MULTIUSER TIMING ACQUISITION OVER MULTIPATH FADING CHANNELS

Akbar M. Sayeed[1] *Behnaam Aazhang*[2*]

[1]Department of Electrical and Computer Engineering
University of Wisconsin–Madison
akbar@engr.wisc.edu

[2]Department of Electrical and Computer Engineering
Rice University
aaz@rice.edu

ABSTRACT

Existing subspace-based multiuser timing acquisition algorithms for code-division multiple access systems do not account for the multipath channel effects satisfactorily. We present a new timing acquisition framework that leverages a canonical representation of the mobile wireless channel to fully exploit the underlying signal structure. The proposed approach promises improved performance based on three key advantages. First, it fully accounts for channel effects in a parsimonious fashion. Second, it is noncoherent in nature since it only requires estimates of second-order channel statistics, as opposed to the actual channel coefficients. Finally, it facilitates accurate modeling of the signal structure of the desired user, thereby yielding powerful interference-suppression algorithms. The framework is applicable in both slow and fast fading scenarios and facilitates maximal channel diversity via joint multipath-Doppler processing. Preliminary results suggest that the resulting algorithms can deliver significantly improved performance at a lower complexity compared to existing methods.

1. INTRODUCTION

Successful demodulation of data in multiuser communication critically depends on accurate knowledge of timings of different users. Multiuser interference and multipath propagation encountered in mobile wireless communication make the timing acquisition problem particularly challenging.

Code division multiple access (CDMA) has emerged as one of the leading technologies for multiuser wireless communications. In the context of timing acquisition, multiuser effects have to be taken into account in both the uplink and downlink — in the uplink due to asynchronous transmission, and in the downlink due to multipath effects. Existing CDMA techniques for multiuser timing acquisition do not incorporate the multipath channel effects adequately [1, 2, 3, 4, 5, 6]. Neither do current methods fully exploit the signal structure of the desired user. Moreover, the state-of-the-art subspace-based (MUSIC) methods have been reported to perform unsatisfactorily at realistic signal-to-noise ratios (SNRs) [7].

In this paper, we develop a new multiuser timing acquisition framework for CDMA systems that fully accounts for the multipath channel effects by leveraging a recently-developed representation of the mobile wireless channel [8, 9]. The representation is based on the wide-sense stationary uncorrelator scatterer (WSSUS) model, and provides a parsimonious description of the channel without any loss of information. The proposed approach only requires *a priori* information about the *desired user*: the spreading code, and certain parameters of the channel representation. Moreover, it is *noncoherent* in nature: it does not require channel state information, only second-order statistics, which can be estimated more reliably. Compared to MUSIC-based approaches, it exploits an appropriate subspace structure for the desired user, as opposed to the signal subspace containing all users, thereby yielding more effective interference suppression. Finally, the resulting algorithms involve the computation of certain eigenvalues, as opposed to eigenspace computation in MUSIC-based methods, which makes them computationally attractive.

The proposed approach is equally applicable in both slow and fast fading scenarios, and, in fact, facilitates maximal diversity under fast fading via joint multipath-Doppler processing [8, 9]. The parsimonious nature of the channel representation also affords robustness to the resulting algorithms. Finally, our methodology is can be tailored to both coherent and noncoherent signaling, making it feasible for both uplink and downlink reception.

In the next section, we present the relevant signal and channel models, and describe the canonical channel representation that lies at the heart of our approach. In Section 3, we develop the basic ideas behind the new timing acquisition framework. Section 4 provides some simulation results, and Section 5 concludes the paper with a discussion of the results and directions for future research.

