CODE ACQUISITION IN DS-CDMA SYSTEMS
OPTIMIZATION AND DSP IMPLEMENTATION

BY

MAYOWA KASSIM AREGBESOLA

A Thesis Presented to the
DEANSHIP OF GRADUATE STUDIES

KING FAHD UNIVERSITY OF PETROLEUM & MINERALS

DHAHRAN, SAUDI ARABIA

In Partial Fulfilment of the
Requirements for the degree of

MASTER OF SCIENCE
In

TELECOMMUNICATION ENGINEERING

KING FAHD UNIVERSITY OF PETROLEUM & MINERALS

DHAHRAN 31261, SAUDI ARABIA

DEANSHIP OF GRADUATE STUDIES

This thesis, written by MAYOWA KASSIM AREGBESOLA under the direction of his thesis advisor and approved by his thesis committee, has been presented to and accepted by the Dean of Graduate Studies, in partial fulfilment of the requirements for the degree of MASTER OF SCIENCE in TELECOMMUNICATION ENGINEERING.

Thesis Committee

Dr. Muhammad Adnan Al-Andalusi
Thesis Advisor

Prof Asrar ul Haq Sheikh
Member

Dr. Abdul Malik Zidouri
Member

Dr. Jamil M. Bakhashwain
Department Chairman

Dr. Mohammad A. Al-Ohali
Dean of Graduate Studies

عمادة الدراسات العليا
DEANSHIP OF GRADUATE STUDIES

Date ٢٨/٤/١٤٢٦هـ

5-6-2005

Dedicated to my parents for their relentless love and care.

رَّبِّ ارْحَمْهُمَا كَمَا رَبَّيَانِي صَغِيرًا

("My Lord! Bestow on them Your Mercy as they did bring me up when I was small.")

ACKNOWLEDGEMENT

All praises are due to Allaah, he who taught man that which he knew not. May His peace and blessings be on the Prophet, his household and all those that follow the truth which he was sent with till the day of resurrection.

Firstly, I like to thank my Parents for always standing by me. I appreciate their efforts and sacrifices more daily as I walk through life. May Allaah reward them abundantly in this life and the hereafter and be merciful to them and accept them to paradise. I am grateful to my sisters for their support.

I am immensely grateful to KFUPM for providing me the opportunity to study at the university. I believe that the University is achieving its goal in becoming a citadel of knowledge.

I am grateful to my committee members, Dr Adnan Al-Andalusi , Prof AUH Sheikh, Dr. AC Zidouri, for their contributions and support. In particular to Dr Al-Andalusi for his supervision and guidance on this thesis that has broadened my horizon of Wireless communication and Spread Spectrum communication to be particular.

I am indebted to all my teachers that have provided guidance and knowledge in all my education endeavours. May Allaah reward all of them and ease their tasks. I also wish to thank all my friends for their friendship and support. We never know how much we mean or help each other till we are no longer there no more. I wish you the very best in your endeavours.

Lastly I pray that Allaah teach us that which will benefit us and benefit us with that which will profit us.

TABLE OF CONTENTS

LIST OF TABLES

LIST OF FIGURES

THESIS ABSTRACT

Name: Aregbesola Kassim Mayowa

Title: Code Acquisition in DS-CDMA Systems Optimization and DSP Implementation

Major Field: Electrical Engineering

Date of Degree:

Code synchronisation is as an attempt to synchronise the receiver clock to the transmitter clock. The process of acquiring the timing information of the transmitted spread spectrum signal is essential to the implementation of any form of spread spectrum technique. The overall system performance is dictated by the code acquisition subsystem since other subsystems rely on a successful code Acquisition

In CDMA systems, the signal to interference plus noise ratio at the receiver depends on the chip pulse shape used in spreading the user signals. The performance of the acquisition system was extended using the Improved Gaussian approximation which gives a better account of the effect of interference on the CDMA system.

The threshold should be set above the noise level, thus an estimate of this figure is required by the detector. Under stable channel conditions the threshold can be fixed but in dynamic channel conditions, where interference is present, adaptive threshold schemes should be used. It is necessary to design an acquisition threshold with its value being set according the effective SINR. A modified threshold setting system is proposed for the Hybrid search scheme. A modified threshold setting scheme was proposed for the Hybrid Search scheme.

Implementation of the algorithms was also carried out on a Texas TMS320C6713.

ملخص الرسالة

الاســـــــــــــم: معاوية قاسم ارجبسولا

عنوان الرسالة: إستملاك رمزٍ في تحقيق أمثلية أنظمةِ DS-CDMA وتطبيق على DSP

التخصــــــــص: الهندسة الكهربائية

تاريخ التخــرج: 05-2005

مزامنة الرموزهي محاولة مزامنة ساعة المستقبل و ساعة المرسل, عملية طلب توقيت اشارة CDMA المرسلة مهمة لتطبيق اي نظام CDMA. الاداء الكلي للنظام يعتمد اساسا على نظام التوقيت للرموزلان كل النظم الاخرى مرتبطة به.

في نظام CDMA, نسبة الاشارة الى التداخل زائد الضجيج عند المستقبل تعتمد على شكل النبضة المستعملة لنشر اشارات المستعمل. اداء نظام التوقيت مدد باستعمال تقريب Gaussian المحسن الذي ياخذ اكثر بعين الاعتبار تاثير التداخل على نظام CDMA.

العتبة يجب تكون اكثر من حد الضجيج, لذا فان تقدير هذه الصورة لازمة للملتقط. تحت ظروف مستقرة للقناة فان العتبة يمكن تثبيتها, لكن في حالة الظروف الديناميكية للقناة اين التداخل موجود فان استعمال عتبات تكيفية تصبح لازمة. يجب تصميم عتبة توقيتية اين يتم تغيير قيمتها تبعا لقيمة نسبة الاشارة الى التداخل زائد الضجيج.

في هذه الرسالة نقترح نظام محسن لتحديد عتبة لنظام البحث الهجين. الخوارزميات طبقت ايضا على *Texas TMS320C6713* .

NOMENCLATURE

PN	Pseudonoise
DSP	Digital Signal Processor
MF	Matched Filter
SNR	Signal to Noise Ratio
MAI	Multiple access interference
P	Signal Power
T_c	Chip Duration
T_c	Sampling Time
ζT_c	Arbitrary receiver code delay
$\hat{\zeta} T_c$	Receiver's estimate of code delay
$\varepsilon = \zeta - \hat{\zeta}$	Delay offset, normalized by chip time
H_1	Hypothesis that the delay corresponds to the correct delay
H_0	Hypothesis that H_1 is not true
P_f	Probability of false alarm
P_d	Probability of detection
P_M	Probability of a miss
N_S	Number of Serial Searches in Hybrid Search Scheme
N_P	Number of Parallel Searches in Hybrid Search Scheme

T_{acq}	Mean Acquisition Time
T_p	Penalty Time
λ	Number of correct H_1 cells
TC	Threshold Crossing
Th	Threshold

CHAPTER 1

INTRODUCTION

In Digital Communications systems, it is necessary for the receiver to be synchronised (in time, frequency) with the received waveform. The subject of this thesis, is the additional task that must be performed by the receiver in a Code Division Multiple Access (CDMA) system, and that is *code synchronisation*, whereby the receiver synchronises its locally generated pseudonoise (PN) code to the PN code of the received signal.

Code synchronisation is an important aspect in a spread spectrum receiver and the performance of a spread spectrum system is often limited by the performance of the code synchronisation subsystem[1]. It is often considered to be composed of two parts: acquisition and tracking. Acquisition involves a search through the region of time-frequency phase uncertainty and it determines that the locally generated and the incoming code are closely aligned. Tracking is the process of maintaining alignment of the two signals using some kind of feedback loop.

Usually, there are two measures of performance of the acquisition procedure, namely, the mean acquisition time and the probability of successfully acquiring the code. The measure of mean acquisition time is well suited for commercial mobile networks where time constrains are crucial.

The goals of this work are to explore initial code acquisition schemes, look at design issues and lastly, the implementation of these schemes on a programmable Digital Signal Processor (DSP).

1.1 SPREAD SPECTRUM

Spread Spectrum Communication is a communication technique wherein the modulated data is spread in bandwidth prior to transmission over the channel and then despread in bandwidth by the same amount at the receiver[2]. This modulation produces a transmitted spectrum much wider than the minimum bandwidth required.

There are many ways to generate spread spectrum signals; *direct sequence* (DS), *frequency hop* (FH), *time hop* (TH), and *Multi-carrier* (MC), and of course, hybrid systems, that utilize the advantages of the different techniques.

Spread Spectrum signals used for the transmission of digital information are distinguished by the characteristics that their bandwidths W are much greater than the information rate R in bits/s. The bandwidth expansion factor $B_e = W / R$ for a spread spectrum signal is greater than unity[3]. The large redundancy inherent in spread spectrum signals is required to overcome the severe levels of interference that are encountered in the transmission of digital information over some radio and satellite channels.

Advantages of Spread Spectrum can be summarized as;

- Provision of resistance to interference and jamming,
- Provision of means for masking the transmitted signal in the background noise in order to lower the probability of detection,
- Resistance to signal interference from multiple transmission paths,
- Multiuser access for common communication channel,
- Provision of means of measuring distance between two points,
- Soft Handover in cellular systems.

1.2 DIRECT SEQUENCE SPREAD SPECTRUM CDMA COMMUNICATION SYSTEM

A spread spectrum system, in addition to the elements of a conventional digital communication system, employs two identical pseudorandom sequence generators, one for spreading at the transmitting end and the other for despreading at the receivers. These two pseudorandom sequence generators generate the pseudorandom sequence used for spreading and despreading at the transmitter and receiver. The model of a Spread Spectrum System is depicted in Figure 1.

In direct sequence spread spectrum system, the baseband data is spread by directly multiplying the baseband data pulses with a pseudonoise sequence that is produced by a pseudonoise generator[4]. In other words, a data-modulated signal is modulated a second time using a very wideband spreading signal.

This wideband signal, spreading signal, consists of a sequence of small pseudonoise (PN) chips whose interval is much smaller than the symbol interval T. The sequence is actually a periodic random sequence with period N. The spread spectrum signal can be converted (despread) back to the original signal by simply multiplying the received sequence with the same spreading sequence.

Considering a BPSK signal having a constant power P, radian frequency ω_0 and data phase modulation $\theta_d(t)$ [5].

$$s_d(t) = \sqrt{2P}\cos\left[\omega_0 t + \theta_d(t)\right] \tag{1}$$

A Direct Sequence is achieved by spreading $s_d(t)$ by the spreading signal (code) $c(t)$. The resulting transmitted signal is

$$s_t(t) = \sqrt{2P}c(t)\cos\left[\omega_0 t + \theta_d(t)\right] \tag{2}$$

Demodulation is accomplished in part by first despreading with the appropriately delayed version of the spreading code[1]. The signal component of the output of the despreader is

$$s(t) = \sqrt{2P}c(t - T_d)c\left(t - \hat{T}_d\right)\cos\left[\omega_0 t + \theta_d(t - T_d) + \phi\right] \tag{3}$$

where $T_d = \zeta T_c$, is an arbitrary delay in the received signal. $\hat{T}_d$ is the receiver's best estimate of the delay in the received signal. When correctly synchronised, the signal component of the output despreader is equal to $s_d(t)$ except for a random phase ϕ, and $s_d(t)$ can be demodulated using a conventional phase demodulator[5].

---

[1] There is delay in the received signal due to propagation delays. Thus, the locally generated PN code has to be delayed accordingly.

The spreading sequences used in CDMA Systems to achieve their multiple-access capability are chosen to have three desirable attributes[6]：

1. the autocorrelation have small off-peak values, to allow for rapid sequence acquisition at the receiver and to minimize self interference due to multipath
2. the cross-correlation are small at all delays, to minimize multiple-access interference
3. the sequences are balanced as much as possible so that each element of the sequence alphabet occurs with equal frequency.

All users in a CDMA system are assigned different spreading sequences to distinguish each user's signal. Hence the properties of spreading sequence a major factor to determining the performance of a CDMA system[6].

1.3 CODE ACQUISITION IN DS-CDMA

The pseudonoise (PN) code synchroniser is an essential element of the CDMA communication system because data transmission is possible only after a receiver accurately synchronises the locally generated *PN* sequence with the received *PN* sequence. For a Direct Sequence Spread Spectrum system, if we are off by a single chip duration, despreading the received spread spectrum signal may be impossible, since the spread sequence is designed to have a small out-of-phase autocorrelation magnitude. In essence, the process of acquiring the timing information of the transmitted spread spectrum signal is essential to the implementation of any form of spread spectrum technique.

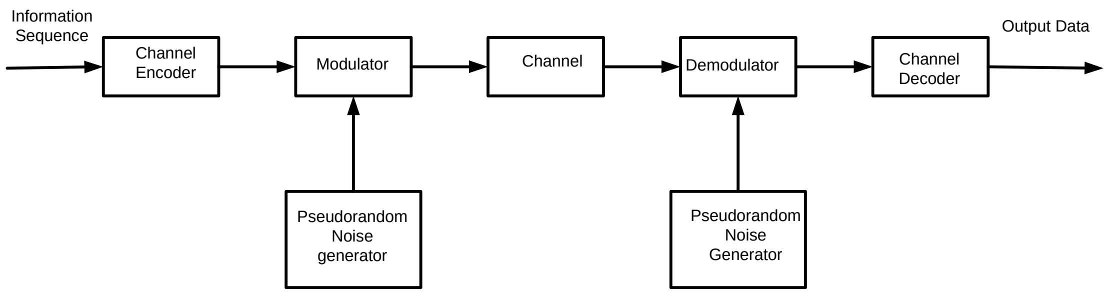


Figure 1 Model of Spread-Spectrum Digital Communications System[7]

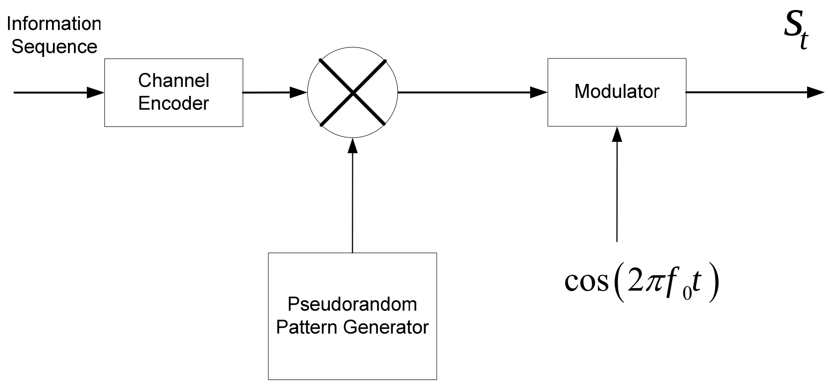


Figure 2 Direct Sequence CDMA Transmitter[7]

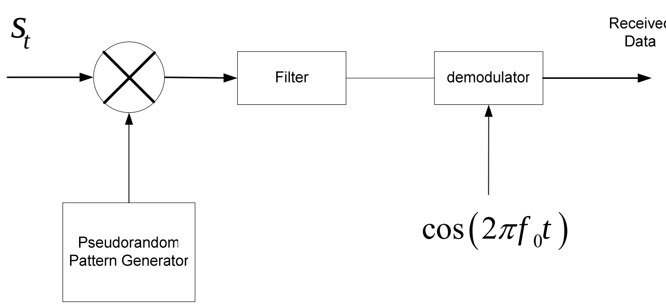


Figure 3 DS-CDMA Receiver[7]

The problem of initial synchronisation (code acquisition) may be viewed as an attempt to synchronise the receiver clock to the transmitter clock. Usually, extremely accurate and stable clocks are used in Spread Spectrum systems to reduce the time uncertainty between the receiver clock and the transmitter clock. Nevertheless, there is always an initial timing uncertainty that is due to propagation delay in the reception of the signal.

1.4 THESIS MOTIVATION

The CDMA air interface is used in both 2G and 3G networks. 2G CDMA standards are branded cdmaOne and it includes both IS-95A and IS-95B. CDMA is the foundation for 3G services: the two dominant IMT-2000 standards, CDMA2000 and WCDMA, are based on CDMA. CDMA are used in Wireless Local Area Networks. Future generation systems and envisioned to use some form of CDMA.

Code acquisition is one of the crucial subsystems in a CDMA system. There have been several techniques proposed in literature to solve the problem of synchronisation. The Hybrid search scheme employed in the acquisition system is considered as one of the best candidates to fulfil the objectives of the code synchronisation subsystem. We choose to study this system in details for our work.

