Return Link Code Acquisition for 1-D and 2-D with DS-CDMA for High Capacity Multiuser Systems

Mark C. Reed*, Leif Hanlen*, and Giovanni E. Corazza†

* National ICT Australia, The Australian National University

Canberra 0200, ACT, Australia

{mark.reed}{leif.hanlen}@nicta.com.au

Ph: +61 2 6125 8803, Fx: +61 2 6125 8623

† DEIS/ARCES - University of Bologna

Viale Risorgimento, 2, 40137 Bologna, Italy

gecorazza@deis.unibo.it

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Abstract

Acquisition of the code timing in a direct-sequence code-division multiple-access system at the base station must take place before signal detection and decoding is possible. Code acquisition under severe multiple access interference conditions with time varying codes makes the task even more difficult. Inefficient designs lead to large number of false alarms and/or missed detections. This requirement is needed for conventional single antenna (one dimensional) designs and also for multi-element antenna (two dimensional) designs. This paper details a powerful code acquisition technique for the uplink of direct-sequence code-division multiple-access systems under high loaded situations for both 1-D and 2-D schemes, where the number of users is greater than the processing gain. Under this high multiple access interference condition the DS-CDMA acquisition problem becomes very difficult and conventional search methods simply fail. The method discussed utilises soft data from a multiuser detector to reduce the interference received by the acquisition unit. Analytical performance is compared to simulation results in terms of the number of users, processing gain, interferer signal power, cancellation factor, antenna configuration, and noise variance. Numerical results validate performance under realistic conditions with amplitude, phase, and frequency impairments.

Index Terms

Synchronization, Code division multiple access, Cochannel Interference, Multiuser Channels, Antenna Arrays

I. INTRODUCTION

The introduction of cellular wireless systems in the 1980s has resulted in a huge demand for personal communication services. This demand has increased the need for efficient systems, which has been partially satisfied by the introduction of second generation digital systems, some of which are based on direct-sequence code-division multiple-access (DS-CDMA). New third generation systems, almost all based on DS-CDMA, are now being deployed and will require even more efficient utilisation of the limited spectrum if the high bandwidth features are to become a reality.

In the uplink of a DS/CDMA cellular communication system it is well known that the limiting factor on performance is interference from other users, i.e. multiple access interference (MAI). To mitigate this effect, a large body of research work has been performed to find receiver methods for minimising the MAI [1]–[4]. A very promising receiver technique, that has been recently developed, is based on the "Turbo Principle" from Turbo Codes [5]. The technique is known as iterative multiuser detection (IMUD) and iteratively reduces interference. These techniques can achieve single user (no interference) performance even when the number of users is greater than the processing gain. Although there are many variations on the IMUD technique the low complexity interference cancellation based approach was first published in [2] and was analysed in [3]. Although these receiver techniques bring large improvements, further performance improvement are sought. Antenna arrays at the receiver can be included in the iterative MUD scheme and performance has been shown in [6]. System performance in terms of capacity and cell size improvements was shown in [4]. This work describes simulation and analysis for a high performance acquisition scheme for both single and multiple antenna elements are used by the receiver,, while [4] was essentially concerned only with data detection performance

Efficient timing acquisition in 1-D DS-CDMA systems [7]–[9] is essential to minimise false alarms and missed detections. The acquisition function is to determine the code timing to within a half-chip interval prior to passing the timing information to a tracking function such as a delay lock loop. It has been shown that the acquisition task in the presence of MAI can significantly reduce the capacity of the system [10]. In [11] the authors defined the capacity of multiple users while maintaining acceptable acquisition performance. In [10] the authors suggested a multiple dwell solution, consisting of a search mode and a verification mode, in this paper the number of users was significantly less than the processing gain.

Intuitively the sensitivity of the acquisition unit is reduced due to the MAI, which appears as noise-like interference.

In this paper we develop acquisition techniques that work under high interference scenarios. The situations we are interested in is where the number of users is greater than the processing gain. Previous work on 1-D code acquisition with the presence of MAI includes [12] which studied acquisition performance in the presence of MAI and the near-far problem. The authors compared the performance of approximate maximum likelihood, MUSIC and correlator performance, where the number of users was only one third the spreading factor. In [13] a constant modulus algorithm was proposed. However the number of users was also always less than the processing gain. In [14] an investigation into parallel acquisition in DS-CDMA systems with MAI and other effects was performed. However no suggestions on how to reduce MAI were discussed. In [8] an improved acquisition technique was introduced which is useful for channels with low SNR or MAI. The paper treated the MAI as noise and is therefore not in the same class as the technique we propose. A Paper by Moon *et. al.* [15] studied the effect of MAI and an acquisition approach where multiple users exist. This only considered a small number of users where the ratio of users to spreading factor was far less than one.

Efficient timing acquisition in 2-D DS-CDMA systems exploits the spatial dimension to maximise the capacity of a receiver that utilises an antenna array. A technique for single user acquisition was first proposed in [16], a multi-user approach utilising a subspace technique was proposed in [17], however, the authors used a system where the number of users was only equal to the processing gain. The effective processing gain is actually the product of the number of antenna elements with the processing gain of the spreading code. In [18] the authors studied 2-D code acquisition, however they treated the interfering users as noise. In [19] a 2-D code acquisition approach based on antenna arrays and least mean square (LMS) was presented. As we will show, the performance of a system that performs a 2-D acquisition as well as interference cancellation, can significantly outperform a system that treats multiple access interference (MAI) as noise. We show acquisition under loadings equivalent to the product of the number antenna elements with the processing gain, therefore removing the capacity limitations caused by code acquisition.

Our acquisition technique is based on acquiring each path of each user one at a time. In a practical implementation this is what is required. Users, or paths from users, are detected (and lost) one at a time as a user enters (or exits) the particular cell of interest. For example when a base station is switched on there are no users present and one by one they are acquired, detected and received by the base station.

We describe in this paper a data directed acquisition system where the current users' information is used to improve the signal used for detecting new users or new paths from current or new users. The method is independent of acquisition technique. For the analysis and simulations in this paper we use simple correlator techniques [20], however, the technique could be used with both alternative detection methods or alternative acquisition methods, such as MAX/TC methods [21]. The utilization of long sequences has no impact on the performance of the iterative MUD schemes as dicussed in [3]. The technique described has been patented under [22], [23]. We show performance in terms of probability of false alarm and probability of missed detection. This paper does not say anything about the correct design points, however, as this depends on numerous system issues including the amount of correlation integration time used, channel variation and frequency offset specifications.

The paper is organised as follows. In Section II we discuss the system model. Section III describes the data directed acquisition approach. An analytical study for 1-D is detailed in Section IV and 2-D is analysed in Section V. Section VI describes maximum liklihood acquisition complexity and compares it to our scheme. Analytical and simulation results for both 1-D and 2-D are presented in Sections VII and VIII. Section IX shows results when an asynchronous system is tested. Finally conclusions are drawn in Section X.

II. System Model

We assume that K users each transmit data symbols $d_t^{(k)}$ via BPSK modulation of the length N spreading codes $\mathrm{s}'_{c,t} \in \{(\pm)/\sqrt{N}\}^N$, where t is the time index and c is a chip index and bold type represents vector notation. We also assume that the pilot signal is transmitted using BPSK modulation, orthogonal to the data signal with spreading code length N and spreading codes $\mathrm{s}''_{c,t} \in \{(\pm)/\sqrt{N}\}^N$. This is similar to the modulation methods used in 3GPP [24].

The signal of interest, representing the transmitted preamble, uses random complex spreading codes which is a short code (repeated every symbol interval. In this model we assume no modulation or data encoding for the user of interest (i.e. $d_t^{(1)} = \pm 1$), as would be the case for the preamble of 3GPP [24].

