Design of an Efficient RAKE Receiver Architecture for Multiuser Detection with Adaptive Channel Estimation

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ABSTRACT -

In this paper, we propose an efficient implementation scheme of wideband version of parallel interference cancellation (*PIC*) called regenerative *PIC* for multiuser detector with Rayleigh fading multipath channel and multiple access interference (*MAI*). Multiuser detection shall be used in the receiver. The successive *PIC* scheme cancels the interference of all users simultaneously. The multistage *PIC* suppresses the multiple access interference in multiple consecutive steps. Stage *n* cancels the interference by utilizing the hard symbol decision from stage *n*-1. The cancellation operation itself requires knowledge of the user codes, the relative code phases, complex estimates for the multipath taps, and finally hard symbol decisions for each user.

1. Introduction

Channel coding is an essential part of wideband CDMA (*W-CDMA*) air interfaces. Multiuser detection (*MUD*) can be either be partitioned or integrated. Partioned approaches treat multiuser interference equalization problem and decoupling separately. Integrated approaches perform the equalization and decoding operations together [1]. Among the many multiuser detectors, interference cancellation (*IC*) techniques are relatively computationally efficient and can be used in systems utilizing aperiodic spreading sequences (i.e., spreading sequences with periods longer than the symbol interval) without incurring additional complexity. Basically, there are two IC architectures, both of them have a multistage structure: the parallel interference canceler (*PIC*) (e.g., [2]) and the successive interference canceler (*SIC*) ([3]-[4]). In each stage, the *PIC* simultaneously regenerates and cancels from each user the *MAI* due to the other users based on the detected symbols in the preceding stage. The *SIC* sequentially removes the *MAI* due to the stronger users before detecting the weaker one (i.e., only one user is detected and canceled in a stage); as a result, the near-far problem is less pronounced in the *SIC* than in the conventional Matching Filter *(MF)* detector.

In order to regenerate and cancel the *MAI*, accurate parameter estimation is necessary. The conventional *SIC* (CSIC) employs the magnitude of the output of the *MF* bank as the received amplitude estimate. An adaptive multiuser decorrelator, which minimizes a least-squares cost function of the received signal and its estimate, in combination with an IC structure, was proposed in [5] to jointly estimate BPSK-modulated data and the user amplitudes. However, this method cannot be applied is a straightforward manner to systems with higher dimensional modulation formats in fading channels. More recently, decision-driven amplitude estimates have been used in conjunction with the PIC structure [6].

Partitioned approaches can be further divided into hard and soft decision approaches. In the soft decision approach a multiuser detection algorithm feeds the soft symbols after detection into a channel decoder. Third generation *W-CDMA* systems will have multiple data rates. Recently, multiuser detectors for multirate *CDMA* systems have been proposed in [9-11]. Another linear receiver is the linear minimum mean square error (*LMMSE*) [13], which, unlike the decorrelator, doesn't enhance the noise. Even though *LMMSE* has also been proposed for centralized receivers in *AWGN* and known fading channels, the main interest has been focused on the architecture of implementation for multiuser detection/ interference cancellation. The state-of-the-art digital *DSP* and *ASIC* technologies already provide an opportunity for building some kind of interference cancellation unit, although optimal implementation (*MLSE*) is still not too far.

2. Preliminaries

A. System Model and MUD Algorithm Formulation

The received signal is time-discretized by anti-alias filtering and sampling at the rate $T_s^{-1} = \frac{SG}{T}$, where S is the number of samples per chip. If the received signal is processed in blocks of N_b symbols, the received discrete-time signal can be written as

$$\mathbf{r} = \sum_{k=1}^{K} \mathbf{S}_k \mathbf{C} \mathbf{A}_k \mathbf{b}_k + \mathbf{n} \quad \in \mathbb{C}^{SGN_b}$$

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(1)

where $\mathbf{r}=[\mathbf{r}^{T(0)},\cdots,\mathbf{r}^{T(N_b-1)}]^T$ is the input sample vector with $\mathbf{r}^{T(n)}=[r(T_s(nSG+1)),\cdots,r(T_s(n+1)SG)]$, $\mathbf{S}=[\mathbf{S}^{(0)},\cdots\mathbf{S}^{(N_b-1)}]\in\Re^{SGN_b\times KLN_b}$ is the spreading code matrix with $\mathbf{S}^{(n)}=[\mathbf{s}_{1,1}^{(n)},\cdots,\mathbf{s}_{1,L}^{(n)},\mathbf{s}_{K,L}^{(n)}]$, where $\mathbf{s}_{k,l}^{(n)}=[\mathbf{0}^T_{(nSG+\tau_{k,l})\times 1},\mathbf{s}_k^T,\mathbf{0}^T_{((N_b-n-1)SG-\tau_{k,l})\times 1}]$, where $\tau_{k,l}$ is a time-discretized delay in samples ($\tau_{1,1}=0$) and $\mathbf{s}_k=[s_k(T_s),\cdots,s_k(T_sSG)]^T$ is the sampled signature sequence of the kth user, $\mathbf{C}=diag[\mathbf{C}^{(0)},\cdots,\mathbf{C}^{(N_b-1)}]\in\mathrm{C}^{KLN_b\times KN_b}$ is the channel coefficient matrix with $\mathbf{C}^{(n)}=diag[\mathbf{c}_1^{(n)},\cdots,\mathbf{c}_K^{(n)}]\in\mathrm{C}^{KL\times K}$, $\mathbf{c}_k^{(n)}=[c_{k,1}^{(n)},\cdots,c_{k,l}^{(n)}]^T\in\mathrm{C}^L$, $\mathbf{A}=diag[\mathbf{A}^{T(0)},\cdots,\mathbf{A}^{T(N_b-1)}]=\mathbf{I}_{N_b}\otimes\mathbf{A}^{(n)}\in\Re^{KN_b\times KN_b}$ is the matrix of the total received energies with $\mathbf{A}^{(n)}=diag[A_1,\cdots,A_K]$, $\mathbf{b}=[b^{T(0)},\cdots;b^{T(N_b-1)}]^T\in\{-1,1\}^{KN_b}$ is the data vector with $\mathbf{b}^{(n)}=[b_1^{(n)},\cdots,b_K^{(n)}]$, and $\mathbf{n}\in\mathrm{C}^{SGN_b}$ is the channel noise vector. For the synchronous downlink, the channel for all users is the same, i.e., $\mathbf{c}_k=\mathbf{c}\quad\forall k\in\{1,\cdots,K\}$.