The interference from other users in the CDMA network (Multiple Access Interference) affects the performance of CDMA system. Most studies on the effect of this interference focus on the error rate performance of the system. This interference is dependent on the chip waveform used in the spreading. Consequently the system performance which is dependent on the Signal to noise plus interference ratio is affected. Pulse shaping is essential especially in Mobile communications systems where these systems operate with a large number of users given a minimal bandwidth constraint.

The issues of Pulse shaping relating to PN code synchronisation has not received as much attention despite their major impact on the overall DS-CDMA system performance. To date, there is little work on the impact of chip waveform selection on the synchronisation performance of CDMA systems. In [40], conventional pulses (such as, rectangular, sinusoidal, raised-cosine, etc), were considered to show the effect on synchronisation and tracking using the parallel search scheme. In [41], half-sine and triangular pulses are found to yield better tracking performance when compared with a rectangular pulse at the transmitter. We extend these results in our work to include a more realistic analysis of the interference contribution using the improved Gaussian Approximation. The improved Gaussian Approximation has been shown to give results closer to "exact" scenarios from a probability of error perspective. By correctly modelling the system we can adequately plan for system capacity.

The decision of correct acquisition is based on decision thresholds set at the receiver. Acquisition systems decision threshold is crucial to the system performance. The threshold should be set above the noise level, thus an estimate of this figure is required by the detector. Under stable channel conditions the threshold can be fixed but in dynamic channel conditions, where for instance interference is present, adaptive threshold schemes are used. The channels in Mobile Systems are dynamic channels and thus the need for Adaptive Threshold Setting Schemes

Digital Signal Processors (DSPs) are well suited for signal processing in communication applications and are extensively used in developing products like Cellular phones and Base-stations. In this work we used a Texas C6713 as our test bed to implement our acquisition algorithm.

1.5 CONTRIBUTION OF THIS THESIS

- Modifications were made to the analysis of the noncoherent Hybrid search schemes to reflect a more reflection performance measure. The Improved Gaussian MAI approximation was applied in the analysis of the noncoherent Hybrid search receivers instead of the traditional Gaussian approximation. Numerical results are presented to illustrate the differences in performance between both approximations.
- The effect of chip pulse shaping on the synchronisation performance of the hybrid noncoherent receivers was investigated. Results on the system perforce with different pulse shapes are found.
- Adaptive Threshold Setting

  A modified Adaptive threshold setting scheme was proposed for the hybrid noncoherent receivers. Threshold setting is crucial to acquisition schemes, the system performance is dependent on it. Performance improvements of the scheme over other schemes are presented.
- DSP implementation of the Hybrid Search Scheme

  The Hybrid Search Algorithm was implemented on a Texas C6713 DSK.

1.6 THESIS OVERVIEW

The remainder of the thesis is organized as follows. Code acquisition in Direct Sequence Spread Spectrum is discussed in the next chapter, Chapter 2. Effects of user environments are also discussed. Our approach to code acquisition, the matched filter Hybrid Search Scheme acquisition is then presented in details.

The pseudorandom sequences used in CDMA systems are shaped by a waveform to reduce the effects of the channel. The Multiple access interference, MAI varies according to the chip pulse shape used; consequently the probability of detection and mean acquisition time is affected. Effects of pulse shaping are elaborated on in chapter 3. Improved Gaussian Approximation is applied in the analysis of the effect of MAI on the acquisition scheme. Numerical results are presented.

Threshold setting, its importance and implications are discussed in chapter 4. A modified Adaptive threshold scheme was proposed. Performance improvements of the scheme are included.

Discussions on implementation of a programmable DSP are presented in chapter 5. Some implementation issues are discussed as trade offs.

CHAPTER 2

CODE ACQUISITION IN DS-CDMA

The *pseudonoise (PN)* code synchroniser is an essential element of the CDMA communication system because useful data reception is possible only after a receiver accurately synchronises the locally generated PN sequence with the received PN sequence. For a DS-SS system, if we are off by a single chip duration, despreading the received spread spectrum signal may be impossible, since the spread sequence is designed to have a small out-of-phase autocorrelation magnitude. In essence, the process of acquiring the timing information of the transmitted spread spectrum signal is essential to the implementation of any form of spread spectrum technique.

The problem of initial synchronisation may be viewed as an attempt to synchronise the receiver clock to the transmitter clock. Usually, extremely accurate and stable clocks are used in SS systems to reduce the time uncertainty between the receiver clock and the transmitter clock. Nevertheless, there is always an initial timing uncertainty that is due to propagation delay in the received signal.

The problem of code acquisition is usually solved via a two-step approach:[5, 7, 8]

- *Initial code acquisition (coarse synchronisation)* which synchronises the transmitter and receiver to within an uncertainty of a chip period i.e $\pm T_C$,
- *Code tracking* which maintains fine synchronisation between the transmitter and receiver.

Given the initial acquisition, code tracking is a relatively easy task and is usually accomplished by a delay lock loop (DLL). The tracking loop keeps on operating during the whole communication period.

2.1 ACQUISITION TECHNIQUES, PARAMETERS AND PERFORMANCE MEASURES.

The essential operative constituents of code acquisition are *search strategy* and *detector structure* employed to identify the presence or not of alignment[7]. The set of design parameters for the acquisition procedure includes threshold settings, correlation length, number of tests per code chip, and system complexity as manifested by the choice of search strategy, verification logic, etc[9]. Also implicit is the knowledge of the design SNR, code rate, code length, code uncertainty region, reset penalty time and others. On the other hand, specification of satisfactory performance measures for the overall system is a more complicated task and very dependent on the particular application. However, the dominant parameters of interest in most cases are, the time which elapses prior to acquisition $Tacq$ (Mean Acquisition Time) and the probability of detection P_D.

The receiver hypothesizes a code phase of the spreading sequence and attempts to despread the received signal using the hypothesized phase. In general, both the phase and the frequency of the received spread-spectrum signal will be unknown to the receiver. An energy detector at the despreader output measures the signal plus noise energy in a narrow bandwidth at a known frequency. If the hypothesized phase matches the sequence in the received signal, the wide-band spread spectrum signal will be despread correctly to give a narrowband data signal. In this case, the receiver decides a coarse synchronisation has been achieved and activates the tracking loop to perform fine synchronisation.

On the other hand, if the hypothesized phase does not match the received signal, the despreader will give a wideband output and the Band Pass Filter will only be able to collect a small portion of the power of the despread signal. Based on this, the receiver decides this hypothesized phase is incorrect and other phases will be tried.

The position in which code sequences are in-phase, henceforth leading to the acquisition state (ACQ), is referred to as a *synchro cell*. The remaining out-of-phase positions between codes correspond to *nonsynchro cells*.

Due to the presence of thermal noise, the operation of the detector is not perfect. Two parameters associated with the detector performance are the *detection probability* P_D and the *false alarm probability* P_{FA}. The Mean acquisition time is a function of the probabilities P_D and P_{FA}. The actual relationship between these detector probabilities and the dwell times will depend upon the type of detector which is used.

Acquisition time is the duration taken to lock up the receiver from the start of the search and is an important measure of the receiver's performance. In many instances, the time required to achieve synchronisation is sufficiently important to impose severe limitations upon the communications system. Under such circumstances, it is desirable to obtain an optimum sync search procedure [10] such that:

- the average time required to obtain sync is minimized, and
- for a given synchronisation time, the probability of obtaining sync is maximized.

Given the received signal $r(t)$ and the locally generated code replica $c(t)$, the receiver will a apply search strategy to determine the position in which code alignment occurs. Each relative position between the codes is called a cell[11]. The uncertainty region is defined as the total number of cells to be searched. Cells are tested by correlating the received and locally generated codes over a dwell time T_d, i.e. the time it takes to test a cell..

One additional parameter may also be required. This parameter is the *penalty time*, T_p, for a false alarm[12]. When a false alarm does occur it is assumed that it can be recognized in some manner[1] and the search for the proper cell will resume after a delay of T_p seconds. Moreover, the search will continue at the cell following the one producing the false alarm. It can also be assumed that the penalty time T_p can be approximated as an integer multiple of the dwell time [12]

$$T_p = KT_d \tag{4}$$

According to [14], the sequence timing can be acquired and tracked accurately, and hence the spreading can be done , without the knowledge of carrier phase and with only a rough estimate of carrier frequency.. The timing and frequency uncertainty

[1] One common way to identify a false alarm is the failure of the fine synchronization system to lock, the indication of which is provided by a lock detector [13].

region may be graphically depicted by a rectangle with dimensions $\Delta T \times \Delta F$ as in Figure 4 . It will be assumed that a priori information is available to the receiver which bounds the time uncertainty to a range of ΔT seconds and the frequency uncertainty to a range ΔF rad/s [5].This rectangle can be subdivided into smaller rectangles whose dimensions Δt and Δf are the range around the correct phase/frequency over which the system will yield a *hit*.

Their total number of uncertainties is equal $q = \left(\frac{\Delta T}{\Delta t}\right)\left(\frac{\Delta F}{\Delta f}\right)$. For accurate estimates the sizes of Δt and Δf should be small as possible, however should not be too small as that would overburden the receiver complexity or extend search time.

2.1.1 DETECTOR STRUCTURES

In order to determine whether a cell corresponds to the synchro position or not, the received signal $r(t)$ is correlated with the locally generated delay-controllable version of the same code[7, 15, 16]. The operation is denoted as

$$\lambda\left(\hat{\zeta}\right) = \int_0^{\tau_d} r(t)\,c\left(t + T_c\hat{\zeta}\right)dt \tag{5}$$

which is the time domain correlation operation between the observed waveform and the local code, positioned at the candidate offset $\hat{\zeta}T_c$. Where τ_d is the *dwell time*. This correlation result is thus a sufficient statistic for estimating ζ .

In principle, two basic approaches detection are possible, namely *coherent* or *noncoherent detection*. Code acquisition in the absence of carrier phase information is designated as noncoherent code acquisition. In some cases there is enough SNR before dispreading to activate the tracking loop, in the case the coherent code acquisition (knowledge of θ_c and/or ω_D) can be used.

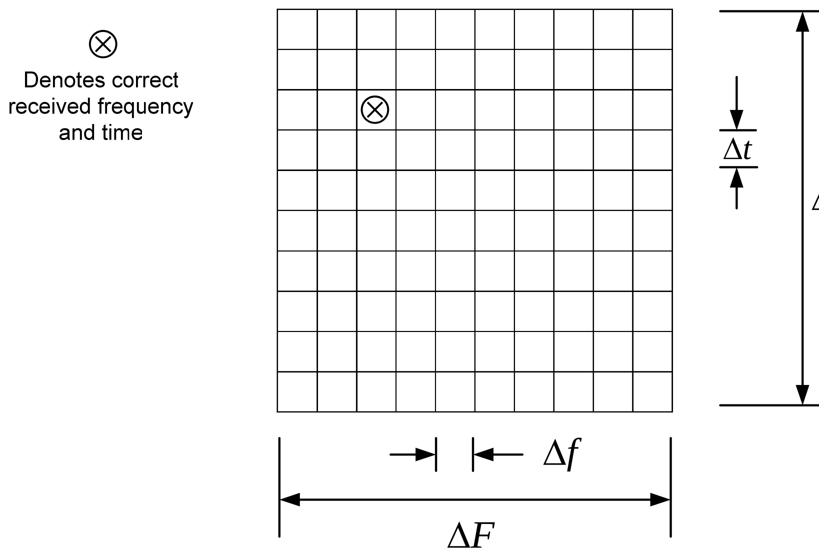


Figure 4 Phase and Frequency uncertainty region

In general, coherent detection is not used in the context of code acquisition due to this requirement of carrier phase information for the operation of coherent correlation. Another class of receivers are the differentially coherent receivers [17]. Like the noncoherent receivers, the differentially coherent receivers do not require prior knowledge of the carrier phase.

For the noncoherent receiver the ML estimator of ζ , the code phase , given that the received signal *r(t)* is observed during τ_d seconds, is the $\hat{\zeta}$ maximizing

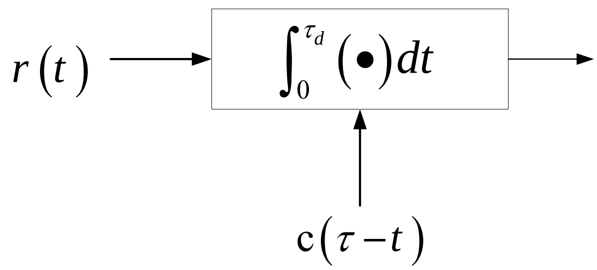
$$\lambda(\zeta) = \sqrt{\lambda_c^2(\zeta) + \lambda_s^2(\zeta)} \tag{6}$$

where

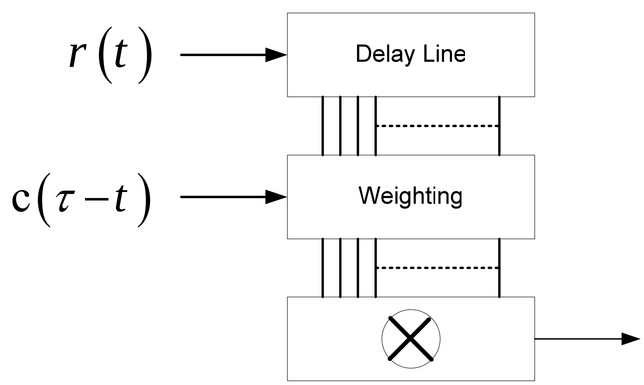
$$\begin{aligned} \lambda_c(\zeta) &= \sqrt{2}\int_0^{\tau_d} r(t)\, c(t;\zeta)\cos(\omega_c t + \omega_D t)\, dt \\ \lambda_s(\zeta) &= \sqrt{2}\int_0^{\tau_d} r(t)\, c(t;\zeta)\sin(\omega_c t + \omega_D t)\, dt \end{aligned} \tag{7}$$

2.1.1.1 ACTIVE AND PASSIVE DETECTORS

The basic unit in any acquisition receiver is the decision making device (detector). An energy detector is the structure used to carry out the correlating operation defined in equations (5) and(6). The detector plays a fundamental role in the performance of the acquisition process and its task is to detect with a high degree of reliability the presence of synchro or nonsynchro cells. If the phase and frequency of the local generated code are correct, the received signal will be despread, and translated to the central frequency of the bandpass filter, and the energy detector will detect the presence of signal.



(a)



(b)

Figure 5 (a) Active Correlating unit (b) Passive correlating unit (Code Matched Filter)

The acquisition receiver can be implemented using either an *active correlator*, where the correlation is performed serially on a *chip-by-chip* basis, or a *passive correlator* (*matched filter*), where the correlation is performed on chips in parallel[18].

In active acquisition, the received signal is correlated with a replica of the code generated by the receiver while in the passive method, a filter matched to the spreading code is used[1].

Active Correlator

The active correlator can be seen as a minimum complexity approach where only a simple correlating unit is employed[11]. As before, the receiver hypothesizes a phase of the spreading process. The despread signal is bandpass filtered with a bandwidth roughly equal to that of the narrowband data signal. The output of the bandpass filter is squared and integrated for a duration of τ_D to detect the energy of the despread signal.

Matched Filter Energy Detector

The receiver hypothesizes a phase of the spreading sequence to generate a reference signal $c\left(t+\hat{\zeta}T_C\right)$ for despreading. The PN *matched filter* is a passive device that stores a block of N PN code chips of duration T_C, and coherently integrates over a signal period equal to the equivalent block length, NT_C. A matched filter can be implemented such that the incoming code phases move (slides) with respect to a fixed local code. Thus by continuously observing the matched filter output, we can effectively evaluate different hypothesized phases.

The overall acquisition time required for the Matched filter is much shorter than that of active correlators. However, the performance of the matched filter acquisition technique is severely limited by the presence of frequency uncertainty. As a result, matched filter acquisition can only be employed in situations where the frequency uncertainty is very small.

Though both active and passive correlating units materialize the correlation operation of (7), however there are some differences in terms of speed. If the sampling rate is N/T_C and the correlation length of the matched filter spans M_C chips, the matched filter requires $M_C N$ multiplications in every interval. On the contrary, if we use an active correlator, the same number of multiplications is required in every T_C/N interval. Whereas for the active correlator is used, the same number of multiplications is required every $M_C T_C$ interval[19]. In terms of speed, the superiority of the MF scheme is clear for large values of N, but on the other hand, its inherent complexity makes its implementation feasible only for low to moderate values of N[11].

Matched filter can be implemented either in continuous-time or discrete-time mode using charge-coupled devices, surface acoustic wave convolvers, and discrete-time correlators. Modern matched filter synchronisation systems are usually implemented digitally.

The envelope of the matched filter output is compared with a threshold after each sampling interval T_S. If the sampling interval T_S equals T_C/N the search is conducted at the rate of N/T_C sample positions (or, $1/N$ chip positions) per second after some initial latency time. However, the matched filter scheme requires more computations than active correlator does[15, 20].

