BLIND 2D RAKE RECEIVER FOR CDMA INCORPORATING CODE SYNCHRONIZATION AND MULTIPATH TIME DELAY ESTIMATION. *

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ABSTRACT

A novel wideband beamforming technique for cellular CDMA systems is presented in this paper. The proposed algorithm asymptotically provides the optimum combination of space-time samples to maximize the SINR for the *Signal with the Desired Code* (SDC) by optimally combining its multipath and canceling strong *Multi-User Access Interference* (MUAI). In contrast to previously proposed techniques, code synchronization for the SDC is not required. The algorithm presented herein asymptotically provides the exact time of arrivals of the multipaths within a bit period, and subsequently the optimum space-time weights for combining the fingers across both space and time. The instrumental property exploited in this technique is the fact that although the respective spectra of the SDC and MUAI components at the output of the matched filter are statistically equal, the respective spectra of their squared values differ.

1. INTRODUCTION

The problem investigated is that of combatting both multipath and multi-user access interference in a cellular CDMA system. In order to achieve the best performance relative to probability of bit error, the goal is to cancel strong MUAI while simultaneously combining the multipath for the SDC via a RAKE-type receiver. The ability to cancel strong MUAI improves the overall capacity of the cellular system [2]. It also eliminates the near-far problem that occurs when the SDC transmits from the outer edges of the cell while another co-channel user transmits simultaneously much closer to the base [1].

2D RAKE receivers have been proposed in recent years for optimally combining the multipath for the SDC across both space and time, while simultaneously canceling strong MUAI [3]. An algorithm is presented herein for determining the optimum space-time sample combination that has several major advantages over previous methods [5]. First, it does not require bit synchronization for the desired user. Using only knowledge of the spreading waveform for the desired user, the algorithm provides estimates of the time of arrivals of the RAKE fingers. Finally, synchronization amongst the multi-users is not assumed, knowledge of the codes for the other users is not required, no model is assumed for the antennas, and no pilot symbols are needed.

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2. DATA MODEL

The scenario assumed is that of J users in a CDMA multiuser access system sharing a common frequency band. Discrimination amongst users is achieved by different spreading signatures or PN (Pseudo Noise) codes. The J coexisting signals are collected by each of N antennas. There is no constraint on either the antenna locations or their respective radiation patterns. The signal transmitted from the k-th user, $(1 \leq k \leq J)$, arrives at the antenna array via L_k paths. The direct path of the user k is received with amplitude σ_k. The remaining $L_k - 1$ multipaths of the k-th user are attenuated and phase shifted by the complex number ρ_k^l, $2 \leq l \leq L_k$. Therefore, without lost of generality the following notation assumes $|\rho_k^1| = 1$ for each of the J users. The l-th path of the k-th user arrives with a delay of τ_k^l and through a direction $\boldsymbol{\theta}_k^l$, where the boldface on θ denotes the fact that no constraint on the array geometry is assumed so that the direction is 2-D, in general. Under the narrowband assumption, the 2-D angle $\boldsymbol{\theta}_k^l$ influences only a complex scale factor between antennas which is characterized by the array manifold $\mathbf{a}(\boldsymbol{\theta}_k^l)$. Using the above notation, the vector of the received signals $\mathbf{x}(t)$ is given by

$$\mathbf{x}(t) = \sum_{k=1}^{J} \sigma_k \sum_{l=1}^{L_k} \rho_k^l \mathbf{a}(\boldsymbol{\theta}_k^l) \sum_{n=-\infty}^{\infty} b_k(n) c_k(t - nT_b - \tau_k^l) + \mathbf{n}(t), \quad (1)$$

where $b_k(n)$ is the $n-$th symbol (bit) value, $c_k(t)$ is the code assigned to the k-th user and T_b is the bit period. The code is composed of L_c chips of duration T_c. $\mathbf{n}(t)$ is the noise vector with $E[\mathbf{n}(t)\mathbf{n}^H(t)] = \sigma_n^2 \mathbf{I}$, σ_n^2 is the noise power, and $\mathbf{I}$ the $N \times N$ identity matrix.