2. CHANNEL AND SIGNAL MODEL

The complex baseband signal $r(t)$ at a mobile wireless receiver is given by

$$r(t) = x(t) + n(t), \qquad (1)$$

where $n(t)$ is complex additive white Gaussian noise with power spectral density $\mathcal{N}_0$, and the signal $x(t)$ is the channel output to the transmitted complex baseband signal $s(t)$

$$x(t) = \int_0^{T_m} \int_{-B_d}^{B_d} H(\theta, \tau) s(t-\tau) e^{j2\pi\theta t} d\theta d\tau. \qquad (2)$$

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The channel is represented by the time-frequency spreading function $H(\theta, \tau)$ [10, 11]. The time variable τ corresponds to the multipath delays encountered in transmission, and T_m denotes the multipath spread of the channel. Similarly, the frequency variable θ corresponds to the Doppler shifts caused by the motion of the users, and B_d denotes the Doppler spread of the channel.

In the WSSUS channel model, $H(\theta, \tau)$ is a Gaussian process, uncorrelated for different (θ, τ) values, and thus characterized by [10]:[1] $\Psi(\theta, \tau) \stackrel{def}{=} \mathrm{E}\left\{|H(\theta, \tau)|^2\right\}$. The function $\Psi(\theta, \tau) \geq 0$ is called the scattering function and denotes the power in the different (independent) multipath-Doppler channel components.

A spread-spectrum signaling waveform used in CDMA systems has the form

$$s(t) = \sum_{n=0}^{N-1} a[n] v(t - nT_c) \ , \quad 0 \leq t < T, \tag{3}$$

where T is the signaling duration, $v(t) = I_{[0,T_c)}(t)$ is the rectangular chip waveform of duration T_c, and $a[n]$ is the pseudorandom spreading code corresponding to $s(t)$ [10]. The signal bandwidth $B \approx 1/T_c$, and the parameter $N = T/T_c \approx TB \gg 1$ is called the spreading gain of the CDMA system. We assume negligible intersymbol interference ($T_m \ll T$), which is typically the case.

2.1. Channel Representation

Our timing acquisition approach is based on the following fundamental channel representation [8, 9].

Theorem 1. For a spread-spectrum input, the output signal $x(t)$ in (2) admits the representation

$$x(t) \approx \frac{T_c}{T} \sum_{l=0}^{L} \sum_{p=-P}^{P} \widehat{H}\left(\frac{p}{T}, lT_c\right) u^{p,l}(t) \tag{4}$$

where $L = \lceil T_m/T_c \rceil \approx \lceil T_m B \rceil$, $P = \lceil TB_d \rceil$, and $\widehat{H}(\theta, \tau)$ is a smoothed version of $H(\theta, \tau)$. The waveforms $u^{p,l}(t)$'s are time-frequency shifted copies of $s(t)$

$$u^{p,l}(t) \stackrel{def}{=} s(t - lT_c) e^{j\frac{2\pi pt}{T}}, \tag{5}$$

and are roughly orthogonal

$$\left\langle u^{p,l}, u^{p',l'} \right\rangle \approx \|s\|^2 \delta_{p-p'} \delta_{l-l'} \ , \tag{6}$$

where δ_m denotes the Kronecker delta function. □

The (θ, τ)-sampling in (4) is depicted in Figure 1. It is worth noting that there is virtually no loss of information due to the sampling [8, 9]. In fact, for a sufficiently smooth scattering function $\Psi(\theta, \tau)$, (4) is a Karhunen-Loève-like expansion of the received signal $x(t)$: the $\widehat{H}(p/T, lT_c)$'s are uncorrelated (independent) random variables, and the waveforms $u^{p,l}(t)$'s are (roughly) orthogonal. The orthogonality

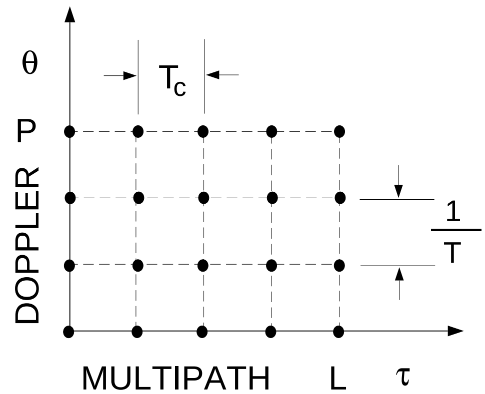

Figure 1. Sampling of the time-frequency plane inherent in the canonical channel representation.

of the $u^{p,l}(t)$'s follows from the correlation properties of pseudorandom sequences [9, 10].