2.1.2 SEARCH STRATEGIES

The search strategy is the manner by which the receiver searches through the uncertainty region. This sweep can be carried out continuously or in discrete steps. The time uncertainty region is usually quantized into a finite number of elements (cells), through which the receiver is stepped. Which particular search strategy is selected by the receiver is dependent on the nature of the uncertainty region, available prior information, statistical quality of the tests performed, availability of stepping and rewinding mechanisms, etc. [9]. See [7, 9, 13, 21, 22] for more description of the different search strategies.

Code Acquisition search schemes can be classified as serial, parallel or the hybrid.

2.1.2.1 PARALLEL SEARCH

Parallel search makes use of a larger number of correlating elements. In one extreme the receiver could use p correlating elements to simultaneously search the q cells composing the uncertainty region. This will largely reduce the acquisition time, but on the other hand, implementation complexity of such a receiver will increase with q, being unpractical for long spreading codes. The maximum likelihood (ML) approach to code acquisition can be seen as a method in which the timing information is obtained from the received signal by a concurrent testing of all possible cell positions[11, 15]. A detector performs simultaneous correlation between the received signal and each of the locally generated realizations of the code sequence. Such an approach is practically feasible in cases where uncertainty regions, q are considerable short, as with short spreading sequences.

In general, when the uncertainty region is small to moderate in size, parallel acquisition brings only little improvement in performance as compared to serial-search whereas the difference becomes more marked for large uncertainty regions.

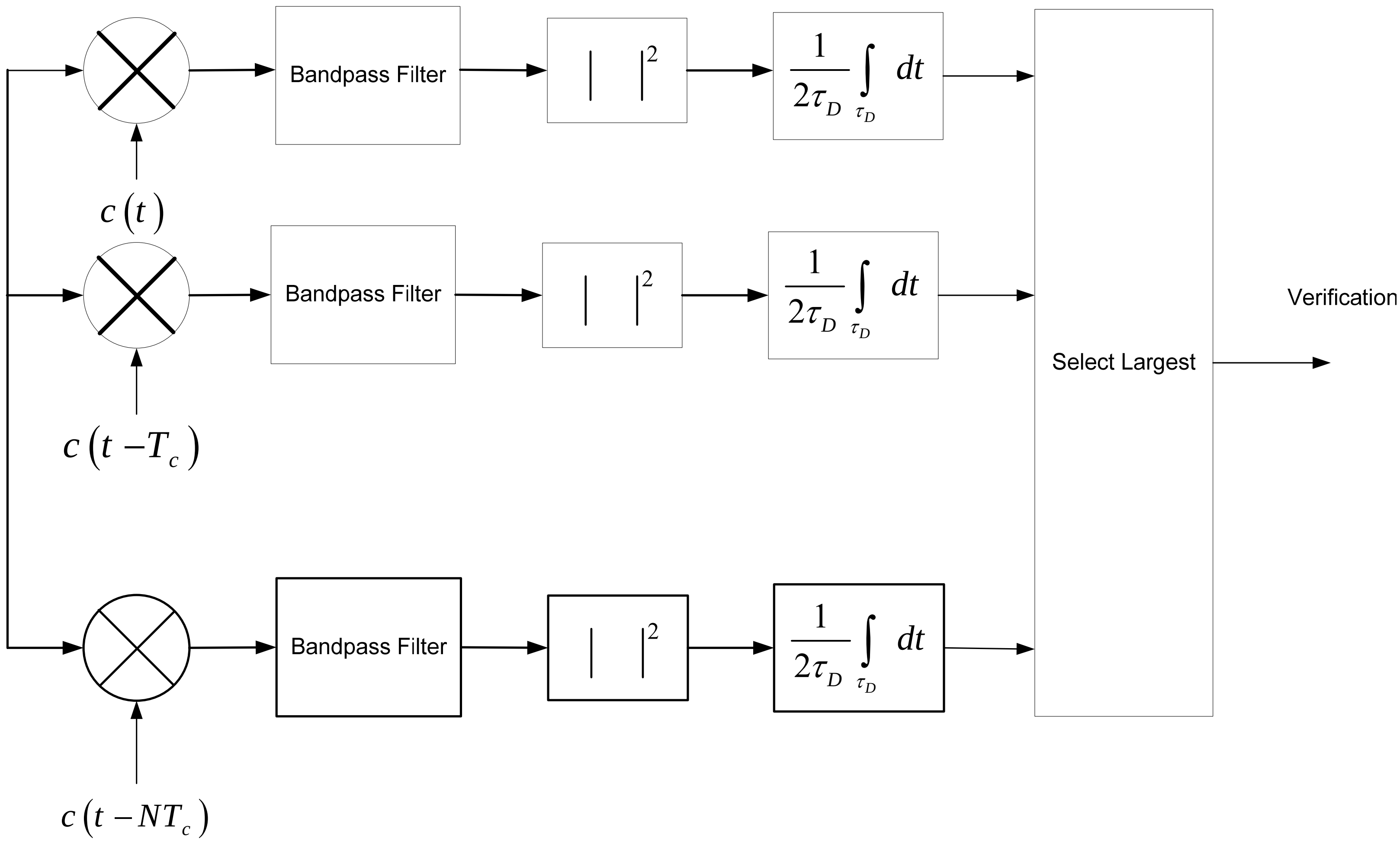


Figure 6 Parallel Search Circuit[23]

2.1.2.2 SERIAL SEARCH

The practical implementation of the parallel scheme is usually prohibitive, especially when the uncertainty region is large. A simpler alternative is to serially test all the possible code positions and then choose the position with the largest corresponding detector output. This reduces implementation complexity however, at the expense of longer periods to make decisions.

This acquisition strategy is the *serial search* (Figure 7). In this method, the acquisition circuit attempts to cycle through and test all possible phases one by one [24] (serially) starting from an arbitrary cell or by some prior knowledge about the synchro cell and proceeds in a specified manner. At each phase, the detector output compared with a threshold and the process is repeated until a correct phase alignment is detected. The serial search techniques are by far the most commonly used spread-spectrum techniques[5].

The penalty time associated with a missing the correct phase (miss) is large. Therefore we need to select a larger integration (dwell) time to reduce the miss probability. This, together with the serial searching nature, gives a large overall acquisition time (i.e., slow acquisition).

In cases where the probability density function of the synchro cell is known. If the probability distribution function of the synchro cell is known, then the most likely phase cells should be searched first and then the less likely. The main serial search strategies are the straight search, the ζ–search and the expanding window search. A structured classification of serial-search strategies and their analysis, including ζ–search and expanding window approaches, were studied in [9, 16, 18, 22, 24-27].

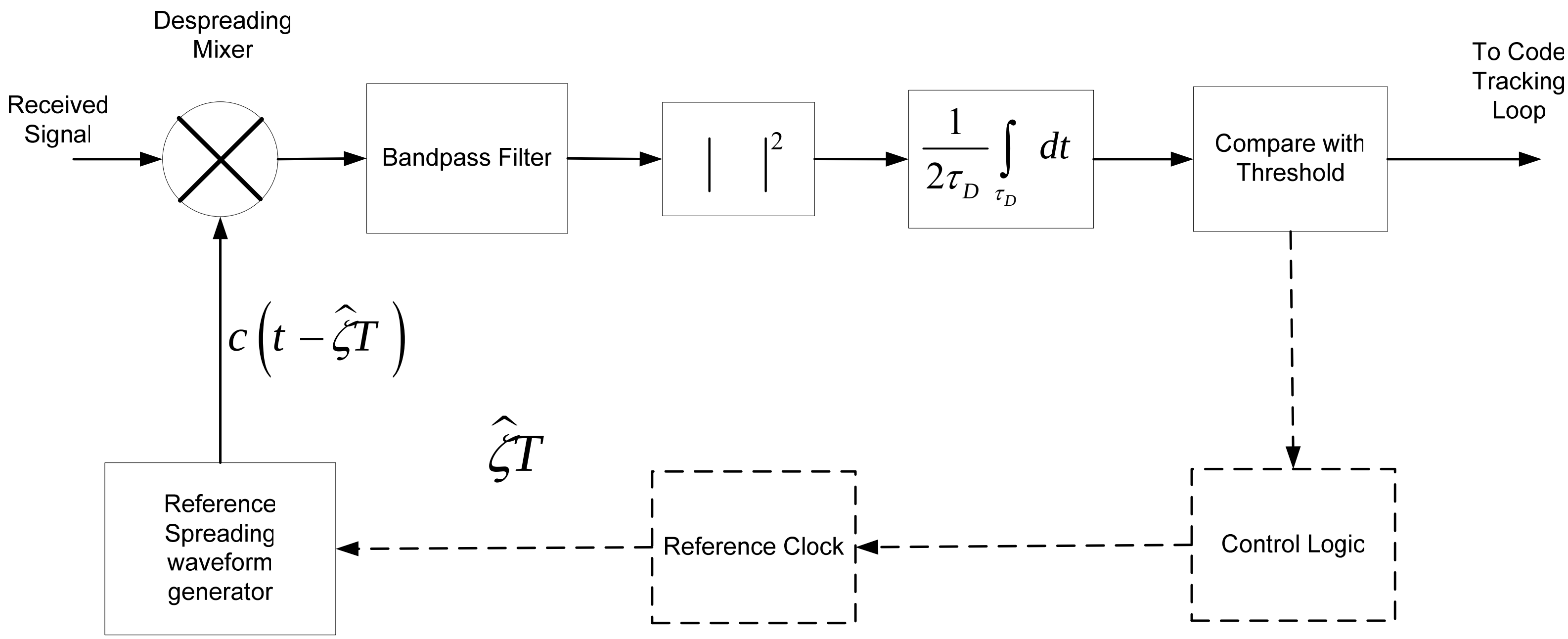


Figure 7 Serial Search Acquisition Scheme [7]

2.1.2.3 HYBRID SEARCH

A good compromise between the two previous methods is the serial-parallel search (Hybrid Search). For a total number of q cells to be searched, a block of $p : p \leq q$ cells is searched in parallel. If no synchronisation occurs the search phase is updated by p cells, until synchronisation occurs. This scheme is better than the serial search and the hardware complexity is lower compared to the full-parallel case. Hybrid serial-parallel approaches have been proposed as an attractive solution for the trade-off between acquisition speed and implementation complexity [28-31].

2.1.2.4 SINGLE AND MULTIDWELL DETECTION

Fixed Integration Time detectors can be further divided into single dwell and multiple dwell detectors. In a typical system there are far more nonsynchro cells than synchro cells. Thus, most of the time is spent in testing cells corresponding to nonsynchro positions[11]. In addition, since a false alarm state is associated with every nonsynchro cell, the time to acquire could be excessively long.

Different approaches based on repeated observation of the cells have been developed to reduce the acquisition time. The methods considered so far make a cell decision based on a single-dwell or integration. In many practical code acquisition systems, to avoid any false alarm, a second dwell, usually characterized by longer integration time, could be used upon synchro cell detection by the first dwell, to verify the correctness of the first (or tentative) decision, thus avoiding occurrences of false alarms. The presence of a verification mode is usually denoted by the nomenclature "multiple-dwell-time detectors" as opposed to "single-dwell-time detectors" for no verification[9].

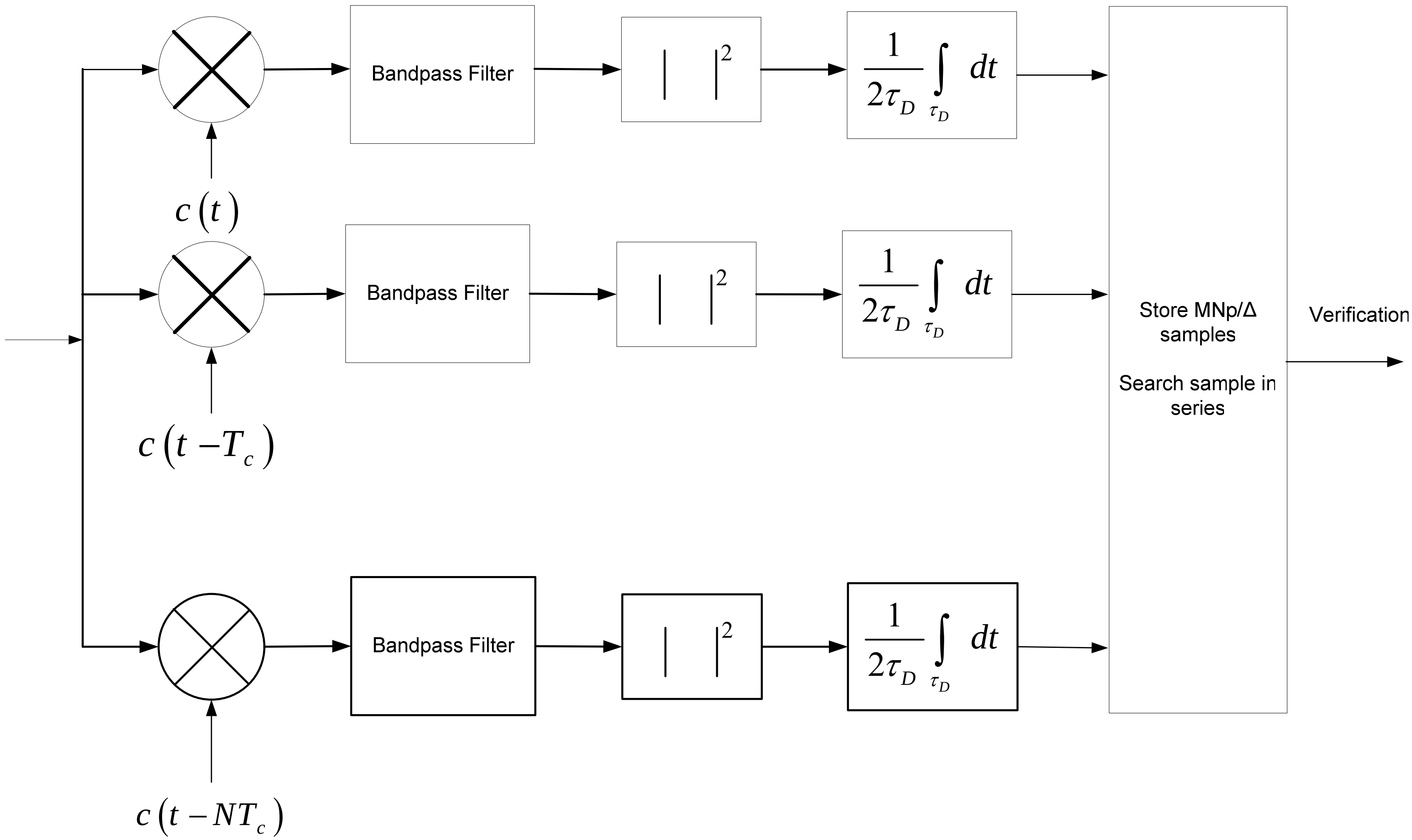


Figure 8 Hybrid Search Circuit[28]

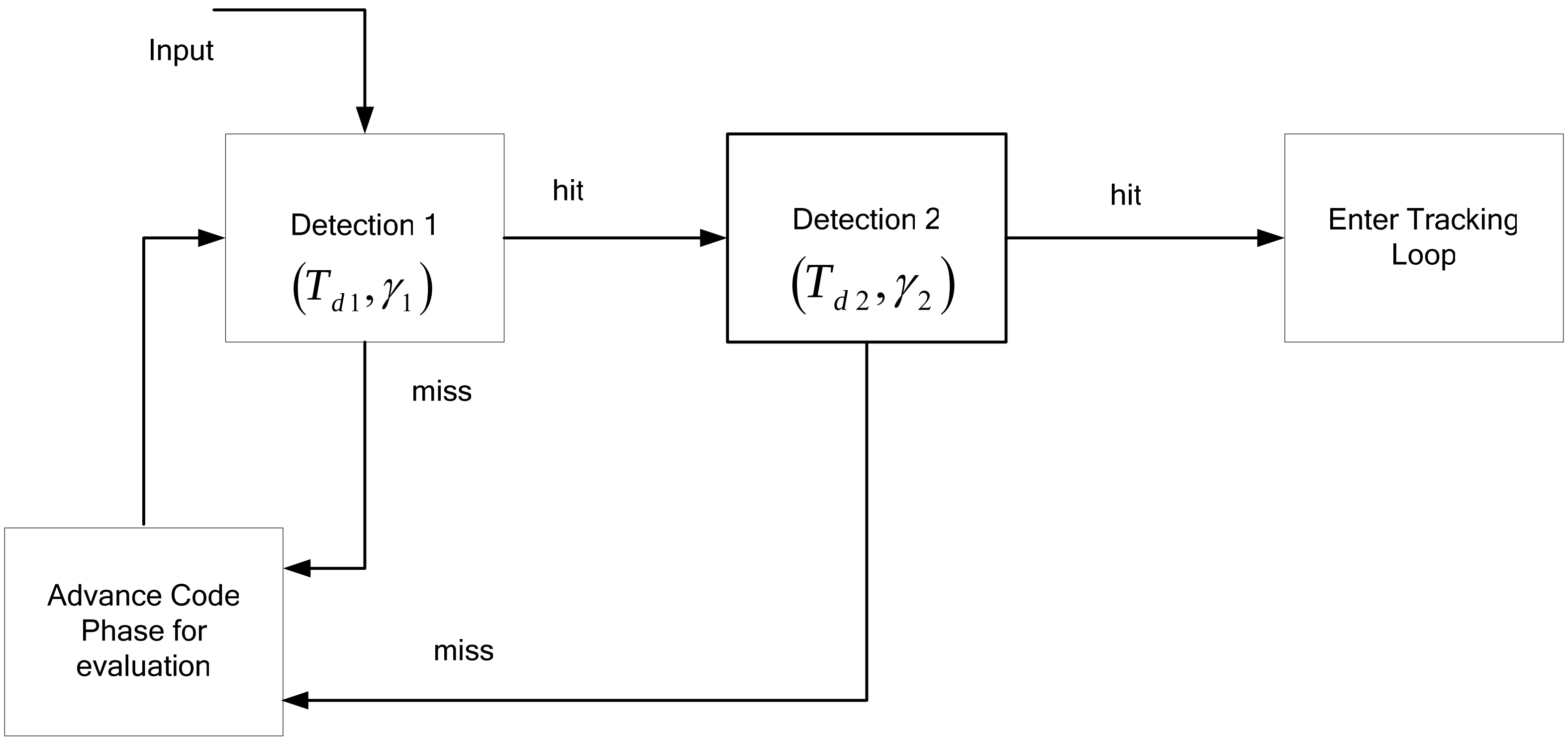


Figure 9 Multidwell search circuit[5]

The verification process alternates with the search process: it is started whenever acquisition is declared; the search is in rest during verification, and is resumed when false alarm is declared. The verification can reduce the mean acquisition time especially when the probability of false alarm of the search mode is high.