The output signal from each antenna is passed through a matched filter based on the spreading waveform of the desired user. The impulse response of the filter is $h(t) = c_1^*(-t)$, assuming the signal enumerated as $j = 1$ is the desired user, the SDC. The signal vector at the output of these filters is

$$\mathbf{y}(t) = \sum_{j=1}^{J} \sigma_j \sum_{l=1}^{L_j} \rho_j^l \mathbf{a}(\boldsymbol{\theta}_j^l) \cdot \sum_{n=-\infty}^{\infty} b_j(n) c_j(t - nT_b - \tau_j^l) * c_1(-t) + \mathbf{n}_y(t), \quad (2)$$

where the operator $*$ denotes convolution and $\mathbf{n}_y(t)$ is an

$N \times 1$ vector containing i.i.d. random processes obtained as $\mathbf{n}(t) * c_1(-t)$.

With regard to each of the terms in (2), note that ideally we would like to construct PN codes that satisfy $c_j(t) * c_i(t) = \delta_{ji}\delta(t)$[1], where δ_{ji} is the Kronecker delta and $\delta(t)$ is the Dirac delta function. If this condition was strictly satisfied only L_1 terms in (2) would be non-zero and they would take non-zero values for $t = \tau_1^l$, $1 \leq l \leq L_1$. These terms are called *fingers*. However perfect orthogonality amongst codes cannot be achieved for each and every arbitrary time delay with finite length codes. Thus, residual cross-correlation terms arise in practical systems. These spurious terms can be stronger than the fingers when the MUAI is received with much higher power levels than the SDC, hence, the *Near-Far problem*.

3. SPACE-TIME PROCESSING.

The algorithm proposed herein proceeds as follows. First, it estimates the locations of the fingers $t = \tau_1^l$, $1 \leq l \leq L_1$. Second, for each instant $t = \tau_1^l$ it finds the beamformer that extracts that finger and cancels the rest of the paths present with a significant power level at that particular time. It cancels paths corresponding to both MUAI's and the other $L_1 - 1$ paths from the SDC as well. The weight vector that achieves both of these conditions is obtained from a generalized eigenvalue decomposition. Once the fingers are extracted, the third step of the algorithm determines how they should be optimally combined. Note that after the MUAI's are canceled, the remaining noise is Gaussian.

3.1. Arrival Time and Extraction of Fingers

Let the $N \times N$ *time-varying* correlation matrix $\mathbf{R}_y(t)$ be defined as

$$\mathbf{R}_y(t) \stackrel{\text{def}}{=} E[\mathbf{y}(t)\mathbf{y}^H(t)\text{rect}\left(\frac{t}{T_b}\right)], \tag{3}$$

where $E[\bullet]$ is the expectation operator, $(\cdot)^H$ is the conjugate transpose, and $rect\left(\frac{t}{T_b}\right)$ is unity over an interval of width T_b centered at $t = 0$ and zero elsewhere. Further, let $\mathbf{S}_y(f)$ be defined as $\mathbf{S}_y(f) \stackrel{\text{def}}{=} \mathcal{F}[\mathbf{R}_y(t)]$, where $\mathcal{F}$ represents the Fourier Transform operating element-wise on $\mathbf{R}_y(t)$.

Substituting the signal model in (2) and assuming

1. the symbol values are uncorrelated for a given user as well as between different users, i.e., $E\{b_k(n)b_p^*(m)\} = \sigma_b^2\delta_{kp}\delta_{nm}$, where σ_b^2 is the average energy per symbol,
2. the noise is stationary, Gaussian and i.i.d. (independent and identically distributed) for all the antennae with variance σ_n^2,
3. noise is uncorrelated with each user's signal, and
4. $c_k(t)$ are PN codes, therefore the phase of the Fourier Transform (FT) of each code $C_k(f)$ may be well modeled as a random variable uniformly distributed over $[0, 2\pi)$

the following proposition can be made.

