Weighted RLS Channel Estimators for DS/CDMA Signals in Multipath

Giuseppe Caire, Urbashi Mitra

Abstract— **Channel estimation and equalization is explored in the context of Direct-Sequence/Code-Division Multiple-Access signals transmitted through multipath channels. The focus is on adaptive and recursive methods which exploit pilot symbols or pilot signals, resulting in algorithms which are alternatively appropriate for the uplink or the downlink. Moderate complexity, linear estimators and receivers are investigated which optimize the minimum-mean squared error or a weighted least-squares criterion, respectively. Different weight estimators are thus investigated. The proposed algorithms are explored via Monte Carlo simulations and are shown to be robust and offer near-optimal performance in certain scenarios. Fast convergence is also achieved. In addition, the channel estimation algorithms do not require coordination amongst the active users in terms of the transmission of pilots. Finally, the effects of mismatch in synchronization data are explored and methods to compensate for such errors are examined.**

Keywords— **CDMA, adaptive algorithms, multiuser receivers, frequency selective fading channels.**

I. Introduction

In this paper we address the construction of linear receivers for Direct Sequence Code Division Multiple Access (DS-CDMA) environments where pilot signals are available to aid in channel estimation. Furthermore, we investigate the effects of timing errors on such receivers and develop modified receiver structures to compensate for such errors. In particular, we consider methods based on a weighted least-squares criterion. Both centralized and decentralized receivers are designed. Centralized receivers, appropriate for a base-station-type receiver, can exploit side information about all active users, whereas decentralized receivers only have access to the timing information, spreading code, and potentially the channel of the desired user. The limited information available to a decentralized receiver will be due to the difficulty in estimating such information or due to security reasons. Centralized receivers include the jointly optimal receiver, the non-adaptive MMSE receiver, as well as receivers based on feedback or interference cancellation (see *e.g.* [1]).

Several commercial DS/CDMA systems [2], [3] make use of continuously transmitted pilot signals in order to perform channel estimation and enable coherent detection. Either single or multiple pilot signals will be transmitted depending on the transmission scenario (downlink or uplink, for example). Regardless of the transmission scenario, the pilot signals are viewed as additional virtual users whose data sequences are known to the receiver. As a result, such signals exacerbate the *dimension crowding* effect experienced by such linear receivers as the MMSE or decorrelating receivers [1]. This problem stems from the scenario where the dimensionality of the signal space is larger than the length of the receiving filter; thus the interference rejection properties of the receiver are lost. Our proposed receivers mitigate the dimension crowding effect through active pilot signal cancellation (APSC), *i.e.* explicit cancellation of the pilots from the received signal, *before* data detection. Therefore, fairly large power can be devoted to the pilots in order to achieve good channel estimation without affecting the interference level and without increasing the dimensionality of the signal space.

We start from a general discrete-time finite-memory channel representation that does not assume discrete multipath. As the decentralized and centralized channel estimation schemes are based on weighted least-squares, both the channel(s) for the user(s) of interest and the inverse covariance matrix of the interference. Thus the algorithms can be coupled with an adaptive linear MMSE receiver without additional complexity.

The current work extends our prior results [9] through the consideration of the effects of timing mismatch on the adaptive receivers. In addition, we develop schemes for compensating for such synchronization errors.

II. Discrete-time finite-memory signal model

For brevity we begin with the filtered and sampled received signal. An asynchronous multi-user system is considered. The relative time delays, τ_k, can be expressed as $\tau_k = q_k/W + \epsilon_k$ where q_k is the *integer* part of the delay and ϵ_k is the *fractional* part. In our signal representation, the effects of q_k and ϵ_k are exhibited in the spreading code matrix and effective channel vector, respectively.