We generalise the model to include antenna arrays with a signal received by an L element Uniform Linear Array (ULA) (where $L = 1$ for a single antenna solution) with half-wavelength element spacing as described by

$$\mathbf{y}_{c,t} = \sum_{k=2}^{K+1} \mathbf{e}^{(k)} (s_{l,c,t}^{'(k)} P_d d_t^{(k)} + j s_{l,c,t}^{''(k)} P_p p_t^{(k)}) + \mathbf{e}^{(1)} s_{c,t}^{(1)} + \mathbf{n}_{c,t} \quad (1)$$

$$= \sum_{k=2}^{K+1} \mathbf{e}^{(k)} g_{c,t} + \mathbf{e}^{(1)} s_{c,t}^{(1)} + \mathbf{n}_{c,t} \quad (2)$$

where l is the element number $l = 1, \ldots, L$. P_d and P_p are constant and are the amplitude of the data and pilot signal, respectively. Using vector notation to represent the antenna array the steering vector is

$$\mathbf{e}^{(k)} = \begin{bmatrix} 1 & \exp\left\{-j\pi l \sin(\theta^{(k)})\right\} & \cdots & \exp\left\{-j\pi(L-1)\sin(\theta^{(k)})\right\} \end{bmatrix}^T \quad (3)$$

for the k'th user and $\mathbf{n}_{c,t} = [n_{1,c,t}, \cdots, n_{L,c,t}]$ is a vector of L independent, identically distributed Gaussian noise terms, with variance σ_n^2. We shall measure angles relatively to the broadside direction of the array.

Equation (1) shows the user of interest separate from the K interfering users while (2) shows the MAI as $g_{c,t}$.

Initially, synchronous chip and symbol timing are assumed when comparing results with analysis, where the phase of all signals are uniformly distributed $0 - 2\pi$. These constraints are later removed to show simulation results for a chip and symbol asynchronous system. Our system performs non-coherent detection and frequency offsets are included.

We assume we will be testing a finite number of code epochs obtained by discretizing the time uncertainty region into cells, where correct detection results when the H_1 cell is above the threshold and is the maximum. Likewise a missed detection is when the H_1 cell is below the threshold. A false alarm occurs when a H_0 cell (all other cells except the H_1 cell) is above the threshold. The false alarm probability (P_{fa}) is defined as the probability of the H_0 cell being above the threshold per symbol interval. The missed detection probability (P_{md}) is defined as the probability of the H_1 cell being below the threshold per symbol interval.

III. Acquisition with Iterative MUD under High Multiple Access Interference

For given acquisition requirements the system capacity has been shown to be limited by the acquisition techniques [10]. This means that as the ratio of number of users to processing gain increases traditional code acquisition techniques fail. This is because the system is interference limited and no amount of integration time will allow correct detection of the timing position. In the uplink of a mobile cellular system the base-station receiver is assumed to use a high performance multi-user receiver. This receiver contains information about the signals from the users that are currently being detected. These signals cause the MAI in the acquisition unit which is looking for new (unknown) users and/or new paths for currently known users in the mult-path case.

A. 1-D Acquisition Concept

Initially the base station is turned on and no terminals are connected, over time callers connect and disconnect. The receiver detects the data for all users connected. The iterative MUD techniques [3], [6] possess a very good estimate of the baseband spread signal at the receiver input. If the receiver cancels the input baseband signal against the receiver estimate for every connected user then what is left is system noise, noise from incorrect estimation of interfering users, and signals from unknown users, or unknown paths of currently tracked users.

If the signal estimates from the currently tracked users are "good" then the remaining cancelled signal may be processed with a conventional correlator to find new users or new paths of currently tracked users. The definition of a "good" signal from the receiver is subjective, however: the received signal must be of a high enough signal to noise ratio after interference cancellation that the frame error rate conditions (QoS conditions) are met. If this were not the case this user would be dropped by the receiver as part of the Radio Resource Controller (RRC) function. Figure 1 shows a block diagram of the technique, here soft information from the multi-user receiver (which is the receiver's best guess at the input signals) is subtracted from the input signal. A correlator may then be used to acquire these signals. Figure 1 also highlights the need for a delay between the input signal and the cancellation process. The delay is needed because the IMUD requires a certain amount of time to determine its estimate of the received signal.

We now show the formulation of the technique. We assume the received spread signal, $y_{c,t}$, consists of signals that are being tracked by the receiver $g_{c,t}$ plus one new signal $s_{c,t}^{(1)}$. The decision-directed approach then computes

$$w_{c,t} \quad = \quad y_{c,t} - \tilde{g}_{c,t} \tag{4}$$

where $\tilde{g}_{c,t}$ is the sum of estimates of the spread signal sent by users $k = 2 \ldots K+1$. The iterative multiuser detection technique discussed in [4], [6] is well suited to this decision-directed approach as the receiver has available the spread signal estimates as they are needed for the receiver implementation. The added complexity of the implementation is therefore very low as shown in Section VI. This paper says nothing about the architecture of the correlating searcher unit only that it should be implemented following the cancellation of the input signal against the receiver estimate of the input signal.

In the correlator the received signal is convolved with the time reversed complex conjugate of the desired transmit sequence such that the timing position can be determined. This method is based on the conventional correlation approach. The resultant signal is

$$r_{c,t} = \left| \sum_{c=1}^{N} w_{c,t} \big(s_{c,-t}^{(1)}\big)^{T} \right| \tag{5}$$

where the absolute value is taken as the phase of the signal is unknown.

B. 2-D Acquisition Concept

We now show the formulation of the technique for the 2-D case. The configuration is similar, except the variables are now vectors due to the antenna array. We assume the received spread signal, $\mathbf{y}_{c,t}$, consists of signals that are being tracked by the receiver $\mathbf{g}_{c,t}$ plus one new signal $\mathbf{s}_{c,t}^{(1)}$. The decision-directed approach then computes

$$\mathbf{w}_{c,t} \;=\; \mathbf{y}_{c,t} - \tilde{\mathbf{g}}_{c,t} \tag{6}$$

where $\tilde{\mathbf{g}}_{c,t}$ is the sum of an estimate of the spread signal sent by users $k = 2 \rightarrow K+1$ on each antenna element.

Beamforming is performed prior to correlation. This is achieved by multiplying by the transpose of the steering vector, and is equivalent to maximum ratio combining [25]

$$w_{c,t} \;=\; \big(\mathbf{e}^{(p)}\big)^{T} \mathbf{w}_{c,t} \tag{7}$$

where $e^{(p)}$ is the steering vector is for the p^{th} sub-sector of interest. The cell is divided into sub-sectors, where time correlation is performed before moving to the next sub-sector.

The same conventional correlation process can be performed then as in the 1-D case

IV. 1-D ANALYSIS

In this section the goal is to develop an analytical result for the performance of the proposed system in terms of false alarm probability and missed detection probability. This result will be in terms of the number of users, processing gain, noise variance and the cancellation factor of the system. In [3] an analytical method was developed for computing the second order statistics of the interference. This was used successfully to predict the performance of the receiver. In this section we extend this result and analytically determine the density functions of the interference terms.

