_{l,\mathrm{ni}}^{(k)}. \tag{A.2}$$

If we assume that the angles of arrival of the multipath components are uniformly distributed over $[0, \pi)$, the total interference vector $\mathbf{I}_{l,\mathrm{si}}^{(k)} + \mathbf{I}_{l,\mathrm{mai}}^{(k)}$ will be spatially white [2, Chapter 6]. In this case, the variance of the undesired signal vector is calculated as

$$\begin{aligned} E\left\lfloor \mathbf{u}_l^{(k)} \cdot \mathbf{u}_l^{(k)^H} \right\rfloor &= \sigma_{s,l}^{(k)^2} \cdot \mathbf{I} \\ &= \left(\sigma_{\mathrm{mai},l}^{(k)^2} + \sigma_{\mathrm{si},l}^{(k)^2} + \sigma_{\mathrm{ni},l}^{(k)^2}\right) \cdot \mathbf{I}, \end{aligned} \tag{A.3}$$

where $\sigma_{\mathrm{mai},l}^{(k)^2}$ and $\sigma_{\mathrm{si},l}^{(k)^2}$ are the noise variance of MAI and SI in one-dimension antenna system. For the RLSTT model [7], all active users are synchronous in the first branch. Therefore, we can obtain the different variance of the total interference-plus-noise for $l = 0$ and for $l \geq 1$, conditions on the value of λ_k, respectively, expressed as follows:

$$\begin{aligned}
&\sigma_{s,0}^{(k)^2}(\lambda_1, \lambda_I) \\
&\quad = E_b T \Omega_0 \left(\frac{(2N-3)\lambda_I \{q(L_r, \delta) - 1\}}{12N(N-1)} \right. \\
&\qquad \left. + \frac{\lambda_1 \{q(L_r, \delta) - 1\}}{4N} + \frac{\eta_0}{4E_b \Omega_0} \right) \quad \text{for } l = 0, \\
&\sigma_{s,l}^{(k)^2}(\lambda_1, \lambda_I) \\
&\quad = E_b T \Omega_0 \left(\frac{(N-1)\lambda_I q(L_r, \delta)}{6N^2} \right. \\
&\qquad \left. + \frac{\lambda_1 \{q(L_r, \delta) - 1\}}{4N} + \frac{\eta_0}{4E_b \Omega_0} \right) \quad \text{for } l \geq 1.
\end{aligned} \tag{A.4}$$

Using the Hermite polynomial approach, we can evaluate the average total interference-plus-noise power per AA element.

With these assumptions, the optimal beamformer weight of the kth user at the lth multipath can be shown to be $\mathbf{W}_l^{(k)} = \sigma_{s,l}^{(k)^{-2}} \cdot \mathbf{V}(\theta_l^{(k)})$. Therefore, between the array response vector of the mth user at the hth multipath and the weight vector of the kth user's lth path, the spatial correlation can be expressed as

$$C_{lh}^{(k,m)} = \frac{\mathbf{V}^H\left(\theta_l^{(k)}\right)\mathbf{V}\left(\theta_h^{(m)}\right)}{\sigma_{s,l}^{(k)^2}} = \frac{CR_{lh}^{(k,m)}}{\sigma_{s,l}^{(k)^2}}, \tag{A.5}$$

where

$$\begin{aligned}
CR_{lh}^{(k,m)} &= \sum_{i=0}^{M-1} \exp\left(j\pi \,\mathrm{si} \cos\left(\theta_l^{(k)}\right)\right) \exp\left(-j\pi \,\mathrm{si} \cos\left(\theta_h^{(m)}\right)\right), \\
s &= \frac{2d}{\lambda}.
\end{aligned} \tag{A.6}$$

The second-order characterization of the spatial correlation is calculated as

$$\zeta_{lh}^{(k,m)^2} = E\left[\left\{C_{lh}^{(k,m)}\right\}^2\right] = \frac{E\left[\left\{CR_{lh}^{(k,m)}\right\}^2\right]}{\sigma_{s,l}^{(k)^4}}, \tag{A.7}$$

where

$$\begin{aligned}
&\left\{CR_{lh}^{(k,m)}\right\}^2 = A\left(\theta_l^{(k)}, \theta_h^{(m)}\right) \\
&\quad = \sum_{i=0}^{M-1} (i+1) \exp\left(j\pi \,\mathrm{si} \cos\theta_l^{(k)}\right) \exp\left(-j\pi \,\mathrm{si} \cos\theta_h^{(m)}\right) \\
&\qquad + \sum_{i=M}^{2(M-1)} (2M-i-1) \exp\left(j\pi \,\mathrm{si} \cos\theta_l^{(k)}\right) \\
&\qquad\qquad \times \exp\left(-j\pi \,\mathrm{si} \cos\theta_h^{(m)}\right).
\end{aligned} \tag{A.8}$$

The mean angles of arrival $\theta_l^{(k)}$ and $\theta_h^{(m)}$ have uniform distribution in $[0, \pi)$ independently. So,

$$\begin{aligned}
&E\left[\left\{CR_{lh}^{(k,m)}\right\}^2\right] \\
&\quad = \int_0^\pi \int_0^\pi A\left(\theta_l^{(k)}, \theta_h^{(m)}\right) d\theta_l^{(k)} d\theta_h^{(m)} \\
&\quad = \begin{cases}
\displaystyle\sum_{i=0}^{M-1} (i+1) J_0(\pi \,\mathrm{si}) J_0(-\pi \,\mathrm{si}) \\
\displaystyle + \sum_{i=M}^{2(M-1)} (2M-i-1) J_0(\pi \,\mathrm{si}) J_0(-\pi \,\mathrm{si}), & k \neq m \text{ or } l \neq h, \\
M^2, & k = m, l = h,
\end{cases}
\end{aligned} \tag{A.9}$$

where $J_0(x)$ is the zero-order Bessel function of the first kind.

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