The front-end baseband receiver is an idealized low-pass filter with bandwidth $[-W/2, W/2]$ and gain $1/\sqrt{W}$ followed by sampling at rate W with an arbitrary sampling epoch. We assume that an integer number of samples per chip, $N_c = WT_c$ is collected. We assume that there exist integers P and Q such that the filtered and sampled channel impulse response $c_k[i;j]$ and chip pulse $\psi(j/W)$ are negligible for $j \notin [0, P]$ $\forall i$ and $j \notin [-Q, Q]$ respectively.

The vector channel model of the received signal is given

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by,

$$\mathbf{y}[n] = \sqrt{\gamma_1}\sum_{k=1}^{K}\sum_{m=-B_1}^{B_2}\mathbf{S}_k[m]\mathbf{c}_k[n]b_k[n-m] + \sqrt{\gamma_2}\sum_{g=1}^{G}\sum_{m=-B_1}^{B_2}\mathbf{S}_g^{(p)}[m]\mathbf{c}_g[n]d_g[n-m] + \boldsymbol{\nu}[n] \quad (1)$$

The variables γ_i define the powers of the data and pilot signals. The transmitted data for user k is denoted by $b_k[n]$ and the known pilot symbols are given by $d_g[n]$. The matrices $\{\mathbf{S}_k[m] : m = -B_1, \ldots, B_2\}$ and $\{\mathbf{S}_g^{(p)}[m] : m = -B_1, \ldots, B_2\}$ both of size $(M_1 + M_2 + 1) \times P$ are constructed from the spreading sequences of the active users, ($\mathbf{s}_k = (s_{k,0}, \cdots, s_{k,L-1})^T$) and the pilots ($\mathbf{s}_g^{(p)} = (s_{g,0}^{(p)}, \cdots, s_{g,L-1}^{(p)})^T$) and the integer part, q_k, of the user delays. The (i,j)-th element of $\mathbf{S}_k[m]$ is given by

$$[\mathbf{S}_k[m]]_{i,j} = \frac{1}{\sqrt{W}} s_k((mLN_c - q_k + M_2 - i - j)/W) \quad (2)$$

for $i = 0, \ldots, \widetilde{L} - 1$ and $j = 0, \ldots, P$. A finite processing window size is considered $\widetilde{L} = M_1 + M_2 + 1$. The matrices $\mathbf{S}_g^{(p)}[m]$ are similarly derived as functions of $\mathbf{s}_g^{(p)}$.

The effects of pulse-shaping, $\psi(t)$, and the fractional part of the user delay ϵ_k are incorporated into the effective channel response, $\mathbf{c}_k[n] = (c_k[nLN_c; 0], \cdots, c_k[nLN_c; P])^T$, which is modeled as a finite impulse response filter that is constant over a single symbol under the condition that $B_d/W \ll 1$. The additive white Gaussian noise process is denoted by $\nu[n]$ and has variance N_0. The summation limits B_1 and B_2 are obtained by noticing that $\mathbf{S}_k[m]$ is not identically zero over all possible $q_k \in [-LN_c/2, LN_c/2)$ if and only if $-B_1 \leq m \leq B_2$. Thus each user contributes with *at most* $B_1 + B_2 + 1$ symbols to the vector $\mathbf{y}[n]$.

In a cellular type architecture, the number of pilot signals at the uplink is matched to the number of active users ($G = K$). For the downlink scenario, only a single pilot signal is necessary ($G = 1$). We shall also consider a decentralized *ad hoc* network scenario where each user transmits a pilot ($G = K$), although side information about the other users will not be exploited. Thus, $\mathbf{c}_k[n] = \mathbf{c}_g[n]$ if the k-th user belongs to the g-th group.

III. Pilot-aided adaptive channel estimation

We focus on the joint estimation of channels $\{\mathbf{c}_g[n] : g \in \mathcal{S}\}$, where $\mathcal{S} = \{g_1, \ldots, g_S\}$ is a subset of size S of $\{1, \ldots, G\}$. We assume that timing q_g, pilot spreading sequence $\mathbf{s}_g^{(p)}$ and pilot symbol sequence $\{d_g[n]\}$ are known for all $g \in \mathcal{S}$ and unknown for all $g \notin \mathcal{S}$. The delay fractional part ϵ_g is implicitly handled by estimating the channel vector $\mathbf{c}_g[n]$.