Generalization to *multiple-dwell* detectors, that is, consecutive tests to the same cell with successively increased dwell times, is straightforward. The synchro cell is declared only after all the stages result in synchro cell detection. A failure to detect a synchro cell at any dwell stage results in advancing the phase of the local code and repeating the multiple-dwell testing.

2.1.3 OTHER CODE ACQUISITION SCHEMES

Several acquisition receiver structures have been proposed and studied taking into consideration the particular performance degradation resulting from the MAI signal. A Least Mean Square (LMS) based Finite-Impulse Response (FIR) adaptive filter is used for code acquisition instead of a conventional MF in [32]. MAI is also taken into account by the LMS algorithm when computing FIR filter coefficients, acquisition–based capacity is considerably higher for that case than for the MF approach (MF is matched to the spreading code but not to the MAI signal)[11]. A robust adaptive receiver for code acquisition is proposed and studied in [33]. As a filter adaptation the authors consider Recursive Least Square (RLS) and LMS algorithms. The RLS approach is found to provide much better resistance to MAI than the LMS algorithm. In [34], a blind code acquisition scheme using adaptive linear filtering based on a linearly constrained constant modulus algorithm (CMA) is proposed.

A narrow-band interference suppression scheme for code acquisition is considered in [35]. Code acquisition in a CDMA overlay scenario where the spectrum is shared with some narrowband users is studied in [36]. A linear prediction filter previous to the MF is used to reject carrier energy from the narrowband users. Considerable gains in terms of reduced acquisition times are obtained by interference suppression filters. Gains are reduced for increased bandwidth of the interference signals. The method exploits transform-domain techniques by which the received signal previous correlation is Fourier transformed, filtered in the frequency domain and inverse Fourier transformed. Analytical results show that the presence of the suppressor filter helps improving both probabilities of detection and false alarm. An interference-cancellation aided code acquisition in asynchronous CDMA networks was studied in [37]. Two approaches exploiting iterative interference cancellation are considered, based on removing at each stage either the strongest interference component or all the estimated components. The removed signals in a given stage are not estimated again in the following stage. The authors showed that both schemes are effective to combat the degrading near-far effects, though the second approach provides better performance. A parallel multistage interference cancellation approach for enhancing synchronization performance is proposed in [38], where it is shown that acquisition times are reduced as the number of interference cancellation stages is increased. A case where the algorithm adaptation uses a known training sequence is studied in [39] whereas a blind approach is considered in [40, 41]. It is found that the minimum mean square error approach is also effective to combat the near-far problem at the initial acquisition stage. In [42] use a parallel interference canceller (PIC) to suppress the MAI contribution of all synchronized users from the received signal. The authors show that this scheme allows the accommodation of more users in the network, increasing thus capacity.

Acquisition receivers robust to interference have also received considerable attention. Based on subspace decomposition techniques, the problem of delay estimation in scenarios affected by the near-far effect were considered in [43, 44] [41]. In [45], the problem of blind and joint acquisition and demodulation was considered, results show that the proposed receiver is near-far resistant. A maximum likelihood delay estimation approach where timing information is obtained from the sample mean and sample covariance matrix of the received signal is studied in [41]. A maximum-likelihood interference tolerant delay estimation scheme particularly suitable for CDMA networks with long code sequences in proposed and analysed in [46].

2.1.4 NON-COHERENT DETECTION

For the non-coherent case, under the H_0, y is zero mean and it is χ^2 distribution with two degrees of freedom, whose probability density function is given as [47]

$$p\left(y|H_0\right)=\int_{Th}^{\infty}\frac{1}{2\sigma^2}\exp\left(-\frac{x}{2\sigma^2}\right)dx=\exp\left(-\frac{Th}{2\sigma^2}\right) \tag{8}$$

In the case of H_1, the random variable is the sum of the squares of two independent Gaussian variables of variances σ^2, in this case y has a non zero mean and it is χ^2 distributed

$$p\left(y|H_1\right)=\int_{Th}^{\infty}\frac{1}{2\sigma^2}\exp\left(-\frac{\left(y+x\right)}{2\sigma^2}\right)I_0\left(\frac{\mu\sqrt{y}}{\sigma^2}\right)dx=Q\left(\frac{y}{\sigma},\frac{Th}{\sigma}\right) \tag{9}$$

Q is the Marcum's Q-function.

Where y is the decision variable, μ and σ are the mean and the is the variance of the noise respectively.

2.1.5 PERFORMANCE MEASURES

A number of sources contribute to the randomness of the acquisition process[9]

1. initial uncertainty about the code phase offset
2. channel distortion
3. possible presence of random data
4. unknown carrier phase and carrier offset (Doppler)
5. front-end receiver noise.
6. partial correlation between the received code and the local replica.

In investigating T_{acq}, one can distinguish two basic scenarios[9].

1. The case where no absolute time limit or termination time exists on the acquisition time T_{acq}. This situation arises when data are always present in the received waveform.
2. In certain systems where data transmission starts after a certain time interval from the initial system turn-on, it is imperative that code acquisition be performed in that interval with very high probability; otherwise, communication is impossible.

The above two possibilities constitute a first partition of the class of acquisition receivers. Of course, in both scenarios, the measure of performance is statistical in nature and is imbedded in the knowledge of the probability distribution function $F_{acq}(t)$) of the random time T_{acq}. Having $F_{acq}(t)$, one can then derive and optimize any meaningful performance parameter, such as the mean acquisition time $\overline{T}_{acq}$ (case 1) or the probability of prompt acquisition $\Pr\{T_{acq} < T\}$ (case 2).

Another important measure of system performance is the overall probability of code detection P_d. If the acquisition process terminates only when the acquisition cell is identified, the detection probability P_d is one. Another deadlock can sometimes be created due to a false-alarm situation (FA). By false alarm, we indicate the case where the acquisition mechanism erroneously decides that code synchronisation has occurred and fine tracking via a code-tracking loop is initiated. The time required for the tracking loop to indicate false lock and the acquisition system to resume search is a random variable and can be modelled as such. In this case, we refer to the false alarm state as a *returning state* associated with a random Penalty time T_p.[1]. In Some other cases, however, the delay involved in initiating the tracking loop is catastrophic for the system operation. *False Alarm* then corresponds to an *absorbing state*, whereupon reaching it results in complete loss of code acquisition (in other words, a final miss)[7].

In the presence of both *system-deadlock* situations, namely, finite acquisition stopping time, and absorbing false-alarm state, P_M is the sum of the two probabilities:

$P_M = \Pr\{FA\ before\ T\} + \Pr\{neither\ FA\ nor\ acq\ before\ T\}$ where ACQ indicates the correct acquisition-absorbing state.

Similarly, for case 1 (no stopping time), but with absorbing False Alarm, it follows that $P_M = \Pr\{FA\ occurs\ some\ time\}$

[1] For computational ease, T_p can also be modelled as a fixed, known time[12].

2.2 CODE ACQUISITION ANALYSIS TOOLS

Acquisition time is a random variable which depends on the starting position of the code, the number of false alarms, and the number of times the true code phase is missed. Although the mean and variance of the acquisition time may provide a description of the acquisition process[8] a complete evaluation of the acquisition performance must utilize the probability distribution of the acquisition time[48]. Two approaches have been previously used to obtain statistics of the acquisition time: time-domain techniques and transform-domain techniques (or circular-state diagram). The fact that code acquisition performance can be characterized by a number of cell parameters allows in general independent optimization of the detector structure and selection of the search strategy[16].

Transform-domain techniques exploit the duality existing between the state transition diagram of a discrete time Markov process and the flow graph of electrical systems [9]. Historically, the flow graph technique was first applied to statistically characterize the code acquisition process in [13].

A different technique for the analysis of code acquisition is employed by time-domain analysis, also known as the *direct approach* [5, 7, 15, 16]. The same results can be obtained by considering code acquisition as a combinatorial process, as shown in [5, 49]. In general, time domain analysis is relatively simple but results become cumbersome rather quickly [16].

Flow graphs methods attain considerable dimensions, modelling and analysing a particular code acquisition scheme is usually a cumbersome task. Any changes in the system require an overall reworking of the model; hence flow graphs techniques are not appropriate for generalizations. A more systematic approach was proposed by [9] by introducing the concept of a circular diagram. Exploiting the fact that the acquisition process repeats itself after all the cells have been visited, in [9], the flow diagram was arranged in a circular fashion. In this manner systems with different search strategies and detector structures can be modelled with a similar scheme.

As compared to the flow graph approach, the circular diagram method offers a more systematic and less tedious procedure to theoretically evaluate code acquisition performance.

2.3 EFFECTS OF NOISE ON CODE ACQUISITION

After despreading at receiver, the overall received signal will have three components: the useful signal, the multiple access interference (MAI), and Gaussian noise[50]. The useful signal will be proportional to the autocorrelation function of the code and will have high value if the input code and the locally generated replica are synchronised; otherwise the value of the autocorrelation is close to zero (for the ideal PN codes). The MAI is proportional to the cross correlations between the despreading code and the codes of all other users in the network.

The presence of noise causes two different kinds of errors in the acquisition process:

1. A *false alarm* occurs when the integrator output exceeds the threshold for an incorrect hypothesized phase.
2. A *miss* occurs when the integrator output falls below the threshold for a correct hypothesized phase.

A false alarm will cause an incorrect phase to be passed to the code tracking loop which, as a result, will not be able to lock on to the DS-SS signal and will return the control back to the acquisition circuitry eventually. However, this process will impose severe time penalty to the overall acquisition time. On the other hand, a miss will cause the acquisition circuitry to neglect the current correct hypothesized phase. Therefore a correct acquisition will not be achieved until the next correct hypothesized phase comes around.

The time penalty of a miss depends on acquisition strategy. In general, we would like to design the acquisition circuitry to minimize both the false alarm and miss probabilities by properly selecting the decision threshold and the integration time.

2.4 CODE ACQUISITION IN FADING CHANNELS

In most practical radio channels, due to multipath propagation and the relative movement between transmitter and receiver, the attenuation and phase shift imposed by the channel cannot be considered as time invariant.

The signal $r(t)$ at the front end of the acquisition receiver contains the modulated code (possibly with data) plus noise,

$$r(t) = \sqrt{2P}\alpha(t)d(t)c(t+\zeta T_c)\cos(\omega_c t + \omega_D t + \theta_c) + n(t) \tag{10}$$

ω_c and θ_c are the nominal carrier radian frequency and random phase, respectively, P is the transmitter signal power, $n(t)$ represents AWGN with one-sided power spectral density N_0, $d(t)$ is the data sequence, α is a coefficient representing the fading process of the channel, ω_D is the Doppler radian frequency.

Performance of code acquisition in time-varying channels is computed by shaping the results obtained for the static channel model with the statistical behaviour of the channel. A common way of categorizing fading channels is according to their speeds is to compare the channel coherence time to the duration of a symbol[3, 4]. However, from the synchronisation problem standpoint, it is more reasonable to compare the coherence time T_{coh} against the duration of the acquisition processing itself T_{acq}.

Temporal fluctuations in the channel have a direct impact on the signal correlation process. This is particularly true at the synchro position, where due to the effect of fading, signals appear to be less correlated and consequently the probability of detecting that cell (P_D) will be reduced. Since the correlation at nonsynchro positions is inherently low (ideally zero), this value is not affected much by the time-varying channel, hence the effect of fading on P_{FA} is in general minor. Due to the mentioned reasons the fading process tends to degrade acquisition process performance.

This is an essential condition for applying the analysis techniques discussed previously. For instance, homogeneous Markov chain theory, applied by the flow-graph techniques, is based on the assumption that transition probabilities associated with the cells remain constant during the acquisition process. Some new analytical approaches considering the correlation incurred by fading (e.g., channel memory) have been proposed and are discussed later.

2.5 THRESHOLD SETTING

The decision whether a particular cell corresponds to the synchro or nonsynchro position is carried out by comparing the detector output to a threshold value T_h. The threshold is a reference value used as the cell acceptance (or refusal) criterion. Depending of the statistical test approach employed by the detector, the threshold could be set fixed for all the cells or could be cell-dependent. The former case, based on the Neyman-Pearson decision criterion, attempts to keep P_{FA} fixed and is used when no *a priori* information on the synchro cell is available. The latter, applying the Bayes test criterion, is used when *a priori* information on the synchro cell is available.

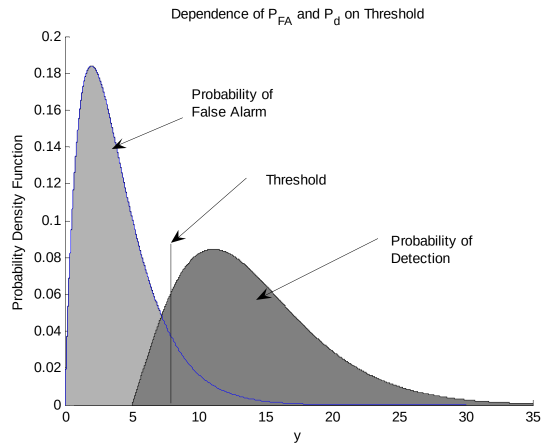


Figure 10 Dependence of P_{FA} and P_D on threshold value Th

Since the probabilities P_{FA} and P_D depend both on the T_h value, special care should be taken when setting the threshold to obtain a desired acquisition performance. In fact the average time to acquire depends, among others, on the level of prevailing *SNR*, P_{FA} and P_D. If the threshold is set too high, P_{FA} will get smaller but so will P_D. Too low of a threshold, on the other hand, will beneficially increase P_D but will result also in a higher P_{FA}. It is significant to note that P_D and P_{FA} are important only in how they affect the more meaningful measures of performance such as the mean time to acquisition and the mean hold-in time[8].

2.6 HYPOTHESIS TESTING IN CODE ACQUISITION

Acquiring synchronisation in CDMA systems involve testing binary hypotheses indicating whether the synchronisation is achieved or not[47]. The model we consider assumes that the received signal *r(t)* is observed in some time interval.

It also assumes that the conditional probability distributions of the process *r(t)* are known for the observations under the two hypotheses. In the case of the acquisition problem we denote these hypotheses H_1 (synchronisation is achieved) and H_0 (synchronisation is absent).

The simple acquisition problem arises from "genie-aided" acquisition in the DS-CDMA system. It is assumed that the desired user sends a known periodic BPSK signal to the acquisition receiver. The receiver searches serially through all potential time offset shifts of ζ_k. If synchronisation is not established (hypothesis H_0), $E[z_n]=0$; otherwise, (hypothesis H_1) $E[z_n]\neq 0$. The variance $\operatorname{var}(z_n)$ is in both cases determined by background noise and the other user interference signals.

