ROBUST DETECTION IN ASYNCHRONOUS CDMA MULTIPATH FADING CHANNELS

Yong-Suk Moon and Mostafa Kaveh and Laurie B. Nelson

Department of Electrical and Computer Engineering
University of Minnesota
Minneapolis, MN 55455

ABSTRACT

This paper presents a robust detection approach in the presence of transmission delay estimation errors in asynchronous CDMA frequency selective Rayleigh fading channels. The transmission delays in asynchronous channels need to be estimated in practice and are subject to estimation errors, which may cause severe performance degradation for most detectors assuming perfect synchronization. As the fundamental framework of the robust detector, an asynchronous subspace detection scheme based on oblique projection is introduced and examined in comparison with the conventional RAKE receiver and the multipath decorrelator. The performance degradation of the subspace detector due to delay estimation errors is also investigated. The robust detector shows reliable performance even when the errors are much larger than those of delay estimation algorithms like MUSIC, and it is not affected by the near-far problem.

1. INTRODUCTION

In many CDMA applications, such as cellular mobile radio, indoor wireless communications and personal communication networks, multipath fading characteristics of channels often severely limit the system performance. The conventional approach to combat this problem, and in fact take advantage of resolvable paths is to employ a single-user RAKE receiver which consists of a matched filter banks with a diversity technique, like maximal ratio combiner, for each user [6]. The RAKE system, however, suffers from the near-far problem, though it is optimal for single-user transmission in a multipath fading channel. Several multisuer detection schemes for multipath fading channels have been derived to alleviate the near-far problem inherent in RAKE receivers. Zvonar presented the optimum receiver [9] with the assumption of perfect estimation of all channel coefficients, and also addressed a linear detector [10] based on the asynchronous decorrelator [3] to reduce the complexity of the optimum detector. The RAKE-based decorrelating detector which consists a bank of conventional RAKE receivers for all the users was presented in [2], and was shown to have better performance over the multipath decorrelator.

Most receivers in multipath fading channels assume perfect synchronization for all users' signals. The transmission delays in asynchronous channels should be acquired before the detection in practice, and thus, estimation errors are involved in the acquisition process. It has been reported that even a small estimation error may cause severe degradation of the detection performance and the near-far resistance of the decorrelating detector [7]. This effect may be much worse in multipath fading channels since whole paths of all users act like multiple access interferences (MAIs). The goal of this paper is to present a detector which is robust to delay estimation errors in an asynchronous CDMA multipath fading channel. First, we introduce an asynchronous near-far resistant subspace detector based on oblique projection with the assumption of perfect synchronization, which is the extension of [5]. Then, the perturbation effect of the subspace detector due to delay estimation errors is investigated, and finally, a robust version of the subspace detector is presented.

This work was supported by the National Science foundation under Grant # NSF/NCR-9625085

2. SIGNAL MODEL

Consider an asynchronous K-user frequency selective Rayleigh multipath fading channel, where the bandwidth of the spread-spectrum signals W is much greater than the coherence bandwidth $(\Delta f)_c$ of the channel. The number of resolvable paths for each multiuser signal in this channel is given by $L = \lceil WT_m \rceil$ [6], where $T_m = 1/(\Delta f)_c$ is the channel multipath spread. Hence, the time-varying frequency selective channel for each user can be modeled as a tapped delay line with tap spacing $T_c = \frac{1}{W}$ and weight coefficients which are independent complex Gaussian random variables due to the Rayleigh fading nature of the channel. The received signal of the channel can be represented as

$$r(t) = \sum_{i=-M}^{M} \sum_{k=1}^{K} b_k^i h_k(t - iT - \tau_k) + n(t), \tag{1}$$

where b_k^i and τ_k are the data symbol and the transmission delay of k_{th} user respectively; $n(t)$ is zero mean complex Gaussian random noise with power spectral density N_0; $h_k(t)$ is defined by $h_k(t) = \sum_{j=0}^{L-1} g_{k,j}(t) A_k \mathbf{c}_k(t - jT_c) e^{j\phi_k}$, where A_k, ϕ_k, and $g_{k,j}(t)$ denote user k's transmission signal amplitude, the phase and the channel coefficient of j_{th} path respectively, and $\mathbf{c}_k(t)$ is the normalized signature waveform with support of $[0, T]$. When the coherence time of the channel is much greater than the data symbol interval, the channel is slowly fading, and we can assume that the channel fading coefficients remain constant during a symbol interval. Perfect estimation of the fading gains only for the desired user is assumed throughout this paper.