One of the salient features of the above representation is that it parsimoniously describes the channel via the $(L + 1) \times (2P + 1)$ multipath-Doppler samples on the uniform grid in Figure 1. Thus, once the *global* timing offset is known, the locations of all the required channel samples are automatically known. This rigidity of the channel representation reduces the number of parameters in the channel description, and provides robustness from a timing acquisition perspective. Moreover, as we will see, only the average power in the channel samples needs to be known in the proposed approach, which results in algorithms with lower complexity compared to existing methods.

Another attractive aspect of the channel representation (4) is that it facilitates maximal exploitation of inherent channel diversity via joint multipath-Doppler processing [8, 9, 12]. More specifically, the channel is decomposed into independent fading channels by the (approximately) orthogonal multipath-Doppler shifted copies $u^{p,l}(t)$ of the transmitted spreading waveform. Diversity processing is achieved via a time-frequency generalization of the RAKE receiver that results in a substantially higher level of diversity. Several system modalities can benefit from joint multipath-Doppler processing under fast fading [9, 12].

2.2. Multiuser Signal Model

For simplicity of exposition, we will develop the multiuser timing acquisition framework in the context of BPSK signaling — extensions to M-ary and orthogonal signaling are possible. For asynchronous BPSK reception, the received signal is given by

$$r(t) = \sum_{k=1}^{K} x_k(t - \tau_k) + n(t) \tag{7}$$

where $x_k(t - \tau_k)$ is the received signal of the k-th user, and τ_k is the corresponding unknown delay relative to some fixed reference. Without loss of generality, we assume that $0 \leq \tau_k < T$, for all k. In terms of the channel representation

[1] We assume a zero-mean (Rayleigh fading) channel. Extension to non-zero mean situations (Rician fading) is straightforward.

(4), the signal of the k-th user can be expressed as

$$x_k(t) \approx \frac{T_c}{T}\sum_i b_k(i)\sum_{l=0}^{L_k}\sum_{p=-P_k}^{P_k}\widehat{H}_k^{p,l}(i)u_k^{p,l}(t-iT) \quad (8)$$

where $b_k(i) \in \{-1,1\}$ denotes the i-th bit, $\widehat{H}_k^{p,l}(i)$ are the sampled channel coefficients corresponding to the i-th bit, and the $u_k^{p,l}(t)$'s are time-frequency shifted copies of the spreading waveform $s_k(t)$, as defined in (5). In general, all the bits should be jointly decoded for optimal performance. However, in practice, finite blocks of bits are used. Evidently, the duration of the observation interval should be at least $2T$ in order to observe one complete symbol waveform for each user. Without loss of generality, we consider demodulation of the $i=0$-th bit, and for simplicity we consider the observation interval $[0,T)$ which includes influence of the $i=-1$ and $i=0$ bits. Longer observation intervals can be incorporated straightforwardly.

3. TIMING ACQUISITION FRAMEWORK

In contrast to existing methods, our framework is inspired by the observation that timing acquisition can be thought of as a problem of estimating the parameters of a Gaussian process. For BPSK signaling, the correlation function of $r(t)$ is independent of the bit sequence. Moreover, conditioned on the bit sequence $b_k(i)$, the received signal $r(t)$ is a zero-mean Gaussian process, and it can be shown that under sufficiently fast fading (or perfect interleaving), the statistics of the process are independent of the bit sequence. Thus, for all practical purposes, the received signal can be modeled as a Gaussian process, and timing acquisition is equivalent to estimating the parameters (τ_k) of the process.