For all n and all $g \in \mathcal{S}$, we define

$$\mathbf{H}_g[n] = \sqrt{\gamma_2}\sum_{m=-B_1}^{B_2}\mathbf{S}_g^{(p)}[m]d_g[n-m] \quad (3)$$

and define the matrix: $\mathbf{H}[n] = [\mathbf{H}_{g_1}[n], \cdots, \mathbf{H}_{g_S}[n]]$. The total channel vector is given by: $\mathbf{c}[n] = (\mathbf{c}_{g_1}[n]^T, \ldots, \mathbf{c}_{g_S}[n]^T)^T$. Note that $\mathbf{c}[n]$ is of length $S(P + 1)$. Then, (1) can be rewritten as

$$\mathbf{y}[n] = \mathbf{H}[n]\mathbf{c}[n] + \mathbf{w}[n] \quad (4)$$

where $\mathbf{w}[n]$ is uncorrelated with $\mathbf{H}[n]\mathbf{c}[n]$ and contains all user data signals, noise and all pilot signals not in the subset $\mathcal{S}$. With our assumptions, the sequence of matrices $\mathbf{H}[n]$ is known.

We shall design a Weighted Least-Squares (WLS) channel estimator minimizing the cost function

$$J(\mathbf{c}) = \sum_{i=1}^{n}\alpha^{n-i}(\mathbf{y}[i] - \mathbf{H}[i]\mathbf{c})^H\mathbf{M}[i](\mathbf{y} - \mathbf{H}[i]\mathbf{c}) \quad (5)$$

where $0 < \alpha \leq 1$ is an exponential forgetting factor. and $\{\mathbf{M}[n]\}$ (size $\widetilde{L}\times\widetilde{L}$) is a sequence of non-singular matrices. If $\widetilde{L} \geq S(P + 1)$ and the matrices $\mathbf{H}[i]$ have full column-rank, the solution is easily obtained as

$$\widehat{\mathbf{c}}[n] = \left[\sum_{i=1}^{n}\alpha^{n-i}\mathbf{H}[i]^H\mathbf{M}[i]\mathbf{H}[i]\right]^{-1}\left[\sum_{i=1}^{n}\alpha^{n-i}\mathbf{H}[i]^H\mathbf{M}[i]\mathbf{y}[i]\right] \quad (6)$$

In this way, we obtain jointly the channel estimates for all users in the group $g \in \mathcal{S}$. The implementation of the solution falls into the class of Kalman filters with vector state and vector observation [6], for which recursive computation is possible.

The proper choice of the sequence $\{\mathbf{M}[n]\}$ still remains. A simple choice is $\mathbf{M}[n] = \mathbf{I}$ for all n. Then, (5) becomes the classical exponentially-weighted Least-Squares cost function. A different sensible choice is $\mathbf{M}[n] = \mathbf{R}_w^{-1}$, where $\mathbf{R}_w = E[\mathbf{w}[n]\mathbf{w}[n]^H]$ is the "interference+noise" covariance in the the channel model (4) [10]. With this choice, (6) becomes an exponentially weighted version of the Best Linear Unbiased Estimator (BLUE) [6], that coincides with the maximum-likelihood estimator if $\mathbf{w}[n]$ were a Gaussian vector process. Unfortunately, the receiver has no knowledge of $\mathbf{R}_w$; therefore $\mathbf{R}_w$ must also be estimated recursively. We can write

$$\widetilde{\mathbf{R}}_w[n] = \sum_{i=1}^{n}\beta^{n-i}\widetilde{\mathbf{w}}[n]\widetilde{\mathbf{w}}[n]^H \quad (7)$$

where $0 < \beta \leq 1$ is an exponential forgetting factor (not necessarily equal to α) and where $\widetilde{\mathbf{w}}[n] = \mathbf{y}[n] - \mathbf{H}[n]\widehat{\mathbf{c}}[n]$. Then, with the choice $\mathbf{M}[n] = \widetilde{\mathbf{R}}_w[n-1]^{-1}$, we can approximate (6) by the following recursion (we omit the derivation for space limitations):