We perform chip matched filtering followed by chip-rate sampling at the receiver to yield an equivalent discrete version of the received signal vector. The chip matched filter output samples during the i_{th} symbol interval is obtained by

$$\mathbf{r_i} = \sum_{k=1}^{K} \sum_{j=0}^{L-1} \{\alpha_{k,j}^{i-1} b_k^{i-1} \mathbf{u}_{k,j}^r + \alpha_{k,j}^i b_k^i \mathbf{u}_{k,j}^l\} + \mathbf{n}_i, \tag{2}$$

where $\alpha_{k,j}^i = A_k g_{k,j}(i) e^{j\phi_k}$; $\mathbf{n}_i$ is a complex Gaussian noise vector with mean $\mathbf{0}$ and covariance matrix $\sigma^2 \mathbf{I}_N$; $\mathbf{u}_{k,j}^r$ and $\mathbf{u}_{k,j}^l$ are the chip matched filter outputs of the delayed spreading codes associated with the previous and current data symbols respectively, which are given by

$$\begin{aligned} \mathbf{u}_{k,j}^r &= (1-\delta_k)\mathbf{c}_k^r(p_{k,j}) + \delta_k \mathbf{c}_k^r(p_{k,j+1}) \\ \mathbf{u}_{k,j}^l &= (1-\delta_k)\mathbf{c}_k^l(p_{k,j}) + \delta_k \mathbf{c}_k^l(p_{k,j+1}), \end{aligned} \tag{3}$$

where $p_{k,j}$ denotes the chip delay of j_{th} path of user k, and the relation, $p_{k,j+1} = p_{k,j} + 1$, is established from the property of the tapped delay line model of the channel; $\delta_{k,j} \in [0, T_c)$ denotes the interchip delay of user k, and thus, total delay is represented as $\tau_k = p_{k,0} T_c + \delta_k$. Note that a sample for a non-zero interchip delay δ is represented as a convex combination of the two adjacent chips. $\mathbf{c}_k^r(p_{k,j})$ and $\mathbf{c}_k^l(p_{k,j})$ denote the $N \times 1$ shifted versions of spreading codes for the previous and current data symbols respectively, which are defined by $\mathbf{c}_k^r(p_{k,j}) \triangleq 1/\sqrt{N}[\beta_{N-p_{k,j}}^k \cdots \beta_N^k \; 0 \cdots 0]^T$ and $\mathbf{c}_k^l(p_{k,j}) \triangleq 1/\sqrt{N}[0 \cdots 0 \; \beta_0^k \cdots \beta_{N-p_{k,j}-1}^k]^T$, where N is the spreading gain, and β_r^k denotes the r_{th} chip of the k_{th} user's spreading code.

3. ASYNCHRONOUS DETECTION FOR A FADING CHANNEL

Recently, we introduced subspace detection approaches based on oblique projection which resolves a signal space into desired signal and interference subspaces [4, 5]. The oblique projection operator gives a geometrical explanation for a decentralized detector in multiuser channel, which is given by [1]

$$\mathbf{L}_S = \mathbf{S}(\mathbf{S}^H \mathbf{P}_G^{\perp} \mathbf{S})^{-1} \mathbf{S}^H \mathbf{P}_G^{\perp}, \tag{4}$$

where $< \mathbf{S} >$ and $< \mathbf{G} >$ denote the desired signal and the interference subspaces respectively, and $\mathbf{P}_G^{\perp}$ is the null-steering operator which nulls all interferences in a signal vector.

Reconfigure the signal model in (2) with subspace parameters as follows :

$$\mathbf{r}_i = \mathbf{S}^r \boldsymbol{\Theta}_{i-1}^r + \mathbf{S}^l \boldsymbol{\Theta}_i^l + \mathbf{G}^r \boldsymbol{\Psi}_{i-1}^r + \mathbf{G}^l \boldsymbol{\Psi}_i^l + \mathbf{n}_i, \tag{5}$$