We need to determine the second order statistics of both the MAI and AWGN. The variance of $w_{c,t}$ is

$$\begin{aligned} \text{var}\{w_{c,t}\} &= E\left\{(y_{c,t} - \tilde{g}_{c,t})^2\right\} \\ &= \sigma_x^2 E\left\{\sum_{k=1}^{K+1}(s_{c,t}^{(k)})^2\right\} + \sigma_n^2 \\ &= \sigma_x^2 \frac{P_d^2 + P_p^2}{2}\frac{K+1}{N} + \sigma_n^2 \qquad (8) \\ \sigma_w^2 &= \sigma_{MAI}^2 + \sigma_n^2 \qquad (9) \end{aligned}$$

since the $s_{c,t}^{(k)}$ are i.i.d., with variance $1/N$ and the variance of the cancellation factor is $\sigma_x^2 = E\left\{(d_t - \tilde{d}_t)^2\right\}$ where $\tilde{d}_t = P(d_t = +1) - P(d_t = -1)$. The distribution of the MAI is therefore $\mathcal{N}(0, \sigma_{MAI}^2)$ and the noise is $\mathcal{N}(0, \sigma_n^2)$. Although the individual values of $\tilde{d}$ are not Gaussian the sum of a large number of these values will be and we utilise this assumption throughout our work. It turns out from comparing the simulation to the analysis that this assumption is very close to correct. The value of $\tilde{d}$ is provided by the decoders on subsequent iterations, as per the turbo receiver principles, this does not necessarily equal the distribution of the source data.

A. Analysis of Off-time (H_0) Density for 1-D Analysis

We now develop a method to determine the density of the off-time (H_0) signal, that is, where the correlator sequence is not time matched with the transmitted sequence. As the codes are random this is equivalent to determining the density of any random code convolved with (4). Utilising the real valued distributions, we first need to compute the distribution of $w_{c,t}$ which is the sum of two independent random variables that are both Gaussian distributed. Utilising a result in [26] that the distribution of the sum of independent random variables, normally distributed has a mean value equal to sum of the mean of the individual densities, with a variance equal to the sum of the variances of the individual densities, the resulting Gaussian distribution is

$$f_z(z) = \frac{C_1}{\sqrt{2\pi}\sigma_w} \exp \frac{-z^2}{2\sigma_w^2} \tag{10}$$

where C_1 is a constant not required as the result is later normalised.

The next task is to convert this result to a density which represents the random variables after convolution. The convolution process is the multiplication by $s_{c,t}^{(1)} = \frac{\pm 1 \pm j}{\sqrt{N}}$, therefore the convolution process is a phase rotation and some constant multiplication, which can be absorbed in the constant C_1. Converting the Gaussian density to the density associated with the absolute value we then have a Rayleigh distribution [27],

$$p_r(r) \quad = \quad \frac{C_1 r}{2\sigma_w^2} \exp \left\{ \frac{-r^2}{4\sigma_w^2} \right\} \tag{11}$$

We have now developed the so-called off-time(H_0) density and along with (8) we have now the result as a function of the number of users K, the spreading factor N, the noise variance σ_n^2, the signal level for the interfering pilot and data signals, and a given x-axis value r (used as the threshold point later). This density will be later used to determine the missed detection probability (P_{md}) where this is found by integrating from the threshold value to infinity.

B. *Analysis of On-time(H_1) Density for 1-D Analysis*

We now derive the on-time (H_1) density. The real valued Gaussian distribution is similar to (10), however now there is a non-zero mean due to the on-time (H_1) correlation peak. Therefore the density is

$$f_e(e) = \frac{C_2}{\sqrt{2\pi}\sigma_e} \exp \frac{-(e - m_e)^2}{2\sigma_e^2} \tag{12}$$

where $\sigma_e^2 = \sigma_x^2 \frac{(P_d^2 + P_p^2)}{2} \frac{K+1}{N} + \sigma_n^2$, $m_e = \sqrt{2}$, and C_2 is a constant not required as the result is later normalised. The variance is reduced by $1/N$ as there is only interference from K users due to the code of interest generates the real/imag. value of $\sqrt{2}$.

This density represents the random variable in either the real or imaginary axis, to convert this to a density which represents the absolute of the signal we utilise the Rician distribution where the statistically independent Gaussian random variables can have different means and a common variance [27]. Therefore the density is

$$p_r(r) \;=\; \frac{C_3 r}{2\sigma_e^2}\exp\left\{-\frac{r^2+s^2}{4\sigma_e^2}\right\} I_0\left(\frac{rs}{2\sigma_e^2}\right) \tag{13}$$

where $s^2 = m_1^2 + m_2^2 = 4$, and $I_0(.)$ is the zero order modified Bessel function of the first kind.

C. Integrating densities to generate Probabilities for 1-D Analysis

Now that we have derived the densities we wish to integrate these functions to determine the probabilities of missed detection and false alarm. This means we need to integrate (11) from $a \rightarrow \infty$ to determine the false alarm probability. We also need to integrate (13) from $0 \rightarrow a$ to determine the missed detection probability. Starting first with (11) we have

$$\begin{aligned} p_r(r) &= \frac{C_1 r}{2\sigma_w^2}\exp\left\{\frac{-r^2}{4\sigma_w^2}\right\} dr \\ &= 2C_4 C_1 r \exp\left\{-r^2 C_4\right\} \end{aligned} \tag{14}$$

where $C_4 = \frac{1}{4\sigma_w^2}$

$$\begin{aligned} P_{fa} &= Pr(R > a) \\ &= C_1 \int_a^\infty r\exp\left\{-r^2 C_4\right\} dr \\ &= \frac{-2C_4 C_1}{2C_4}\int_a^\infty e^u du \quad\; u = -C_4 r^2 \quad\; du = -2C_4 r dr \\ &= C_1 \exp\left\{-C_4 a^2\right\} \end{aligned} \tag{15}$$

We now derive the probability of missed detection by integrating the Rician distribution from (13), therefore

$$\begin{aligned} P_{md} &= Pr(0 < R < a) \\ &= \int_0^a \frac{C_3 r}{2\sigma_e^2}\exp\left\{-\frac{r^2+s^2}{4\sigma_e^2}\right\} I_0\left(\frac{rs}{2\sigma_e^2}\right) dr \end{aligned} \tag{16}$$

where $C_3 = 1$ as (16) integrated $0 \rightarrow \infty$ must equal 1. The probability of missed detection (P_{md}) is therefore the CDF of the density which is detailed in [27] as

$$F_R(r) \;=\; P_{md} = 1 - Q_1\left(\frac{s}{\sigma_e}, \frac{a}{\sigma_e}\right) \quad\; a \geq 0 \tag{17}$$

where $Q_1(.)$ is the generalised Marcum's Q function and $m = n/2 = 1$ as the degrees of freedom is $n = 2$.

V. 2-D ANALYSIS

The 2-D analysis is similar to that generated for the 1-D case, the biggest difference is in determining the variance and the mean of the densities used. These parameters need to be extended to include the antenna configuration and cross-correlations between users.

A. Analysis of Off-time (H_0) Density for 2-D Analysis

We now develop a method to determine the pdf of the off-time (H_0) signal, *ie.*, where the correlator sequence is not time matched with the transmitted sequence. Each user's signal will be weighted by the spatial filtering of the L element array according to their location. We assume that the signal angle of arrival of each user is distributed uniformly at random over the interval $[-\pi/6, \pi/6]$. The effect on MAI is to adjust the equivalent noise variance by a factor a_p which corresponds to the received power from each (interfering) user subject to the location of each user.

The correlation is carried out after beamforming, where the beamforming is performed for sub-sectors and each sub-sector is approximately equal to the beamwidth of the antenna array. This implies that the correlation between steering vectors for distinct sub-sectors is at least 3 dB less than for the same sub-sector.