B. Precombining and Postcombining LMMSE Receiver

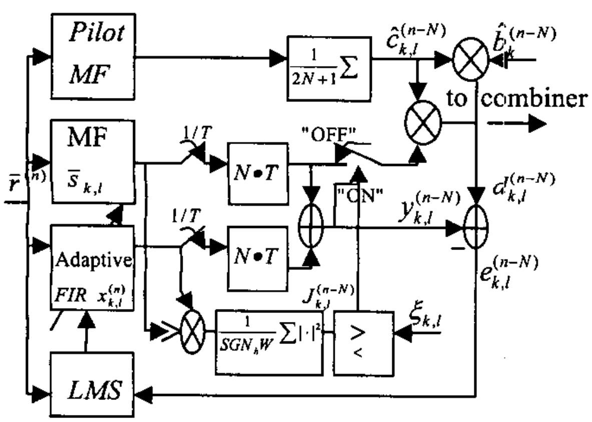

Fig.1. System Model of adaptive LMMSE-Rake Receiver and Its Implementation of Parallel IC (PIC) Scheme

The received signal is first passed through the matched filter bank. The standard approach in deriving *LMMSE* receiver is to minimize the cost function $\mathbf{E}\{|\mathbf{b}-\hat{\mathbf{b}}|^2\}$ element wise, where $\hat{\mathbf{b}}=\mathbf{L}^{\mathbf{H}}\mathbf{r}$ and $|\mathbf{x}|$ is the vector of absolute values of the vector $\mathbf{x}$. It is easy to show [14, p.391] that the standard *LMMSE* receiver can be written as

$$\mathbf{L}=\mathbf{SCA}(\mathbf{AC^HRCA}+\sigma^2\mathbf{I})^{-1}\in\mathrm{C}^{SGN_b\times KN_b},\tag{2}$$

where $\mathbf{R}=\mathbf{S^HS}\in\mathrm{R}^{KLN_b\times KLN_b}$ is the signature sequence cross-correlation matrix. The output of the postcombining LMMSE receiver is

$$\mathbf{y}_{[\mathbf{L}]}=(\mathbf{AC^HRCA}+\sigma^2\mathbf{I})^{-1}(\mathbf{SCA})^{\mathbf{H}}\mathbf{r}\quad\in\mathrm{C}^K\tag{3}$$

where $(\mathbf{SCA})^{\mathbf{H}}\mathbf{r}$ is the multipath combined matched filter bank output. Thus, the postcombining *LMMSE* detector clearly leads to a postcombining interference suppression type of receiver. If the channel is a nonfading *AWGN* channel, i.e., all nonzero elements in C are equal to one, the detector can be written in the standard form [12]. It can be seen from (2) that the standard *LMMSE* receiver in fading channel depends on the complex coefficients of all users and paths. Thus, the adaptive versions of the *LMMSE* receivers have increasing convergence problems as the fading rate increases. To circumvent such adaptation problems, one could minimize each element of $\mathbf{E}\{|\mathbf{b}-\hat{\mathbf{b}}|^2\}$, where $\mathbf{h}=\mathbf{CAb}$ and $\hat{\mathbf{h}}=\mathbf{M^Hr}$ This leads to the modified *LMMSE* receiver:

$$\mathbf{M}=\mathbf{S}(\mathbf{R}+\sigma^2\Sigma_h^{-1})^{-1}\in\mathrm{R}^{SGN_b\times KLN_b},\tag{4}$$

where $\Sigma_\mathbf{h}=diag[A_1^2\Sigma_{\mathbf{c}_1},...,A_K^2\Sigma_{\mathbf{c}_K}]\in\mathrm{R}^{\mathbf{KLN_b\times KLN_b}}$ is a diagonal matrix and it consists of transmitted user power and the average channel tap powers, with $\Sigma_{\mathbf{c}_K}=diag[\mathbf{E}[|\mathbf{c}_{\mathbf{K},1}|^2],\cdots,\mathbf{E}[|\mathbf{c}_{\mathbf{K},\mathbf{L}}|^2]\in\mathrm{R}^{L\times L}$ where $E[|c_{k,l}|^2]$ is the average energy of the kth user lth propagation path. As it can be seen from (4), the modified receiver no longer depends on the instantaneous values of complex channel coefficients but on the average power profiles of the channels. The adaptation requirements are now much milder and the receiver can be made adaptive even in the fading channels. The output of the precombining *LMMSE* receiver is

$$\mathbf{y}_{[\mathbf{M}]}=(\mathbf{R}+\sigma^2\Sigma_\mathbf{h}^{-1})\mathbf{S^Tr}\quad\in\mathrm{C}^{KL}\tag{5}$$

where $\mathbf{S^Tr}$ is the matched filter bank output vector without multipath combining.

