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A Beamspace-Time Blind RAKE Receiver for Sectored CDMA Systems ⋆

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Abstract. A beamspace-time (BT) RAKE receiver is proposed for multiple access interference (MAI) suppression and multipath diversity reception in sectored wireless CDMA communications. The scheme involves three stages. First, a set of adaptive beamformers encompassing a prescribed angular sector is constructed on an antenna array, each providing effective suppression of out-of-sector MAI and reception of in-sector signal. Second, a set of adaptive correlators is attached to each beam to combat in-sector MAI. Finally, the beamspace correlator output data are combined to capture the signal multipaths coherently.The above three-stage operation is performed in a blind mode in that no training signal is needed. The only information required is the signature, timing and a rough estimate of the angle of arrival (AOA) of the desired signal.

Keywords: CDMA, RAKE receiver, array beamforming, interference cancellation.

1. Introduction

Two major limiting factors in a CDMA system are the multiple access interference (MAI) and multipath fading phenomenon. In the worst case, MAI causes the near-far effect and multipath induces deep fading and/or inter-symbol interference (ISI). To combat the MAI problem, advanced detectors have been developed which provide full or partial immunity to the near-far effect [1]. On the other hand, the RAKE receiver [2] is proposed for coherent combining of multipath signals. The RAKE receiver is a temporal matched filter that exploits the temporal signature of the multipath channel to enhance the SNR. In general, the path signals arrive at the receiver with different angles of arrival (AOA). To exploit this spatial signature, an antenna array is employed. A receiver that exploits both the temporal and spatial signatures to constructively combine the multipath components is referred to as the space-time (ST) RAKE receiver [3, 4]. In ST RAKE receivers, the combining of correlator outputs at different antennas can be done either jointly or separately in the spatial-temporal domain [4]. On the other hand, the combining weight vectors can be determined via the maximum ratio (MR) or minimum-mean-square error (MMSE) criterion, with the latter offering better MAI suppression [3]. In either case, a training signal is necessary for channel estimation.

The linear adaptive CDMA receiver is an improved version of the conventional RAKE receiver [5]. The improvement lies in its enhanced MAI suppression via either post- or pre-

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despread processing. The post-despread receiver works with the outputs of a bank of matched filters, one for each user (including signal and MAI). The pre-despread receiver works directly with the chip-sampled data, and is usually operated in a decentralized manner without the knowledge of MAI's information [5]. It can be regarded as an adaptive matched filter that performs temporal signature matching and MAI suppression simultaneously. A simple and popular pre-despread receiver is the minimum output energy (MOE) receiver proposed in [6]. The MOE receiver is constructed by minimizing the output energy subject to a unit response constraint for the desired signal. It is similar to the generalized sidelobe canceller (GSC) in array processing [7], and can offer the performance of the optimal minimum-mean-square-error (MMSE) receiver without the aid of a training sequence [6], provided that an accurate estimate of the signal's signature is available. With a certain mismatch in the signal's signature, however, the MOE receiver exhibits severe performance degradation due to the effect of signal cancellation [5]. To remedy this, auxiliary processing is incorporated to alleviate the sensitivity problem [6].

In this paper, an adaptive ST RAKE receiver is proposed for sectored CDMA systems [8]. In a sectored system, the entire field-of-view of the receiver is divided into several angular sectors, with each sector responsible for a distinctive set of users. With an antenna array incorporated, sectorization can be done adaptively to meet the following two requirements. First, multiple beams are formed to collect desired signal multipath components in the designated angular sector. Second, strong MAI from outside the sector are suppressed in the sidelobe of these beams. These can be achieved by performing adaptive nulling on a set of beams steered to different look directions. To avoid signal cancellation incurred with coherent multipaths or mismatch of steering vectors in adaptive nulling, a modified GSC is employed to construct a set of linearly constrained minimum variance (LCMV) beamformers [7]. The output of these beamformers are processed by a bank of adaptive correlators, which can be regarded as a set of LCMV combiners in the temporal domain. A modified GSC is again employed to collect the multipath components and suppress the in-sector MAI. The beamformers and correlators together constitute a pre-despread beamspace-time (BT) processor, which performs the function of a RAKE receiver. With MAI successfully suppressed, a simple maximum ratio combining (MRC) criterion can be used to determine the weights of the BT RAKE receiver. Compared to the conventional ST receiver, beamspace sectored processing can potentially increase the system capacity by suppressing out-of-sector MAI, and also lower the computational complexity by reducing the spatial dimension. The proposed BT receiver is blind in that the construction of adaptive beamformers, correlators, and MRC is done without the aid of a training signal. The only information required is the spreading sequence, timing and a rough estimate of the AOA of the desired signal for sector selection.

2. Data Model

Consider a DS-CDMA system in which there are K active users. The kth user's contribution to the received signal is given by

$$d_k(t) = \sqrt{P_k} \sum_i b_k(i) s_k(t - iT) , \tag{1}$$

where P_k is the transmit power, $b_k(i)$ is the ith information symbol taking on ± 1 with equal probability, T is the symbol duration, and $s_k(t)$ is the spreading waveform given by

$$s_k(t) = \sum_{m=0}^{M-1} c_k[m] p(t - mT_c) , \tag{2}$$

where $c_k[m]$ is the spreading (signature) sequence of the kth user, M is the spreading factor, $p(t)$ is the chip waveform, and T_c is the chip duration. Suppose that a D-element antenna array is employed at the basestation. In this case, the baseband data observed at the array output can be expressed in the vector form:

$$\mathbf{x}(t) = \sum_{k=1}^{K} \sqrt{P_k} \sum_{j=1}^{J} \alpha_{k,j} \mathbf{a}(\theta_{k,j}) d_k(t - \tau_{k,j}) + \mathbf{n}(t) , \tag{3}$$

where J is the path number (assumed the same for all users), and $\theta_{k,j}$, $\tau_{k,j}$ and $\alpha_{k,j}$ are the AOA, delay and complex gain, respectively, of the jth path of the kth user. $\mathbf{a}(\theta)$ is the steering vector which represents the gain/phase response of the array to a signal from θ. It can be regarded as the spatial signature of the signal as viewed by the antenna array. Finally, $\mathbf{n}(t)$ is the noise vector with its entries being independent, identically distributed complex Gaussian random variables with variance σ_n^2.

In order to facilitate digital processing, and fully exploit the temporal signature of the multipath channel, $\mathbf{x}(t)$ is chip matched filtered and then chip rate sampled at $t = (i-1)T + mT_c + T_c/2$ over the ith symbol duration, i.e., $m = 0, 1, \ldots, M+L-2$, where L is the number of RAKE fingers. Assuming user 1 to be the desired signal, the chip-sampled pre-despread ST data can be written as a $D \times (M + L - 1)$ matrix:

$$\begin{aligned} \mathbf{X}(i) &= [\mathbf{x}(0), \mathbf{x}(1), \cdots, \mathbf{x}(M+L-2)] \\ &= \sqrt{P_1} \sum_{j=1}^{J} \alpha_{1,j} \mathbf{a}(\theta_{1,j}) \mathbf{c}_{\tau_{1,j}}^T b_1(i) + \mathbf{I}(i) + \mathbf{N}(i) \\ &= \mathbf{H}_1 b_1(i) + \mathbf{I}(i) + \mathbf{N}(i) , \end{aligned} \tag{4}$$

where $\mathbf{c}_{\tau_{1,j}}$ is the augmented signature vector of the jth path of user 1. Depending on the delay $\tau_{1,j}$, $\mathbf{c}_{\tau_{1,j}}$ is given by one of the columns of the $(M + L - 1) \times L$ augmented signature matrix:

$$\mathbf{S}_1 = \begin{bmatrix} c_1[0] & 0 & \cdots & 0 \\ \vdots & c_1[0] & \ddots & \vdots \\ c_1[M-1] & \vdots & \ddots & 0 \\ 0 & c_1[M-1] & \ddots & c_1[0] \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & c_1[M-1] \end{bmatrix} \tag{5}$$