For each sub-sector p, we compute the sum of cross-correlations between each user angle and the sector angle. This provides a spatial weighting of the variance for each interferer. For each sub-sector p, this weighting equals

$$a_p = E\left\{\sum_{k=2}^{K+1} \left|\mathbf{r}_p{}^T \mathbf{e}_k\right|^2\right\} \tag{18}$$

where $\mathbf{e}_k$ is the $\mathbb{C}^{1\times L}$ steering vector for the k^{th} user and $\mathbf{r}_p$ is the $\mathbb{C}^{1\times L}$ steering vector for the p^{th} sub-sector with $p = 1, \ldots, P$. The total variance – including MAI and noise – for sub-sector p equals

$$\sigma_p^2 = \frac{a_p \sigma_x^2}{N} \frac{(P_d^2 + P_p^2)}{2} + L\sigma_n^2 \tag{19}$$

As the codes are random this is equivalent to determining the density of any random code convolved with (7). Utilising the real valued distributions, we first need to compute the distribution of $w_{c,t}$ which is the sum of two independent random variables that are both Gaussian distributed. Utilising the same result as earlier, the resulting Gaussian distribution is

$$f_{z,p}(z,p) = \frac{C_1}{\sqrt{2\pi}\sigma_p} \exp \frac{-z^2}{2\sigma_p^2} \tag{20}$$

where C_1 is a constant which is later normalised. The next task is to convert (20) to a density which represents the random variables after convolution. Converting the Gaussian density to the density associated with the absolute value we get a Rayleigh distribution [27] for a single sub-sector

$$f_{r,p}(r,p) \;=\; \frac{C_1 r}{2\sigma_p^2} \exp \frac{-r^2}{4\sigma_p^2}, \qquad r = |z| \tag{21}$$

To compute the result for multiple sub-sectors we average the $f_{r,p}(r,p)$ over all sub-sectors. We have now developed the off-time (H_0) density as a function of the number of users K, the spreading factor N, the noise variance σ_n^2, the number of antenna elements (L), user and interferer position, and a given x-axis value r (used as the threshold point later). This density will be later used to determine the missed detection probability (P_{md}) where this is found by integrating from the threshold value to infinity.

B. Analysis of On-time(H_1) Density for 2-D Analysis

We now derive the on-time (H_1) density. The real valued Gaussian distribution is similar to (10), however now there is a non-zero mean due to the on-time (H_1) correlation peak. The mean value is given by the square root of a_p, (18) for $k = 1$.

The density is only calculated for the sector which has the on-time (H_1) user. Therefore the density is

$$f_e(e) = \frac{C_2}{\sqrt{2\pi}\sigma_p} \exp \frac{-(e-m_e)^2}{2\sigma_p^2} \tag{22}$$

where C_2 is a constant which is later normalised.

The density (22) represents the random variable in either the real or imaginary axis, to convert this to a density which represents the absolute value of the signal we utilise the Rician distribution where the statistically independent Gaussian random variables can have different means and a common variance [27]. Therefore the density is

$$f_r(r) \;=\; \frac{C_3 r}{2\sigma_p^2} \exp\left\{-\frac{r^2+2m_e^2}{4\sigma_p^2}\right\} I_0\left(\frac{rs}{2\sigma_p^2}\right), \quad r = |z| \tag{23}$$

$I_0(.)$ is the zero order modified Bessel function of the first kind. The integration of the densities is the same as that for the 1-D case.

VI. Maximum Likelihood Acquisition with Multiuser Interference Cancellation

To understand the complexity and size of the acquisition problem it is beneficial to describe the Maximum Likelihood (ML) problem where there is multiuser interference. In the following subsections the ML solution is described. This will be compared to the computational complexity of our proposed method.

A. One Dimensional Case

The maximum likelihood solution for timing synchronization of user $K+1$ given K users exist in the system already will be described. This analysis is independent of the search algorithm used and only describes the complexity based on the search space. The assumption is that hypotheses are taken for all users over all data symbols that are currently being detected. The assumption is that the data hypotheses are hard, therefore the data to correlate against is based on $\hat{w}_{c,t} = y_{c,t} - \hat{g}_{c,t}$, where $\hat{g}_{c,t} = \sum_{k=2}^{K+1} s_{c,t}^{(k)} P_d \hat{d}_t^{(k)}$ and $\hat{d} \in \{+1, -1\}$. We assume the pilot signal is known at the receiver and already cancelled.

This is described by the following equation

$$\{\tau, \hat{\mathbf{d}}^{k=1:K}\} = \arg \max_{\tau, \mathbf{d}^{k=1:K}} \sum_{t=1}^{T} \left| \sum_{c=1}^{N} \hat{w}_{c,t} (s_{c,-t}^{(1)})^T \right| \tag{24}$$

here the solution is to maximise a correlation for the known spread sequence and data set, where all combinations of timing offset and data symbol possibilities for the other users is searched. Where $\tau = c + Nt$ is the chip time epoch that provides the maximum of the function. If the data symbols involved in the correlation are T then the order of the complexity of this search is $O(N_s D.2^{TK})$, where the first $N_s T$ is for the number of timing positions that need to be checked, and 2^{TK} are the number of different data combinations of data combinations that need to be hypothesised.

B. Two Dimensional Case

For the two dimensional case the ML solution for timing synchronization is extended in the spatial dimension. Assuming there are P subsectors and the p^{th} subsector is searched sequentially.

$$\{\tau, p, \hat{\mathbf{d}}^{k=1:K}\} = \arg \max_{p, \tau, \mathbf{d}^{k=1:K}} \sum_{t=1}^{T} \left| \sum_{c=1}^{N} \hat{w}_{c,t} (s_{c,-t}^{(1)})^T \right| \tag{25}$$

The order of the complexity for this is then $O(PN_sT.2^{TK})$, therefore increasing the search size by P. The complexities of both the 1-D and 2-D solutions are exponentially greater than that of our proposed approach.

C. Proposed Solution Computational Complexity for 1-D and 2-D Acquisition

In this section we describe the computational complexity of the proposed solution as discussed. The optimisation cannot be described as the maximisation of the argument as soft values are taken from the detector and cancellation is performed. Therefore the computational complexity for the 1-D system is simply $O(N_s.T.K)$ cancellations and for the 2-D case the complexity is $O(P.N_s.T.K)$. The use of soft values in our solution rather than hard values, will at most lead to a constant increase in complexity, i.e. the number of bits required to represent the soft value.

The computational complexity is therefore linear with the spreading factor and the number of symbols for the 1-D case and for 2-D this is extended to include the number of antenna elements. The complexity is trivial compared to the exponential search over the number of symbols and the number of users for the ML solution.

Figure 2 shows the computational complexity as a function of the number of users. For this example the integration time was $T = 3$, the number of Antenna sub-sectors for the 2-D case was $P = 6$, the spreading factor was $N = 100$, and the number of users was $K = 150$. As can be expected the maximum liklihood (ML) computational complexity increases exponentially with the number of users, while the interference cancellation (IC) computational complexity only increases linearly with the number of users. For more than about 10 users the ML approach is not feasible.

VII. 1-D Analytical and Numerical Performance Results

In this section we show results from both our analytical development and from our simulation. The simulation consists of transmitting T symbols where the codes for the multiple access interference are randomly selected for each symbol interval and for each user. The system is chip and symbol-synchronous and all users have equal power.

Although this result is based on the assumption of a single cell we have considered the loading that can be supported in multi-cell environments. This was studied in detail in [28] where a similar interference canceling receiver was considered. The author discusses the problem of noise rise in a multi-cell situation. The author found that the system load could support up to $K/N = 1.7$, this means the load that we use of 1.5 is under this level and therefore realistic for the multi-cell environment.