The decision variable of the precombining *LMMSE* receiver or conventional *RAKE* receiver (M = S) for user k after maximal ratio combining can be expressed in the form

$$y_{[\mathbf{M},MRC]k}^{(n)}=\mathbf{c}_k^{\mathbf{H}(n)}\mathbf{y}_{[\mathbf{M}]k}^{(n)}\tag{6}$$

where $c_k^{(n)}=[c_{k,1}^{(n)},\cdots,c_{k,L}^{(n)}]^T\in\mathrm{C}^L$ is the combining vector and

$$\mathbf{y}_{[\mathbf{M}]}^{(n)}=\mathbf{M}^T\bar{\mathbf{r}}^{(n)}=[\mathbf{y}_{[\mathbf{M}]1}^{T(n)},\cdots,\mathbf{y}_{[\mathbf{M}]K}^{T(n)}]^T\in\mathrm{C}^L$$

includes the receiver output vector for user k.

Let

$Q = \frac{1}{2}\begin{pmatrix} 0_L & I_L \\ I_L & 0_L \end{pmatrix} \in \left\{0, \ \frac{1}{2}\right\}^{2L\times 2L}$ and $v = [\mathbf{c}_k^{T(n)}, \mathbf{y}_{[\mathbf{M}]k}^{T(n)}]^T \in \mathbb{C}^{2L}$.

By rewriting (6), the decision variable can be expressed in the form

$$y_{[\mathbf{M},MRC]}^{(n)} = v^{\mathbf{H}} Q v. \tag{7}$$

The receiver output vector $\mathbf{y}_{[\mathbf{M}]k}^{(n)}$ conditioned on the data vector $\overline{\mathbf{b}}^{(n)}$ (where $\overline{\mathbf{b}}^{(n)} = [\overline{\mathbf{b}}_1^{T(n)}, \cdots, \overline{\mathbf{b}}_K^{T(n)}]^T$, and $\overline{\mathbf{b}}_k^{T(n)} = [b_k^{(n-D)}, \cdots; b_k^{(n)}, \cdots; b_k^{(n+D)}]^T$ is a complex Gaussian random vector). Since the weight vector $c_k^{(n)}$ is also Gaussian, the probability of bit error for user k conditioned on $\overline{\mathbf{b}}^{(n)}$ can be expressed in the case of the *BPSK* modulation as [15].

$$P_r\{error \mid \overline{\mathbf{b}}^{(n)}\} = \sum_{\substack{i=1 \\ \lambda_i < 0}}^{2L} \prod_{\substack{j=1 \\ j \neq i}}^{2L} \frac{1}{1 - \frac{\lambda_j}{\lambda_i}} \tag{8}$$

where $\lambda_i, i = 1, \cdots, 2L$ are the eigenvalues of the matrix $\Sigma_{v|\overline{b}^{(n)}} Q$ and

$$\Sigma_v \Big| \overline{\mathbf{b}}^{(n)} = \begin{pmatrix} \Sigma_{\mathbf{c}_k^{(n)}} & \Sigma_{\mathbf{c}_k^{(n)}, \mathbf{y}_{[\mathbf{M}]k}^{(n)} | \overline{b}^{(n)}} \\ \Sigma^{\mathbf{H}}_{\mathbf{c}_k^{(n)}, \mathbf{y}_{[\mathbf{M}]k}^{(n)} | \overline{b}^{(n)}} & \Sigma_{\mathbf{y}_{[\mathbf{M}]k}^{(n)} | \overline{\mathbf{b}}^{(n)}} \end{pmatrix} \tag{9}$$

is the covariance matrix of the vector v. The covariance matrix (9) depends on the channel estimation filter. Finally, the BEP for the kth user is expressed as

$$P_k = \frac{1}{2^{MK-1}} \sum_{\substack{\overline{\mathbf{b}}^{(n)} \in \{-1,1\}^{MK-1} \\ b_k = 1}} P_r\{error \mid \overline{\mathbf{b}}^{(n)}\}. \tag{10}$$

According to the optimal Maximum Likelihood Soft Decision (*MLSD*) receiver derived by Kailath [16], the maximum likelihood sequence estimate makes its decision as

$$\hat{\mathbf{b}}_{[MLSD]} = \arg \max_{\overline{\mathbf{b}} \in \Xi^{NK}} P_r(\mathbf{y} \mid \overline{\mathbf{b}}) \tag{11}$$

where $P_r(\mathbf{y} \mid \overline{\mathbf{b}})$ is the probability density function (pdf) of the MF bank output vector conditioned on the data vector $\overline{\mathbf{b}}$. Assume the covariance matrix $\Sigma_{\mathbf{C}}$ of the channel coefficient vector $\mathbf{c}$ known in the detector derivation. From the result of Juntti and Latva-aho [9], the *MLSD* rule can be expressed as

$$\hat{\mathbf{b}}_{[MLSD]} = \arg \max_{\overline{\mathbf{b}} \in \Xi^{NK}} \mathbf{y}^{\mathbf{H}} \Sigma_{\mathbf{y}|\overline{\mathbf{b}}}^{-1} \mathbf{y}. \tag{12}$$

By applying the matrix inversion lemma [14, p.571] to (12), the *MLSD* rule (12) becomes

$$\hat{\mathbf{b}}_{[MSLD]} = \arg \max_{\overline{\mathbf{b}} \in \Xi^{NK}} \hat{\mathbf{h}}^{\mathbf{H}}_{[MMSE]} \mathbf{y} \tag{13}$$

where

$$\hat{\mathbf{h}}_{[MMSE]} = \left(\mathbf{R} + \sigma^2 \Sigma_{\mathbf{h}|\overline{\mathbf{b}}}^{-1}\right) \mathbf{y} \tag{14}$$

is the *MMSE* estimate of the vector $\boldsymbol{h}$ ([9], [14]) conditioned on the data. In other words, the optimal *MLSD* receiver estimates the received noiseless signal, and multiplies (correlates) the matched filter bank output with the estimated complex amplitude vector.