where

$$\begin{aligned} \mathbf{S}^r &= [\mathbf{u}_{1,0}^r \cdots \mathbf{u}_{1,L-1}^r], \quad \mathbf{S}^l = [\mathbf{u}_{1,0}^l \cdots \mathbf{u}_{1,L-1}^l], \\ \boldsymbol{\Theta}_{i-1}^r &= [\alpha_{1,0}^{i-1} \cdots \alpha_{1,L-1}^{i-1}]^T b_1^{i-1}, \; \boldsymbol{\Theta}_i^l = [\alpha_{1,0}^i \cdots \alpha_{1,L-1}^i]^T b_1^i, \\ \mathbf{G}^r &= [\mathbf{u}_{2,0}^r \cdots \mathbf{u}_{2,L-1}^r \cdots \mathbf{u}_{K,0}^r \cdots \mathbf{u}_{K,L-1}^r], \\ \mathbf{G}^l &= [\mathbf{u}_{2,0}^l \cdots \mathbf{u}_{2,L-1}^l \cdots \mathbf{u}_{K,0}^l \cdots \mathbf{u}_{K,L-1}^l], \\ \boldsymbol{\Psi}_{i-1}^r &= [\alpha_{2,0}^{i-1} b_2^{i-1} \cdot \alpha_{2,L-1}^{i-1} b_2^{i-1} \cdot \alpha_{K,0}^{i-1} b_K^{i-1} \cdot \alpha_{K,L-1}^{i-1} b_K^{i-1}]^T, \\ \boldsymbol{\Psi}_i^l &= [\alpha_{2,0}^i b_2^i \cdots \alpha_{2,L-1}^i b_2^i \cdots \alpha_{K,0}^i b_K^i \cdots \alpha_{K,L-1}^i b_K^i]^T. \end{aligned}$$

More than one observation interval is required for decision of symbols in asynchronous channels, since the transmission delays shift the spreading codes to the next symbol intervals. Considering one more consecutive observation vector $\mathbf{r}_{i+1}$, we can define a new observation vector set $\mathbf{z}_i = [\mathbf{r}_i^T \; \mathbf{r}_{i+1}^T]^T$, that is

$$\mathbf{z}_i = \mathbf{S}\boldsymbol{\Theta}_z^i + \mathbf{G}\boldsymbol{\Psi}_z^i + \mathbf{n}_z^i, \tag{6}$$

where $< \mathbf{S} >$ and $< \mathbf{G} >$ denote the subspaces for the desired signal and interferences respectively, which are given by

$$\mathbf{S} = \begin{bmatrix} \mathbf{S}^r \; \mathbf{S}^l \; \mathbf{0} \\ \mathbf{0} \; \mathbf{S}^r \; \mathbf{S}^l \end{bmatrix}, \quad \mathbf{G} = \begin{bmatrix} \mathbf{G}^r \; \mathbf{G}^l \; \mathbf{0} \\ \mathbf{0} \; \mathbf{G}^r \; \mathbf{G}^l \end{bmatrix};$$

$\boldsymbol{\Theta}_z^i = [(\boldsymbol{\Theta}_{i-1}^r)^T \; \boldsymbol{\Theta}_i^T \; (\boldsymbol{\Theta}_{i+1}^l)^T]^T$, $\boldsymbol{\Psi}_z^i = [(\boldsymbol{\Psi}_{i-1}^r)^T \; \boldsymbol{\Psi}_i^T \; (\boldsymbol{\Theta}_{i+1}^l)^T]^T$ are $3L \times 1$ parameter vectors of each space to be estimated; $\mathbf{n}_z^i$ denotes the noise vector for two consecutive observation interval. Recall that perfect synchronization is required to build the subspaces.

The ML solution of $\hat{\boldsymbol{\Theta}}_z^i$ is given by [1]

$$\hat{\boldsymbol{\Theta}}_z^i = (\mathbf{S}^T \mathbf{P}_G^{\perp} \mathbf{S})^{-1} \mathbf{S}^T \mathbf{P}_G^{\perp} \mathbf{z}_i. \tag{7}$$

Note that $\mathbf{S}\hat{\boldsymbol{\Theta}}_z^i = \mathbf{L}_S \mathbf{z}_i$, which is the oblique projection of the received signal vector $\mathbf{z}_i$ onto the desired signal subspace $< \mathbf{S} >$. It is easily seen that $\hat{\boldsymbol{\Theta}}_z^i$ has Gaussian distribution such that

$$\hat{\boldsymbol{\Theta}}_z^i \sim N(\boldsymbol{\Theta}_z^i, \sigma^2 (\mathbf{S}^T \mathbf{P}_G^{\perp} \mathbf{S})^{-1}). \tag{8}$$