Theorem 1

$$\mathbf{S}_y(f) = \sum_{k=1}^{J}\sum_{l=1}^{L_k} p_k^l \mathbf{a}(\theta_k^l)\mathbf{a}^H(\theta_k^l) S_k^{ll}(f) + \sigma_n^2 \mathbf{I}\ sinc(fT_b), \tag{4}$$

where

$$S_k^{ll}(f) = \begin{cases} S_c(0)\delta(f) & \text{if } k \neq 1 \\ e^{j2\pi\tau_1^l f} S_c(f) & \text{if } k = 1 \end{cases} \tag{5}$$

and $p_k^l = \sigma_k^2 |\rho_k^l|^2 \sigma_b^2$,

$$S_c(f) = \left[\left(|C_p(f)|^2 * |C_p(f)|^2\right) T(f)\right] * sinc(fT_b), \tag{6}$$

$C_p(f)$ is the Fourier transform of the chip pulse waveform and $T(f) = \sum_{n=-\infty}^{\infty} e^{j2\pi f n T_b} = \sum_{n=-\infty}^{\infty} \delta(f - \frac{n}{T_b})$.

Proof: Using the definitions of $\mathbf{R}_y(t)$, $\mathbf{S}_y(f)$ and the FT, $\mathbf{S}_y(f)$ may be expressed as

$$\mathbf{S}_y(f) = \int_{-\infty}^{\infty} E[\mathbf{y(t)y(t)}^H \text{rect}\left(\frac{t}{T_b}\right)] e^{j2\pi t f} dt \tag{7}$$

Substituting the signal model assumed for $\mathbf{y}(t)$ in (2) into (7) and invoking the assumptions listed previously as 1 and 3, (7) simplifies as

$$\mathbf{S}_y(f) = \sigma_b^2 \int_{-\infty}^{\infty} \left\{ \sum_{k=1}^{J} \sigma_k^2 \sum_{l=1}^{L_k}\sum_{q=1}^{L_k} \rho_k^l \rho_k^{q*} \mathbf{a}(\theta_k^l)\mathbf{a}^H(\theta_k^q) \right.$$

$$\sum_{n=-\infty}^{\infty} \{c_k(t-nT_b-\tau_k^l) * c_1(-t)\}\{c_k(t-nT_b-\tau_k^q) * c_1(-t)\}^* +$$

$$\left. E\left[\mathbf{n}_y(t)\mathbf{n}_y^H(t)\right] \right\} \text{rect}\left(\frac{t}{T_b}\right) e^{-j2\pi t f} dt \tag{8}$$

Under these assumption number 2 (8) simplifies as

$$\mathbf{S}_y(f) = \sigma_b^2 \sum_{k=1}^{J} \sigma_k^2 \sum_{l=1}^{L_k}\sum_{q=1}^{L_k} \rho_k^l \rho_k^{q*} \mathbf{a}(\theta_k^l)\mathbf{a}^H(\theta_k^q) S_k^{lq}(f)$$

$$+\sigma_n^2 \mathbf{I}\ \text{sinc}(fT_b) \quad \text{where} \tag{9}$$

$$S_k^{lq}(f) = \int_{-\infty}^{\infty} \sum_{n=-\infty}^{\infty} \{c_k(t - nT_b - \tau_k^l) * c_1(-t)\}$$

$$\cdot \{c_k(t - nT_b - \tau_k^q) * c_1(-t)\}^* \text{rect}\left(\frac{t}{T_b}\right) e^{-j2\pi t f} dt. \tag{10}$$