C. Suboptimal PIC Receivers and Delay Tracking

The decision variable of the precombining *LMMSE* receiver or conventional *RAKE* receiver (M = S) for user k after maximal ratio combining can be expressed in the form

$$\begin{aligned} \mathbf{y}_{[\mathbf{M}]k}^{(n)} &= \mathbf{M}\overline{\mathbf{r}}^{(n)} = [y_{[M]1}^{T(n)}, \cdots, y_{[M]k}^{T(n)}]^T \\ &= S^H S A b + S^H n \\ &= \mathbf{RAb} + \tilde{\mathbf{n}} = (\mathbf{R} - \mathbf{I}_{KLN_b} + \mathbf{I}_{KLN_b})\mathbf{Ab} + \tilde{\mathbf{n}} \\ &= \mathbf{Ab} + (\mathbf{R} - \mathbf{I}_{KLN_b})\mathbf{Ab} + \tilde{\mathbf{n}} \overset{\Delta}{=} \mathbf{Ab} + \Psi + \tilde{\mathbf{n}}. \end{aligned} \tag{15}$$

where $[\mathbf{R}]_{ii} = 1$, and

$\mathbf{R}^{(n,n)} = S^{T(n)}(0)S^{(n)}(0) + \cdots + S^{T(n)}(D)S^{(n)}(D)$,

$\mathbf{R}^{(n,n-D)} = S^{T(n)}(0)S^{(n-1)}(D)$, $\mathbf{R}^{(n-D,n)} = \mathbf{R}^{T(n,n-D)}$, an d $\mathbf{\Psi} = [\mathbf{\Psi}^{T(0)}, \cdots, \mathbf{\Psi}^{T(N_b-1)}]^T$ is (KLN_b) × 1 multiple access interference vector.

$$\mathbf{R} = \mathbf{S^H S} = \begin{bmatrix} \mathbf{R}^{(0,0)} & \cdots & \mathbf{R}^{(0,D)} & \mathbf{0}_{KL} & \mathbf{0}_{Kl} \\ \vdots & \ddots & \vdots & \ddots & \vdots \\ \mathbf{R}^{(D,0)} & \ddots & \ddots & \ddots & \mathbf{0}_{KL} \\ \mathbf{0}_{KL} & \ddots & \ddots & \ddots & \mathbf{R}^{(N_b-D,N_b-1)} \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ \mathbf{0}_{KL} & \cdots & \mathbf{0}_{KL} & \cdots & \mathbf{R}^{(N_b-1,N_b-1)} \end{bmatrix}.$$

The matched filter output during the nth symbol interval can be written as

$$\mathbf{y}^{(n)} = \mathbf{A}^{(n)}\mathbf{b}^{(n)} + \mathbf{\Psi}^{(n)} + \tilde{\mathbf{n}}^{(n)}, \tag{16}$$

with

$$\mathbf{\Psi}^{(n)} = \sum_{i=-D}^{D} (\mathbf{R}^{(n,n+i)} - \delta_{i,0}\mathbf{I}_{KL})\mathbf{A}^{(n+i)}\mathbf{b}^{(n+i)} + \tilde{\mathbf{n}}^{(n)}, \tag{17}$$

where $\delta_{i,0}$ is the Kronecker delta and $\boldsymbol{R}, \boldsymbol{A}, \boldsymbol{b}$ are all-zeros $\forall\, n$ or $n + i \notin \{0, 1, \cdots, N_b\}$.

A simple way to suppress *MAI* is to estimate and subtract it from matched filter outputs before synchronization and detection, which results in *PIC* based algorithms. The *PIC* based detectors and channel estimators require tentative data decisions to form the *MAI* estimates. Tentative decisions are obtained from the earlier stages of the multistage detector [2]. The data decisions in the case of *BPSK* modulation at the pth stage can be written as

$$\hat{\mathbf{b}}^{(n)}(p+1) = \mathrm{sgn}[\Re\{\mathbf{C}\hat{\mathbf{A}}^{\mathbf{H}(n)}(p)(\mathbf{z}^{(n)} - \hat{\mathbf{\Psi}}^{(n)}(p))\}] \tag{18}$$

where $\mathbf{C} = \mathbf{I}_K \otimes \mathbf{1}_L^T$ is the multipath combing matrix,