Under sufficiently fast fading (or perfect interleaving), the correlation function of the received signal $r(t)$ is given by

$$\begin{aligned} R(t_1,t_2) &\stackrel{def}{=} \mathrm{E}\{r(t_1)r^*(t_2)\} \\ &= \sum_{k=1}^{K} R_k(t_1-\tau_k,t_2-\tau_k)+\mathcal{N}_0\delta(t_1-t_2) \quad (9)\end{aligned}$$

where $R_k(t_1,t_2)$ is the correlation function of $x_k(t)$

$$R_k(t_1,t_2) \approx \frac{T_c^2}{T^2}\sum_i\sum_{l=0}^{L_k}\sum_{p=-P_k}^{P_k}\widehat{\Psi}_k^{p,l}u_k^{p,l}(t_1-iT)u_k^{*p,l}(t_2-iT) \quad (10)$$

which follows from (8) where $\widehat{\Psi}_k^{p,l} \stackrel{def}{=} \widehat{\Psi}_k\left(\frac{p}{T},lT_c\right)$ are the samples of the scattering function for the k-th user.[2] On the space of functions on $[0,T)$, denote by $\mathbf{R}_k(\tau_k)$ the operator defined by $R_k(t_1-\tau_k,t_2-\tau_k)$:

$$\left(\mathbf{R}_k(\tau_k)x\right)(t) \stackrel{def}{=} \int_0^T R_k(t-\tau_k,u-\tau_k)x(u)du\ . \quad (11)$$

[2] Note that under the WSSUS assumption, the channel *statistics* do not change over time.

It follows that on the observation interval $[0,T)$, the correlation function $R(t_1,t_2)$ admits the operator representation

$$\mathbf{R}(\tau) \stackrel{def}{=} \sum_{k=1}^{K}\mathbf{R}_k(\tau_k)+\mathcal{N}_0\mathbf{I}, \quad (12)$$

where $\mathbf{I}$ denotes the identity operator, and $\tau \stackrel{def}{=} [\tau_1,\tau_2,\cdots\tau_k]$ denotes the unknown vector of timing offsets.[3]

3.1. Timing Estimation: Two Extreme Cases

Our approach to timing acquisition is motivated by two maximum likelihood (ML) estimators that strike two extreme complexity-versus-performance tradeoffs. The log-likelihood function for the observation waveform $r(t)$, $0 \le t < T$, is given by [13, 14]

$$L^{\tau}(r) = \left\langle \mathbf{R}^{-1}(\tau)r,r\right\rangle - \log\left(\det\left(\mathbf{R}(\tau)\right)\right) \quad (13)$$

where $\det(\cdot)$ denotes the determinant of the operator (product of the eigenvalues). Thus, the ML estimate of the timing vector τ is given by

$$\tau^{ML} = \arg\max_{\tau\in[0,T)^K} L^{\tau}(r). \quad (14)$$

From (13), it is clear that finding τ^{ML} is a complicated problem since the nonlinear dependence of $\mathbf{R}^{-1}(\tau)$ on τ cannot be functionally characterized in general. Moreover, the likelihood surface is typically plagued by many local maxima, making a direct ML approach infeasible.

The other extreme estimator is obtained under a weak signal assumption,[4] which simplifies the problem. In this case, the locally-optimal[5] ML estimates are given by

$$\tau_k^{LO} = \arg\max_{\tau_k\in[0,T)} \langle \mathbf{R}_k(\tau_k)r,r\rangle\ \ ,\ \ k=1,2,\cdots K\ . \quad (15)$$

The estimator in (15) is simple: for the k-th user, compute the quadratic forms $\langle \mathbf{R}_k(\tau_k)r,r\rangle$ for different values of $\tau_k \in [0,T)$, and choose the τ_k corresponding to the largest value. However, this "decoupled" estimator is clearly not near-far resistant since it does not account for other users.

3.2. Interference-Resistant Estimator Design

Both the estimators (14) and (15) are *nonnegative definite quadratic forms* in the observed signal $r(t)$ and represent two extremes: (14) is optimal, requires statistics for all users ($\mathbf{R}(\tau)$), and is computationally intractable, whereas (15) is simple, requires the statistics of only the desired user ($\mathbf{R}_k(\tau_k)$), but is not near-far resistant. Our approach exploits the canonical channel model to strike a balance between this complexity versus performance trade-off by modifying (15) to suppress interference.