In order to prove (4) and (5) we only have to guarantee that $S_k^{lq}(f)$ may be expressed as

$$S_k^{lq}(f) = \begin{cases} S_c(0)\delta(f) & \text{if } k \neq 1 \\ \delta_{lq} e^{j2\pi\tau_1^l f} S_c(f) & \text{if } k = 1 \end{cases} \tag{11}$$

Assuming $c_k(t)$ is real-valued and using elemental FT properties $S_k^{lq}(f)$ in (8) may be expressed as

$$S_k^{lq}(f) = \sum_{n=-\infty}^{\infty} \left(C_k(f) C_1^*(f) e^{j2\pi f n T_b} e^{j2\pi f \tau_k^l} \right) *$$

$$\left(C_k(f) C_1^*(f) e^{j2\pi f n T_b} e^{j2\pi f \tau_k^q} \right) * \text{sinc}(fT_b) \tag{12}$$

Note that by construction PN codes approximately satisfy: i) $|C_k(f)| = |C_p(f)|$ where $C_p(f)$ is the FT of the chip waveform, and ii) the phase of $C_k(f)$ is a white random process with a first-order p.d.f. uniformly distributed in $[0, 2\pi)$.

[1]The codes are normalized to have unit energy.

Let us analyze the cases $k = 1$ and $k \neq 1$.

a) $k = 1$

When $l = q$, (12) becomes

$$S_1^{ll}(f) = e^{j2\pi f \tau_1^l}\left[\left(|C_p(f)|^2 * |C_p(f)|^2\right) T(f)\right] * \text{sinc}(fT_b) \quad (13)$$

where $T(f) = \frac{1}{T_b}\sum_{n=-\infty}^{\infty} e^{j2\pi f n T_b} = \sum_{n=-\infty}^{\infty}\delta(f - \frac{n}{T_b})$

Otherwise, if $l \neq q$, the integral $\int_{-\infty}^{\infty}|C_p(s)|^2|C_p(f-s)|^2 e^{j2\pi s(\tau_1^l - \tau_1^q)}ds$ is negligible whenever $\tau_1^l - \tau_1^q \geq T_c$, where T_c is the chip period. In particular, for rectangular waveform chips and $\tau_1^l - \tau_1^q = T_c$, the integral value is exactly 0. For other relative delays and chip waveforms (e.g., a chip waveform having a raised cosine spectrum) simulations confirm this property holds.

b) $k \neq 1$

MUAIs belonging to the same system have the same chip pulse shape and $C_k(f) = |C_p(f)|e^{jr_k(f)}$ where $r_k(f)$ is a white random process uniformly distributed over $[0, 2\pi)$. Also the PN codes are such that the phases of any two of them are independent. Using these properties in (12) it follows that

$$S_k^{lq}(f) = \sum_{n=-\infty}^{\infty}\left(|C_p(f)|^2 e^{j(r_k(f)-r_1(f))}e^{j2\pi f n T_b}e^{j2\pi f \tau_k^l}\right) *$$

$$\left(|C_p(f)|^2 e^{j(r_k(f)-r_1(f))}e^{j2\pi f n T_b}e^{j2\pi f \tau_k^q}\right) * \text{sinc}(fT_b) \quad (14)$$

$e^{jr_k(f)}$ and $e^{jr_1(f)}$ are complex independent random processes with phase uniformly distributed over $[0, 2\pi)$ and constant amplitude. Therefore, $e^{j(r_k(f)-r_1(f))} = e^{jr_k'(f)}$ is also a complex independent random variables with the phase uniformly distributed over $[0, 2\pi)$. Thus, (14) is non-zero only for $f = 0$ and $l = q$. Moreover, (14) simplifies as

$$S_k^{lq}(0) = \int_{-\infty}^{\infty}|C_p(s)|^4 ds = S_c(0) \quad (15)$$

■

In a practical implementation $\mathbf{R}_y(t)$ is computed from a sampled version of $\mathbf{y}(t)$ and $\mathbf{S}_y(f)$ as the DFT of $\mathbf{R}_y(n)$. Since $\mathbf{R}_y(t)$ is only nonzero for $-\frac{T_b}{2} \leq t \leq \frac{T_b}{2}$ the DFT of its sampled version evaluates $\mathbf{S}_y(f)$ at $f = \frac{\kappa}{T_b}$.