$\otimes$ denotes the Kronecker product, and

$$\mathbf{\Psi}^{(n)}(p)=\sum_{i=-D}^{D}(\mathbf{R}^{(n,n+i)}-\delta_{i,0}\mathbf{I})\hat{\mathbf{A}}^{(n+i)}(p)\hat{\mathbf{b}}^{(n+i)}(p), \quad (19)$$

where $\hat{\mathbf{A}}^{\mathrm{H}(n)}(p)$ denotes the channel coefficient estimates at the pth stage. It should be emphasized, that depending on the channel estimation and multistage schemes, the data estimates and channel estimates can be obtained either form the previous stage or following stages. The *PIC* based complex channel coefficient estimator can be written in general as

$$\widetilde{\mathbf{a}}^{(n)}=\hat{\mathbf{A}}^{\mathrm{H}(n)}(\mathbf{z}^{(n)}-\hat{\mathbf{\Psi}}^{(n)}), \quad (20)$$

where $\widetilde{\mathbf{A}}^{(n)}=diag[\widetilde{a}^{(n)}]$, $\widetilde{\mathbf{a}}^{(n)}=[a_{1,1}^{(n)},\cdots,a_{K,L}^{(n)}]^T$, are the rough channel estimates obtained from the canceler output by removing the effect of data modulation. Thus, the matched filter outputs of a multistage *PIC* receiver at the mth stage can be represented in the form [9]

$$\mathbf{y}_{(MUD)}^{(n)}(m)=\mathbf{y}^{(n)}-\hat{\mathbf{\Psi}}^{(n)}(m) \quad (21)$$

with

$$\mathbf{\Psi}^{(n)}(m)=\sum_{i=-P_{PIC}}^{P_{PIC}}(\mathbf{R}(-i)-\delta_{i,0}\mathbf{I}_{KL})\hat{\mathbf{C}}^{(n+i)}(m-1)\times\mathbf{A}^{(n+i)}\hat{\mathbf{b}}^{(n+i)}(m-1), \quad (22)$$

where $P_{PIC}=[(T+T_m)/T]$. The result is a decision-feedback *PIC* receiver of the form presented in [9].

The rough channel estimates can be enhanced by using adaptive channel estimation filters, which are adapted according to user energies and the rate of fading. An adaptive *LMS FIR* predictor will be used for updating the channel estimation filter weights. The output of the first order *FIR* predictor with filter weights $\omega_{k,l}^{(m)}$ and $q_{k,l}^{(n)}$ can be written by

$$\hat{\alpha}_{k,l}^{(n)}=\omega_{k,l}^{(n)}\widetilde{a}_{k,l}^{(n-2)}+q_{k,l}^{(n)}\hat{\alpha}_{k,l}^{(n-1)}, \quad (23)$$

where $\widetilde{a}_{k,l}^{(n)}$ is the rough channel estimate. The *LMS* algorithm for updating the coefficients $\omega_{k,l}$ and $q_{k,l}$ can be expressed as [17]

$$\begin{bmatrix}\omega_{k,l}^{(n+1)}\\ q_{k,l}^{(n+1)}\end{bmatrix}=\begin{bmatrix}\omega_{k,l}^{(n)}\\ q_{k,l}^{(n)}\end{bmatrix}+2\mu_{k,l}e_{k,l}^{(n)}\begin{bmatrix}\widetilde{a}_{k,l}^{*(n-2)}\\ \hat{\alpha}_{k,l}^{*(n-1)}\end{bmatrix}, \quad (24)$$

where $\mu_{k,l}$ is the small constant determining the convergence of the algorithm and $e_{k,l}^{(n)}=\widetilde{a}_{k,l}^{*(n-1)}-\hat{\alpha}_{k,l}^{(n)}$ is the predictor error signal (see Fig. 2). The filter weights should be constrained due to stability problems of adaptive *IIR* filters.

The data decisions and channel coefficient estimate from the last stage can be used in *MAI* estimation in the delay tracking algorithm by adding extra delay in front of the delay tracker. The resulting *MAI* estimator is a smoother for delay tracker. The *PCI* based delay estimator (Fig. 3) in multiuser time-discretized receiver for the kth user's path can be written as

$$\hat{d}_{k,l}^{(n+1)}=\arg\max_{\widetilde{d}_{k,l}}\left\{\Re\left\{\sum_{i=n-LF+1}^{n}\hat{\alpha}_{k,l}^{*(i)}\hat{b}_k^{*(i)}\left[z_{k,l}^{(i)}(\widetilde{d}_{k,l})-\hat{\Psi}_{k,l}^{(i)}(\widetilde{d}_{k,l})\right]\right\}\right\}. \quad (25)$$

where LF is the observation interval, $\hat{d}_{k,l}^{(n+1)}$ is the time-discretized delay estimate and $\hat{\Psi}_{k,l}^{(i)}(\widetilde{d}_{k,l})$ is the *MAI* estimate with the trial delay $\widetilde{d}_{k,l}$. The maximization problem (25) can be solved e.g. by using the delay-lock loop techniques. Especially so-called *SCCL* loops [18] lead to quite simple delay tracking algorithms. The signals at the output of the *LMS* prefilter can be written as

$$\hat{z}_{k,l}^{(i)}(\widetilde{d}_{k,l})=w_{k,l}^{(i)}\breve{z}_{k,l}^{(i)}(\widetilde{d}_{k,l})+q_{k,l}^{(i)}\breve{z}_{k,l}^{(i-1)}(\widetilde{d}_{k,l}), \quad (26)$$

where

$$\breve{z}_{k,l}^{(i)}(\widetilde{d}_{k,l})=\hat{b}_{k,l}^{*(i)}(z_{k,l}^{(i)}(\widetilde{d}_{k,l})-\hat{\Psi}_{k,l}^{(i)}(\widetilde{d}_{k,l})), \quad (27)$$