Recursive WLS channel estimator. Let $\widehat{\mathbf{c}}[0] = \mathbf{0}$, $\mathbf{M}[1] = \delta\mathbf{I}$ and $\mathbf{\Phi}[0] = \delta\mathbf{I}$, with $\delta > 0$. Then, for $n = 1, 2, \ldots$, let

$$\begin{aligned}
\mathbf{\Phi}[n] &= \alpha\mathbf{\Phi}[n-1] + \mathbf{H}[n]^H\mathbf{M}[n]\mathbf{H}[n] \\
\widehat{\mathbf{c}}[n] &= \widehat{\mathbf{c}}[n-1] + \mathbf{\Phi}[n]^{-1}\mathbf{H}[n]^H\mathbf{M}[n] \\
&\quad \cdot(\mathbf{y}[n] - \mathbf{H}[n]\widehat{\mathbf{c}}[n-1]) \\
\widetilde{\mathbf{w}}[n] &= \mathbf{y}[n] - \mathbf{H}[n]\widehat{\mathbf{c}}[n] \\
\mathbf{M}[n+1] &= \frac{1}{\beta}\left[\mathbf{I} - \frac{\mathbf{M}[n]\widetilde{\mathbf{w}}[n]\widetilde{\mathbf{w}}[n]^H}{\beta + \widetilde{\mathbf{w}}[n]^H\mathbf{M}[n]\widetilde{\mathbf{w}}[n]}\right]\mathbf{M}[n] \quad (8)
\end{aligned}$$

□

In the above recursion, the explicit computation of the inverse of $\mathbf{\Phi}[n]$ is needed. Unfortunately, this is an unavoidable feature of Kalman filters with vector observations [6]. $\mathbf{\Phi}[n]$ has dimension $S(P+1) \times S(P+1)$. Therefore, the number of groups S that can be estimated jointly determines also the algorithm computational complexity.

If $\widetilde{L} < S(P+1)$ or if the complexity of (8) is too large, we propose a suboptimal implementation of the channel estimator based on a parallel bank of individual estimators for each $g \in \mathcal{S}$. This can be obtained directly from (8) by constraining $\mathbf{\Phi}[n]$ to be block-diagonal, with S blocks $\mathbf{\Phi}_g[n]$ of size $(P+1) \times (P+1)$. The resulting algorithm is given by:

Parallel bank of Recursive WLS estimators. For each $g \in \mathcal{S}$ let $\widehat{\mathbf{c}}_g[0] = \mathbf{0}$, $\mathbf{M}[1] = \delta\mathbf{I}$ and $\mathbf{\Phi}_g[0] = \delta\mathbf{I}$, with $\delta > 0$ and let $\widehat{\mathbf{c}}[0] = (\widehat{\mathbf{c}}_{g_1}^T[0], \ldots, \widehat{\mathbf{c}}_{g_S}^T[0])^T$. Then, for $n = 1, 2, \ldots$:

1. For all $g \in \mathcal{S}$ let

$$\begin{aligned}
\mathbf{\Phi}_g[n] &= \alpha\mathbf{\Phi}_g[n-1] + \mathbf{H}_g[n]^H\mathbf{M}[n]\mathbf{H}_g[n] \\
\widehat{\mathbf{c}}_g[n] &= \widehat{\mathbf{c}}_g[n-1] + \mathbf{\Phi}_g[n]^{-1}\mathbf{H}_g[n]^H\mathbf{M}[n] \\
&\quad \cdot(\mathbf{y}[n] - \mathbf{H}[n]\widehat{\mathbf{c}}[n-1]) \quad (9)
\end{aligned}$$