$\mathbf{I}(i)$ is the ST interference matrix consisting of ISI and MAI:

$$\begin{aligned}\mathbf{I}(i) \; = \; & \sum_{j=1}^{J} \alpha_{1,j} \mathbf{a}(\theta_{1,j}) \mathbf{c}_{\tau_{1,j}}^{(-)T} b_1(i-1) \\ & + \sum_{j=1}^{J} \alpha_{1,j} \mathbf{a}(\theta_{1,j}) \mathbf{c}_{\tau_{1,j}}^{(+)T} b_1(i+1) \\ & + \sum_{k=2}^{K} \sum_{j=1}^{J} \alpha_{k,j} \mathbf{a}(\theta_{k,j}) \mathbf{c}_{\tau_{k,j}}^{T} b_k(i) \\ & + \sum_{k=2}^{K} \sum_{j=1}^{J} \alpha_{k,j} \mathbf{a}(\theta_{k,j}) \mathbf{c}_{\tau_{k,j}}^{(-)T} b_k(i-1) \\ & + \sum_{k=2}^{K} \sum_{j=1}^{J} \alpha_{k,j} \mathbf{a}(\theta_{k,j}) \mathbf{c}_{\tau_{k,j}}^{(+)T} b_k(i+1)\end{aligned} \tag{6}$$

where $\mathbf{c}_{\tau_{1,j}}^{(-)}$ and $\mathbf{c}_{\tau_{1,j}}^{(+)}$ are the augmented signature vector associated with the $(i-1)$th and $(i+1)$th symbols of user 1, respectively, corresponding to the ISI. $\mathbf{c}_{\tau_{k,j}}$, $\mathbf{c}_{\tau_{k,j}}^{(-)}$ and $\mathbf{c}_{\tau_{k,j}}^{(+)}$ are defined similarly for the MAI. Finally, $\mathbf{N}(i)$ is the ST white noise matrix. From (4), $\mathbf{H}_1$ is the ST composite signature matrix of user 1:

$$\mathbf{H}_1 = \sqrt{P_1} \sum_{j=1}^{J} \alpha_{1,j} \mathbf{a}(\theta_{1,j}) \mathbf{c}_{\tau_{1,j}}^{T} \, . \tag{7}$$

The structure of the ST data in (4) suggests that a receiver for user 1 should be designed to identify and remove $\mathbf{H}_1$ to retrieve $b_1(i)$ from $\mathbf{I}(i)$ and $\mathbf{N}(i)$. For example, the conventional ST RAKE receiver works with the $D \times L$ post-despread ST data matrix:

$$\begin{aligned}\tilde{\mathbf{X}}(i) \; &= \; \mathbf{X}(i)\mathbf{S}_1^* \\ &= \; \mathbf{H}_1 \mathbf{S}_1^* b_1(i) + \mathbf{I}(i)\mathbf{S}_1^* + \mathbf{N}(i)\mathbf{S}_1^* \, ,\end{aligned} \tag{8}$$

where $\mathbf{S}_1$ post-multiplies $\mathbf{X}$ to extract the multipath signals with delays in $[0, (L-1)T_c]$. Let $\tilde{\mathbf{x}}(i)$ and $\tilde{\mathbf{h}}_1$ be the $DL \times 1$ vectors obtained by concatenating the columns of $\tilde{\mathbf{X}}(i)$ and $\mathbf{H}_1\mathbf{S}_1^*$, respectively. The ST RAKE receiver performs a linear combination on $\tilde{\mathbf{x}}(i)$ to extract $b_1(i)$:

$$\hat{b}_1(i) = dec\{\mathbf{q}^H \tilde{\mathbf{x}}(i)\} \, , \tag{9}$$

where $\mathbf{q}$ is the combining weight vector, dec denotes the decision operator, and H denotes the conjugate transpose. The weight vector can be chosen in accordance with the coherent or MMSE criterion [3]. For coherent combining

$$\mathbf{q} = \tilde{\mathbf{h}}_1 \tag{10}$$

and for MMSE combining,

$$\mathbf{q} = \tilde{\mathbf{R}}^{-1} \tilde{\mathbf{h}}_1 \, , \tag{11}$$

where

$$\tilde{\mathbf{R}} = E\{\tilde{\mathbf{x}}(i)\tilde{\mathbf{x}}^H(i)\} \tag{12}$$

is the post-despread ST data correlation matrix. The main difference between the two receivers lies in the MAI suppression capability. The coherent RAKE receiver is an ST matched filter that matches itself to the ST channel $\tilde{\mathbf{h}}_1$ of the signal to achieve the maximum SNR, but ignores the presence of MAI. The MMSE RAKE receiver is an ST Wiener filter that achieves a compromise between signal reception and MAI suppression. For both receivers, the channel vector $\tilde{\mathbf{h}}_1$ is unknown, and must be estimated beforehand. Typically, channel estimation is done with the aid of a training sequence $b_1(i)$ as follows:

$$\tilde{\mathbf{h}}_1 = E\{\tilde{\mathbf{x}}(i)b_1(i)\}\,. \tag{13}$$

Alternatively, blind channel estimation methods that does not require a training sequence can be employed to improve the system's efficiency [3, 9].

3. Development of BT RAKE Receiver

The proposed BT RAKE receiver is developed via the following three-stage procedure. First, a set of adaptive diversity beamformers is constructed for each finger to collect in-sector signals and suppress out-of-sector MAI. Second, a set of adaptive correlators is attached to each beamformer to perform despreading and in-sector MAI suppression. Finally, the correlator outputs are combined coherently for an optimal detection of signal symbols. For the ease of notation, the subscript 1 will be omitted in the expressions of data associated with user 1.

3.1. Construction of Adaptive Beamformers

Suppose that the field-of-view of the receiver is divided into several sectors, and that the AOA's of the signal multipaths are roughly known such that a sector can be chosen for signal reception. With an antenna array employed, the sectorization can be achieved by forming a set of N diversity beams for each of the L fingers, with the beam patterns encompassing the designated sector. Specifically, the beamformers for the lth finger act on the post-despread data vector given by

$$\begin{aligned} \tilde{\mathbf{x}}_l(i) &= \mathbf{X}(i)\mathbf{S}_1^*(:,l) \\ &= \mathbf{H}_1\mathbf{S}_1^*(:,l)b_1(i) + \mathbf{I}(i)\mathbf{S}_1^*(:,l) + \mathbf{N}(i)\mathbf{S}_1^*(:,l) \end{aligned} \tag{14}$$

with $\mathbf{S}_1(:,l)$ being the lth column of the augmented signature matrix in (5). After despreading, $\tilde{\mathbf{x}}_l(i)$ contains essentially the strong MAI and multipath signals of the delay corresponding to $\mathbf{S}_1(:,l)$. To ensure an effective suppression of strong out-of-sector MAI, adaptive nulling is performed for each of the diversity beamformers. A popular nulling scheme is based on the LCMV criterion [7], which says that the beamforming weight vector should be chosen in accordance with:

$$\begin{aligned} &\min_{\mathbf{w}_{l,n}} \quad \mathbf{w}_{l,n}^H\tilde{\mathbf{R}}_l\mathbf{w}_{l,n} \\ &\text{subject to:} \quad \mathbf{w}_{l,n}^H\mathbf{a}(\theta_n) = 1 \end{aligned} \tag{15}$$