is the *MAI* free and weighted correlator output with trial delay $\widetilde{d}_{k,l}=\hat{d}_{k,l}^{(m')}+\Delta_u$, $\Delta_u\in\{-1,0,+1\}, m'=\left\lfloor\frac{m}{LF}\right\rfloor^1$, and $w_{k,l}^{(i)}, q_{k,l}^{(i)}$ are the prefilter weights (24). The delay estimate for the next LF symbols is

$$\hat{d}_{k,l}^{(m'+1)}=\arg\max_{\Delta_u}(\Re\{\hat{z}_{k,l}^{(m'+1)}(\hat{d}_{k,l}^{(m')}+\Delta_u)\}), \quad (28)$$

where

$$\hat{z}_{k,l}^{(m'+1)}=\frac{1}{LF}\sum_{i=m-LF=1}^{m}\widetilde{z}_{k,l}^{(i)}=\frac{1}{LF}\sum_{i=m-LF=1}^{m}\hat{\alpha}_{k,l}^{(i)}\breve{z}_{k,l}^{(i)}. \quad (29)$$

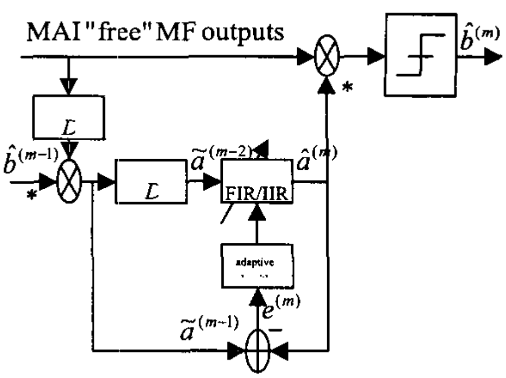

Fig.2 Structure of adaptive predictors

D. Switching of the adaptive receiver branch

The adaptive implementation of the *LMMSE-RAKE* requires the estimation of the channel coefficients and the tentative decisions from the conventional *RAKE* to adapt the finger weights. Due to the imperfect reference signal and channel estimation errors, the adaptive *LMMSE-RAKE* gives sometimes worse performance than the conventional *RAKE* receiver. The adaptive filter should be switched off during the periods when the conventional *RAKE* receiver would give better performance.

The simplest one from the receiver point of view is a network assisted scheme. Another approach is to use the short term mean square error as a quality factor. The short term *MSE* can be estimated as

$$J_{k,l}^{(n)} = \frac{1}{W} \sum_{i=n-W-1}^{n} | e^{(i)} |^2, \qquad (30)$$

where W is the summation interval for the short term *MSE*.

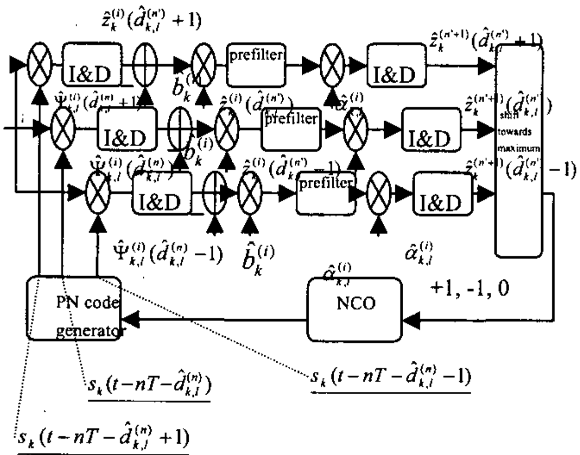

Fig. 3. PIC based SCCL delay tracker.

An alternative approach is based on the projection of the adaptive interference suppression filter to the space spanned by the signature sequence of the desired user. The projection is small if the number of users is low and *SNR* high. At low *SNR*s or high system loads the projection gets larger. Therefore, the projection can be used as a quality indicator in the adaptive receiver branch switching. The filtered squared projection can be expressed as

$$J_{k,l}^{(n)} = \frac{1}{W} \sum_{i=n-W-1}^{n} | x_{k,l}^{H(i)} \bar{s}_{k,l} |^2, \qquad (31)$$

When $J_{k,l}^{(n)}$ exceeds a predetermined threshold ($\xi_{k,l}$) the adaptive part is switched off and only the conventional (*MF*) receiver branch is used in *RAKE* combining (see Fig.1). The adaptive part is switched on again when $J_{k,l}^{(n)}$ is smaller than the threshold:

$$w_{k,l}^{(n)} = \begin{cases} \bar{s}_{k,l}, & J_{k,l}^{(n)} \geq \xi_{k,l} \\ \bar{s}_{k,l} + x_{k,l}^{(n)}, & < \xi_{k,l}. \end{cases} \qquad (32)$$

It should be noted that the adaptive interference suppression filter is updated continuously regardless of switching it on and off in the multipath combining.

3. Suboptimum Detector: Pipelined Multistage

Physical implementation of the Narrowband-PIC detector depends on the symbol rate, spreading factor, type of spreading, number of stages, and the total number of multipaths involved in the IC process. The interference estimation is likely to need an ASIC-based solution, while the actual cancellation could be done with *DSP*s. Therefore, multiple parallel instance in pipeline mode can be used either serial computation or, if the processing rate is too high.

Wideband-*PIC* has the advantage that it can be utilized with both short-and long- code-based systems. In addition, the residual signal provides an interference-resistant input for multipath profile estimator. Although the actual interference cancellation operation requires a higher operation rate than in the *NB-PIC* case, the lack of cross correlation calculation makes the total processing/ resource requirements smaller.