2. Let $\widehat{\mathbf{c}}[n] = (\widehat{\mathbf{c}}_{g_1}[n]^T, \ldots, \widehat{\mathbf{c}}_{g_S}[n]^T)^T$.
3. Update the inverse covariance matrix

$$\begin{aligned}
\widetilde{\mathbf{w}}[n] &= \mathbf{y}[n] - \mathbf{H}[n]\widehat{\mathbf{c}}[n] \\
\mathbf{M}[n+1] &= \frac{1}{\beta}\left[\mathbf{I} - \frac{\mathbf{M}[n]\widetilde{\mathbf{w}}[n]\widetilde{\mathbf{w}}[n]^H}{\beta + \widetilde{\mathbf{w}}[n]^H\mathbf{M}[n]\widetilde{\mathbf{w}}[n]}\right]\mathbf{M}[n]
\end{aligned}$$

□

With this receiver, one needs to compute the inverses of S $(P+1) \times (P+1)$ matrices at each step. Moreover, because the channel spread $P+1$ is normally much less than the processing window size $\widetilde{L}$, so that $\mathbf{\Phi}_g[n]$ is always invertible and S is not limited by P and $\widetilde{L}$, as in the case of (8).

IV. Pilot-aided adaptive receiver with APSC

Without loss of generality, we focus on the detection of user 1, assuming that it belongs to the user group 1 and that $1 \in \mathcal{S}$. We constrain the receiver to be formed by a linear (time-varying) FIR filter with response $\mathbf{h}_1[n]$ of length $\widetilde{L}$ (i.e., equal to the receiver processing window), followed by some (non-linear) detection algorithm based on the filter ouput sequence. Since all pilot signals in $\mathcal{S}$ are known, they can be removed from the received signal vector without the need for decision-feedback. The symbol-rate samples output by the receiver filter are given by

$$z_1[n] = \mathbf{h}_1[n]^H\widetilde{\mathbf{w}}[n] \quad (10)$$

where $\widetilde{\mathbf{w}}[n] = \mathbf{y}[n] - \mathbf{H}[n]\widehat{\mathbf{c}}[n]$ is already provided by the algorithms (8) and (9). We refer to this scheme as *Active Pilot-Signal Cancellation* (APSC). The signal-to-interference plus noise ratio (SINR) for the output $z_1[n]$ is given by

$$\mathrm{SINR}[n] = \left[\frac{\mathbf{h}_1[n]^H\mathbf{R}_{\widetilde{w}}[n]\mathbf{h}_1[n]}{\gamma_1|\mathbf{h}_1[n]^H\mathbf{S}_1[0]\mathbf{c}_1[n]|^2} - 1\right]^{-1} \quad (11)$$

where $\mathbf{R}_{\widetilde{w}}[n]$ is the covariance matrix of $\widetilde{\mathbf{w}}[n]$ given the the channel vectors $\mathbf{c}_k[n]$ and their estimates $\widehat{\mathbf{c}}_k[n]$.

The baseline receiver is the single-user matched filter (SUMF) $\mathbf{h}_1[n] = \mathbf{S}_1[0]\mathbf{c}_1[n]$. This can be approximated by using the channel estimator (8), as

$$\mathbf{h}_1^{\mathrm{sumf}}[n] = \mathbf{S}_1[0]\widehat{\mathbf{c}}_1[n] \quad (12)$$