To substantiate the approximation in (5), Figure 1 shows $|S_k^{ll}(f)|$ for a set of maximal length PN codes of length $L_c = 127$ sampled twice per chip and a raised cosine chip waveform with roll-off coefficient $\beta = .9$. Figure 1 (a) shows $|S_1^{ll}(f)|$ corresponding to SDC and figure 1 (b) shows $|S_2^{ll}(f)|$ corresponding to the l-th path of a MUAI.

Let the matrices $\mathbf{S}_y^{(\kappa)}$ and $\mathbf{S}_y^{(0)}$ be defined as $\mathbf{S}_y(f)$ evaluated at $f = \frac{\kappa}{T_b}$ and $f = 0$, respectively. For $f = 0$, define the "cleaned" version of $\mathbf{S}_y^{(0)}$ as $\mathbf{C}_y^{(0)} = \mathbf{S}_y^{(0)} - \lambda_{min}\mathbf{I}$, where λ_{min} is the smallest eigenvalue of $\mathbf{S}_y^{(0)}$ ideally equal to σ_n^2 in accordance with (4). Consider the matrix pencil $\left\{\mathbf{S}_y^{(\kappa)} - \lambda_i \mathbf{C}_y^{(0)}\right\}\mathbf{w}_i = 0$. Substituting (4) and (5) yields

$$\{S_c\left(\frac{\kappa}{T_b}\right)\sum_{l=1}^{L_1}p_1^l\mathbf{a}(\theta_1^l)\mathbf{a}^H(\theta_1^l)e^{j2\pi\kappa\frac{\tau_1^l}{T_b}}$$

$$-\lambda_i S_c(0)\sum_{k=1}^{J}\sum_{l=1}^{L_k}p_k^l\mathbf{a}(\theta_k^l)\mathbf{a}^H(\theta_k^l)\}\mathbf{w}_i = \mathbf{0} \quad (16)$$

Note that there are L_1 non-zero generalized eigenvalues equal to $\lambda_i = \frac{s_c(\kappa)}{s_c(0)}e^{j2\pi\kappa\frac{\tau_1^i}{T_b}}$ and that each of the corresponding eigenvectors satisfy $\mathbf{w}_i^H\mathbf{a}(\theta_k^l) \propto \delta_{k1}\delta_{il}$, i.e., $\mathbf{w}_i$ is orthogonal to each $\mathbf{a}(\theta_k^l)$, $1 \leq k \leq J$ and $1 \leq l \leq L_k$, except $\mathbf{a}(\theta_1^i)$. Thus, the time location of the l-th finger may be extracted from the phase of the l-th eigenvalue, while the corresponding eigenvector provides the beamformer for extracting the l-th finger, $1 \leq l \leq L_1$.

$\kappa = 1$ is required if there is no synchronization information available. In this case, the proposed algorithm uses the knowledge of the desired user's code to estimate the absolute time locations of the fingers within a bit interval. Higher values of κ lead to an indeterminacy in the time location estimates, but at the same time yield a greater separation in the phase of the eigenvalues thereby causing the generalized eigenvalue problem to be better conditioned. Simulations have revealed that the latter leads to faster convergence. For cold start-up, one can proceed by first solving the problem for $\kappa = 1$ to roughly determine which portion of the post-correlation bit interval the fingers lie in, and then use a value of κ equal to the ratio of the bit duration to the maximum multipath time delay spread.