The *PIC* receiver detects all users at the same time and then cancels interference simultaneously. It applies the multistage principle, and the received signal at the stage n is given by

$$r_k^{(n)} = r^{(n-1)} - \sum_{\substack{i=1 \\ i \neq k}}^{K} S_i A_i \hat{b}_i \qquad (33)$$

where r is the received signal, S_i is the spreading waveform matrix for ith user, A_i is attention coefficients matrix and $\hat{b}_i$ is the vector of data estimates.

For multirate users, one possibility is to cancel only the self-interference from different parallel codes of the same user. Coherent multiuser detectors require the knowledge of the complex channel coefficients, the carrier phases, and the propagation delays. Even though conventional methods for parameter estimation can be used, better results would be achieved with the methods that take interference from other users into account.

The configuration of *RAKE* receiver is shown as follows:

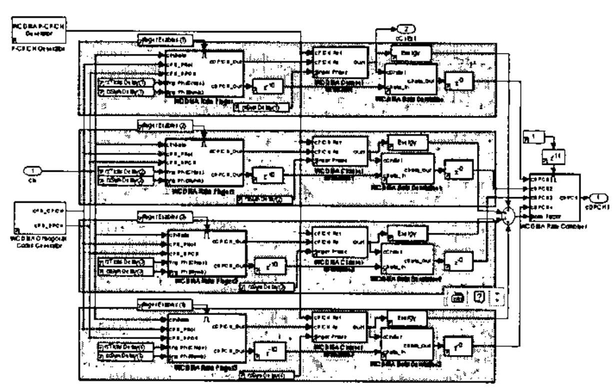

Fig. 4a. WCDMA-Rake-Receiver

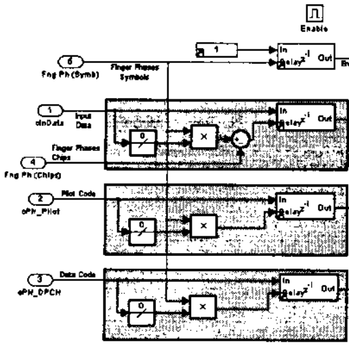

Fig. 4b. WCDMA-Rake-Finger.

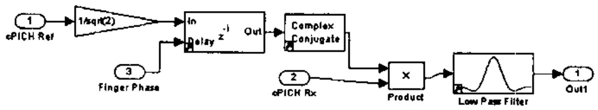

Fig. 4c. WCDMA-Rake-Channel-Estimation.

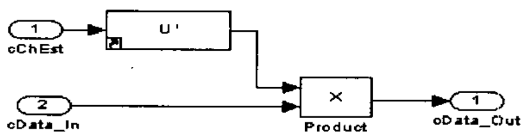

Fig. 4d. WCDMA-data-derotation

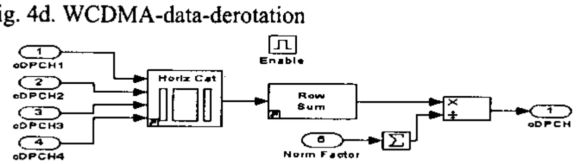

Fig. 4e. WCDMA-Rake-combiner.

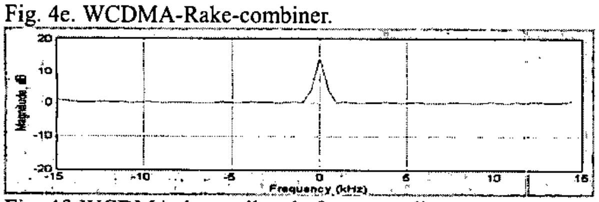

Fig. 4f. WCDMA data collect before spreading spectrum.

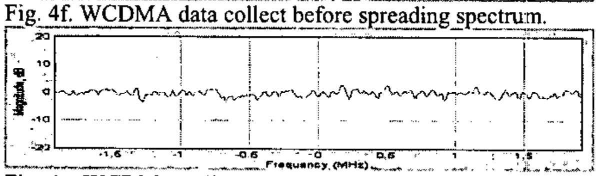

Fig. 4g. WCDMA collect data before pulse shaping.

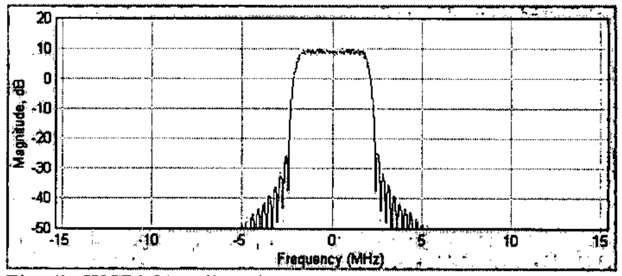

Fig. 4h. WCDMA collect data Tx-Signal.

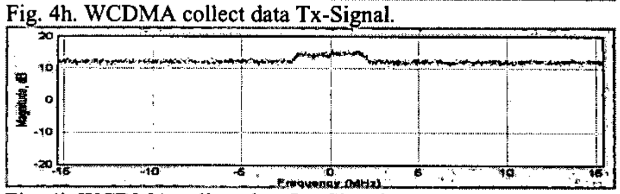

Fig. 4i. WCDMA collect data Rx-Signal.