where $\widehat{\mathbf{c}}_1[n]$ is the first subvector of length $P+1$ of $\widehat{\mathbf{c}}[n]$ provided by (8) or by (9). A receiver with better performance is the minimum mean-square error (MMSE) filter. Provided that APSC is perfect the MMSE filter is given by $\mathbf{h}_1[n] = \mathbf{R}_w[n]^{-1}\mathbf{S}_1[0]\mathbf{c}_1[n]$, where $\mathbf{R}_w[n] = E[\mathbf{w}[n]\mathbf{w}[n]^H]$ is

$$\begin{aligned}
\mathbf{R}_w[n] &= \gamma_1\sum_{k=1}^{K}\sum_{m=-B_1}^{B_2}\mathbf{S}_k[m]\mathbf{c}_k[n]\mathbf{c}_k[n]^H\mathbf{S}_k[m]^H \\
&\quad + N_0\mathbf{I} \quad (13)
\end{aligned}$$

Algorithms (8) and (9) inherently provide a recursive estimate $\mathbf{M}[n+1]$ of $\mathbf{R}_w[n]^{-1}$. The approximation of the MMSE filter is given by

$$\mathbf{h}_1^{\mathrm{mmse}}[n] = \mathbf{M}[n+1]\mathbf{S}_1[0]\widehat{\mathbf{c}}_1[n] \quad (14)$$

In the case of a centralized receiver where $\mathcal{S} = \{1, \ldots, G\}$ (i.e., all user groups are jointly estimated either by (8) or by (9)), the MMSE receiver can be calculated by computing explicitly the inverse of the *structured* covariance estimate by replacing $\mathbf{c}_k[n]$ with $\widehat{\mathbf{c}}_k[n]$ in (13). The resulting filter is

$$\mathbf{h}_1^{\mathrm{mmse}}[n] = \widehat{\mathbf{R}}_w[n]^{-1}\mathbf{S}_1[0]\widehat{\mathbf{c}}_1[n] \quad (15)$$

The latter expression makes use of a good deal of additional information about the structure of the covariance matrix, which will improve receiver performance. Efficient ways of computing (15) when $\widehat{\mathbf{R}}_w[n]$ is given by an identity matrix plus a sum of vector outer products are presented in [11].

V. Synchronization mismatch

In this section, we examine methods for compensating for mismatch in timing information. It has been shown that errors in synchronization as small as a fraction of a chip can significantly affect the performance of MMSE receivers [5]. Assume that $\hat{\tau}_k = \tau_k + \Delta_k$. Recall that the fractional part of the delay is captured in the effective channel description $\mathbf{c}_k[n]$, thus the error we must compensate for is the error in the integer part of the delay. A proposed solution is based on [8]. We construct U receivers in parallel, each synchronized to a different integer part. In [8], it was necessary to approximate the fractional part of the delay, ϵ_k, in order to properly synchronize the receiver in an additive, white Gaussian noise channel; however, for the proposed receivers, the fractional part of the delay is incorporated into the channel model and is thus implicitly estimated.

For the simulations provided in Section VI, for each user there is a receiver matched to: $\hat{\tau}_k - 1$,$\hat{\tau}_k$, $\hat{\tau}_k + 1 (U = 3)$. Three related schemes are considered to determine the best delay:

1. **SINR-based**
   Choose the receiver with the largest instantaneous SINR as determined by (11) by substituting $\mathbf{R}_{\widetilde{w}}[n]$ and $\mathbf{c}_k[n]$ by estimates.
2. **MSE-based**
   Choose the receiver with the smallest instantaneous MSE [8]: $\widehat{\text{MSE}} = 1 - \mathbf{h}_k[n]^H \mathbf{S}_k[0] \widehat{\mathbf{c}}_k[n]$.
3. **Power-based**
   Choose the receiver with the largest soft estimate energy for estimating the pilot signals: $\widehat{\mathbf{E_g}} = |\mathbf{h}_g[i]^H \mathbf{y}[i]|^2$.