In [6] we prove the phase of the generalized eigenvalues of the pencil $\{\mathbf{S}_y^{(\kappa)}, \mathbf{S}_y^{(0)}\}$ is the same as those satisfying (16). Moreover, the the eigendecomposition of $\{\mathbf{S}_y^{(\kappa)}, \mathbf{S}_y^{(0)}\}$ is a better conditioned problem with a more stable solution.

3.2. Optimum Combination of the Beamformer Outputs

In accordance with previous sections, the inner product $\beta_l(n) = \mathbf{w}_l^H\mathbf{y}(nT_b + \tau_1^l)$ follows the expression

$$\beta_l(n) = \mathbf{w}_l^H\mathbf{y}(nT_b + \tau_1^l) = b_1(n)\sigma_1\rho_1^l\mathbf{w}_l^H\mathbf{a}(\theta_1^l) + \mathbf{n}_l \quad (17)$$

where $\mathbf{n}_l = \mathbf{w}_l^H\mathbf{n}_y(nT_b + \tau_1^l)$. We have used the signal model in (2) and assumed the codes $c_k(t)$ are normalized to have unit energy.

The final step in this time-space filtering technique for CDMA consists of the optimum combination of the L_1 *beamspace-time* samples $\beta_l(n)$ to maximize the SINR. Let $\beta(n)$ and $\mathbf{n}_w(n)$ be the $L_1 \times 1$ vectors formed with the entries $\beta_l(n)$ and $\mathbf{n}_l$, $l = 1, \cdots, L_1$, respectively. Now the objective is to find the $L_1 \times 1$ vector γ such that $\hat{b}(n) = \gamma^H\beta(n)$ maximizes the SINR.

The beamspace-time vector $\beta(n)$ is formed with the outputs of the weight vectors $\mathbf{w}_l$ which were proved to be asymptotically orthogonal to all the interferences. Therefore, there is no interference contribution to the vector $\beta(n)$ as shown in equation (17). Under these conditions the SINR is maximized by maximizing the SNR. Moreover the SINR maximization can be expressed as

$$\max_{\gamma}\frac{E[\gamma^H\beta(n)\beta^H(n)\gamma]}{E[\gamma^H\mathbf{n}_w\mathbf{n}_w^H\gamma]} \quad (18)$$

$E[\mathbf{n}_w\mathbf{n}_w^H] = \mathbf{I}$ because the entries of $\mathbf{n}_w$ are samples of noise taken at different snapshots. Thus, the denominator in (18) becomes $\gamma^H\gamma$. The value of γ that maximizes the SINR is the eigenvector corresponding to the largest eigenvalue of the matrix

$$\mathbf{R}_\beta = E[\beta(n)\beta^H(n)] \quad (19)$$

4. SIMULATIONS

We here present a simulation to analyze the effectiveness of the technique described herein. A scenario where the SDC arrives through 3 paths was simulated. The angle of arrival of the paths are 0, 3 and -2 degrees, respectively. Each ray is delayed by one chip respect the previous one. The second ray is 1.5 dB weaker than the direct ray (the first one) and it arrives with a phase shift of 45 degrees at the center of the array. The third ray is 2 dB weaker than the direct path and phase shifted -90 degrees. The SWNR of the direct paths is -5 dB before correlation with the desired code at each of the antennae. The array is an Uniform Linear Array of 8 elements with half wavelength spacing. The chip waveform is raised cosine with $\beta = 0.5$.

Because the angular spread of the multipath is small compared to the beamwidth associated with the array employed, the algorithm forms a unique common beam to extract the 3 paths. The value of $\kappa = 12$. 15 bits are averaged to estimate $\mathbf{S}_y^{(0)}$ and $\mathbf{S}_y^{(\kappa)}$. Figure 2 shows the Bit Error Rate (BER) when the output of the Space-Time processor is fed to a slicer that decides each bit independently – no coding. Figure 2 correspond to a MonteCarlo simulation of a scenario where new MUAIs appear sequentially. The first MUAI arrives at -90 degrees. The following MUAIs arrive at angles in increments of 20 degrees (-90, -70,...). All the MUAIs are 20 dB above the SDC. Thus only 1 MUAI causes the BER to degenerate to 0.5 in the case of a single antenna and sampling at the peak of the strongest finger.