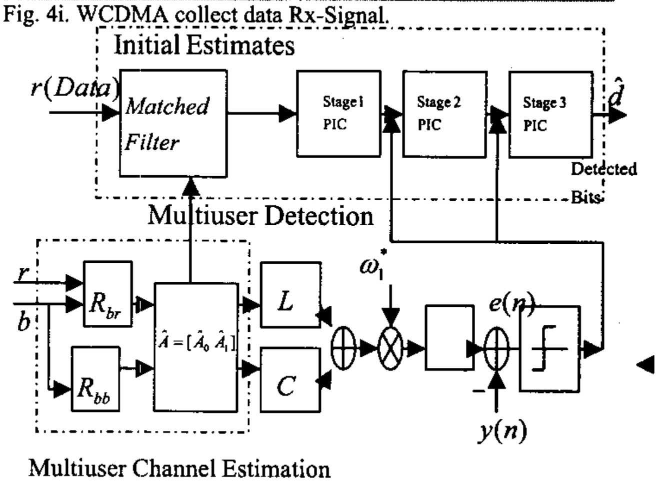

Fig. 5. Task decomposition for joint multiuser channel estimation and detection. The Multistage *PIC* detector can be pipelined into for different tasks [8].

4. PIC for RAKE Receiver

The performance of different multiuser detection schemes for the *PIC* and decorrelating receiver's scheme has been studied in [9]. The most promising scheme was groupwise *SIC (GSIC)*, where users are grouped according to their spreading factor were detected first, and their multiple access interference was subtracted from the Matching Filter outputs of the other users, which are then further detected by the *PIC* or decorrelator [4]. Since all multiuser are in practice near-far limited, power control is required even if *MUD* is used. In the uplink, fast power control improves the performance in three ways: by equalizing the user powers, the detrimental near-far effect is mitigated; by compensating the channel fading, E_b/N_0 performance is improved; and by minimizing of the transmission power, the battery life of mobile stations is increased and intercell interference reduced. Thus, even with multiuser detection it is important to use power control.

The *PIC* scheme shown in Fig. 4 performs best when the user powers are different. To relieve the implementation complexity of the *PIC* receiver with respect to the power ranking, it has been proposed to apply Signal-to-Interference Ratio (SIR)-based power control.

The most portable transmitted signals of all users are regenerated using the symbol estimates from the previous stage. Based on that, the *MAI* estimates are composed for each user and are subtracted from the received signal. The input of *PIC* stage *(i)* is the tentative symbol $\tilde{d}_v^{(i-1)}[n]$ from the output of the previous stage (*i*-1), weighted by the tentative decision function $f_v^{(i)}$ [19]

$$\tilde{d}_v^{(i)}[n] = f_v^{(i)}(\tilde{d}_v^{(i-1)}[n]) \tag{34}$$

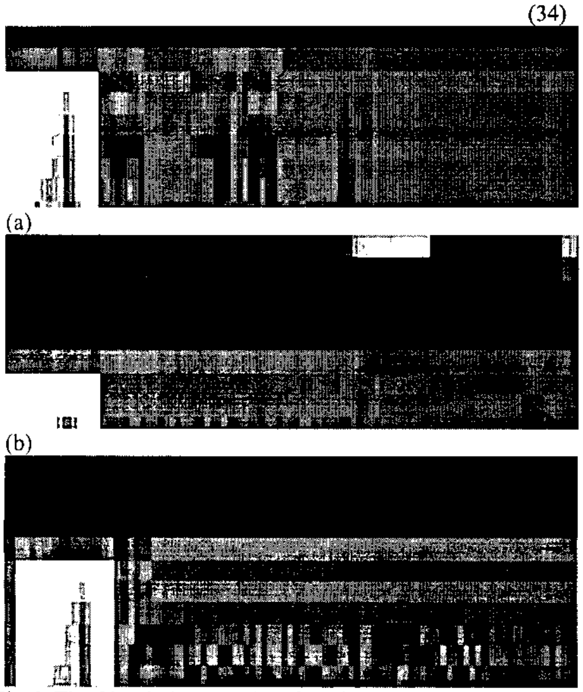

Fig.6. The Simulation Results of *W-CDMA* Rake Receiver (a) Real Part (b) Imaginary Part (c) Output of Channel Estimator.

The matched filter provides the initial estimator for *PIC*. The paper in [14] has shown that differential *PIC* exploits the multistage character of *PIC* to avoid unnecessary double calculations of certain terms in consecutive stages. The *PIC* basic results are revised for future analyzing computational complexity.

1) Initial *PIC* stage: The data symbols at the output of the *RAKE* combiner are

$$\begin{aligned}\tilde{d}_v[n] &= \frac{1}{LK}\sum_{f=0}^{F-1} h_v^*[p_{f,v}] \sum_{m=0}^{LK-1} r[m+(nk+p_f)L]\cdot c_v^*[m] \\ &= \sum_{f=0}^{F-1} \left|h_v[p_{f,v}]\right|^2 \cdot d[n] + I_{A_v[n]} + I_{C_v}[n] + \eta_v[n]\end{aligned} \tag{35}$$

with the *v*th user's self-interference term $I_{Av}[n]$ resulting from non-ideal autocorrelation, and the multiuser interference $I_{Cv}[n]$.

5. Conclusion

In this paper we propose an efficient implementation scheme of wideband version of parallel interference cancellation (*PIC*) called regenerative *PIC* for multiuser detector with Rayleigh fading multipath channel and multiple access interference (*MAI*). The proposed *LMS* adaptive algorithm and its scheme is to replace the matrix inversion instead of the heavy load of matrix computation. The multistage *PIC* suppresses the multiple access interference in multiple consecutive steps. Stage *n* cancels the interference by utilizing the hard symbol decision from stage *n*-1.

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