The Neyman–Pearson criterion gives the optimal solution for these two problems.

Neyman–Pearson Hypothesis Testing

In the binary hypothesis problem we know that either one or the other hypothesis is true. The probabilities of the decision errors depend on the decision criterion we select. In the acquisition problem there is the probability to accept the hypothesis H_1 when H_0 is true, P_F - the *false alarm probability* and to accept the hypothesis H_0 when H_1 is true, P_M - the probability of a miss. The probability $P_D = 1 - P_M$ is the probability of detection. In general, it is desired to make P_F as small as possible and P_D as large as possible. An obvious criterion is to constrain one of the probabilities and maximize (or minimize) the other. The Neyman–Pearson criterion maximizes P_D (or minimizes P_M) under the constraint $P_F \leq \alpha$, where α is a pre determined constant. The solution is obtained easily by using Lagrange multipliers[47]. Say the function Φ

$$\Phi = P_M + \lambda\left[P_F - \alpha\right] \tag{11}$$

Or

$$\Phi = \int_{Z_0} p_1(z)\,dz + \lambda\left[\int_{Z_1} p_0(z)\,dz - \alpha\right] \tag{12}$$

Here the total observation space Z is divided into two parts, Z_0 and Z_1.Whenever an observation falls into Z_0 we say H_0; whenever an observation falls into Z_1 we say H_1. Clearly, if $P_F = a$, then minimizing Φ minimizes P_M and

$$\Phi = \lambda(1-\alpha) + \int_{Z_0} \left[p_1(z) - \lambda p_0(z) \right] dz \tag{13}$$

The Neyman–Pearson criterion leads us to the likelihood ratio test

$$\Lambda(z) = \frac{p_1(z)}{p_0(z)} \underset{H_0}{\overset{H_1}{\gtrless}} Th \tag{14}$$

where Th is a numerical threshold that depends on the constraint on P_F. If the ratio exceeds ψ, the hypothesis H_1 is chosen, whereas if it does not, the hypothesis H_0 is chosen. In fact, for any positive λ the likelihood ratio test will minimize Φ.

This follows directly because to minimize Φ we assign a point z to Z_0 only when the term in the bracket of (13) is negative. This is equivalent to the test (14) with $Th = \lambda$. To satisfy the constraint we choose λ such that $P_F = \alpha$.

2.7 CODE ACQUISITION IN CDMA NETWORKS

The effect of the multiple access interference contribution from the $K_u - 1$ users sharing the network, distinctive of a CDMA network, has to be incorporated in the analysis of code acquisition. As a result, the decision variable at the detector output contains three main terms: an autocorrelation term, product of the correlation between received and locally generated codes, a cross correlation term, containing the sum of particular cross correlations between other users and the local code, and a noise term

If a large number of users K with the same bit-rate is assumed then the Central Limit Theorem can be applied to characterize the Multiple Access Interference (cross correlation term) with a Gaussian distribution with zero mean and variance proportional to K. In this fashion the effect of the Multiple Access Interference (MAI) is modelled as a net increase of the noise power affecting the acquisition process. This is the philosophy used in most literature [28, 51-53] dealing with this subject.

The received signal in may be written as

$$r(t)=\sum_{k=1}^{K}\sqrt{2P}d\left(t+\zeta T_c\right)c\left(t+\zeta T_c\right)\cos\left(\omega_c t+\omega_D t+\theta_c\right)+n(t) \qquad (15)$$

where K is the number of users and as before we assume that the receiver is trying to acquire synchronism on c_1, and that $d(t)=1$. Using a noncoherent matched filter detector, the test variable for the generic cell is given by

$$z=x_c^2+x_s^2 \qquad (16)$$

where the $x_i, i=c,s$, are the outputs from the matched filters in the in-phase and quadrature branches, respectively. It holds

$$x_i=\int_0^{LpTc} r(t)c\left(t+\hat{\zeta}T_c\right)\cos\left(\omega_c t+p_i\frac{\pi}{2}\right)=y_i+I_i+\eta_i \qquad (17)$$

where $i=c,s$ and $c\left(t+\hat{\zeta}T_c\right)$ is the locally generated sequence. y_i is the desired signal, I_i is the MAI contribution, and η_i is additive noise.

The total variance of the disturbance on each receiver branch is therefore

$$\sigma^2=\sigma_\eta^2+\sigma_I^2 \qquad (18)$$

In essence, we can model the signal out of the matched filter in each branch as given by a useful part y_i and an additive Gaussian noise. More discussions on this contribution are given in chapter 5.

2.8 DECISION CRITERIA

Two criteria can be used in determining if the correct phase has been achieved; *threshold crossing* criterion and *MAX* criterion[51].

An estimate $\hat{\zeta}$ can be obtained by comparing the test variables from all cells in the uncertainty region and selecting the maximum. The MAX criterion, which can be specified as:

$$\text{choose cell } q(j) \text{ if } y(j) > y(i), \text{ for } i = 1, 2, \cdots N_c$$

However, testing all cells serially takes time. Therefore, it may be convenient to perform the decision on cells, tested in a specified order. In this case, the threshold crossing (TC) criterion can be used, for which the test variable is compared to a threshold T_h and, if the threshold is crossed, the hypothesis of having acquired synchronism is made (and possibly subsequently verified). The TC Criterion can be defined as

$$\text{choose cell } q(j) \text{ if } y(j) \geq T_h, \text{ else go to the next cell}$$

The threshold crossing criteria is simple to implement, but setting the threshold level is problematic; in a fading channel or in the presence of high noise levels, the acquisition algorithm can either overlook a user or falsely acquire when no user is present.

While a search for maximum over a single symbol window can recover the user's spreading code alignment, this result is valid only if the desired user is transmitting.

Since the maximum search algorithm will always return a peak location irrespective of the presence of a user, a single maximum search has no means of distinguishing a user peak from a peak caused by noise. Even with a transmitting user, noise could produce an alternate peak. A noise source, however, will not produce a peak that consistently appears at the same code phase over several symbols. By considering the peak position over several symbols, the persistent peak algorithm features reliable user detection while providing an accurate estimate of the user's symbol phase.

A hybrid criterion was analyzed in [51], in this case, the uncertainty region is divided into N_S sectors with S cells each, and inside a sector a cell is selected according to the MAX criterion and the test variables selected in each sector are compared to a threshold according to the TC criterion.

MAX/TC criterion:

in a sector, select cell $q(j)$ if $y(j) > y(i)$, for $i = 1, 2, \cdots S$;

choose cell $q(j)$ if $y(j) > Th$, else go to next sector.

Note that for $S = N_c$, $N_S = 1$, $Th = -\infty$ the *MAX/TC* criterion reduces to the *MAX* criterion, while for $S = 1$, $N_S = N_C$ it reduces to the *TC* criterion. It follows that the formulation found in analyzing the *MAX/TC* criterion is most general, since other cases are obtained through simple substitutions. The usefulness of dealing with the MAX/TC criterion lies in the fact that the sector size S is a very important parameter to be optimized in order to achieve minimum acquisition time for a serial architecture, or reasonable complexity for a parallel architecture.

2.9 HYBRID SEARCH SCHEME

Let $c(t+\zeta T_c)$ be a ± 1 valued L-chip-long spreading code with chip time T_c, seconds, delayed by ζT_c. The signal $r(t)$ at the front end of the acquisition receiver contains the modulated code (possibly with data) plus noise,

$$r(t)=\sqrt{2P}d(t)c(t+\zeta T_c)\cos(\omega_c t+\omega_D t+\theta_c)+n(t) \tag{19}$$

In (19) ω_c and θ_c are the carrier radian frequency and random phase, respectively, P is the transmitter signal power, $n(t)$ represents additive white Gaussian noise with one-sided power spectral density N_o, $d(t)$ is the data sequence, and ω_d is the Doppler radian frequency.

$$c(t)=\sum_{k=-\infty}^{\infty} c_k \psi(t-kT_C) \tag{20}$$

ψ is the pulse shaping filter[1]. $\{c_k\}$ is the pseudonoise code for the k^{th} user and T_C is the chip duration. For analysis purpose, it is assumed that $d(t)=1$ (data not present) and $\omega_d=0$ (no Doppler effects).

As stated earlier, the objective of code acquisition is to make the absolute phase offset (normalized by T_c) less than 1 i.e. $|\varepsilon| \triangleq |\zeta-\hat{\zeta}|<1$.

A hybrid search scheme combines features from both the parallel and serial search approaches. The search mode is best described by referring to Figure 12. It consists of a bank of N_P parallel correlators. Let Θ represent the total number of chips in the uncertainty region of T seconds, i.e. $\Theta \triangleq T/T_c$. The uncertainty region Θ, is

[1] Pulse shape will be discussed later in the next chapter.

divided into M sub sequences, where $\Theta = L/\left(N_S \times N_P\right)$. N_P and N_S are the number of parallel and serial searches respectively. The total delay of each subcode is $T_M = MT_C$ where T_c is the chip duration and its total number of taps is $q = M/\Delta$ with ΔT, delay between successive taps. The value of Δ is 2^{-n} where is n a positive integer. The value of Δ is typically ½ and this is the value that was used in this thesis.

Each correlator is matched to each of the S length subcode. The incoming code phase uncertainty region is searched in discrete steps. In T_M seconds, qN_P samples are collected and stored from the N_P parallel Matched filters, each sample corresponds to one possible phase in the uncertainty region.

Each new input data sample is collected and, together with previous $qN_P - 1$ input samples, is correlated with the N_P subcodes loaded in the N_P parallel correlators simultaneously. The process repeats q times, each time with a unique code phase offset between the incoming PN code and the subcode loaded in any correlator, until all the possible PN code phases corresponding to the N_P subcodes are tested once. The correlators generate qN_P decision variables over the period, corresponding to the qN_P possible phases, respectively. In the next period of duration T, the N_P noncoherent correlators are loaded with a new group of PN subcodes corresponding to another qN_P possible input code phases, and the correlation process continues until a coarse code phase alignment is sensed. In this way, over a period of $N_S T_M$, the N_P parallel noncoherent correlators generate Θ/Δ decision variables corresponding to all possible discrete PN code phases of the input signal. The time duration $T_M - \Delta$ for collecting the initial $qN_P - 1$ input signal samples is negligible when compared with the mean acquisition time.

The Max/Tc criterion is used as the decision criteria. If the largest of the qN_P samples exceeds a threshold Th_1, the corresponding phase is assumed, tentatively, to be the correct phase of the received signal and the acquisition system moves to the verification mode. If Th_1 is not exceeded, new qN_P samples are collected and the process is continued.

The function of the verification mode is to avoid a costly false alarm that can supply the tracking system with a wrong phase. When a phase is selected by the search mode, it is correlated for a longer time, if the value exceeds a threshold Th_2 acquisition is declared and the tracking system is enabled, otherwise the system goes back to the search mode.

2.9.1.1 SIGNAL FLOW DIAGRAM ANALYSIS OF THE HYBRID SCHEME

Definitions

P_{Di}^{s} detection probability of the ith H_1 cell in the search mode

P_{Di}^{v} detection probability of the ith H_1 cell in the verification mode.

P_{D1}^{s} detection probability in the search mode

P_{M}^{s} probability of a miss in the search mode

P_{F1}^{s} false alarm probability of the search mode in the H_1

P_{F0}^{s} false alarm probability of the search mode in the H_0

P_{F0}^{v} false alarm probability of the search mode in the H_0

Assuming that there are λH_1 cells in the search space. For the H_1 cells, the probability of detection of the ith H_1, P_{Di}^s is the probability that the output of the detector is larger than the threshold. From (8) and (9)

$$P_{Di}^s = \int_{Th}^{\infty} P\left(y\middle|H_{1i}\right)\left[\left[\int_0^y P\left(x\middle|H_0\right)dx\right]^{qN_P-1} \prod_{\substack{l=1\\ i\neq l}}^{\lambda}\left[\int_0^y P\left(z\middle|H_{1l}\right)dz\right]\right]dy \tag{21}$$

And the probability of a miss is the probability that all samples are less than Th_1 for the entire H_1 cells.

$$P_M = \left[\int_0^{Th} P\left(x\middle|H_0\right)dx\right]^{qN_P-1} \bullet \prod_{l=1}^{\lambda}\left[\int_0^{Th} P\left(y\middle|H_{1l}\right)dy\right] \tag{22}$$

The false Alarm probability in the search mode for the H_1 is

$$P_{F1}^s = 1 - P_M^s - P_D^s \tag{23}$$

The false alarm probability of the search mode in the H_0 is

$$P_{F0}^s = 1 - \left[\int_0^y P\left(x\middle|H_0\right)dx\right]^{qN_P} \tag{24}$$

The transfer function from an initial subset which is Q branches counter-clockwise from the ACQ (acquisition) state to the final destination (correct) subset is given by

$$U_i(z) = \frac{H_0{}^i(z)H_D(z)}{1 - H_M(z)H_0{}^{Q-1}(z)} \tag{25}$$

The total transfer function $U(z)$ is the moment generating function, is

$$U(z) = \frac{1}{Q}\frac{H_D(z)}{1-\left[H_M(z) - H_{F1}(z)\right]H_0{}^{Q-1}(z)}\sum_{i=1}^{Q-1}\pi_i H_0{}^i(z) \tag{26}$$

$$
\begin{aligned}
&H_D(z) = \sum_{i=1}^{\lambda} P_{Di}^{s} P_{D}^{v} z^{K_v+1} \\
&H_M(z) = P_{M1}^{s} z^{1} + \sum_{i=1}^{\lambda} P_{Di}^{s}\left(1-P_{D}^{v}\right) z^{K_v+1} + P_{F1}^{s} P_{F}^{v} z^{K_p+K_v+1} \\
&H_0(z) = \left(1-P_{F}^{s}\right) z + P_{F}^{s}\left(1-P_{F}^{v}\right) z^{K_v+1} + P_{F}^{s} P_{F}^{v} z^{K_p+K_v+1} \\
&H_{F1}(z) = P_{F1}^{s} P_{F}^{v} z^{K_v+1} \\
&H_{F0}(z) = P_{F}^{s} P_{F}^{v} z^{K_v+1}
\end{aligned}
\tag{27}
$$

Assuming a uniform apriori probability, $\pi_i = 1/Q$

$$
U(z) = \frac{H_D(z)\left[1-H_0^{\,Q-1}(z)\right]}{Q\left[1-H_M(z)H_0^{\,Q-1}(z)-H_{f1}(z)\right]\left(1-H_0(z)\right)}
\tag{28}
$$

The mean acquisition time is

$$
E\left[T_{acq}\right] = \left(\left[\frac{d}{dz}U(z)\right]\Bigg|_{z=1} + S\right)Tc
\tag{29}
$$

where S accounts for the time it takes the matched filter to get filled.