For the k-th user, the estimator (15) consists of an array of nonnegative definite quadratic processors, $\langle \mathbf{R}_k(\tau_k)r,r\rangle$, each "matched" to particular value of the delay τ_k. To

[3] After appropriate sampling, these operators are represented by matrices over finite dimensional vector spaces.

[4] That is, the SNR for each user is sufficiently low.

[5] Based on the first term in the Taylor expansion of the likelihood function as a function of SNR.

suppress the contribution due to other users, we define the optimal nonnegative definite quadratic processor $\mathbf{Q}_k(\tau_k)$, corresponding to the delay τ_k of k-th user, as the solution to the following constrained optimization problem

$$\begin{aligned} \text{For } \tau_k &\in [0, T): \\ \mathbf{Q}_k(\tau_k) &= \arg\min_{\mathbf{Q}\geq 0} \mathrm{E}\{\langle \mathbf{Q}r, r\rangle\} = \arg\min_{\mathbf{Q}\geq 0} \mathrm{Tr}\left(\mathbf{QR}(\tau)\right) \\ & \quad \text{subject to } \mathrm{Tr}\left(\mathbf{QR}_k(\tau_k)\right) = 1 \ . \end{aligned} \tag{16}$$

Then, the delay estimate of τ_k is given by

$$\widehat{\tau}_k = \arg\max_{\tau_k \in [0,T)} \mathrm{Tr}\left(\mathbf{Q}_k(\tau_k)\mathbf{R}(\tau)\right). \tag{17}$$

The intuitive motivation for (16) is that the optimal quadratic processor $\mathbf{Q}_k(\tau_k)$ should pass, at a fixed gain, the signal components in the "direction" of the delay τ_k of the k-th user, while minimizing the contribution due to other signal components (interference). The constraint fully exploits the signal structure of the desired user by incorporating the canonical channel model via (10). The optimal delay estimate is then given by the value of τ_k at which the output power of the processor $\mathbf{Q}(\tau_k)$ is minimum (minimum mismatch with the true correlation function of the desired user). Note that the above optimization problem is similar in spirit to the concept of constrained beamforming in array processing [15], and the linear minimum-mean-squared-error receiver design in [16, 17].

The solution to the optimization problem (16) is given by the following result, which we state without proof [18].

Theorem 2. The solution to

$$\begin{aligned} \mu_k &= \min_{\mathbf{Q}\geq 0} \mathrm{Tr}\left(\mathbf{QR}\right) \\ & \quad \text{subject to } \mathrm{Tr}\left(\mathbf{QR}_k\right) = 1 \end{aligned} \tag{18}$$

is given by the smallest generalized eigenvalue of

$$\mathbf{Rc} = \lambda \mathbf{R}_k \mathbf{c}. \tag{19}$$

That is, $\mu_k = \min \mathrm{eig}(\mathbf{R}, \mathbf{R}_k) = 1/\max \mathrm{eig}(\mathbf{R}_k, \mathbf{R})$. □

It follows that the estimator in (17) is given by

$$\begin{aligned} \widehat{\tau}_k &= \arg\max_{\tau_k \in [0,T)} \min \mathrm{eig}(\mathbf{R}(\tau), \mathbf{R}_k(\tau_k)) \\ &= \arg\min_{\tau_k \in [0,T)} \max \mathrm{eig}(\mathbf{R}_k(\tau_k), \mathbf{R}(\tau)), \end{aligned} \tag{20}$$

where the second characterization is preferrable due to numerical robustness. It can be shown that in the noise-free case ($\mathcal{N}_0$=0), $\nu_k(\tau_k) = 1/\mu_k(\tau_k) = \max \mathrm{eig}(\mathbf{R}_k(\tau_k), \mathbf{R}(\tau)) = 1$ at the true underlying value of delay τ_k. Moreover, the unit generalized eigenvalue corresponds to an $L_k \times P_k$-dimensional eigenspace, which is consistent with the signal model (8). On the other hand, the value of $\nu_k(\tau_k)$ is typically significantly larger at mismatched values of τ_k. With noise, of course, the range of values of ν_k is reduced. Figure 2 shows a typical plot of $\nu_k(\tau_k)$ as a function of τ_k at an SNR of 20dB.