for $l = 1, \ldots, L$ and $n = 1, \ldots, N$, where $\mathbf{w}_{l,n}$ is the weight vector of the nth beamformer at the lth finger,

$$\tilde{\mathbf{R}}_l = E\{\tilde{\mathbf{x}}_l(i)\tilde{\mathbf{x}}_l^H(i)\} \tag{16}$$

is the post-despread space-only data correlation matrix at the lth finger. Finally, θ_n is the look angle of the nth beam. A major problem of the LCMV beamformer is the phenomenon of desired signal cancellation [10] due to the mismatch between the look angle steering vectors $\mathbf{a}(\theta_n)$'s and the actual steering vectors associated with the multipath signals. By signal cancellation, it is meant that with a certain mismatch, the weight vector in (15) will tend to treat as interference the signal not exactly matched to $\mathbf{a}(\theta_n)$ and attempt to cancel it in order to minimize the output power. The sensitivity of the LCMV beamformer to steering vector mismatch increases as the array size D or SNR increases, in which case the beamformer will put more efforts to cancel the signal. An effective remedy suggested herein is to use a modified GSC to block the signal before beamforming [7]. The GSC is essentially an indirect but simpler implementation of the LCMV beamformer. In GSC, the weight vector is decomposed as $\mathbf{w}_{l,n} = \mathbf{a}(\theta_n) - \mathbf{B}\mathbf{v}_{l,n}$ into two orthogonal components which lie in the range and null space of the constraint, respectively. The matrix $\mathbf{B}$ is a pre-designed "signal blocking" matrix which removes the signal (and MAI) in the sector. The goal is then to choose the adaptive weight vector $\mathbf{v}_{l,n}$ to cancel the out-of-sector MAI. Following the standard procedure of GSC, the adaptive weight vectors are determined by solving the MMSE problem:

$$\min_{\mathbf{v}_{l,n}} \quad E\{|\mathbf{a}^H(\theta_n)\tilde{\mathbf{x}}_l(i) - \mathbf{v}_{l,n}^H \mathbf{B}^H \tilde{\mathbf{x}}_l(i)|^2\} \tag{17}$$

or equivalently

$$\min_{\mathbf{v}_{l,n}} \; [\mathbf{a}(\theta_n) - \mathbf{B}\mathbf{v}_{l,n}]^H \tilde{\mathbf{R}}_l [\mathbf{a}(\theta_n) - \mathbf{B}\mathbf{v}_{l,n}] \,. \tag{18}$$

Taking the gradient of (18) with respect to $\mathbf{v}_{l,n}$ and setting to zero, we have

$$\mathbf{B}^H \tilde{\mathbf{R}}_l [\mathbf{a}(\theta_n) - \mathbf{B}\mathbf{v}_{l,n}] = \mathbf{0} \tag{19}$$

which gives

$$\mathbf{v}_{l,n} = (\mathbf{B}^H \tilde{\mathbf{R}}_l \mathbf{B})^{-1} \mathbf{B}^H \tilde{\mathbf{R}}_l \mathbf{a}(\theta_n) \,. \tag{20}$$

Substituting this in $\mathbf{w}_{l,n} = \mathbf{a}(\theta_n) - \mathbf{B}\mathbf{v}_{l,n}$ and putting in matrix form, we get

$$\begin{aligned} \mathbf{W}_l &= [\mathbf{w}_{l,1}, \ldots, \mathbf{w}_{l,N}] \\ &= [\mathbf{I} - \mathbf{B}(\mathbf{B}^H \tilde{\mathbf{R}}_l \mathbf{B})^{-1} \mathbf{B}^H \tilde{\mathbf{R}}_l][\mathbf{a}(\theta_1), \, \ldots, \, \mathbf{a}(\theta_N)] \end{aligned} \tag{21}$$

for $l = 1, \, \ldots, \, L$.

The choosing of the blocking matrix $\mathbf{B}$ depends on the sector size and required degree of freedom of the adaptive weight vector $\mathbf{v}_{l,n}$. Let $\mathbf{v}_{l,n}$ be a $G \times 1$ vector. Then $\mathbf{B}$ can be chosen to be a full rank $D \times G$ matrix with columns orthogonal to a set of steering vectors $\{\mathbf{a}(\theta_1^t), \mathbf{a}(\theta_2^t), \ldots, \mathbf{a}(\theta_{D-G}^t)\}$ well representing the angular sector. As an alternative, the set of steering vectors can be replaced by the eigenvectors associated with the $D - G$ largest eigenvalues of the matrix:

$$\mathbf{A}_\Theta = \int_\Theta \mathbf{a}(\theta)\mathbf{a}^H(\theta) d\theta \tag{22}$$

leading to the eigenvector constrained method [11]. The advantage of using eigenvector constraints lies in its fuzzy mode of operation, which offers robustness to the variation in multipath scenario of the signal. The choosing of G is a trade-off between the blocking effect and adaptive nulling performance. In general, a small G gives better "mainlobe performance" (reception of desired signal), and a large G gives better "sidelobe performance" (suppression of out-of-sector MAI). A practical criterion is that the ratio $(D-G)/D$ is approximately equal to the ratio of the sector size to entire field-of-view of the antenna array. That is, the degree of freedom $D-G$ for blocking is proportional to the relative size of the sector. However, numerical results show that a smaller G is required to warrant a "clean" blocking effect for better handling the signal cancellation problem. With the assumption that a single sector is responsible for about one third the field-of-view of the antenna array, a suitable choice for G would be $D/2$.

3.2. Construction of Adaptive Correlators

The beamforming matrices $\mathbf{W}_l$'s are applied at the L fingers to convert the $D \times (M+L-1)$ ST chip-sampled data matrix $\mathbf{X}(i)$ into a set of L BT data matrices of dimension $N \times (M+L-1)$:

$$\begin{aligned} \mathbf{Y}_l(i) &= \mathbf{W}_l^H \mathbf{X}(i) \\ &= \begin{bmatrix} \mathbf{y}_{l,1}^T(i) \\ \mathbf{y}_{l,2}^T(i) \\ \vdots \\ \mathbf{y}_{l,N}^T(i) \end{bmatrix} \end{aligned} \tag{23}$$

for $l = 1, \ldots, L$, where $\mathbf{y}_{l,n}(i)$ is the chip-sampled data vector obtained at the nth beam of the lth finger, given by

$$\begin{aligned} \mathbf{y}_{l,n}(i) &= \mathbf{X}^T(i)\mathbf{w}_{l,n}^* \\ &= \mathbf{H}_1^T \mathbf{w}_{l,n}^* b_1(i) + \mathbf{I}^T(i)\mathbf{w}_{l,n}^* + \mathbf{N}^T(i)\mathbf{w}_{l,n}^* \,. \end{aligned} \tag{24}$$