A. *Determining Residual Interference Levels*

The residual interference from the MUD receiver will determine the performance of the acquisition unit under the proposal we have detailed here. The purpose of this section is to determine a typical residual interference value (σ_x^2), which can be used in showing the expected performance of the system. The typical frame error rate (FER) that maximises capacity has been determined to be FER = 0.1 [24][p.270]. We assume that the bit error rate required is $P_e = 10^{-3}$. Using the rate 1/3 memory equals 8 convolutional code from 3GPP [29][p.166] the $E_b/N_0 = 1.8$ dB. Allowing headroom we select $E_b/N_0 = 2$ dB, which equates to an $E_s/N_0 = -2.7$ dB. If the pilot is 6dB lower than the data signal (typical) then the pilot $E_s/N_0 = -8.7$ dB. Assuming the preamble is at $E_s/N_0 = 1$ dB we can therefore set the amplitude of the data/pilot signals based on an amplitude of 1 for the real/imag. value of the preamble. The data and pilot signals amplitude for this configuration are therefore $0.653 = 10^{-3.7/20}$ and $0.327 = 10^{-9.7/20}$, respectively.

To determine σ_x^2 we utilise the variance in/variance out analysis approach [3] of the memory = 8 rate $R = 1/3$ convolutional code, along with the interference canceller function line with user loading of $K/N = 1.5$. As can be seen in Figure 3 the residual variance is where these two curves cross. We can see that this value is $\sigma_x^2 = 0.005$ when the noise variance is $\sigma_n^2 = 0.946$, this value of noise variance is computed for a $E_b/N_0 = 2$ dB for the code rate 1/3 convolutional code.

B. *Performance in Terms of False Alarms and Missed Detections*

Figure 4 shows the on and off-time densities for three different scenarios. In this result an $E_s/N_0 = 10.54$ dB was used[1] and the simulation was run over 2500 symbols. The first plot is the single user scenario, here the on-time density is centred around 2 and is a Rician distribution, while the off-time density is centred around 0.5 and is Rayleigh distributed. For the single user case there is a distinct separation

[1]This is also equivalent to an $E_s/N_0 = 1$ dB with processing gain = 100 and integration over three symbols.

between the two densities, for example, a threshold set at around 1.2 would provide both a low false alarm probability and a low missed detection probability. The probabilities of missed detection and the probability of false alarm are based on the area under the tails of these two densities, which we want to minimize. In practical systems the threshold is determined for a particular fixed P_{fa}, this determines the amount of time the receiver will be occupied processing false alarms. In the last sub plot the densities for a $K = 150$ user system is shown. Here it is clear that a logical threshold point is not achievable as the on and off-time densities are not clearly separated, this will also be seen later in Figure 5. In the second sub plot the densities for $K = 150$ are shown, however here partial cancellation is used where $\sigma_x^2 = 0.04$. As can be seen the resultant densities, although not as distinct as the single user case are substantially separated.

In Figure 5 we show the performance of the full interference, no interference and two partial cancellation cases ($\sigma_x^2 = 0.04$ and $\sigma_x^2 = 0.005$). We compare the analytical result (solid line) to the simulation results (points). As can be seen the performance of the partial cancellation scheme is substantially better than that of the full interference system without cancellation. This is because the distributions are separated, allowing for a threshold value between the distributions which minimises both false alarms and missed detections.

As the terminal removes the large initial frequency offsets [30] the remaining frequency offset is typically less than 571Hz, consisting of 200Hz frequency error from the terminal, and the addition of 371 Hz, which is due to a vehicle traveling at 200 km/hr. In all of the following results the interfering users have a random frequency offset between ± 571 Hz, uniformly distributed, while the signal of interest has a +571Hz frequency offset. The effect of this, as will be shown, is insignificant, as acquisition occurs within a few symbol intervals, where the rotation in this time interval is less than 16 degrees. The system model assumes no multi-path and no time variation during acquisition, such as time drift. It is the authors' belief that similar benefits with this approach would be found with multi-path and time varying channels.

The y-axis describes the probability of missed detection and the x-axis describes the probability of false alarm, where this probability is related to a symbol interval of time, for this case at $E_s/N_0 = 10.54$ dB, or alternatively, at $E_s/N_0 = 1$ dB with three symbols of integration. For example, fixing a $P_{fa} = 10^{-5}$ in each case the single user (no interference) has $P_{md} \approx 0.39$ while the partial cancellation results are $P_{md} \approx 0.4$

and $P_{md} \approx 0.7$ for $\sigma_x^2 = 0.04$ and $\sigma_x^2 = 0.005$, respectively. In comparison to the full interference system where $P_{md} = 1$, ie. 100% of detections would be missed. The scenario where $\sigma_x^2 = 0.005$ represents a realistic cancellation noise factor derived from the variance in/variance out curves for the multiuser detector at a frame error rate of $P_{FER} = 10^{-1}$, or probabiltiy of bit error of $P_e = 10^{-3}$. The $\sigma_x^2 = 0.04$ value is to show the difference in performance with a cancellation value an order of magnitude greater.

This paper does not say anything about the correct design points for P_{fa} and P_{md} as this depends on numerous system issues including correlation integration time, channel variation and frequency offset specifications.

C. 1-D Performance for a Fixed Probability of False Alarm

Typically in communication systems the false alarm probability is fixed to constrain the false alarms and set the performance of the communication system. This limits the amount of time the receiver is occupied detecting and decoding false timing positions. In this sub section we study the performance of the receiver in terms of a fixed P_{fa}, we utilise our analysis to illustrate the advantages of using partial cancellation methods.

In Figure 6 we show the P_{md} for a fixed $P_{fa} = 10^{-6}$ where the x-axis variable is the E_s/N_0. This shows that the improvement in E_s/N_0 only helps to a certain degree before there is a flooring effect or the case with $\sigma_x^2 = 0.04$. This floor is due to the interference from other users that cannot be removed. Notice we are able to cancel the interference from other users for $\sigma_x^2 = 0.005$ and virtually single user performance is achieved.

The same results for Figure 6 are now plotted with integration on the x-axis in Figure 7. The integration amount is the number of symbols that are coherently combined before a decision is made. This is typically used to improve the performance of the acquisition unit. Again the flooring can be observed for $\sigma_x^2 = 0.04$ value. For the realistic value of $\sigma_x^2 = 0.005$ essentially single user performance is attained and no more than five or six integration steps are needed for fast convergence to a very low probability of missed detection.

What is important to note from Figure 6 and 7 is that for the full interference case no amount of integration or increase in signal power will improve the performance of the acquisition unit. These figures also illustrate the performance improvement possible when partial cancellation is used.

VIII. 2-D Numerical and Analytic Results

In this section we show results from our analysis and numerical results for a receiver with an antenna array. The parameter configuration consists of transmitting $T = 2000$ symbols where the codes for the multiple access interference are randomly selected for each symbol interval and for each user. The system is chip and symbol synchronous and the power levels are set according to Section VII-A. The number of users was $K = 500$ and the processing gain of each spreading code was $N = 100$. The $E_s/N_0 = 10.54$ dB. The receiver uncertainty variance was $\sigma_x^2 = 0.04$, and $\sigma_x^2 = 0.005$. The direction of arrival of the interfering users was uniformly distributed between $\pi/3$ and $-\pi/3$ radians, as would be the case for a 3 sector cellular configuration. The user of interest was positioned at $-\pi/18$ radians from boresight, The antenna array is assumed to be a ULA with half wavelength spacing of the antenna elements and with $L = 5$ antenna elements. The beamforming positions are $5\pi/18, 3\pi/18, \pi/18, -\pi/18, -3\pi/18$, and $-5\pi/18$ radians. The phase offset of every user was uniformly distributed between 0 and 2π radians.

In Figure 8 we show the performance of the full interference and partial cancellation cases. It can be seen that the partial cancellation schemes are substantially better than that of the full loaded system. In fact for the realistic value of $\sigma_x^2 = 0.005$ the false alarm probability is virtually zero for all missed detection probabilities below $P_{md} = 10^{-4}$. The simulation results can only be compared for the full interference case and the $\sigma_x^2 = 0.04$ case as determining simulation results for $P_{fa} = 10^{-20}$ is not possible.