The above estimator forms the basis of our timing acquisition framework. To apply (20) in practice, $\mathbf{R}(\tau)$ can be *estimated* directly from the observed data, whereas $\mathbf{R}_k(\tau_k)$ can be computed *analytically* via the canonical channel representation as in (10). It is evident from (10) that only $s_k(t)$ and the average power $\widehat{\Psi}_k^{p,l}$ in the multipath-Doppler channel components is needed to characterize $\mathbf{R}_k(\tau_k)$. In case the powers are not known *a priori*, but only the number (L_k, P_k) of multipath-Doppler components is known, a uniform power distribution may be assumed.

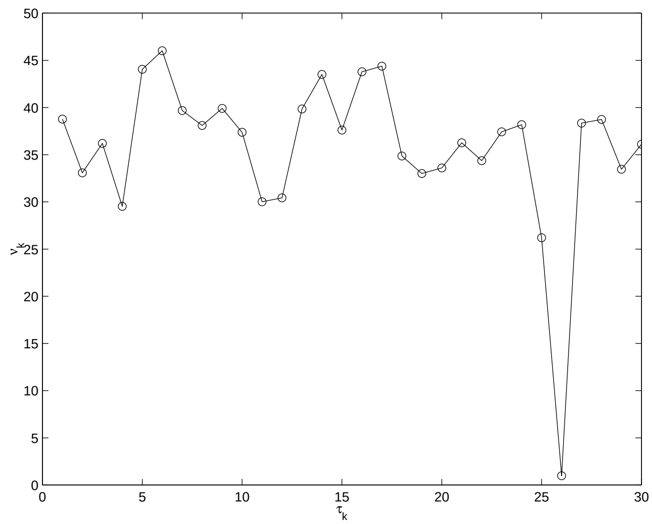

Figure 2. Typical variation of the generalized eigenvalue $\nu_k(\tau_k) = 1/\mu_k(\tau_k)$ for the desired user as a function of τ_k (true underlying value: 26). System parameters: N=31, K=6, SNR of desired user=20dB.

3.3. A Basic Timing Acquisition Algorithm

The above results suggest the following basic algorithm for acquiring the timing of the k-th user.

Step 0. Estimate $\mathbf{R}(\tau)$ from data.
Step 1. Define

$$\tau_k(i) = T_c i\ ,\ \ i = 0, 1, \cdots, N-1. \tag{21}$$

Step 2. For each i, compute

$$\nu_k(i) = \max \mathrm{eig}(\mathbf{R}_k(\tau_k(i)), \widehat{\mathbf{R}}(\tau)) \tag{22}$$

by using the estimate $\widehat{\mathbf{R}}(\tau)$ in Step 1, and the analytical expression for $\mathbf{R}_k(\tau_k(i))$ in (10).
Step 3. Estimate the delay τ_k as

$$\begin{aligned} \widehat{\tau}_k &= \tau_k(i_{min}) \text{ where} \\ i_{min} &= \arg\min_i \nu_k(i). \end{aligned} \tag{23}$$

Notes:

- The above algorithm is based on an observation interval of length T (symbol duration). Longer observation intervals can be easily incorporated, and may be required for desired performance.
- Refined estimates of τ_k may be obtained by using a finer grid for τ_k in Step 1.
- In the case of coherent signaling, phase estimation can be done after timing is acquired. A pilot transmission or training symbols would be required for this purpose.