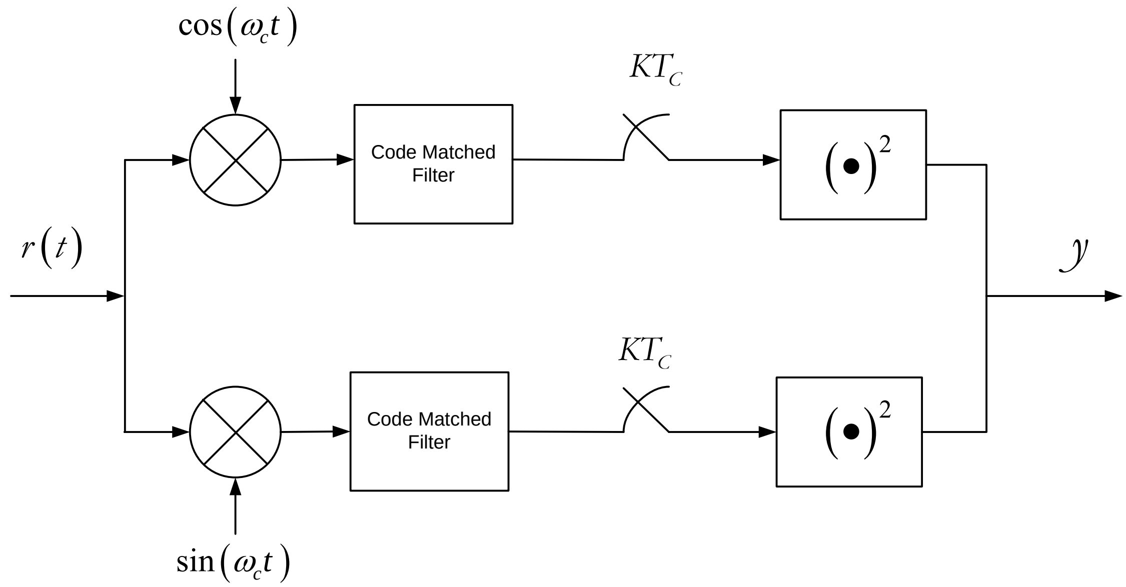


Figure 11 Non-Coherent Receiver

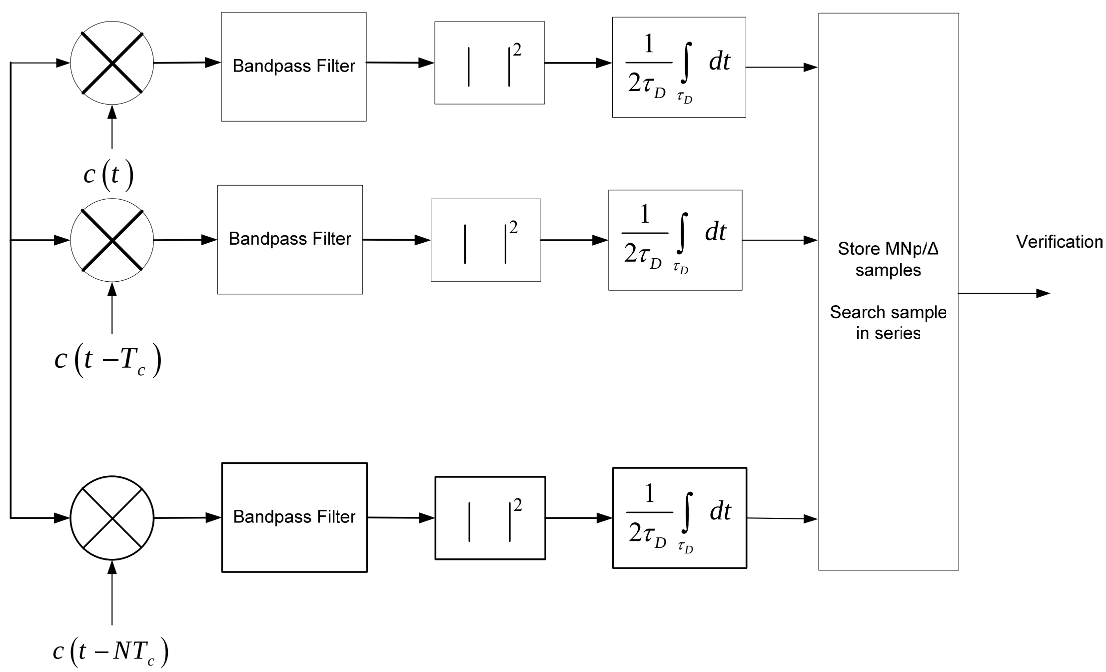


Figure 12 Hybrid Acquisition Scheme

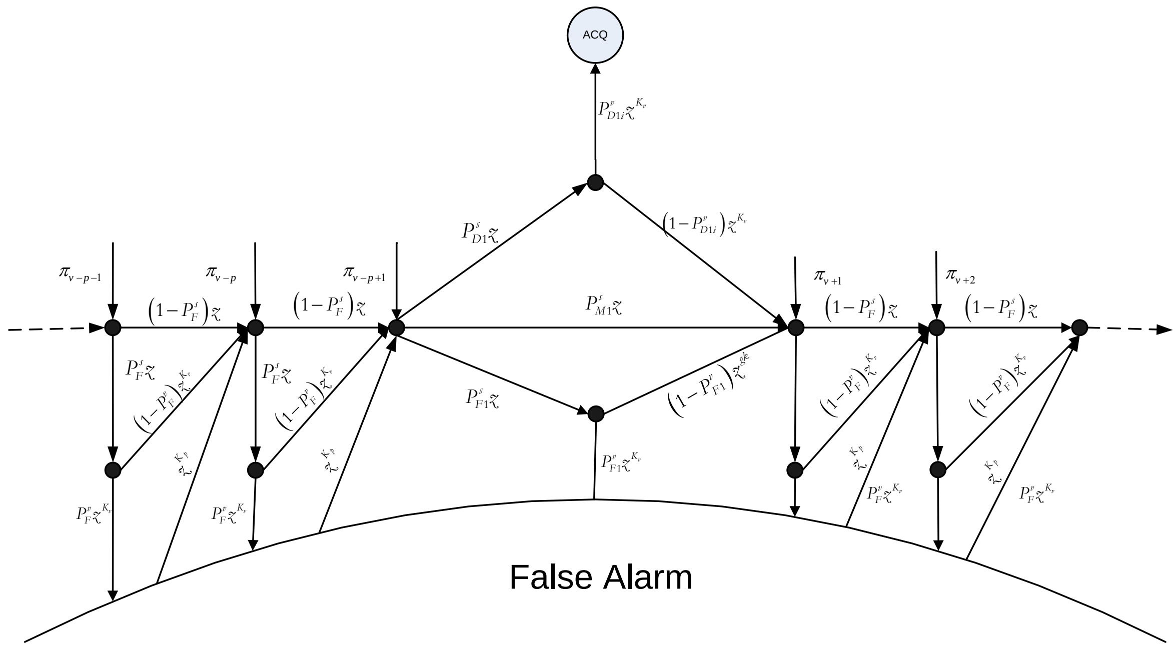


Figure 13 State Diagram of the Hybrid Scheme

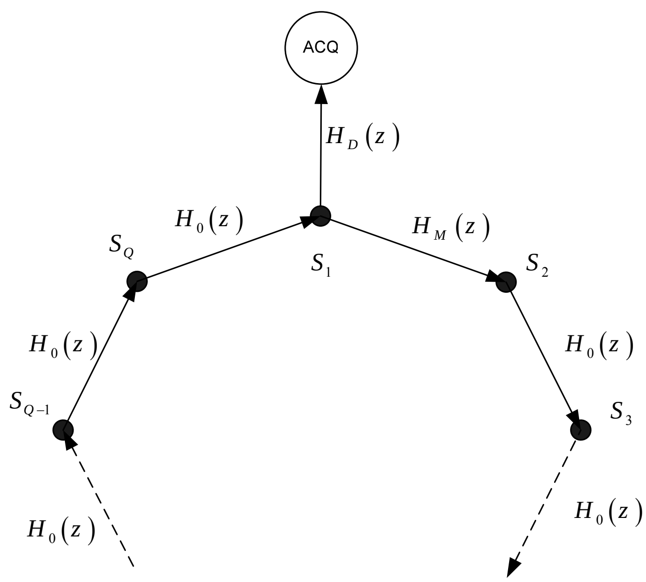


Figure 14 Reduced State Diagram

CHAPTER 3

EFFECT OF PULSE SHAPING IN CDMA CODE ACQUISITION

Pulses are designed to be either band-limited or time limited. When pulses are passed through band-limited channels, the pulses may be distorted. This distortion arises because the channels are bandlimited and hence not all frequency components present in the pulse will experience the same gain. This distortion leads to intersymbol interference or overlaps.

Pulse shaping is essential especially in Mobile communications systems where these systems operate with a large number of users given a minimal bandwidth constraint. Overlapping Pulses should have amplitudes that drop faster so that the tails of these pulses do not cause problems in the presence of jitter error, which leads to problems in synchronisation.

3.1 PULSE SHAPING IN CDMA

Chip waveform shaping has a major impact on DS-CDMA performance. This has been thoroughly analyzed from an error performance standpoint [54, 55]. The selection of chip waveforms affects not only the bandwidth efficiency, but also the performance of a DS-CDMA system[55].

The issues of pulse shaping relating to PN code synchronisation has not received as much attention despite their major impact on the overall DS-CDMA system performance. To date, there is comparatively little work on the impact of chip waveform selection on the synchronisation performance of CDMA systems. In [56] , conventional pulses (such as, rectangular, sinusoidal, raised-cosine, etc), were considered to show the effect on the probability of detection of a parallel search system. In [57], half-sine and triangular pulses are found to yield better tracking performance when combined with a rectangular pulse at the transmitter. Also, in [58] the performance of energy detectors with several known pulses is also discussed. We extend these results in our work to include a more realistic analysis of the interference contribution. We give results for the effect of different pulse shapes on the mean acquisition time and probability of detection for the hybrid search system.

The signal $r(t)$ at the front end of the acquisition receiver contains the modulated code (possibly with data) plus noise,

Following our previous notations, the received signal in may be written as

$$r(t) = \sum_{k=1}^{Ku} \sqrt{2P} d(t + \zeta T_c) c(t + \zeta T_c) \cos(\omega_c t + \omega_D t + \theta_c) + n(t) \tag{30}$$

$$c(t) = \sum_{k=-\infty}^{\infty} c_k p_{T_C}(t - kT_C) \tag{31}$$

Where p_{T_C} is the pulse shaping filter, where K is the number of users and as before we assume that the receiver is trying to acquire synchronism on sequence c_1, and $d(t)=1$. Also we use a noncoherent matched filter detector, the test variable for the generic cell is given by

$$z = x_c^2 + x_s^2 \tag{32}$$

where the $x_i, i=c,s$, are the outputs from the matched filters in the in-phase and quadrature branches, respectively. It holds

$$\begin{aligned} x_i &= \int_0^{NTc} r(t)\, c\left(t+\hat{\zeta}T_c\right)dt \\ &= \int_0^{NTc}\left[\sum_{k=1}^{Ku}\sqrt{2P}c\left(t+\zeta T_c\right)\cos\left(\omega_c t+\theta_c\right)+n(t)\right]c\left(t+\hat{\zeta}T_c\right)\cos\left(\omega_c t+p_i\frac{\pi}{2}\right)dt \\ x_i &= y_i + I_i + \eta_i \end{aligned} \tag{33}$$

where $i=c,s$ and $c\left(t+\hat{\zeta}T_c\right)$ is the locally generated sequence.

$N \le L$ is the number of chips in the correlation, $p_c=0, p_s=1$, y_i is the desired contribution, I_i is the MAI contribution, and η_i is additive noise. The various disturbances can be assumed to be statistically independent random variables[51]. The desired contribution is given by

$$y_i = \sqrt{\frac{P}{2}}\left\{\mathrm{R}_{1,1}\left(\zeta-\hat{\zeta}\right)\cos\left(\phi+p_i\frac{\pi}{2}\right)\right\} \quad i=c,s \tag{34}$$

where $\mathrm{R}_{1,1}(\varepsilon)$ is the autocorrelation function of the user signature sequence.

The noise contribution η_i is given by

$$\eta_i = \int_0^{T_c} n(t)c\left(t+\hat{\zeta}T_c\right)\cos\left(\omega_c t+p_i\frac{\pi}{2}\right)dt \tag{35}$$

and it can be described as a Gaussian random variable., with zero mean and variance equal to

$$\sigma_\eta^2 = \frac{NT_c N_0}{2} \tag{36}$$

Considering the MAI contribution I_i, the MAI results from the cross correlation between the desired users and other users.

$$I_i = \int_0^{NTc} \left[\sum_{k=2}^{Ku} \sqrt{2P} c\left(t + \zeta T_c\right) \cos\left(\omega_c t + \theta_c\right) \right] c\left(t + \hat{\zeta} T_c\right) \cos\left(\omega_c t + p_i \frac{\pi}{2}\right) dt, \quad i = c, s \tag{37}$$

The most straightforward approximation is the standard Gaussian approximation, where the MAI is approximated by a Gaussian random variable[59]. This approximation is simple, however it is not accurate in general. In situations where the number of users is not large, the Gaussian approximation is not appropriate, in depth analysis of must be applied. The improved Gaussian approximation[59, 60], provides a better approximation to the MAI term. The approximation conditions the interference term on the operation condition of each user.

Previous works on acquisition have used the standard Gaussian approximation for the MAI contribution. The Gaussian approximation to the binomially distributed correlations might be somewhat questionable for small numbers of interfering signals if one is considering the tails of the distribution, i.e. very low probabilities of erroneous detection.

The improved Gaussian approximation has been shown to estimate the variance contributed by the multiple users better than the standard Gaussian approximation especially for few number of users[59] from a probability of bit error perspective. The results in [59] were compared with “exact” results and it has been shown that the improved Gaussian approximation gives a result closer to true vales than the standard approximation.

Defining the following terms

$$\begin{aligned}
m_{\psi} &= \frac{1}{T_C^3}\int_0^{T_C} \mathrm{R}_{\psi}^2(\tau)\,d\tau = \frac{1}{T_C^3}\int_0^{T_C} \widehat{\mathrm{R}}_{\psi}^2(\tau)\,d\tau \\
\widehat{m}_{\psi} &= \frac{1}{T_C^3}\int_0^{T_C} \widehat{\mathrm{R}}_{\psi}(\tau)\mathrm{R}_{\psi}(\tau)\,d\tau \\
w_{\psi} &= \frac{1}{T_C^5}\int_0^{T_C} \left[\mathrm{R}_{\psi}^2(\tau)+\widehat{\mathrm{R}}_{\psi}^2(\tau)\right]^2 d\tau \\
\widehat{w}_{\psi} &= \frac{1}{T_C^5}\int_0^{T_C} \widehat{\mathrm{R}}_{\psi}^2(\tau)\mathrm{R}_{\psi}^2(\tau)\,d\tau
\end{aligned} \tag{38}$$

where

$$\begin{aligned}
\mathrm{R}_{\psi}(\tau) &= \int_{T_C-\tau}^{T_C} \psi(t)\psi\left(t-(T_C-\tau)\right)dt \\
\widehat{\mathrm{R}}_{\psi}(\tau) &= \int_{\tau}^{T_C} \psi(t)\psi(t-\tau)dt
\end{aligned} \tag{39}$$

The MAI variance using the standard Gaussian approximation is given as

$$\sigma_{\mu}^2 = \frac{NP(K-1)T_C}{2} m_{\psi} \tag{40}$$

where N is the number of Chips.

Using the Improved Gaussian Approximation, the MAI variance has a mean and variance is given as

$$\begin{aligned}
\varphi_{\mu} &= \frac{N(K-1)PT_C^2}{2} m_{\psi} \\
\varphi_{\sigma} &= \frac{PT_C^2}{2}\left[(K-1)\left(\frac{3}{8}w_{\psi} + \frac{3(N-1)}{2N}\widehat{w}_{\psi} + m_{\psi}^2 + \frac{(K-2)(N-1)}{N^2}\widehat{m}_{\psi}^2\right)\right]^{\frac{1}{2}}
\end{aligned} \tag{41}$$

Both approximations show that the variance of the MAI contribution is dependent on $H(f)$. Consequently, the probability of false alarm and detection are dependent on it.

The simplified Improved Gaussian expressions are based on the fact that a continuous function $f(x)$, may be expressed as

$$f(x) = f(\mu) + (x-\mu) f'(\mu) + \frac{1}{2}(x-\mu)^2 f''(\mu) + .. \tag{42}$$

If x is a random variable and μ is the mean of x, then

$$E\left[f(x)\right] = f(x) + \frac{1}{2}\sigma^2 f''(x) + .. \tag{43}$$

where σ^2 is the variance of x. Further computation gives the approximation

$$E\left[f(x)\right] = f(\mu) + \frac{\sigma^2}{2}\left(\frac{f(\mu+h) - 2f(\mu) + f(\mu-h)}{h^2}\right) \tag{44}$$

It was suggested by Holtzman [60] , that the appropriate choice for h is $\sqrt{3}\sigma$ which yields

$$E\left[f(x)\right] = \frac{2}{3}f(\mu) + \frac{1}{6}f\left(\mu+\sqrt{3}\sigma\right) + \frac{1}{6}f\left(\mu-\sqrt{3}\sigma\right) \tag{45}$$

The corresponding Improved Gaussian approximation to the probability of false alarm and detection can be evaluated using (45).

From Equations(8) &(9),

$$p\left(y \middle| H_0\right) \approx \frac{2}{3}\exp\left(-\frac{Th}{2\varphi_\mu}\right) + \frac{1}{6}\exp\left(-\frac{Th}{2\left(\varphi_\mu + \sqrt{3}\varphi_\sigma\right)}\right) + \frac{1}{6}\exp\left(-\frac{Th}{2\left(\varphi_\mu - \sqrt{3}\varphi_\sigma\right)}\right) \tag{46}$$

$$\begin{aligned} p\left(y \middle| H_1\right) \approx\; & Q\left(\frac{y}{\sqrt{\varphi_\mu}}, \frac{Th}{\sqrt{\varphi_\mu}}\right) + Q\left(\frac{y}{\sqrt{2\left(\varphi_\mu + \sqrt{3}\varphi_\sigma\right)}}, \frac{Th}{\sqrt{2\left(\varphi_\mu + \sqrt{3}\varphi_\sigma\right)}}\right) \\ & + Q\left(\frac{y}{\sqrt{2\left(\varphi_\mu - \sqrt{3}\varphi_\sigma\right)}}, \frac{Th}{\sqrt{2\left(\varphi_\mu - \sqrt{3}\varphi_\sigma\right)}}\right) \end{aligned} \tag{47}$$

Q is the Marcum's Q-function

Equations (46) and (47) are only valid when $\varphi_\mu > \sqrt{3}\varphi_\sigma$ to ensure that the denominator of the third term is positive. Also the improved Gaussian approximation should be used for the case where the number of users, $K < 3$ [4].