The next step is then to perform despreading on $\mathbf{y}_{l,n}(i)$ to restore the processing gain. In order to better handle the MAI, this is done with an adaptive correlator as follows

$$\begin{aligned} z_{l,n}(i) &= \mathbf{g}_{l,n}^H \mathbf{y}_{l,n}(i) \\ &= \mathbf{w}_{l,n}^H \mathbf{X}(i) \mathbf{g}_{l,n}^* \\ &= \mathbf{w}_{l,n}^H \mathbf{H}_1 \mathbf{g}_{l,n}^* b_1(i) + \mathbf{w}_{l,n}^H \mathbf{I}(i) \mathbf{g}_{l,n}^* + \mathbf{w}_{l,n}^H \mathbf{N}(i) \mathbf{g}_{l,n}^* \,, \end{aligned} \tag{25}$$

where $\mathbf{g}_{l,n}$ is the despreading weight vector for the nth beam of the lth finger. As the temporal analogy of the beamforming weight vector $\mathbf{w}_{l,n}$, $\mathbf{g}_{l,n}$ can be determined using the GSC scheme described above with the steering vector $\mathbf{a}(\theta)$ replaced by the augmented signature vector $\mathbf{S}_1(:, l)$. Following the development in (15)-(21), we have $\mathbf{g}_{l,n} = \mathbf{S}_1(:, l) - \mathbf{C}\mathbf{u}_{l,n}$, where $\mathbf{C}$ is the signal blocking matrix which removes user 1's signal. Note that instead of blocking signals with a specific delay, $\mathbf{C}$ must block signals within the entire delay spread in order to avoid signal cancellation due to coherent multipaths. The goal is then to choose the adaptive weight vectors $\mathbf{u}_{l,n}$ to cancel the in-sector MAI (and possibly out-of-sector MAI not canceled by the beamformers). Similar to GSC beamforming, the adaptive weight vectors are determined by

$$\min_{\mathbf{u}_{l,n}} \quad E\{|\mathbf{S}_1^H(:, l)\mathbf{y}_{l,n}(i) - \mathbf{u}_{l,n}^H \mathbf{C}^H \mathbf{y}_{l,n}(i)|^2\} \tag{26}$$

or equivalently

$$\min_{\mathbf{u}_{l,n}} \; [\mathbf{S}_1(:, l) - \mathbf{C}\mathbf{u}_{l,n}]^H \bar{\mathbf{R}}_{l,n} [\mathbf{S}_1(:, l) - \mathbf{C}\mathbf{u}_{l,n}] \,, \tag{27}$$

where

$$\bar{\mathbf{R}}_{l,n} = E\{\mathbf{y}_{l,n}(i)\mathbf{y}_{l,n}^H(i)\} \tag{28}$$

is the pre-despread time-only data correlation matrix at the nth beam of the lth finger. Solving for $\mathbf{u}_{l,n}$ in the same way as for $\mathbf{v}_{l,n}$, and substituting in $\mathbf{g}_{l,n} = \mathbf{S}_1(:, l) - \mathbf{C}\mathbf{u}_{l,n}$, we get

$$\mathbf{g}_{l,n} = [\mathbf{I} - \mathbf{C}(\mathbf{C}^H\bar{\mathbf{R}}_{l,n}\mathbf{C})^{-1}\mathbf{C}^H\bar{\mathbf{R}}_{l,n}]\mathbf{S}_1(:, l) \tag{29}$$

for $l = 1, \ldots, L$ and $n = 1, \ldots, N$.

The choosing of the temporal blocking matrix $\mathbf{C}$ is similar to the choosing of $\mathbf{B}$. That is, $\mathbf{C}$ can be chosen to be a full rank $(M + L - 1) \times (M - 1)$ matrix with columns orthogonal to the set of augmented signature vectors $\{\mathbf{S}_1(:, l), \mathbf{S}_1(:, 2), \ldots, \mathbf{S}_1(:, L)\}$ well representing the multipath delay spread. For a more reliable operation, extra signature vectors can be included to extend the blocking interval to a larger delay spread. This will help to avoid possible signal cancellation due to undetected multipath arrivals. For example, two extra signature vectors

$$\begin{aligned} \mathbf{S}_1^{(-)}(:, 1) &= [c_1[1], c_1[2], \ldots, c_1[M-1], 0, \ldots, 0]^T \\ \mathbf{S}_1^{(+)}(:, L) &= [0, \ldots, 0, c_1[0], c_1[1], \ldots, c_1[M-2]]^T \end{aligned} \tag{30}$$

can be added to the original set to extend by one chip at both ends of the blocking interval.

3.3. Maximum Ratio Combining

Suppose that, after adaptive beamforming and despreading, the MAI's are successfully suppressed, the correlator outputs $z_{l,n}(i)$'s contain essentially the desired signal and colored noise only. In this case, the MRC criterion can be applied to combine these outputs coherently to extract the signal. Let

$$\begin{aligned} \mathbf{z}(i) \;=\; & [z_{1,1}(i), \ldots, z_{1,N}(i), z_{2,1}(i), \ldots, z_{2,N}(i), \ldots, \\ & z_{L,1}(i), \ldots, z_{L,N}(i)]^T \approx \mathbf{h}_z b_1(i) + \mathbf{n}_z(i) \end{aligned} \tag{31}$$

be the beamspace correlator output data vector, with $\mathbf{h}_z$ and $\mathbf{n}_z(i)$ being the corresponding composite signature and noise vectors, respectively, given by

$$\begin{aligned} \mathbf{h}_z \;=\; & \big[\mathbf{w}_{1,1}^H\mathbf{H}_1\mathbf{g}_{1,1}^*, \ldots, \mathbf{w}_{1,N}^H\mathbf{H}_1\mathbf{g}_{1,N}^*, \mathbf{w}_{2,1}^H\mathbf{H}_1\mathbf{g}_{2,1}^*, \ldots, \\ & \mathbf{w}_{2,N}^H\mathbf{H}_1\mathbf{g}_{2,N}^*, \ldots, \mathbf{w}_{L,1}^H\mathbf{H}_1\mathbf{g}_{L,1}^*, \ldots, \mathbf{w}_{L,N}^H\mathbf{H}_1\mathbf{g}_{L,N}^*\big]^T . \end{aligned} \tag{32}$$

$$\begin{aligned} \mathbf{n}_z(i) \;=\; & \big[\mathbf{w}_{1,1}^H\mathbf{N}(i)\mathbf{g}_{1,1}^*, \ldots, \mathbf{w}_{1,N}^H\mathbf{N}(i)\mathbf{g}_{1,N}^*, \mathbf{w}_{2,1}^H\mathbf{N}(i)\mathbf{g}_{2,1}^*, \ldots, \\ & \mathbf{w}_{2,N}^H\mathbf{N}(i)\mathbf{g}_{2,N}^*, \ldots, \mathbf{w}_{L,1}^H\mathbf{N}(i)\mathbf{g}_{L,1}^*, \ldots, \mathbf{w}_{L,N}^H\mathbf{N}(i)\mathbf{g}_{L,N}^*\big]^T . \end{aligned} \tag{33}$$

The final operation of the receiver is then a linear combination on $\mathbf{z}(i)$ using an $NL \times 1$ weight vector $\mathbf{f}$:

$$z_o(i) = \mathbf{f}^H\mathbf{z}(i) \tag{34}$$

The weight vector $\mathbf{f}$ that leads to MRC can be determined by solving the following problem [12]:

$$\max_{\mathbf{f}} \quad \frac{E\{|\mathbf{f}^H\mathbf{z}(i)|^2\}}{E\{|\mathbf{f}^H\mathbf{n}_z(i)|^2\}} \equiv \frac{\mathbf{f}^H\mathbf{R}_z\mathbf{f}}{\mathbf{f}^H\mathbf{Q}_z\mathbf{f}}, \tag{35}$$

where $\mathbf{R}_z = E\{\mathbf{z}(i)\mathbf{z}^H(i)\}$ and $\mathbf{Q}_z = E\{\mathbf{n}_z(i)\mathbf{n}_z^H(i)\}$ are the beamspace correlator output data and noise correlation matrices, respectively. Note that $\mathbf{Q}_z$ can be determined with the knowledge of $\mathbf{w}_{l,n}$'s, $\mathbf{g}_{l,n}$'s, and the whiteness of $\mathbf{N}(i)$. The solution to (35) is well known to be the principle generalized eigenvector of the matrix pair $\{\mathbf{R}_z, \mathbf{Q}_z\}$. As a final remark, we point out that, as opposed to the conventional ST RAKE receiver in which a single beam is responsible for a finger, the proposed BT RAKE receiver requires N beams for a finger. This is because that the BT RAKE receiver does not have the exact AOA or channel information about the signal paths. So the best strategy would be to "collect" the in-sector multipath signals using a set of diversity beams encompassing the entire sector. In fact, using multiple beams for a single finger is the price in complexity paid for not using a training sequence.