- For orthogonal signaling, Step 2 would need to be performed for each signaling waveform.
- In the absence of knowledge about the channel statistics of the desired user, $\mathbf{R}_k(\tau_k(i))$ may be computed by assuming a uniform power distribution ($\widehat{\Psi}_k^{p,l} = 1$).

4. SIMULATION RESULTS

In this section, we present some preliminary simulation results on the performance of the proposed timing acquisition algorithm. The simulated CDMA system has a spreading gain of $N = 31$ and supports $K = 6$ users, each assigned a length-31 M-sequence. The fading channel is represented by Jakes Model with two ($L = 1$) equal-power multipath components. The model parameters are: carrier frequency = 1.8GHz, data rate = 10kb/s, and maximum vehicle speed = 50km/h.

The simulated algorithm is based on chip-rate sampling, and uses an observation interval of one symbol (31 samples). The underlying unknown delays τ_k are uniformly distributed over one symbol period with a resolution of $T_c/2$. Correct acquisition is assumed if the estimated delay is within a chip period of the true delay. The data correlation matrix $\mathbf{R}(\tau)$ is estimated from 200 sample symbols with a separation of 30 symbols between successive samples to facilitate sufficiently uncorrelated samples.[6] $L_k = 1$ and $P_k = 0$ for all users, and a uniform multipath power distribution is assumed for the correlation function of the desired user in Step 2.[7]

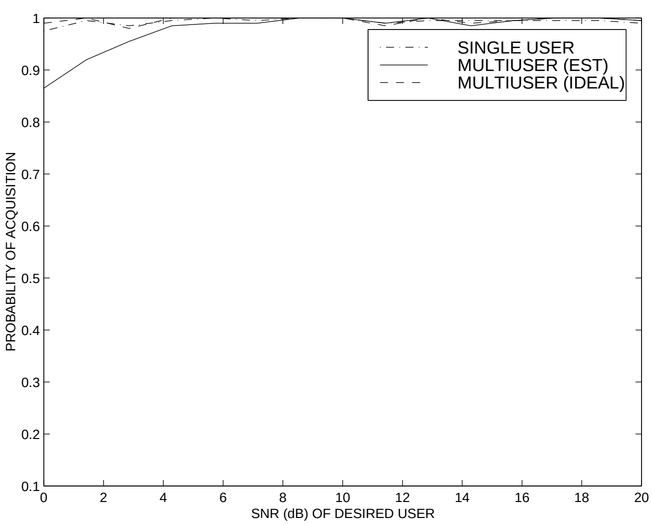

Figure 3. Performance of the three timing estimators as a function of the SNR of the desired user. All users transmitting at equal powers.

The performance of three estimators is compared. The first one is the simple locally optimal estimator (15) which only depends on the *a priori* model for $\mathbf{R}_k(\tau_k)$ in (10). We refer to this estimator as the *single user* estimator since it ignores multiuser interference. The other two estimators are based on the multiuser estimator characterized by (20). The *ideal multiuser* estimator uses the model for data correlation matrix ($\mathbf{R}(\tau)$) given by (9), corresponding to a uniform multipath power distribution for all users. The *estimated multiuser* estimator uses an estimate of the data correlation matrix $\mathbf{R}(\tau)$ based on 200 samples, as explained above.

Figure 3 shows the results for the case when all the users are transmitting at equal powers. All the estimators perform virtually identically — the estimated multiuser estimator shows some degradation at lower SNRs due to less reliable estimates of $\mathbf{R}(\tau)$.

Figure 4 shows the results for the case when the power of the interfering users is 10dB higher than that of the desired user. As expected, the single user estimator performs poorly due to the multiuser interference. In contrast, the ideal multiuser estimator is virtually unaffected over the range of SNR, and the estimated multiuser estimator approaches the ideal performance around 8-10dB.