Numerical results are presented below to show the effect of the better approximation on the acquisition system performance i.e. Probability of detection and Mean Acquisition time. For evaluation five pulse types are used; (a) Rectangular pulse, (b) Half-Sine pulse, (c) truncated Gaussian Pulse, (d) Raised Cosine pulse and (e) the Blackman pulse. The properties of the pulses are given in Appendix - A.

3.2 NUMERICAL RESULTS

Assuming that there is one ACQ cell, from (21) - (24), we have

$$\begin{gathered} P_D^s = \int_{Th}^{\infty} P\left(y\middle|H_1\right)\left[\int_0^y P\left(x\middle|H_0\right)dx\right]^{qN_P-1} dy \\ P_M = \left[\int_0^{Th} P\left(x\middle|H_0\right)dx\right]^{qN_P-1}\left[\int_0^{Th} P\left(y\middle|H_{1l}\right)dy\right] \\ P_{F1}^s = 1 - P_M^s - P_D^s \end{gathered} \tag{48}$$

$$P_{F0}^s = 1 - \left[\int_0^y P\left(x\middle|H_0\right)dx\right]^{qN_P} \tag{49}$$

The decision threshold is computed from(49), by setting the probability of false alarm to a particular value the resulting value, the threshold can be computed.

The mean acquisition time

$$\begin{aligned} E\left[T_{acq}\right] &= \left(\left[\frac{d}{dz}U(z)\right]\Bigg|_{z=1} + S\right)T_c \\ &= \left(\frac{1}{P_D}\left[1 + V\left(1-P_M\right) + K_V P_{F1} + \left(N_S - 1\right)\left(1 + K_V P_{F0} + VP_{F0}^s\right)\right] + M\right)T_c \end{aligned} \tag{50}$$

The Normalised mean acquisition is the Mean acquisition time normalised by the bit period.

Table 1 Simulation Parameters

Filter Length	64
Search Space	512
Post Integration length	320
Pf	10^{-3}
Number of Users	10
Ns	16
Np	4

Table 2 Pulse Parameters

Pulse	m_ψ	$\widehat{m_\psi}$	w_ψ	$\widehat{w_\psi}$
Rectangular	3.3333e-001	1.6667e-001	4.6666e-001	3.3333e-002
Half Sine	2.9332e-001	4.3318e-002	4.2291e-001	3.2365e-003
Gaussian	2.9823e-001	5.4879e-002	4.2631e-001	4.5683e-003
Raised Cosine	2.4055e-001	9.4533e-003	3.4366e-001	1.9039e-004
Blackman	2.0298e-001	2.5002e-003	2.8282e-001	1.4144e-005

Comparison between both approximations

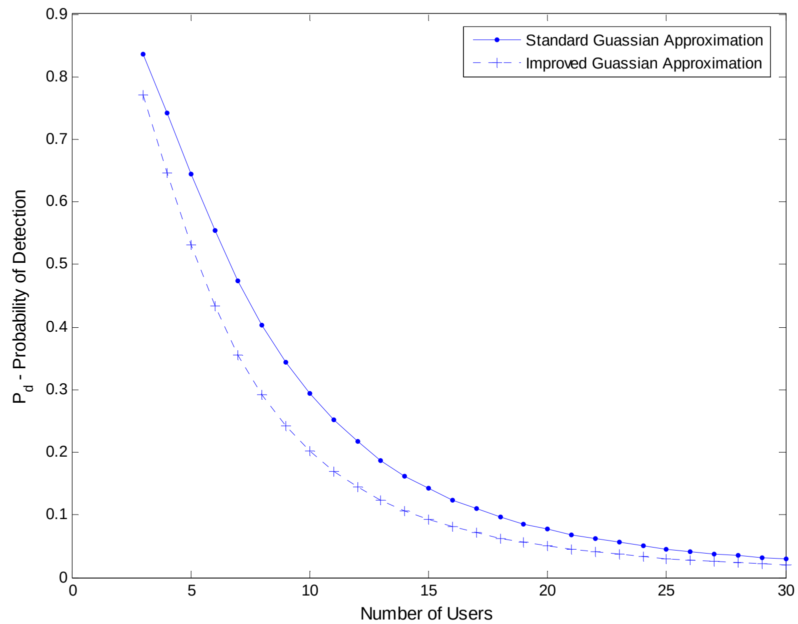


Figure 15 Probability of Detection ($P_f = 10^{-3}$, Half Sine, Ec/No = -5dB)

Comparison between Different Pulses

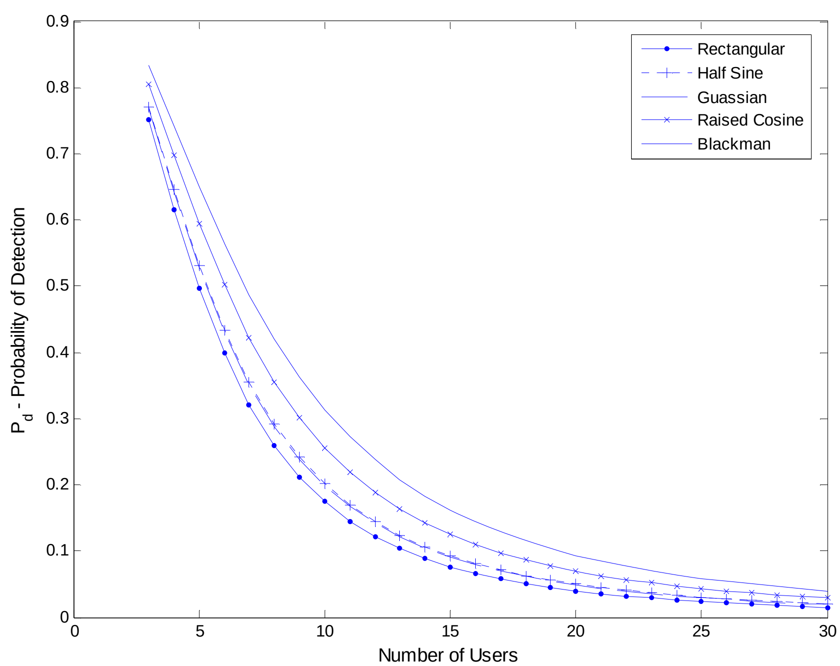


Figure 16 Probability of Detection ($P_f = 10^{-3}$, N = 512, Eb/No = -5dB)

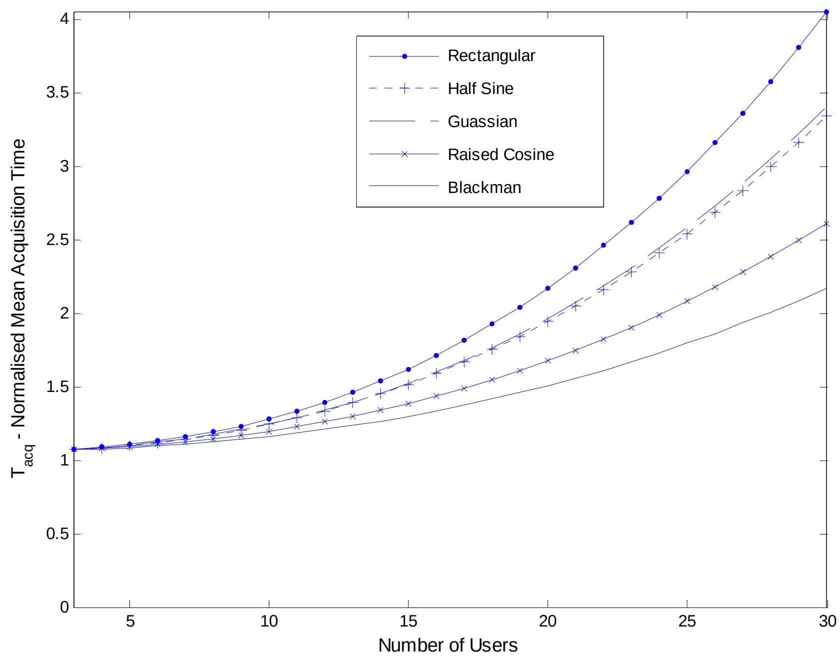


Figure 17 Mean Acquisition Time ($P_f = 10^{-3}$, N = 512, Eb/No = -5dB)

Bandlimited Channel Scenario

Table 3 Simulation Parameters

Chip Rate	3.84 Mcps
Bandwidth	5Mhz
Filter Length	64
Search Space	512
Post Integration length	320
Pf	10^{-3}
Number of Users	10
Ns	16
Np	4

Table 4 Pulse Parameters

Pulse	Inbound Power	m_{ψ}	$\hat{m}_{\psi}$	w_{ψ}	$\hat{w}_{\psi}$
Rectangular	86.89%	3.2840e-001	1.8595e-001	4.3560e-001	4.0408e-002
Half Sine	82.86%	2.6936e-001	9.7249e-002	3.0055e-001	1.5459e-002
Gaussian	84.29%	2.7718e-001	1.0639e-001	3.1714e-001	1.7404e-002
Raised Cosine	73.33%	2.0225e-001	5.3579e-002	1.7233e-001	6.8720e-003
Blackman	66.13%	1.5795e-001	3.3917e-002	1.0615e-001	3.7978e-003

Comparison between Different Pulses

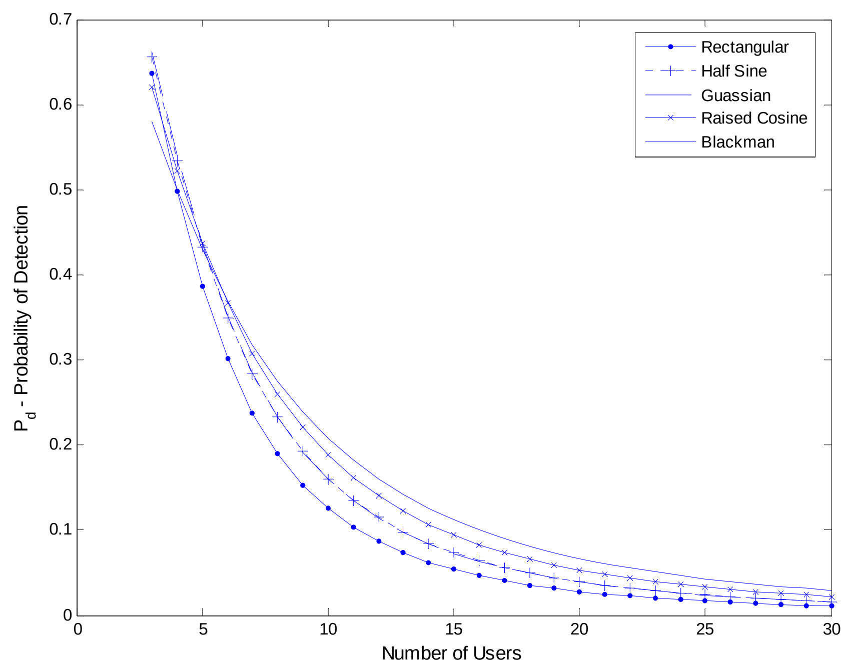


Figure 18 Probability of Detection ($P_f = 10^{-3}$, N = 512, Eb/No = -5dB)

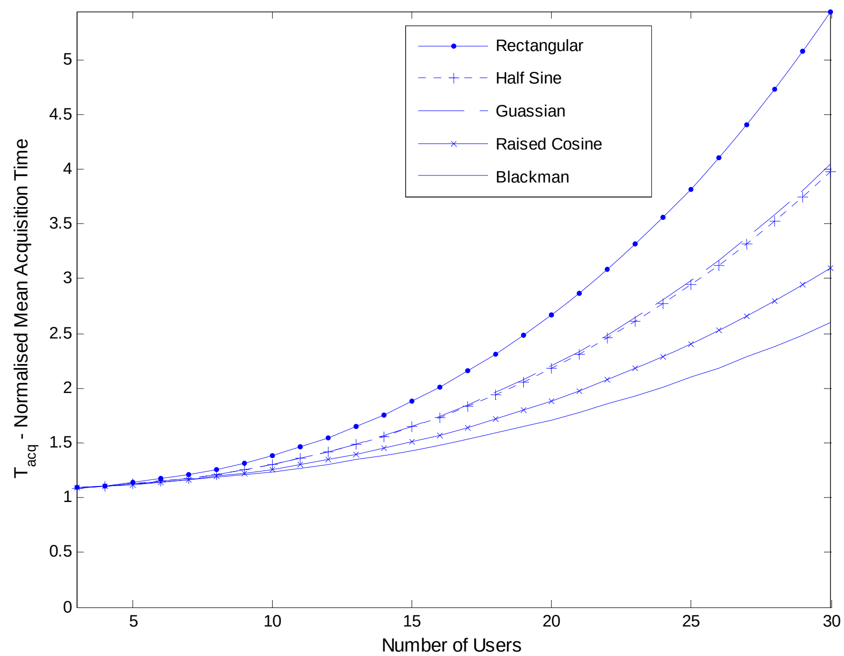


Figure 19 Mean Acquisition Time ($P_f = 10^{-3}$, N = 512, Eb/No = -5dB)

3.3 CONCLUSIONS

The probability of detection using both approximations has been evaluated. For a constant false alarm rate, the improved Gaussian approximation gives a lower probability of detection than the standard approximation. This shows that the standard approximation gives an optimistic result for both performance measures.

The capacity of the system is smaller than what the standard approximation gives. To ensure proper planning for capacity, the improved Gaussian approximation should be used.

The Blackman pulse performed best when compared with other pulse for both performance measures. In order of performance we have, Blackman, Raised Cosine, Half sine, truncated Gaussian and the worst being the Rectangular Pulse.

CHAPTER 4

THRESHOLD SETTING OPTIMIZATION

4.1 INTRODUCTION

For the threshold crossing criterion discussed earlier in chapter 2, the decision whether a particular cell corresponds to the synchro or nonsynchro position is carried out by comparing the detector output to a threshold value T_h. The threshold is a reference value used as the cell acceptance (or refusal) criterion. Depending of the statistical test approach employed by the detector, the threshold could be set fixed for all the cells or could be cell-dependent. The former case, based on the *Neyman-Pearson* decision criterion, attempts to keep probability of flase alarm, P_{FA} fixed and is used when no *a priori* information on the synchro cell is available. The latter, applying the *Bayes* test criterion, is used when *a priori* information on the synchro cell is available.

The probabilities P_{FA} and P_D depend both on the T_h value, thus special care should be taken when setting the threshold to obtain desired acquisition performance. In fact the average time to acquire depends, among others, on the level of prevailing SNR, P_{FA} and P_D. If the threshold is set too high, P_{FA} will get smaller but so will P_D. Too low of a threshold, on the other hand, will beneficially increase P_D but will result also in a higher P_{FA}.

The threshold should be set above the noise level, thus an estimate of this figure is required by the detector. Under stable channel conditions the threshold can be fixed but in dynamic channel conditions, where for instance interference is present, adaptive threshold schemes are used.

For a receiver based on the Neyman-Pearson criterion or constant threshold (CT) criterion, if the acquisition threshold *Th* is set relatively low in favour of receiving weak signals, the false acquisition probability will be too high when strong level signals show up, because the partial correlation sidelobes are likely to exceed the threshold before the main correlation peak is detected. On the other hand, if *Th* is set relatively high in favour of large signals, the probability that receivers miss those relatively weak packets will be increased. Both cause losses of many packets. Such phenomena become more serious in CDMA environment because the desired signal including the sidelobes (off-phase autocorellaion) and main peaks are more likely to fluctuate with the random number of active users in the system. Thus it is necessary to design an acquisition threshold with its value being set according the effective SNR of the main peak and the strengths of sidelobes.

It is significant to note that P_D and P_{FA} are important only in how they affect the more meaningful measures of performance such as the mean time to acquisition and the mean hold-in time[8].