3.4. Algorithm Summary

In practice, the data correlation matrices are usually estimated by the sample average versions:

$$\tilde{\mathbf{R}}_l \approx \frac{1}{N_s}\sum_{i=1}^{N_s}\tilde{\mathbf{x}}_l(i)\tilde{\mathbf{x}}_l^H(i) \tag{36}$$

$$\bar{\mathbf{R}}_{l,n} \approx \frac{1}{N_s}\sum_{i=1}^{N_s}\mathbf{y}_{l,n}(i)\mathbf{y}_{l,n}^H(i) \tag{37}$$

$$\mathbf{R}_z \approx \frac{1}{N_s}\sum_{i=1}^{N_s}\mathbf{z}(i)\mathbf{z}^H(i)\,, \tag{38}$$

where $\tilde{\mathbf{x}}_l(i)$, $\mathbf{y}_{l,n}(i)$ and $\mathbf{z}(i)$ are given by (14), (24) and (31), respectively, and N_s is the number of symbols used during the processing period. With these estimates, the algorithm of the proposed BT RAKE receiver is summarized as follows:

1. Determine working sector Θ, look angles θ_n's and GSC blocking matrices $\mathbf{B}$ and $\mathbf{C}$.
2. Compute in parallel $\mathbf{W}_l$, $l = 1, \ldots, L$, according to (21), with $\tilde{\mathbf{R}}_l$ estimated by (36).
3. Compute in parallel $\mathbf{g}_{l,n}$, $l = 1, \ldots, L$, $n = 1, \ldots, N$, according to (29), with $\bar{\mathbf{R}}_{l,n}$ estimated by (37).
4. Obtain $\mathbf{Q}_z$ and compute $\mathbf{f}$ according to (35), with $\mathbf{R}_z$ estimated by (38).

The corresponding schematic diagram is depicted in Figure 1.

4. Performance and Implementation Issues

4.1. Numerical Stability

The computation of beamforming weight vectors $\mathbf{w}_{l,n}$'s in (21) and despreading weight vectors $\mathbf{g}_{l,n}$'s in (29) involves the inversion of matrices $\mathbf{B}^H\tilde{\mathbf{R}}_l\mathbf{B}$ and $\mathbf{C}^H\bar{\mathbf{R}}_{l,n}\mathbf{C}$, respectively. Numerical instability may arise when there are few strong MAI present, leading to ill-conditioned correlation matrices. To remedy this, pseudo noise terms $\eta_v\mathbf{I}$ and $\eta_u\mathbf{I}$ are added to $\tilde{\mathbf{R}}_l$ and $\bar{\mathbf{R}}_{l,n}$, respectively, to alleviate the sensitivity problem [13]. The pseudo noise can help to improve signal reception in Θ [13, 14], and should be chosen large enough to handle the strong MAI, but not too large to distort the original scenario. A suitable choice which proves to be robust

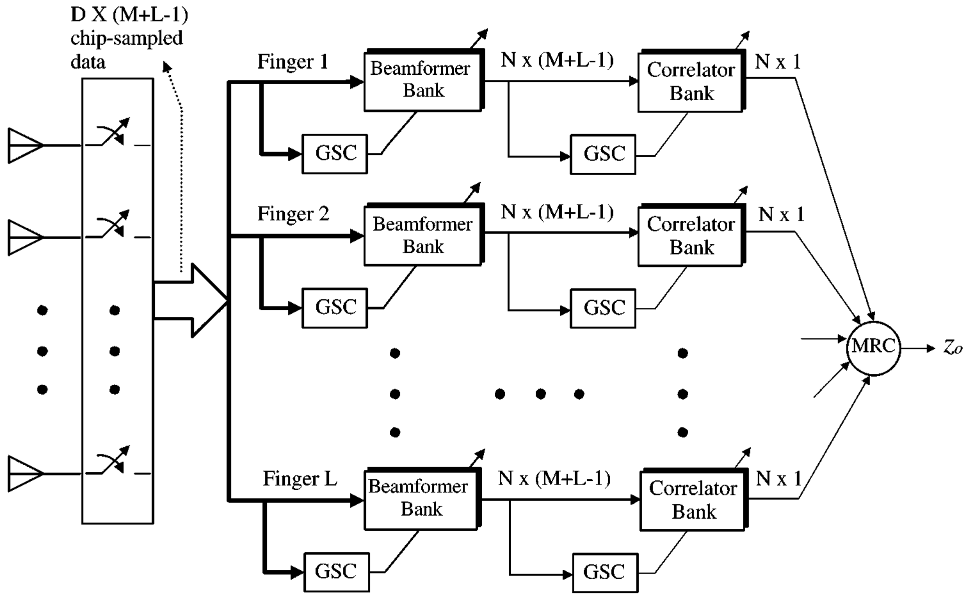


Figure 1. Overall schematic diagram of proposed BT RAKE receiver.

against scenario variations is such that η_v and η_u are approximately equal to the signal power in $\tilde{\mathbf{R}}_l$ and $\bar{\mathbf{R}}_{l,n}$, respectively [14]. This means that the adaptive beamformer or correlator will only put efforts to eliminate MAI significantly stronger than the desired signal.

4.2. Sector Location

To determine the working sector Θ, some kind of location technique is required to obtain a coarse estimate of the signal AOA. For the proposed receiver, a suitable choice is the multi-beam (MB) technique described in [15]. The MB technique works with a bank of beams pointed at different directions, and determines the signal AOA by comparing the signal power levels observed at these beamformer outputs. The beamformer with the maximum output power is likely to be the one pointed at the signal, and its look direction is taken to be the estimate of the signal AOA. Finally, the working sector is chosen to be the one that contains the estimated AOA. To apply the MB technique in the proposed receiver, a set of beams is formed simultaneously whose patterns encompass the entire field-of-view of the antenna array. Power comparison is then performed on the post-despread data observed at these beamformer outputs to determine the signal AOA and sector location. In a nonstationary environment in which the signal source moves with time, it is necessary to keep track of the signal to update the working sector via some prescribed hand-off procedure.

4.3. Recursive Computation of Weight Vectors

For a more efficient and practical implementation, the GSC can be realized in a time-recursive fashion using stochastic gradient algorithms such as LMS [16]. This leads to recursive formulation of the solutions to (17) and (26), respectively:

$$\mathbf{v}_{l,n}(i+1) = \mathbf{v}_{l,n}(i) + \mu_v \left[\mathbf{a}^H(\theta_n)\tilde{\mathbf{x}}_l(i) - \mathbf{v}_{l,n}^H \mathbf{B}^H \tilde{\mathbf{x}}_l(i)\right]^* \mathbf{B}^H \tilde{\mathbf{x}}_l(i) \tag{39}$$

$$\mathbf{u}_{l,n}(i+1) = \mathbf{u}_{l,n}(i) + \mu_u \left[\mathbf{S}_1^H(:,l)\mathbf{y}_{l,n}(i) - \mathbf{u}_{l,n}^H \mathbf{C}^H \mathbf{y}_{l,n}(i)\right]^* \mathbf{C}^H \mathbf{y}_{l,n}(i) \tag{40}$$

for $i = 1, 2, \ldots$, where μ_v and μ_u are the stepsizes of adaptation. It was shown that the convergence of blind adaptive algorithms is slow and noisy compared to the non-blind training signal based algorithm. This is more significant when the dimension of weight vectors and/or SNR is large. In view of this drawback, it is suggested that blind algorithms be employed in the initialization stage, and be switched to the decision-directed mode once the SINR has improved to offer a reliable decision reference [5]. In the decision-directed mode, the detected symbols are treated as correct and fed back as a training signal to "direct" the operation of the MMSE receiver. Of course, blindly detected symbols contain an arbitrary phase rotation which should be removed by incorporating differential encoding.