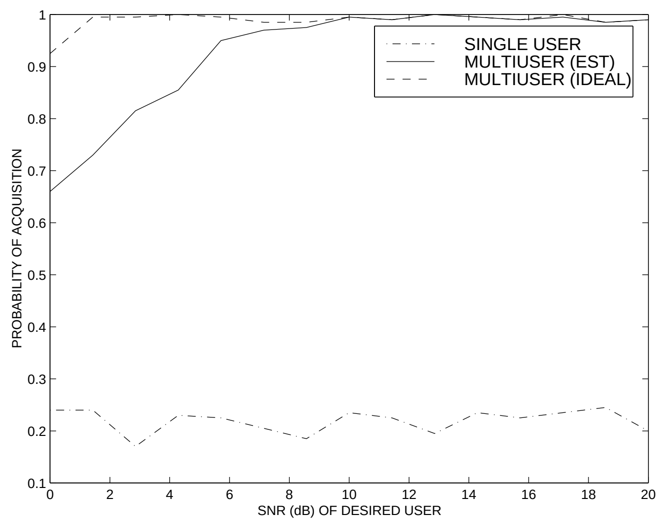

Figure 4. Performance of the three timing estimators as a function of the SNR of the desired user. The five interfering users are 10dB stronger than the desired user.

Figure 5 takes another look at the immunity of the proposed receivers to multiuser interference. It shows the performance of the three estimators as a function of the increasing power of one interfering user relative to the desired user. The SNR of the desired user is fixed at 8dB. As expected, the performance of the single user estimator degrades severely as the power of the interfering user increases. The performance of the multiuser estimators, on the other hand, is virtually identical, and remains unaffected by the increasing interference.

5. CONCLUSIONS

We have presented the basic ideas behind a new multiuser timing acquisition framework that promises improved performance compared to existing subspace-based methods. Our approach has three key advantages over existing methods. First, it fully accounts for multipath channel effects via

[6] This corresponds to a delay of 0.6 seconds.

[7] We note that Doppler diversity ($P_k \geq 1$) may be exploited by employing time-selective signaling and reception [19]. More discussion in [18].

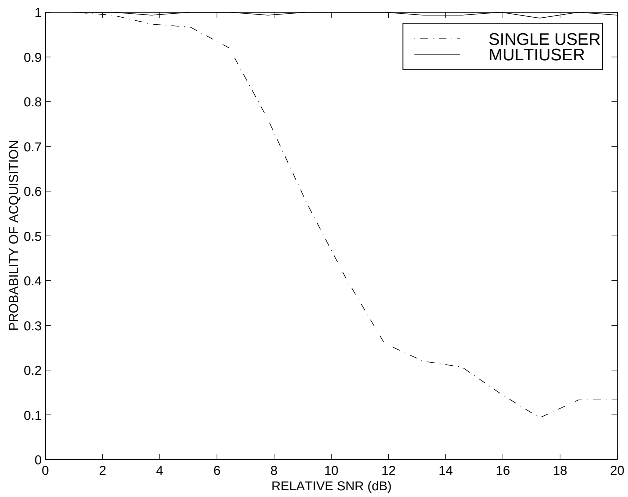

Figure 5. Performance of the three timing estimator as a function of the relative SNR between one interfering user and the desired user. The SNR of the desired user is fixed at 8dB, and four of the interfering users have the same SNR.

a parsimonious model. Second, it is noncoherent in nature: it models the signal structure based on second-order channel statistics, as opposed to channel coefficients. As a consequence, the resulting algorithms require the computation of certain eigenvalues, as opposed to the signal eigenspace, thereby reducing the computational complexity. Finally, the underlying channel representation facilitates more accurate modeling of the signal structure of the desired user, thereby affording effective interference suppression. In view of [7], the preliminary results presented in this paper seem quite promising. A thorough performance analysis of the proposed algorithms is currently under investigation.

In closing we make a few comments about possible extensions of this work. To start with, improved performance may be attained by exploiting joint multipath-Doppler diversity in the proposed approach via time-selective signaling and reception [19]. Moreover, the basic ideas presented here can also be leveraged for noncoherent multiuser detection over multipath fading channels [20]. Blind techniques for joint acquisition and demodulation based on concepts presented in this paper also warrant future investigation.

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