Hence, the parameter vector of interest, $\hat{\boldsymbol{\Theta}}_i$, has the distribution given by

$$\begin{aligned} \hat{\boldsymbol{\Theta}}_i &= [\alpha_{1,0}^i \cdots \alpha_{1,L-1}^i]^T b_1^i \triangleq \alpha_1 b_1^i \\ &\sim N(\boldsymbol{\Theta}_i, \sigma^2 (\mathbf{S}^T \mathbf{P}_G^{\perp} \mathbf{S})_{[L+1]}^{\dagger}), \end{aligned} \tag{9}$$

where $\mathbf{A}_j^{\dagger}$ denotes the $L \times L$ square block matrix which begins with the $(j,j)_{th}$ element of the matrix $\mathbf{A}^{-1} \in R^{3L \times 3L}$.

Let $(\mathbf{S}^T \mathbf{P}_G^{\perp} \mathbf{S})_{[L+1]}^{\dagger} \triangleq \kappa^H \kappa$. Consider that the noise components in L branches of the desired user are correlated. To decorrelate the noise, a whitening filter $(\kappa^H)^{-1}$ obtained by Cholesky decomposition prior to combining is used. The output of the whitening filter within a symbol interval is given by

$$\mathbf{y} = (\kappa^H)^{-1} \alpha_1 b_1^i + \mathbf{n}_y, \tag{10}$$

where $\mathbf{n}_y$ is zero mean complex Gaussian white noise vector with covariance matrix $\sigma^2 \mathbf{I}_L$. Then, the maximal ratio combiner, $\mathbf{q} \triangleq \alpha_1^H \kappa^{-1}$, operates on the output of the whitening filter. It can be easily verified that the conditional detection error probability, given the knowledge of channel gains, is given by

$$P_{e|\mathbf{q}} = Q\left(\frac{A_1}{\sigma}\sqrt{\mathbf{g}_1^H (\kappa^H \kappa)^{-1} \mathbf{g}_1}\right), \tag{11}$$

where $\mathbf{g}_1$ denotes the channel gain vector of L paths of user 1, that is $\mathbf{g}_1 \triangleq [g_{1,0} \cdots g_{1,L-1}]^T$. We can conclude that this detection approach is near-far resistant since the error probability does not depend on other interfering users' signal powers. Note that the subspace detector requires only the desired user's channel gains as the multipath decorrelator does.

4. EFFECTS OF DELAY ESTIMATION ERRORS

The asynchronous subspace detector presented above is based on the assumption of perfect synchronization of all users' signals so that complete subspaces for the desired signal and interferences can be obtained. Perfect rejection of interferences is undoubtedly achieved since the range of null steering operator is orthogonal to all interferences. In practice, it is difficult to have complete subspaces since delay estimation is always subject to some amount of errors, which may cause severe perturbation of the subspaces. The null-steering operator based on the perturbed interference subspace cannot reject the interferences effectively in this situation.

Consider the perturbation of the interference subspace due to delay estimation errors, which is given by $\mathbf{G}_\Delta = \mathbf{G} + \boldsymbol{\Delta}$, where $\boldsymbol{\Delta}$ is the perturbation matrix. It is specifically assumed that the perturbation is mainly due to the estimation errors for interchip delays. The estimated interchip delay $\hat{\delta}_k$ can be represented as a sum of the real transmission delay δ_k and the error γ_k such that $\hat{\delta}_k \in [0, T_c]$. The ML solution of $\hat{\boldsymbol{\Theta}}_z^i$ is similarly obtained by

$$\hat{\boldsymbol{\Theta}}_z^i = (\mathbf{S}^T \mathbf{P}_{G_\Delta}^\perp \mathbf{S})^{-1} \mathbf{S}^T \mathbf{P}_{G_\Delta}^\perp \mathbf{z}_i, \tag{12}$$

and $\hat{\boldsymbol{\Theta}}_z^i$ has Gaussian distribution such that

$$\hat{\boldsymbol{\Theta}}_z^i \sim N(\boldsymbol{\Theta}_z^i + \mathbf{T}, \sigma^2 (\mathbf{S}^T \mathbf{P}_{G_\Delta}^\perp \mathbf{S})^{-1}), \tag{13}$$