On the other hand, the generalized eigenvector required for the computation of the MRC weight vector in (35) can be also obtained via a time-recursive algorithm without the need of a complicated eigenvalue decomposition [9, 12]. The required computational complexity is of the order NL (BT dimension) per iteration.

5. Computer Simulations

Computer simulations were conducted to demonstrate the performance the proposed BT GSC based RAKE (G-RAKE) receiver. The antenna employed was a linear array consisting of $D = 8$ identical elements uniformly spaced by a $1/\sqrt{3}$ wavelength. The inter-element spacing was chosen for the field-of-view $[-60^\circ,\ 60^\circ]$, which represents the effective angular region of operation for a linear array [8]. Note that with the inter-element spacing chosen to be a $1/\sqrt{3}$ wavelength, $D = 8$ orthogonal beams can be formed in the 120° region, with two adjacent beams spaced by a half 3-dB beamwidth. This is in contrast to the conventional linear array with an inter-element spacing of $1/2$ wavelength, in which orthogonal beams are formed in the entire 180° region. For simplicity, the elements were assumed to be ideal omnidirectional antennas, leading to the following steering vector structure:

$$\mathbf{a}(\theta) = \left[1, e^{j\frac{2\pi}{\sqrt{3}}\sin\theta}, e^{j\frac{4\pi}{\sqrt{3}}\sin\theta}, \cdots, e^{j\frac{2(M-1)\pi}{\sqrt{3}}\sin\theta}\right]^T,$$

where θ is measured with respect to the normal to the array axis. In practical systems, however, directional antennas with a suitable front-to-back ratio should be employed to avoid backward radiation [8]. The sector of interest in our simulations was $[-20^\circ,\ 20^\circ]$, and $N = 3$ diversity beams were formed at look directions $\{-12.5^\circ,\ 0^\circ,\ 12.5^\circ\}$ to cover the 40° region. The blocking matrix $\mathbf{B}$ was constructed with $G = 4$ by the eigenvector constrained method described in [7]. It was assumed that the multipaths of all users followed the discrete uniform distribution model [8] with the same angle spread of 10°. That is, J paths were generated with their AOA's evenly distributed in a 10° angular interval centered at the line-of-sight (LOS) angle of the source. Moreover, the LOS angle of the signal was randomly selected in the working sector, and the LOS angles of the MAI's were randomly selected in the entire 120° field-of-view. The path gains $\alpha_{k,j}$'s were assumed independent, identically distributed unit variance complex Gaussian random variables, the path delays $\tau_{k,j}$'s were assumed uniform over $[0, 2T_c]$, and the number of paths was $J = 10$ for all users. All CDMA signals were generated using the Gold code of length $M = 31$. Finally, the number of fingers of the receiver was $L = 3$, and the temporal blocking matrix $\mathbf{C}$ was chosen, according to Section 3.2, to be the matrix whose columns are orthogonal to $\{\mathbf{S}_1(:, 1), \mathbf{S}_1(:, 2), \mathbf{S}_1(:, 3)\}$.

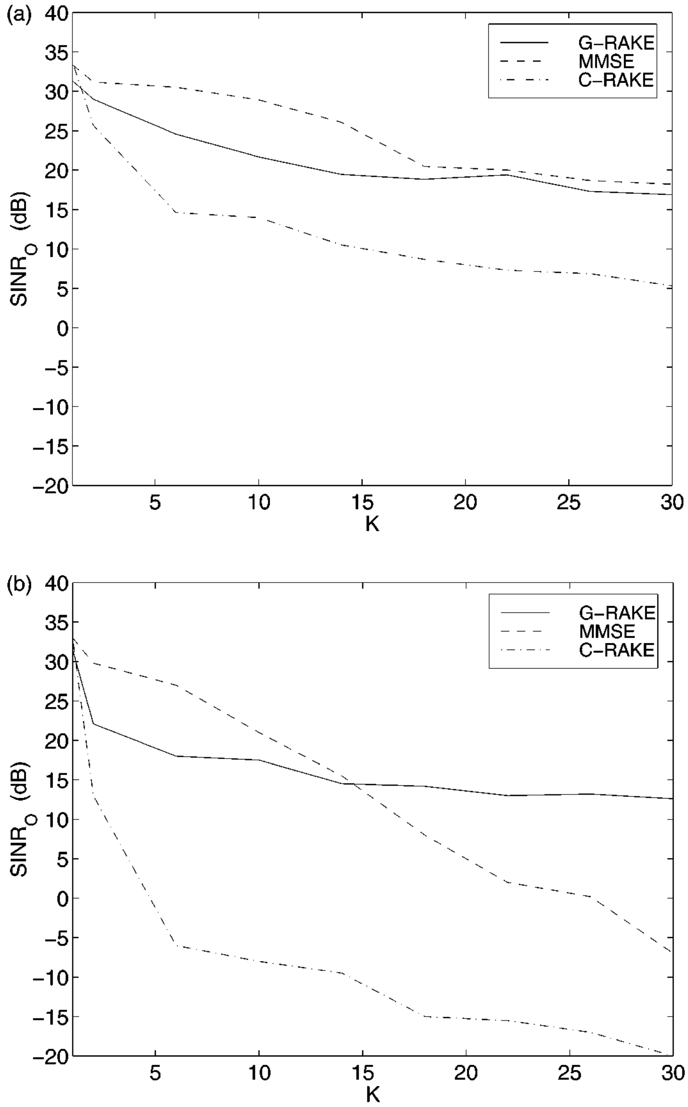


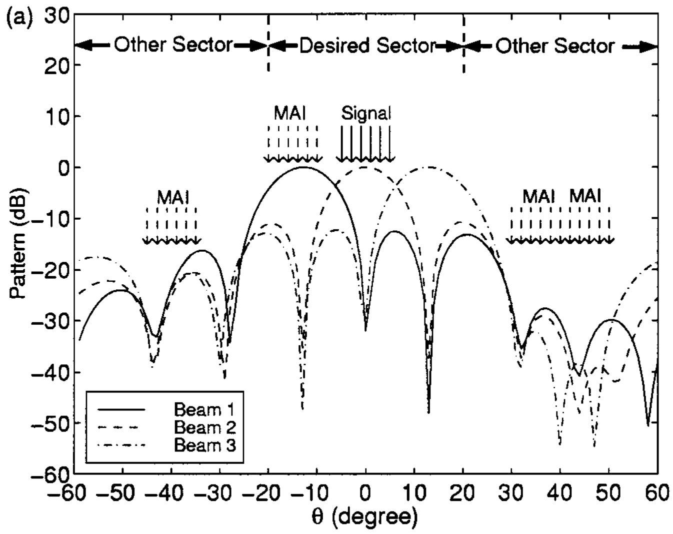
Figure 2. Evaluation of system capacity of G-RAKE, MMSE and C-RAKE receivers with (a) NFR = 0 dB (b) NFR = 20 dB.