IX. Asynchronous Results for the 1-D System

In this section we show results in Figure 9 when the correlation process is allowed to be asynchronous for the 1-D case. Each users timing position is randomly selected with a uniform distribution over the symbol, the spreading codes are random, the phase of each user is uniformly distributed $[0 - 2\pi]$, the oversampling rate is $O_s = 4$, each interfering user has a random frequency offset between ±571 Hz and the signal of interest has a frequency offset of 571 Hz. For this example the number of users was $K = 150$, and the processing gain was $N = 100$. The signal power of the preamble was $E_s/N_0 = 10.54$ dB, or alternatively with integration of three symbols at $E_s/N_0 = 1$ dB. The interference power was as described in Section VII-A. The transmitter and receiver pulse shaping filter is a root raised cosine filter of length 24 chips, providing overall raised cosine filtering. Under these conditions analysis of the system, to the authors' knowledge, is not possible as this would require knowing the distribution of signals that have been low pass filtered.

The results in Figure 9 show the same trend as those for the chip synchronous solution [31]. The results are shown in terms of the false alarm probability (x-axis) P_{fa} and the missed detection probability (y-axis) P_{md}. The performance shows that the partial cancellation is essential to acquire signals under the given conditions, where the conventional approach has 100% missed detections. We also can note as the reliability of the data from the receiver degrades so to does the performance of the acquisition system.

Figure 9 also shows results with impairments "Imp", for amplitude and phase impairments, with the rest of the parameters the same. As expected the acquisition performance degrades but the results are still very good. Here the amplitude error variance of the sum of the multiple access interference is $\sigma_a^2 = 0.005$ and the phase error variance is $\sigma_p^2 = 0.09$ radians. These interference levels are an estimate based on a mean multiple access interference amplitude of 0.3, which equates to a mean square error of MSE = 0.038, this is equivalent to an operating point of $E_b/N_0 = 2$dB in [32] and therefore is a pessimistic value, where [32] evaluates the channel estimation performance in terms of MSE within an iterative multiuser detector.

X. Summary

In this paper we have investigated a low complexity acquisition technique based on a soft data directed assistance from the receiver. This has been performed under severe multiple access interference, where the number of users is greater than the processing gain for a 1-D system, and greater than the product of the processing gain with the number of antenna elements for the 2-D case.

For both the 1-D and 2-D cases we detailed the system model and the concept of using a data directed interference cancellation approach to minimise multiple access interference with acquisition. We analytically derived density functions to represent the so-called on-time (H_1) and off-time (H_0) random variables after convolution and we compared these densities to simulation results. These densities are functions of the number of users, the processing gain, the power of the interference, the cancellation factor, number of antenna elements, cross correlations between user directions of arrival, and the noise variance. We integrated these densities to determine the probability of false alarm and probability of missed detection, as a function of the threshold value.

We illustrated that increasing the E_s/N_0, or the integration amount, will not improve the performance of the non-cancelled result (full interference system), however this does allow the partially cancelled systems to achieve improved results while being able to determine the integration amount and/or E_s/N_0 required to achieve a required performance level. Our results, for the given parameters, showed that acquisition using conventional correlation techniques is not practical, however, when utilising the cancellation approach, a significantly improved performance is possible for both 1-D and 2-D situations. We detailed a complexity study and show the low complexity of our solution. We also showed asynchronous numerical results with and without impairments to indicate the benefits of our technique.

This paper demonstrated that for high performance multiuser detection designs that information sharing between the acquisition unit (searcher) and the receiver is essential to maximise the capacity of the system and achieve acquisition of new users, or new paths of current users. Depending on the reliability of the information from the receiver, this technique significantly improved the capacity of this return link system.

XI. Acknowledgment

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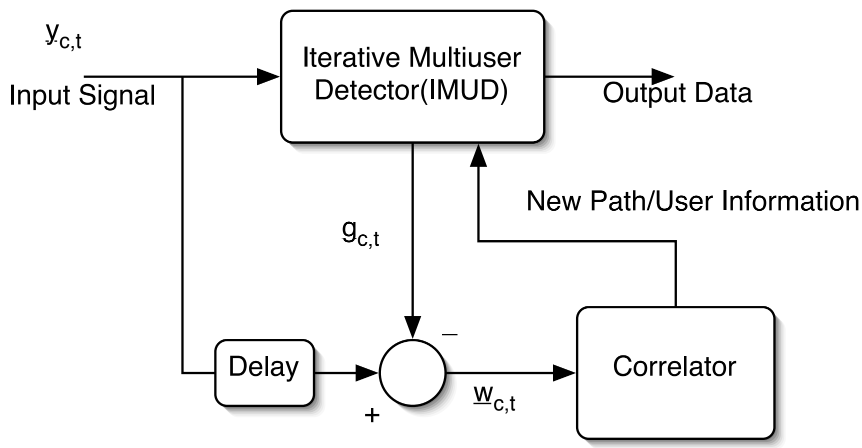

Fig. 1. Decision-Directed Acquisition Unit.

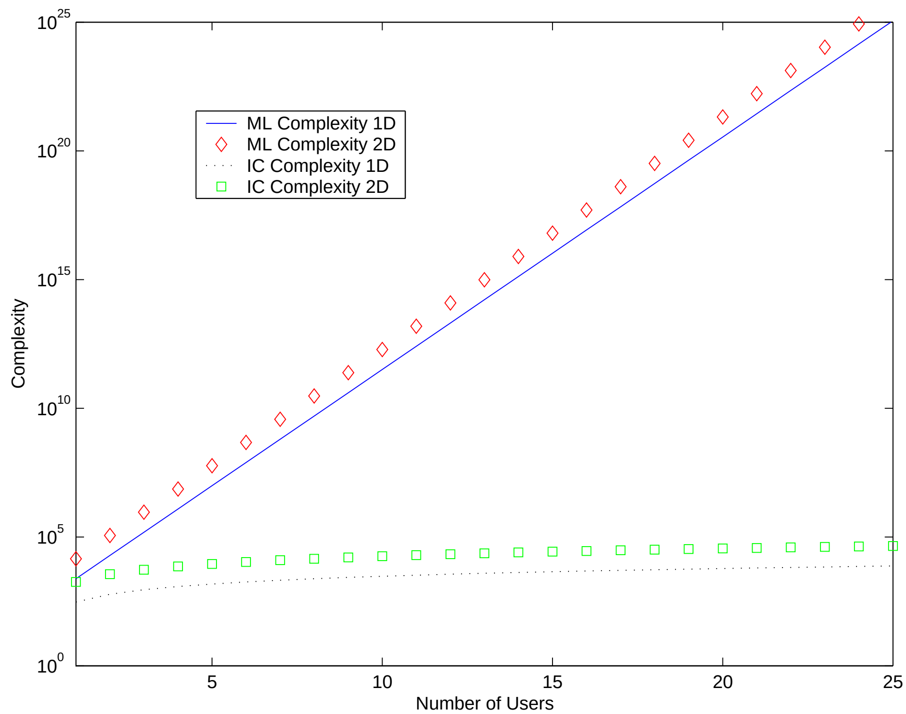

Fig. 2. Computation Complexity Comparison between ML and Interference Cancellation.