where $\mathbf{T} = (\mathbf{S}^T \mathbf{P}_{G_\Delta}^\perp \mathbf{S})^{-1} \mathbf{S}^T \mathbf{P}_{G_\Delta}^\perp \{\mathbf{G}\boldsymbol{\Psi}\}$. The perturbation vector $\mathbf{T}$ results from the fact that the range of the perturbed interference subspace $< \mathbf{G}_\Delta >$ may not include the real space $< \mathbf{G} >$ completely. Hence, this term is strongly affected by the principal angle of two unmatched spaces, i.e., the amount of delay estimation errors, and MAIs. The distribution of the parameter of interest $\hat{\boldsymbol{\Theta}}_i$ is given by

$$\hat{\boldsymbol{\Theta}}_i = \alpha_1 b_1^i + \mathbf{T}_\Theta \sim N(\boldsymbol{\Theta}_i + \mathbf{T}_\Theta, \sigma^2 (\mathbf{S}^T \mathbf{P}_{G_\Delta}^\perp \mathbf{S})_{[L+1]}^\dagger), \tag{14}$$

where $\mathbf{T}_\Theta$ is the $[L+1 : 2L]_{th}$ elements of the vector $\mathbf{T}$. Define $(\mathbf{S}^T \mathbf{P}_{G_\Delta}^\perp \mathbf{S})_{[L+1]}^\dagger$ as $\kappa_\Delta^H \kappa_\Delta$. After employing whitening filter $(\kappa_\Delta^H)^{-1}$ and then maximal ratio combining filter, $\mathbf{q}_\Delta = \alpha_1^H \kappa_\Delta^{-1}$, to the obliquely projected signal, we can verify that the conditional detection probability given channel gains is obtained by

$$P_{e|\mathbf{q}_\Delta} = Q\left(\frac{A_1}{\sigma} \sqrt{\mathbf{g}_1^H (\kappa_\Delta^H \kappa_\Delta)^{-1} \mathbf{g}_1} + \frac{1}{\sigma} \frac{\mathbf{g}_1^H (\kappa_\Delta^H \kappa_\Delta)^{-1} \mathbf{T}_\Theta}{\sqrt{\mathbf{g}_1^H (\kappa_\Delta^H \kappa_\Delta)^{-1} \mathbf{g}_1}}\right). \tag{15}$$

The first term in the argument of Q function is similar in form with (11), while the second one represents the performance degradation of the subspace detector by the delay estimation errors and the powers of the interferences.

5. ROBUST DETECTION BY OBLIQUE PROJECTION

Recall that chip matched filtering outputs can be represented as a convex combination of two consecutive chip delayed spreading codes. One idea for complete rejection of the interferences in presence of delay estimation errors is to build an interference subspace to cover the perturbation due to the errors. Consider the following matrices as the substitutes for $\mathbf{G}^r$, $\mathbf{G}^l$, which are given by

$$\begin{aligned} \mathbf{G}_M^r &= [\mathbf{c}_2^r(p_{2,0}) \cdots \mathbf{c}_2^r(p_{2,L}) \cdots \mathbf{c}_K^r(p_{K,0}) \cdots \mathbf{c}_K^r(p_{K,L})] \\ \mathbf{G}_M^l &= [\mathbf{c}_2^l(p_{2,0}) \cdots \mathbf{c}_2^l(p_{2,L}) \cdots \mathbf{c}_K^l(p_{K,0}) \cdots \mathbf{c}_K^l(p_{K,L})] \end{aligned} \tag{16}$$

Note that all interfering signals can be represented as a linear combination of specific two bases in these matrices. Then, we have a new interference subspace given by

$$\mathbf{G}_M = \begin{bmatrix} \mathbf{G}_M^r & \mathbf{G}_M^l & \mathbf{0} \\ \mathbf{0} & \mathbf{G}_M^r & \mathbf{G}_M^l \end{bmatrix}. \tag{17}$$

With the modified interference subspace, the ML solution is obtained as

$$\hat{\boldsymbol{\Theta}}_z^i = (\mathbf{S}^T \mathbf{P}_{G_M}^\perp \mathbf{S})^{-1} \mathbf{S}^T \mathbf{P}_{G_M}^\perp \mathbf{z}_i, \tag{18}$$