As a performance index, we defined the output SINR to be the ratio of the signal power to the MAI-plus-noise power at the receiver output $z_o(i)$:

$$\text{SINR}_o = 10 \log_{10} \frac{\text{output power of signal in } z_o(i)}{\text{output power of (MAI+noise) in } z_o(i)}\,.$$

The input SNR was defined as:

$$\text{SNR}_i = 10 \log_{10} \frac{P_1}{\sigma_n^2}\,.$$



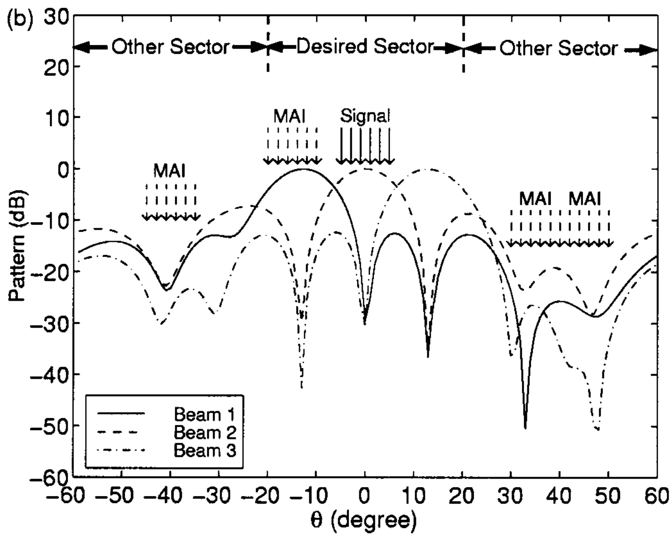


Figure 3. Patterns of diversity beams for (a) finger 1 (b) finger 3, obtained with $K = 5$ and NFR $= 10$ dB.

For simplicity, we assumed that all out-of-sector MAI's had the same power, and defined the near-far-ratio (NFR) to be the ratio of the out-of-sector MAI power to signal power before beamforming and despreading, i.e.,

$$\mathrm{NFR} = 10\log_{10}\frac{P_k}{P_1}$$

with k belonging to the out-of-sector MAI indices. Except for one case, the in-sector MAI was assumed power controlled with the signal, i.e., $P_k = P_1$ for k belonging to the in-sector MAI indices (this is not strictly necessary since in-sector MAI not power controlled can be suppressed by the adaptive correlators). For each result in the simulations, N_s symbols were used to estimate the correlation matrices, and a total of 50 Monte-Carlo runs were executed to obtain an average SINR_o, with each trial using a different set of $\alpha_{k,j}$'s, $\tau_{k,j}$'s and LOS

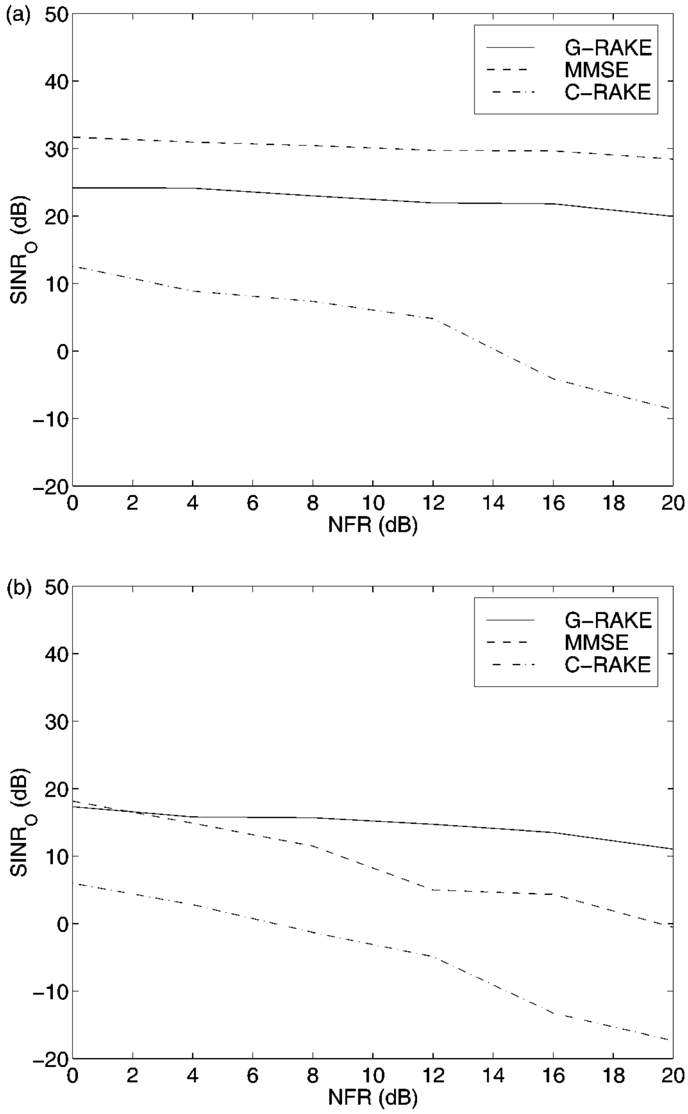


Figure 4. Evaluation of near-far resistance of G-RAKE, MMSE and C-RAKE receivers with in-sector MAI power controlled. (a) $K = 5$ (b) $K = 25$.

angles of the users. For comparison, we also included the results obtained with the non-blind ST coherent RAKE (C-RAKE) and MMSE receivers described in (10) and (11), respectively, with the channel vector $\tilde{\mathbf{h}}_1$ and post-despread ST data correlation matrix $\tilde{\mathbf{R}}$ estimated using the same N_s symbols as the training signal.

First, the system capacity is evaluated in Figure 2 with NFR $=$ 0 dB and 20 dB. The input SNR was set to be 0 dB, and $N_s = 500$.As expected, the MMSE receiver gives the best performance with a small user number K and low NFR, in which case both channel estimation and MAI suppression can be done effectively. On the other hand, the proposed G-RAKE receiver performs reliably for a wide range of K, even in the presence of strong MAI. In fact, the G-RAKE receiver outperforms the MMSE receiver with a moderately large K and high NFR, indicating that the adaptive beamformers and correlators have successfully