In fact, method 2 and method 3 are closely related as the instantaneous MSE is simply one minus the desired signal energy. Method 2 provides an estimate of the average MSE (hence desired signal energy), while Method 3 is simply the instantaneous desired signal energy. The benefit of Method 3 is that for severe near-far environments where intial estimates of the channels may be quite poor, such erroneous estimates can be ignored if they result in incorrect pilot signal estimates. That is one can compare $\text{dec}(\mathbf{h}_g[i]^H \mathbf{y}[i])$ to $d_g[i]$ to determine the fidelity of the linear receiver and hence the channel estimator. The actual instantaneous estimates of each metric proved to be too noisy for use, thus weighted averages were considered: *e.g.* $\text{MSE}[n] = \sum_{i=1}^{n} \rho^{n-i} (1 - \mathbf{h}_k[i]^H \mathbf{S}_k[0] \widehat{\mathbf{c}}_k[i])$. Where $0 < \rho < 1$ is an exponential forgetting factor.

For the centralized receiver schemes, a global search over all combinations of possible receivers would require a search of U^K receiver structures; this would incur prohibitive complexity. A low complexity alternative is to construct the U receivers for user k using (14) (one receiver matched to each time delay estimate for user k only) and choose the desired receiver using one of the schemes above. Thus UK searches are performed. Then, the *centralized* structured receiver is constructed using the results of the search and (15). These multiple receiver strategies will be compared to the effect of overestimating the channel order when the timing is imperfect.

VI. Numerical results

We considered a system with $K = 8$ users and processing gain $L = 16$. Each user transmits the superposition of data and pilot signals (thus $G = K$). Spreading sequences are obtained by chip-wise multiplication of a Walsh-Hadamard (WH) sequence and a pseudo-noise (PN) sequence. Each users has two distinct WH sequences, one for data and the other for pilot, and one PN sequence. Each user is given a distinct PN sequence, but may use the same WH of other users. PN sequences are randomly generated with i.i.d. components over a 4PSK signal set and also the modulation symbols for both the pilot and the data signals are 4PSK. For simplicity, we assume ideal Nyquist chip pulses $\psi(t) = \frac{1}{\sqrt{T_c}} \text{sinc}(t/T_c)$ and we choose the receiver sampling rate $W = 1/T_c$, yielding $N_c = 1$ sample per chip. Without loss of generality, we let $q_1 = 0$ and we independently generate the delays q_k for $k = 2, \ldots, K$, uniformly distributed over the integers in $[-L/2, L/2)$, and ϵ_k for $k = 1, \ldots, K$, uniformly distributed over $[0, T_c)$. The synchronization delays are generated uniformly over the integers $\{-1, 0, 1\}$. A delay estimator which exploits the properties of multi-user signals will produce estimates with errors greater than one chip with negligible probability (see *e.g.* [7]). The channel vectors $\mathbf{c}_k[n]$ are obtained from the filtered and sampled continuous time channel response, where these continuous-time channel responses $c_k(t;\tau)$ are derived from the multipath Rayleigh fading model $c_k(t;\tau) = \sum_{p=0}^{P'} g_p(t) \delta(\tau - \tau_p)$ where $g_p(t)$ are zero-mean mutually independent complex Gaussian WSS random processes with an exponentially decreasing *delay-intensity profile* [4] spanning 5 chips. The resulting channel vectors were scaled in order to achieve the desired user SNRs Fixed channels are considered. The receiver processing window is chosen to span three symbol intervals ($\widetilde{L} = 48$). We considered a near-far SNR assignment (SNR is defined as the ratio of the total data+pilot symbol energy over N_0) where users $k = 1, \ldots, 4$ have SNR= 10 dB and users $k = 5, \ldots, 8$ have SNR= 15 dB. This situation is representative of an uncompensated near-far effect.