It is clear that $Range(\mathbf{G}_M) \supseteq Range(\mathbf{G})$, which means the perturbation term in (13) can be completely rejected from the fact that $\mathbf{P}_{G_M}^\perp \{\mathbf{G}\boldsymbol{\Phi}\} = \mathbf{0}$. After the same whitening and combining steps, the conditional error probability given the channel gains of the desired user is now represented as

$$P_{e|\mathbf{q}_M} = Q\left(\frac{A_1}{\sigma} \sqrt{\mathbf{g}_1^H (\kappa_M^H \kappa_M)^{-1} \mathbf{g}_1}\right), \tag{19}$$

where $(\mathbf{S}^T \mathbf{P}_{G_M}^\perp \mathbf{S})_{[L+1]}^\dagger = \kappa_M^H \kappa$, and $\mathbf{q}_M = \alpha_1^H \kappa_M^{-1}$. The robust detector preserves the near-far resistant characteristics of the asynchronous subspace detector.

The perturbation of the desired signal space due to the delay estimation error, say $\mathbf{S}_\Delta$, has not been considered yet. In this situation, the estimation of the parameter of interest, $\hat{\boldsymbol{\Theta}}$, may not be performed exactly. Consider the expectation value of $\hat{\boldsymbol{\Theta}}_z$ given by

$$E(\hat{\boldsymbol{\Theta}}_z) = (\mathbf{S}_\Delta^T \mathbf{P}_{G_M}^\perp \mathbf{S}_\Delta)^{-1} \mathbf{S}_\Delta^T \mathbf{P}_{G_M}^\perp \{\mathbf{S}\boldsymbol{\Theta}_z\}. \tag{20}$$

The performance is mainly decided by the level of the principal angle between the spaces $< \mathbf{S}_\Delta >$ and $< \mathbf{S} >$. More exact estimation of the desired user's transmission delay is required for better performance. As an example, it has been demonstrated that the MUSIC algorithm produces a good delay estimation in fading channels [8]. Therefore, a combining approach of the robust detector and MUSIC may be a good realization in practice.

6. NUMERICAL RESULTS

We have performed experiments on the bit error rate performance of the asynchronous CDMA subspace detector based on oblique projection and its robust version in frequency-selective Rayleigh fading channels. In our simulated system, there are three active users with 31 chips per bit Gold codes, and three paths are assigned to each user. The fading on all multipaths is normalized to one, that is, $E[|g_{k,l}|^2] = 1$ for $k = 1, \cdots, K$ and $l = 0, \cdots, L-1$. All interfering users are assumed to have equal energy. Each Monte-Carlo run represents a particular realization of the noise, data sequence and channel coefficients.

Fig. 1 and Fig. 2 show the performance comparison of the asynchronous subspace detector with the conventional RAKE receiver and multipath decorrelator when MAIs are equal to 0dB and 40dB respectively. Random setting of transmission delays were assumed to each run. The saturation effect of the RAKE system is observed in Fig. 1, while others definitely outperform the RAKE system as the SNR goes higher. In Fig. 2, the subspace detector shows a near-far resistant nature as we expected, and it has almost the same performance with the multipath decorrelator.

The sensitivity of the subspace detector in the presence of delay estimation errors is shown in Fig. 3. The perturbation level is defined by $|\hat{\tau}_k - \tau_k|$, which is the maximum estimation error distance. The result shows that the performance of the subspace detector is no longer reliable, as the perturbation level and MAIs increase even under high SNR situation. It is an interesting observation that the subspace detector preserves the performance level under perfect power control.

In Fig. 4, the performance of the suggested robust detector for the perturbation by delay estimation errors is presented. $\hat{\tau}_k$ has a uniform distribution between $\tau_k - P_t$ and $\tau_k + P_t$, and it was randomly generated in each run. The result shows reliable performance even when P_t is half-chip interval and MAIs are equal to 40dB. Moreover, we need to note that the robust detector has almost the same performance as the perfect synchronization case within $P_t = 0.1$ chips which is in the range of the delay estimation algorithms like MUSIC in fading channels.

7. CONCLUSION

We have suggested an asynchronous subspace detector based on oblique projection as the fundamental framework of a robust detector in CDMA frequency-selective Rayleigh fading channels. This detector shows a near-far resistant nature and has similar performance as the multipath decorrelating receiver. It was, however, shown that even a small transmission delay estimation error may cause severe performance degradation of the subspace detector as it does to other near-far resistant multiuser receivers. In this paper, we have developed a multiuser receiver which is robust to the errors by employing a new interference subspace which can cover the perturbation due to the estimation errors. It was shown that the robust detector shows reliable performance even when the perturbation and MAI are relatively large.