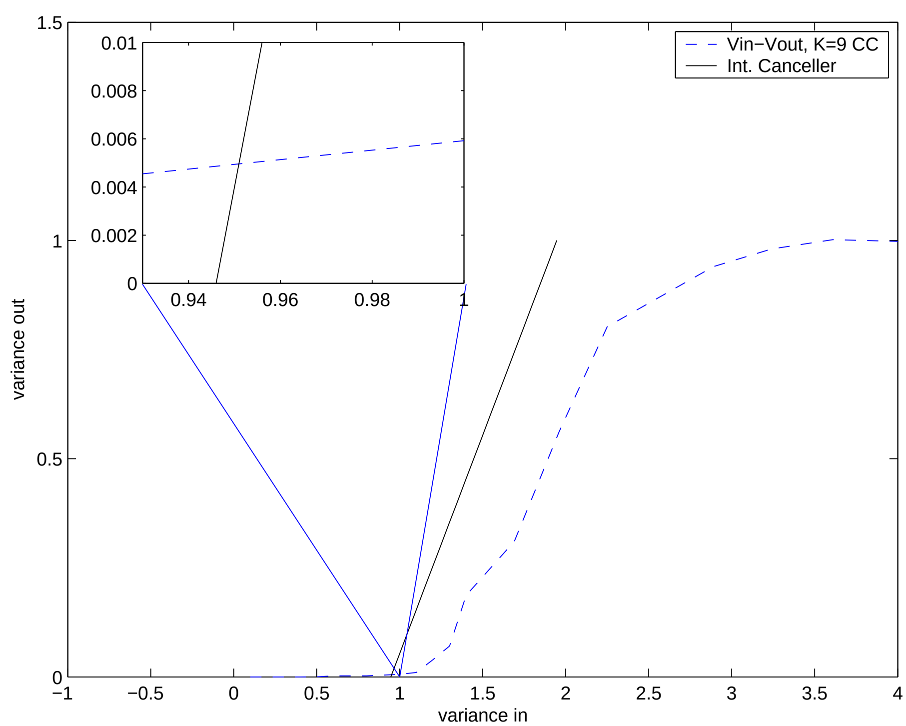

Fig. 3. Variance in vs. Variance Out Plot for Conv. Code and Interference Canceller.

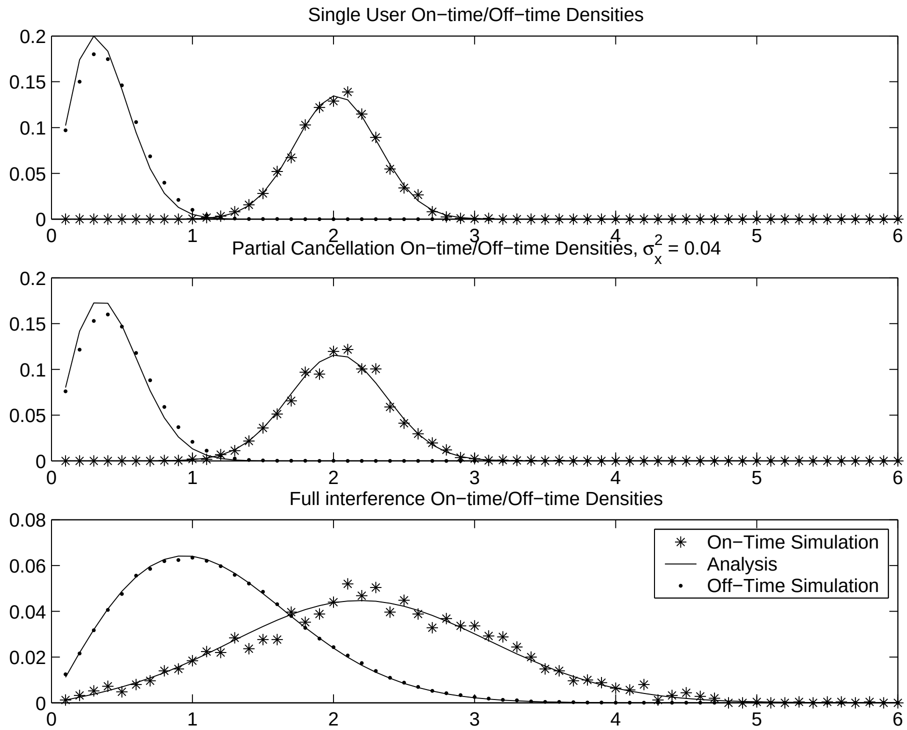

Fig. 4. Densities for different scenarios.

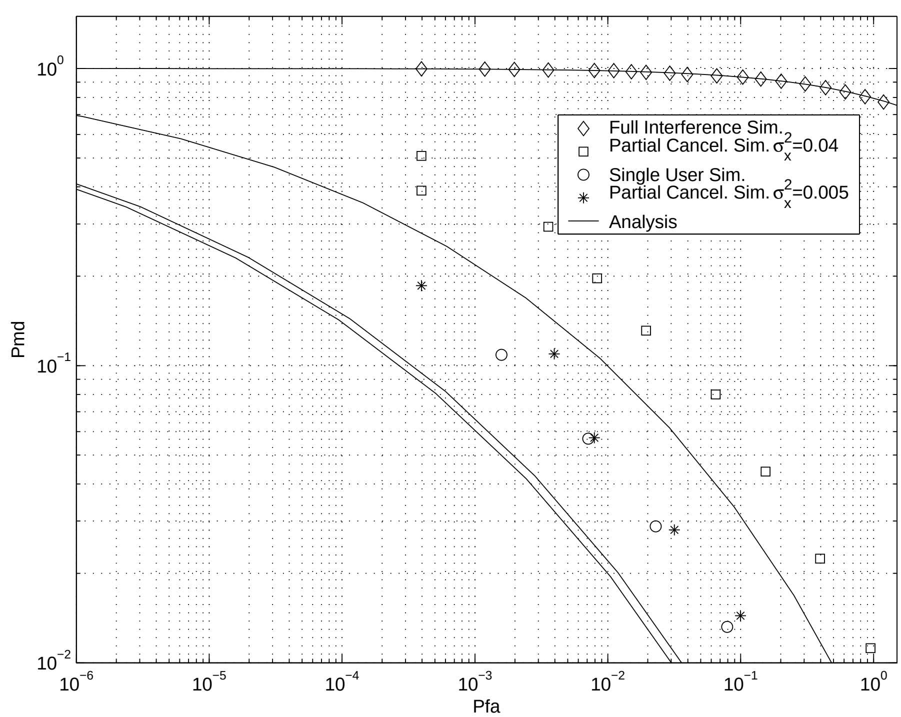

Fig. 5. Performance of the 1-D Acquisition Unit.

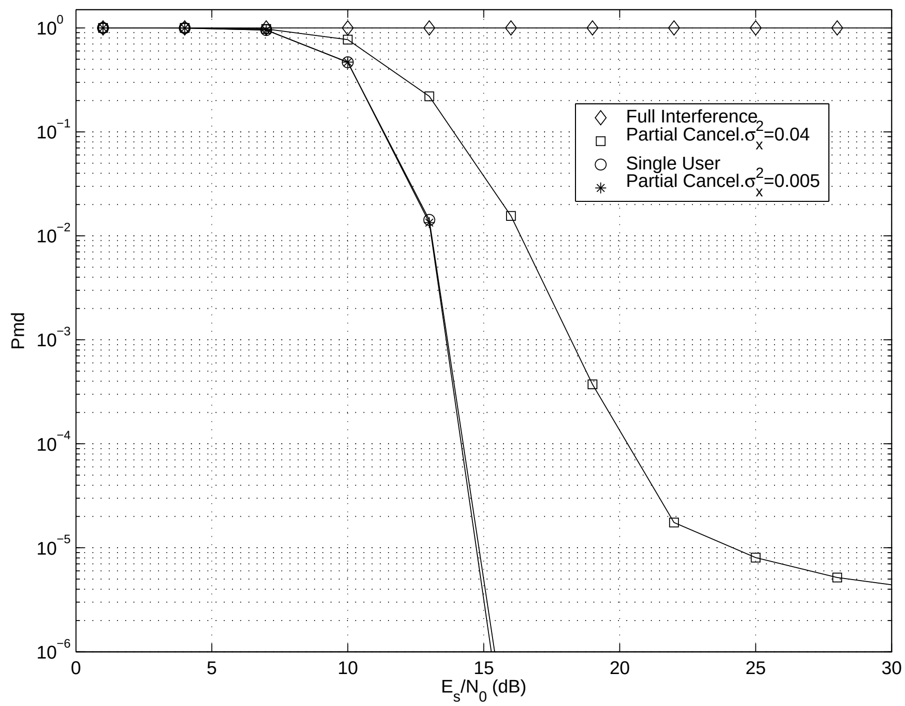

Fig. 6. Missed Detection Performance vs. E_s/N_0. (1-D)