We first consider the decentralized detection scheme operating in an *ad hoc* network where all users transmit pilot signals. Thus APSC only cancels a single pilot signal. Therefore, the best performance that a receiver can achieve will be close to the SINR of the MMSE receiver without APSC. This is observed in the decentralized receiver curves seen in Figure 1. It is noted that the effect of timing mismatch is insignificant. Thus the receiver with

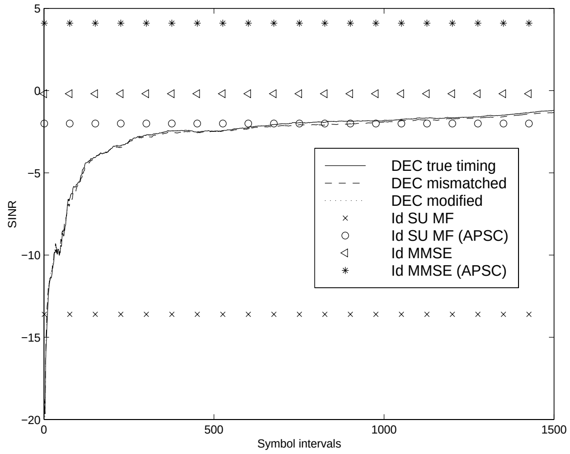

Fig. 1. SINR vs n for the decentralized case.

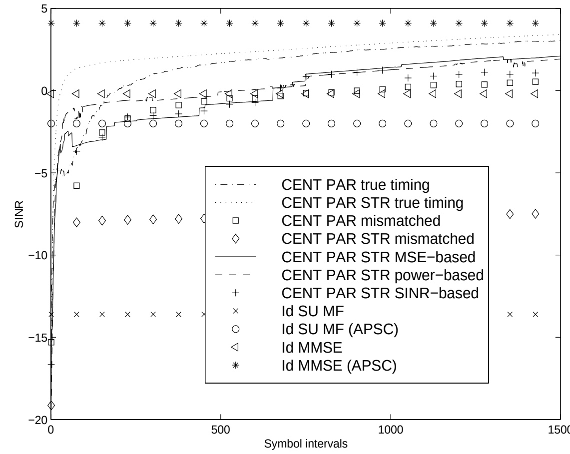

Fig. 2. SINR vs n for the centralized case.

true timing, the mismatched receiver and the modified receivers achieve the same performance. As all of the modified receivers perform comparably, only the plot for the power-based receiver is provided.

However for the centralized receiver schemes, the effect of timing mismatch is a loss in SINR of about 4dB. These plots are provided in Figure 2. This loss is due to the fact that ASPC is attempted for all pilot signals and thus the errors in timing are compounded. Furthermore, if a centralized structured receiver is constructed based on (15) using the imperfect timing information, severe performance degradation is experienced resulting in an SINR that is inferior even to that of the SUMF without APSC. As observed previously, the receiver with structured covariance estimates (based on (15)) in the perfect timing scenario achieves convergence more quickly than the unstructured method (based on (14)).

The modified receivers are able to regain about 2dB in SINR, thus modified receiver schemes with parallel receivers cannot completely track the performance of the true timing case. This is the loss incurred by not performing an optimal search over all U^K possible receiver structures. It is noted that as the receivers approach convergence, the three modified methods perform similarly; however, the method based on power achieves its convergent SINR at a faster rate than the MSE or SINR based receivers. In fact, for both faster convergence and desirable limiting SINR, the power-based receiver appears to offer the best compromise.

VII. Conclusions

In this paper, we have considered the design and study of adaptive linear channel estimators and receivers for DS-CDMA multiuser systems operating in multipath channels. Estimation is facilitated through the use of pilot signals which are then actively canceled in order to improve data detection. Both centralized and decentralized schemes have been investigated. The focus of the current work was on the effects of synchronization errors on the proposed receivers. It was determined that such errors do incur a loss in performance. A multiple receiver structure to compensate for such errors was proposed. Three different metrics were considered for the multiple receiver scheme. It was shown that these methods can result in improved performance in the mismatched timing scenario.

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