8. REFERENCES

[1] R.T. Behrens, L.L. Scharf, "Signal Processing Applications of Oblique Projection Operators", IEEE Trans. on Signal Processing, Vol. 42, No. 6, June 1994.

[2] H.C. Huang, S.C. Schwartz, "A Comparative Analysis of Linear Multiuser Detectors for Fading Multipath Channels", in Proceedings of Globecom 1994, pp. 11-15.

[3] R. Rupas, S. Verdu, "Near-Far Resistance of Multiuser Detectors in Asynchronous Channels", IEEE Trans. on Communications, Vol. 38, No. 4, Apr. 1990.

[4] Y. Moon, M. Kaveh, L.B. Nelson, "Subspace-Based Detection for CDMA Communications", in ICASSP'98, pp. 3201-3204.

[5] Y. Moon, M. Kaveh, L.B. Nelson, "A New Subspace-based Approach for Multiuser Detection in DS-CDMA", Submitted to IEEE Trans. on Vehicular Technology.

[6] J. Proakis, *Digital Communications*, McGraw Hill, 3rd Ed., 1995.

[7] E.G. Strom, S. Parkvall, S.L. Miller, B.E. Ottersten, "Sensitivity Analysis of Near-Far Resistant DS-CDMA receivers to Propagation Delay Estimation Errors", in Proceedings of IEEE Conference on Vehicular Technology, 1994.

[8] E.G. Strom, S. Parkvall, S.L. Miller, B.E. Ottersten, "Propagation Delay Estimation of DS-CDMA Signals in a Fading Environment", in Proceedings of GLOBECOM 1994.

[9] Z. Zvonar, D. Brady, "Optimum Detection in Asynchronous Multiple-Access Multipath Rayleigh Fading Channels", in Proceedings of the 26th Annual Conference on Information Sciences and Systems, Princeton University, pp. 826-831, 1992.

[10] Z. Zvonar, D. Brady, "Coherent and Differentially Coherent Multiuser Detectors for Asynchronous CDMA Frequency-Selective Channels", in Proceedings of Milcom 1992, pp. 17.6.1-17.6.5.

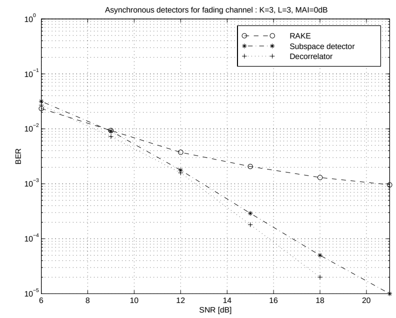

Figure 1: Performance comparison of the asynchronous detectors in a fading channel under perfect power control.

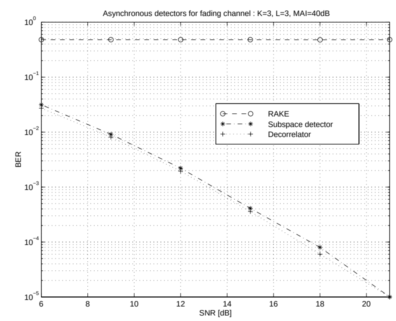

Figure 2: Performance comparison of the asynchronous detectors in a fading channel when MAIs are equal to 40dB.

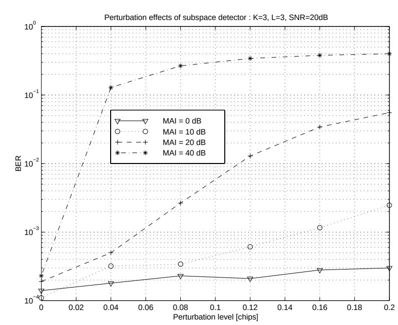

Figure 3: Sensitivity of the subsapce detector to the perturbation of delay estimation errors.

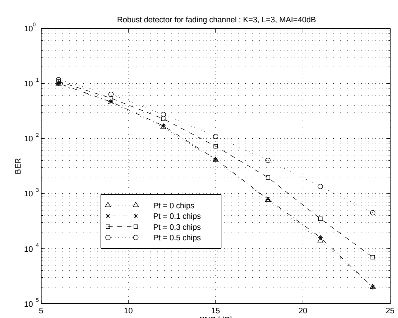

Figure 4: Performance of the robust detector to the perturbation of delay estimation errors.