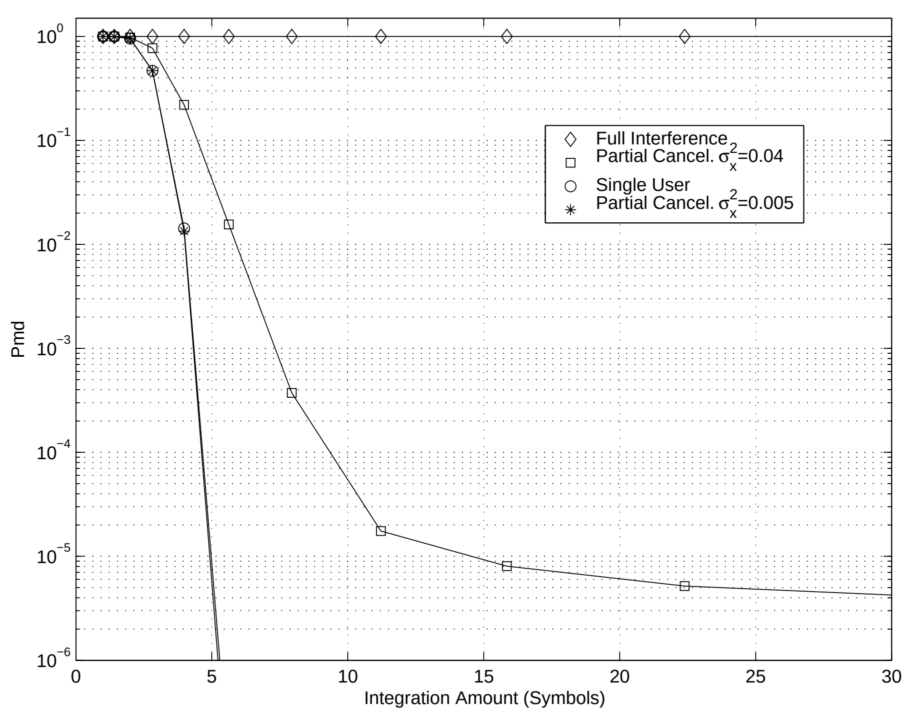

Fig. 7. Missed Detection Performance vs. Integration Amount.(1-D)

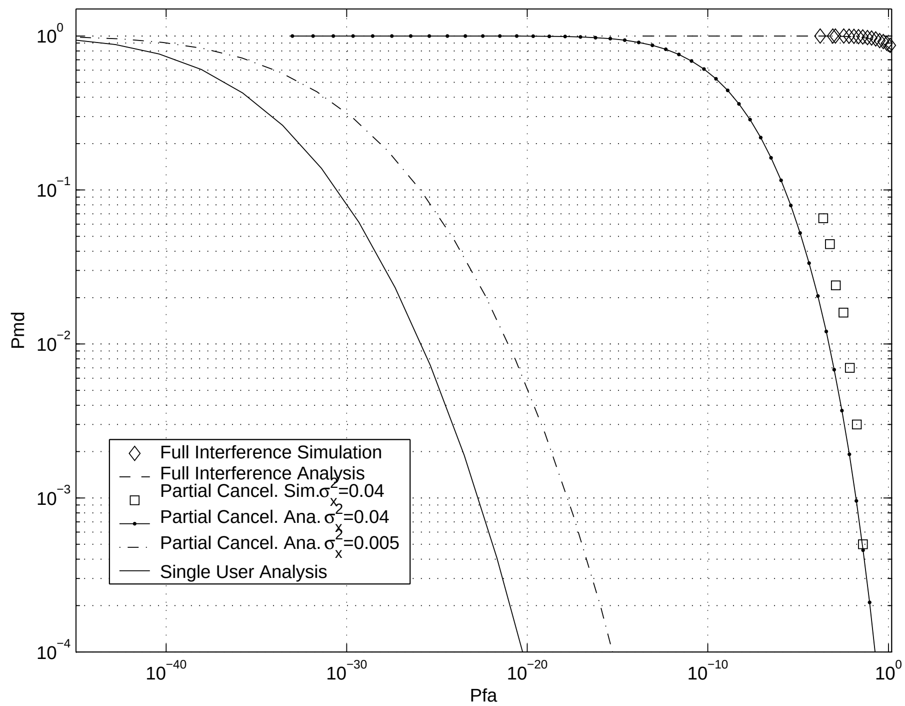

Fig. 8. Performance of the 2-D Acquisition Unit.

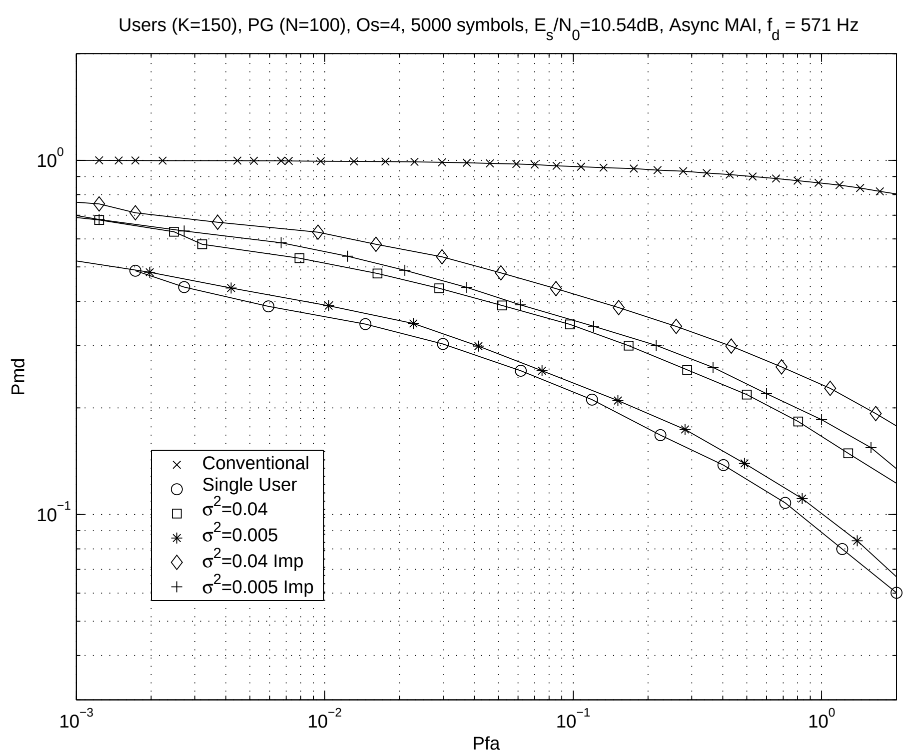

Fig. 9. Acquisition Performance for 1D with 4 times oversampling