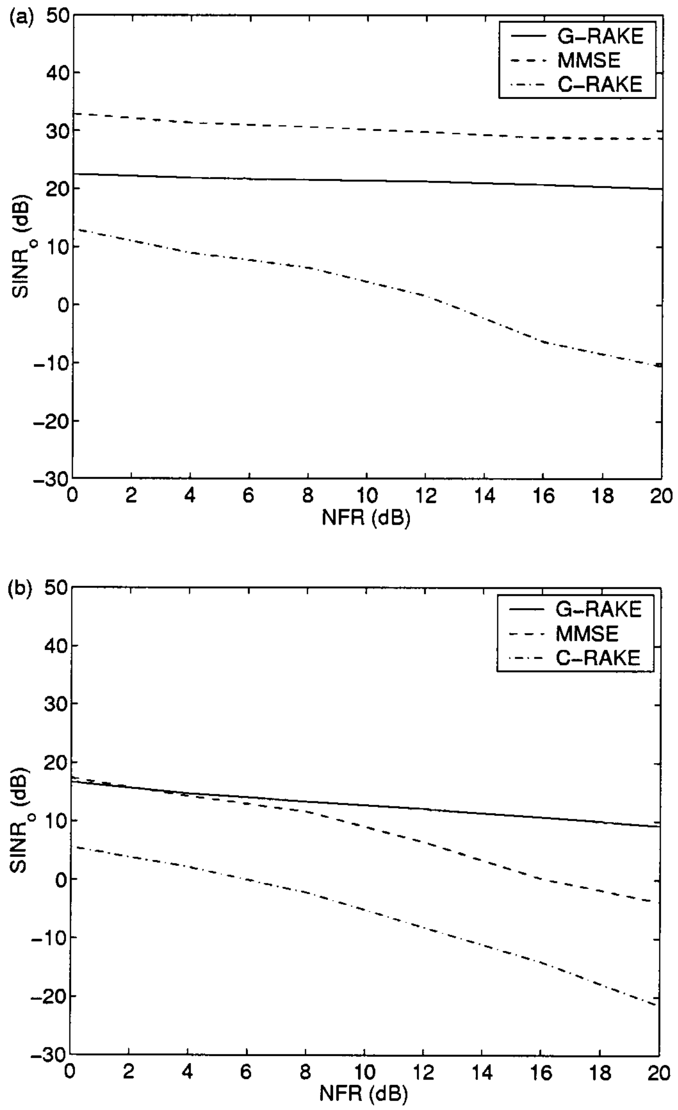


Figure 5. Evaluation of near-far resistance of G-RAKE, MMSE and C-RAKE receivers with in-sector MAI not power controlled. (a) $K = 5$ (b) $K = 25$.

eliminated the strong MAI. The C-RAKE receiver totally fails with NFR = 20 dB due to the lack of MAI suppression.

In Figure 3, the patterns of the diversity beams for the first and third fingers are plotted for the case $K = 5$ users and NFR = 10 dB. The mainlobes and deep nulls confirm that the adaptive beamformers can effectively collect the in-sector signals and suppress out-of-sector MAI.

Next, the near-far resistance of the proposed receiver is evaluated with different NFR values. Figure 4 shows the results obtained with $K = 5$ and 25 users, with the input SNR equal to 0 dB and $N_s = 500$. It is observed that the C-RAKE receiver fails again, and the MMSE receivers loses its near-far resistance with $K = 25$ due to the exhaustion of degree of freedom ($8 \times 3 = 24$) for strong MAI suppression. On the contrary, the G-RAKE receiver

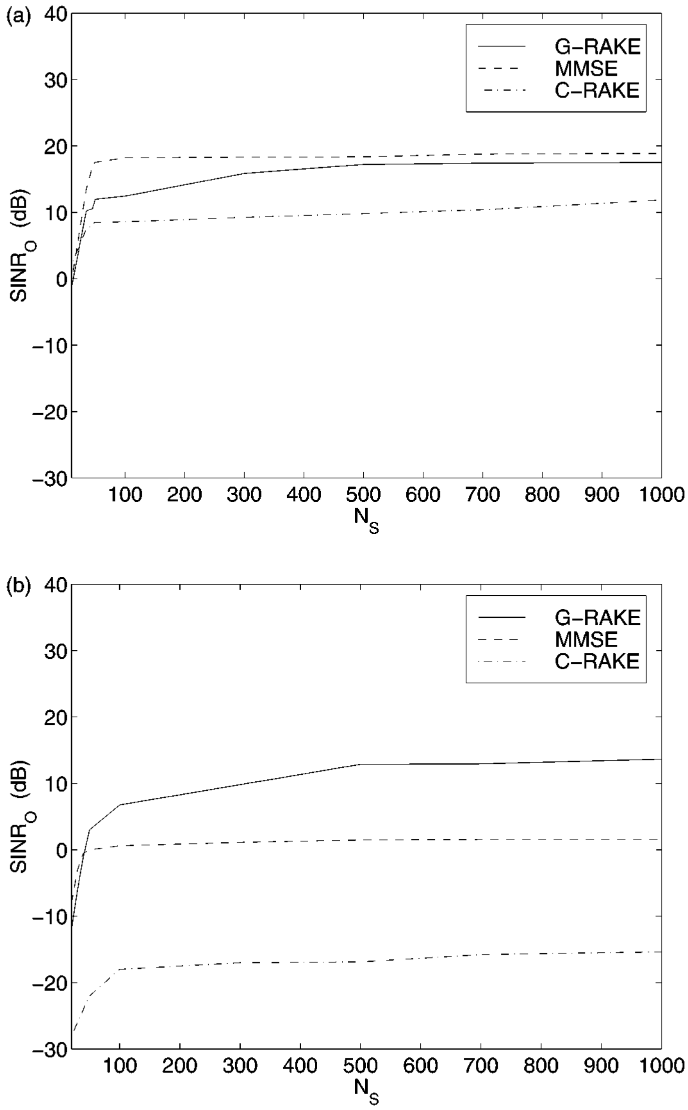


Figure 6. Convergence behaviors of G-RAKE, MMSE and C-RAKE receivers with $K = 25$ and (a) NFR $= 0$ dB (b) NFR $= 20$ dB.

achieves its excellent near-far resistance by successfully canceling the MAI using the temporal degree of freedom $(31 + 3 - 1 = 33)$ offered by the pre-despread chip sampled data.

To demonstrate the efficacy of the adaptive correlators in handling in-sector MAI, we repeated the same simulation with the power control of in-sector MAI "turned off". In this case, the in-sector and out-of-sector MAI had the same power determined by the NFR value. The results shown in Figure 5 confirm that the non-power-controlled in-sector MAI have little effect on the G-RAKE receiver even with $K = 25$ since they can be effectively suppressed as long as a sufficient degree of freedom is available for adaptive processing.

Finally, the convergence behaviors of the three RAKE receivers are evaluated by varying the data sample size N_s. The resulting output SINR are plotted in Figure 6 with NFR $= 0$ dB and 20 dB. The number of users was $K = 25$. As expected, the output SINR increases as N_s

increases for the three receivers. The MMSE receiver converges significantly faster than the G-RAKE receiver with a low NFR, but loses this advantage in the presence of strong MAI. The G-RAKE receiver achieves 95% of its maximum SINR_o in about 500 symbols, and takes another 500 symbols to reach its full performance. This is observed to be due to the errors in blind beamforming operation. It should be mentioned, however, that the relatively slow convergence of G-RAKE receiver does not raise practical problems since the receiver can be switched to the decision directed mode as long as the MAI has been sufficiently suppressed.

6. Conclusion

This paper proposes a blind adaptive 2-D RAKE receiver for sectored CDMA wireless communications. The proposed receiver is designed with a three-stage procedure. First, adaptive diversity beams are formed at each finger to collect in-sector multipath signals, and suppress strong out-of-sector MAI. Second, the output of each beamformer is processed by a set of signature matched adaptive correlators to suppress in-sector MAI. The beamformers and correlators together constitute a joint beamspace-time (BT) processor. Finally, a simple maximum ratio combining (MRC) criterion determines the tap weights of the BT RAKE receiver without the need of a training signal. Compared to the conventional antenna level receiver, beamspace receiver can increase the SINR by suppressing the out-of-sector MAI and lower the computational complexity by reducing the dimension. From simulation results, we have shown that the proposed RAKE receiver is near-far resistant, and performs reliably in a heavily loaded system.

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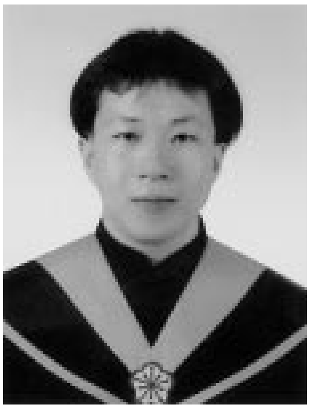

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