The BER is computed as the $Q\left(\frac{d}{2\sigma}\right)$ where $2d$ is the distance between the two points of the constellation free of noise and interferences, σ is the standard deviation of the points of the constellation around the noise and interference free points, and $Q(x) = \frac{2}{\pi}\int_{t=x}^{\infty} e^{-t^2}dt$. $\frac{d}{2\sigma}$ is estimated by a MonteCarlo simulation of 4096 independent trials.

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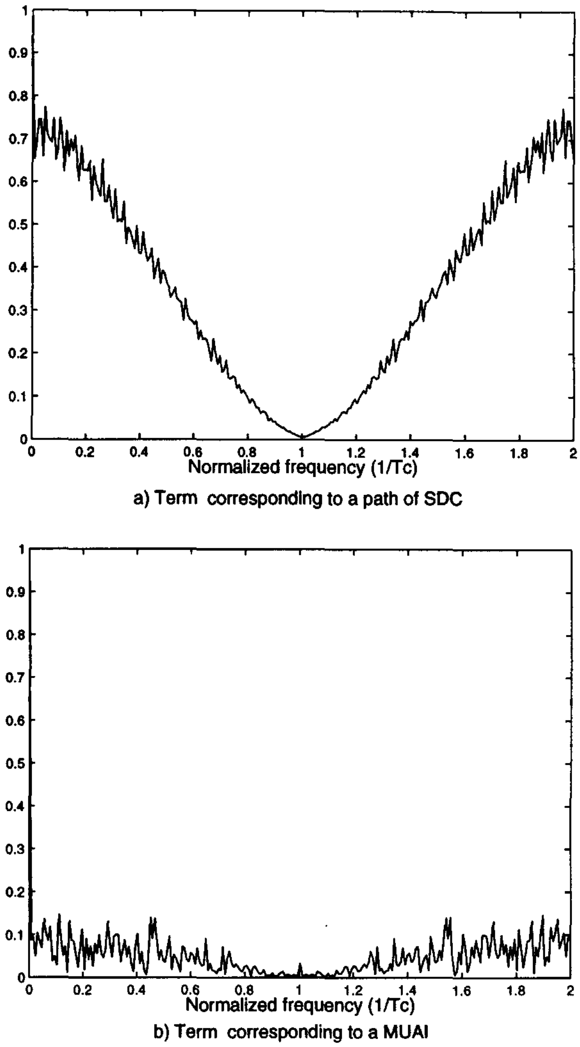

Figure 1. $|S_k^l(k)|$ **for a)** $k = 1$ **and b)** $k \neq 1$.

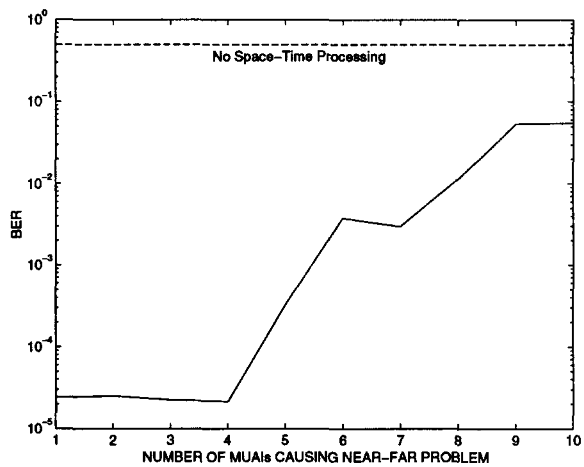

Figure 2. BER vs. Number of MUAIs causing Near-Far Problem (20 dB)