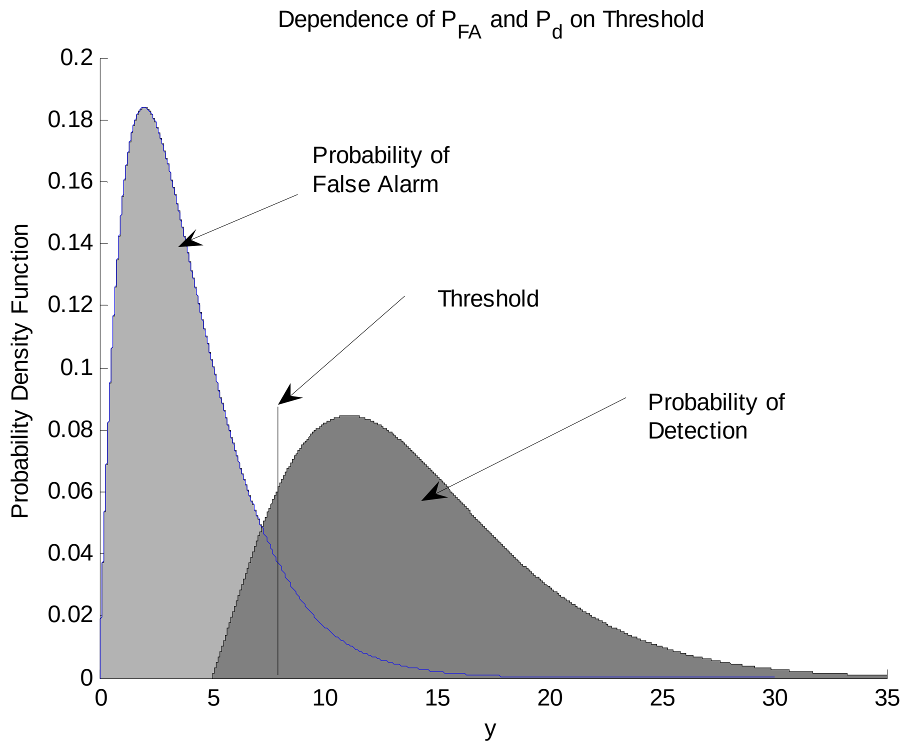


Figure 20 Dependence of PFA and PD with threshold value Th

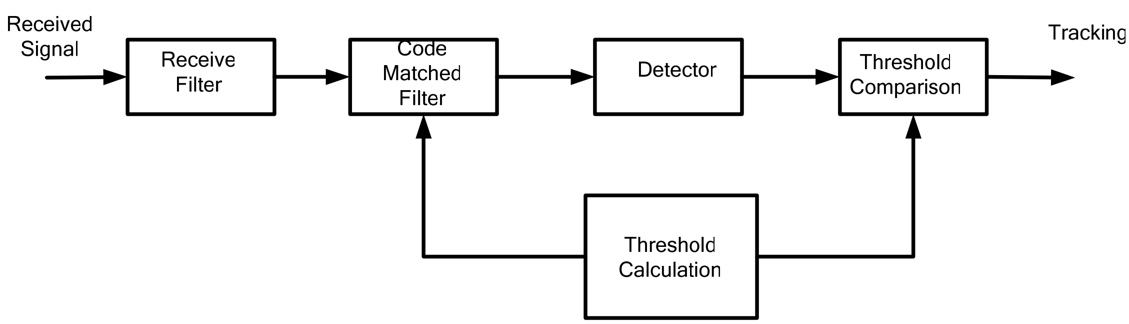


Figure 21 Threshold Setting Block Diagram

A threshold setting scheme based on the SNR or interference-to-signal power ratio is considered by [61]. Adaptive threshold can also be obtained by using reference filters [62], noise power measurements [63], median filters [64], or rank filters[65]. Automatic Decision Threshold Control (ADTLC) scheme based on the threshold crossing statistics during a repeated number of observations is proposed and analyzed in [66]. In the ADTLC approach the threshold value is simply computed from the pdf of the signal at the correlator output, the threshold *Th* resulting in a simple function of P_{FA} and the noise power estimate. A comprehensive comparative study of different threshold setting techniques in single-path and multipath channels is presented in [67, 68].

4.2 THRESHOLD SETTING ALGORITHMS

4.2.1 THRESHOLD SETTING BASED ON POWER AVERAGING

The threshold is set as a proportion of the estimated noise power. By estimating the noise values, the threshold can be computed for a desired probability of false alarm from equation (51) give below

The false alarm probability of the search mode in the H_0 is

$$P_F^s = 1 - \left[\int_0^y P\left(y \middle| H_0\right) dx \right]^{qN_P} \tag{51}$$

where

$$p\left(y \middle| H_0\right) = \int_{Th}^{\infty} \frac{1}{2\sigma^2} \exp\left(-\frac{x}{2\sigma^2}\right) dx = \exp\left(-\frac{Th}{2\sigma^2}\right) \tag{52}$$

4.2.1.1 MEAN LEVEL SCHEME

In the mean level scheme the power-level estimate is obtained by summing the cells of the window defined as

$$X = \frac{1}{M}\sum_{i=1}^{M} y(i) \tag{53}$$

where y are the code Matched filter output. The threshold is determined as

$$Th = kX \tag{54}$$

where k is a factor used to achieve constant false alarm rate. The value of k is determined from the desired probability of false alarm i.e. Equation (51)

4.2.1.2 RANK (ORDER) SCHEME

In the order statistics scheme the data in the window are sorted in increasing order. The resulting ordered sequence is $y_{(1)} \le y_{(2)} \le \cdots\cdots \le y_{(M-1)} \le y_{(M)}$. A predetermined k^{th} value in the order is chosen as the threshold.

4.2.1.3 USE OF REFERENCE FILTER

This is a technique suggested in [69]. A reference filter is loaded with a code orthogonal to the transmitting code. The output of this filter is considered as noise and it is used to estimate the noise.

4.2.2 PROPOSED MODIFIED RANK ACQUISITION SCHEME

In real situations involving mobile communication, there may exist multipath signals in the window of given size. For the Mean Level (ML) scheme, as some multipath components enter the window, a high threshold value will be given. However, the rank scheme has an inherent immunity to this problem. A problem with the rank scheme is how to determine the rank as the rank chosen affects the Pf.

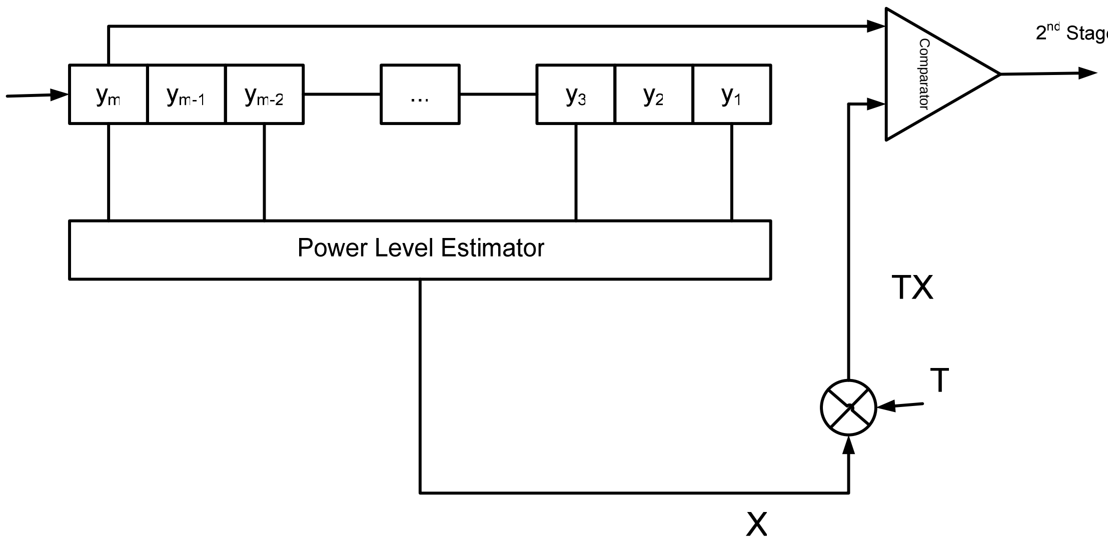


Figure 22 Threshold estimator

We propose here to use a hybrid scheme, this takes the merits of the Mean Level (ML) scheme and the Rank scheme, the estimate is computed as an average of a window of sorted values.

The ordered sequence is $y_{(1)} \leq y_{(2)} \leq \cdots\cdots \leq y_{(M-1)} \leq y_{(M)}$. The threshold is calculated as

$$\begin{aligned} &Th = kX \\ &where\ X = \frac{1}{l_2 - l_1} \sum_{i=l_1}^{l_2} y(i) \end{aligned} \tag{55}$$

This ensures that the high values which can be attributed to multipath are removed. Similarly, very low values that may reduce the noise estimate are removed. For this work l_1 and l_2 are chosen to be 5% and 95%.

4.3 THRESHOLD SETTING SIMULATION RESULTS

Here are some comparison in the probability of detection and normalised[1] mean acquisition of Time for the several search schemes are presented.

The MF is matched to a segment of the spreading code. The length of the filter is 64. The output signal of the filter (MF) is proportional to the partial autocorrelation function (ACF) of the code. Additive white Gaussian noise and the multiple user interference contribution are added to the desired signal. For the verification mode, after the Matched filter, there is an envelope detector for the non-coherent receiver.

[1] Normalised by the Bit period $T_b = NT_c$

4.3.1 AWGN CHANNEL

Table 5 Simulation Parameters

N_S	1
N_P	2
Filter Length	64
Search Space	512
Post Integration length	320
Pf	10^{-4}
Pulse shape	Half Sine
Number of users	8

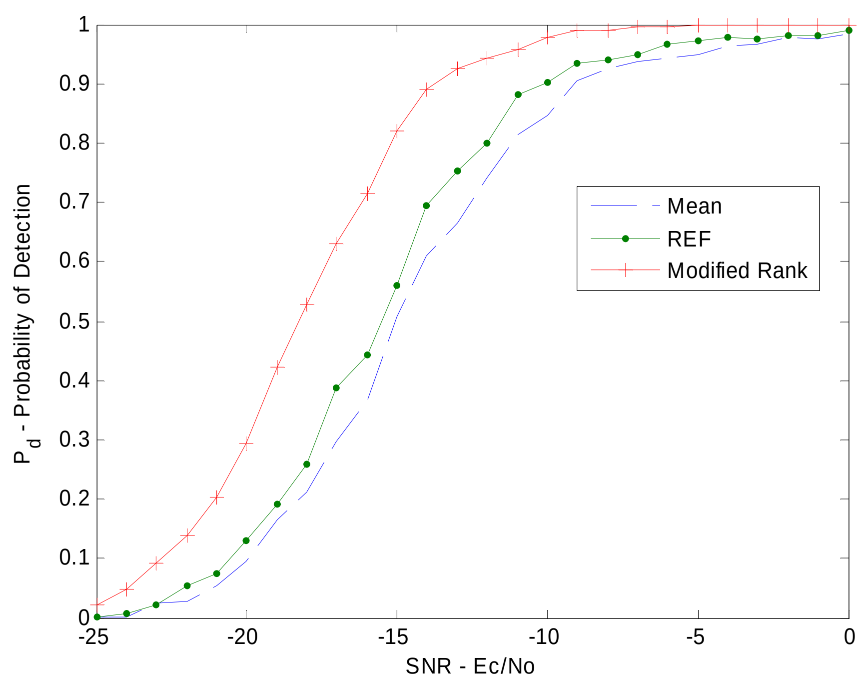


Figure 23 Probability of Detection (single path)

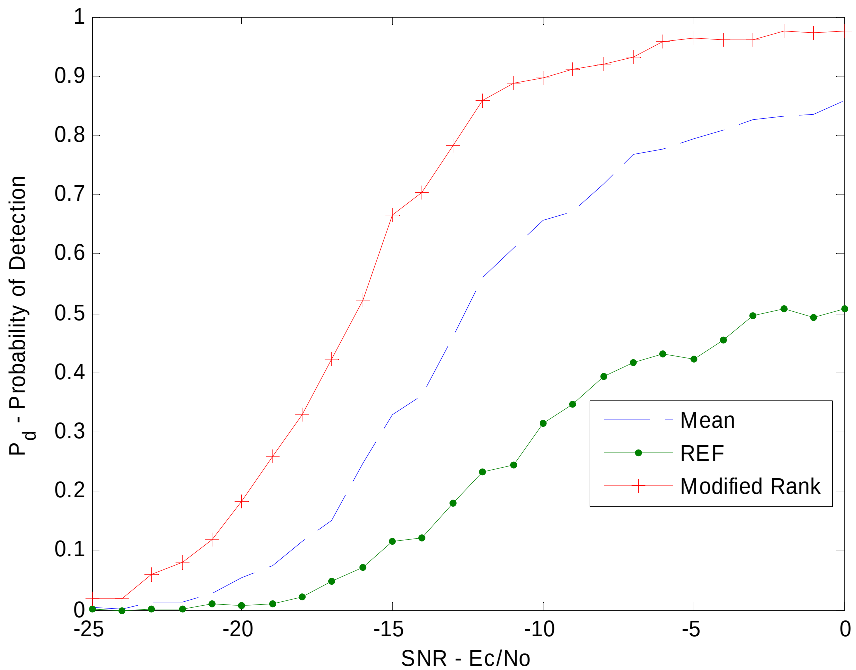


Figure 24 Probability of Detection (3-paths)

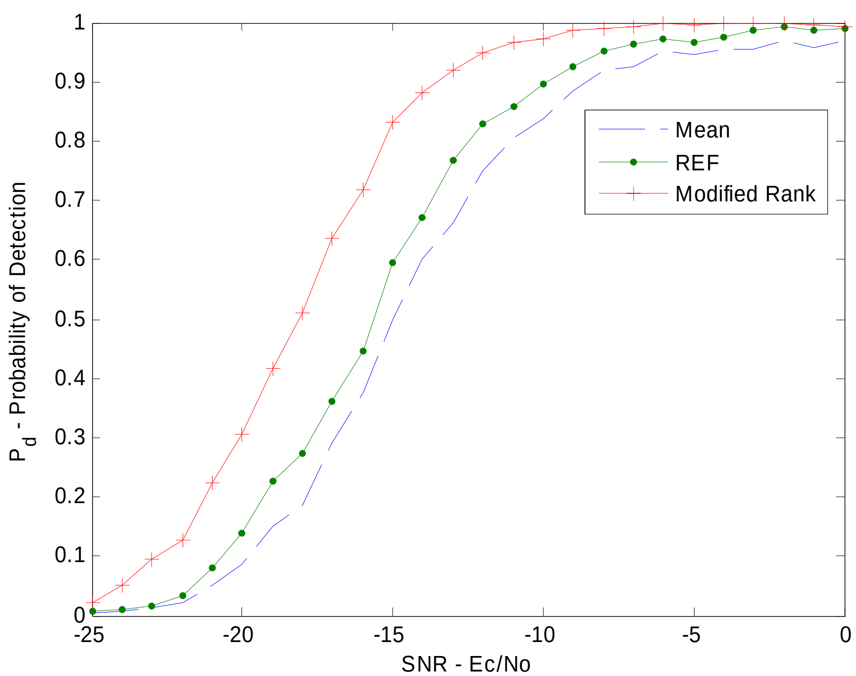


Figure 25 Probability of Detection (half chip sampling & single path)

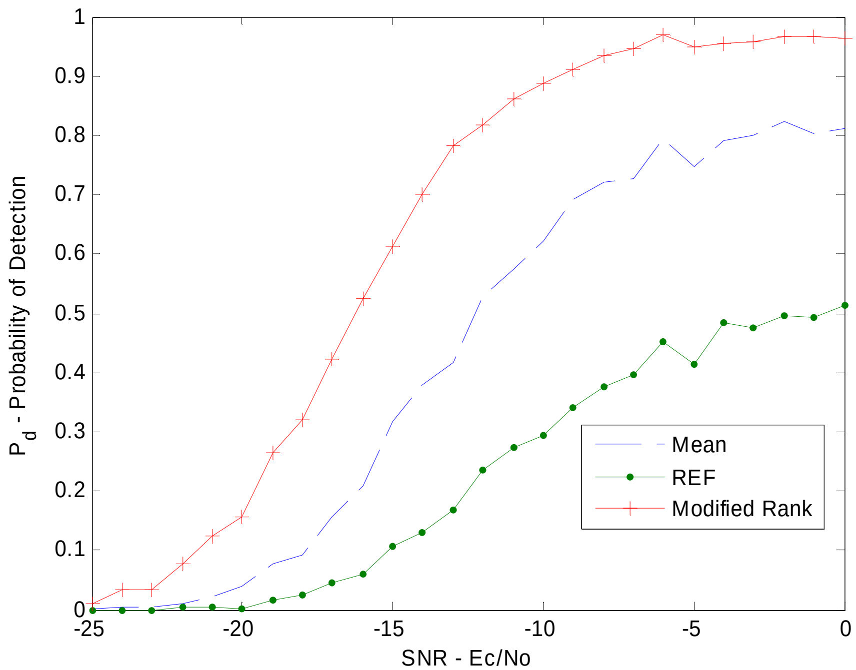


Figure 26 Probability of Detection (3-paths & half chip sampling)

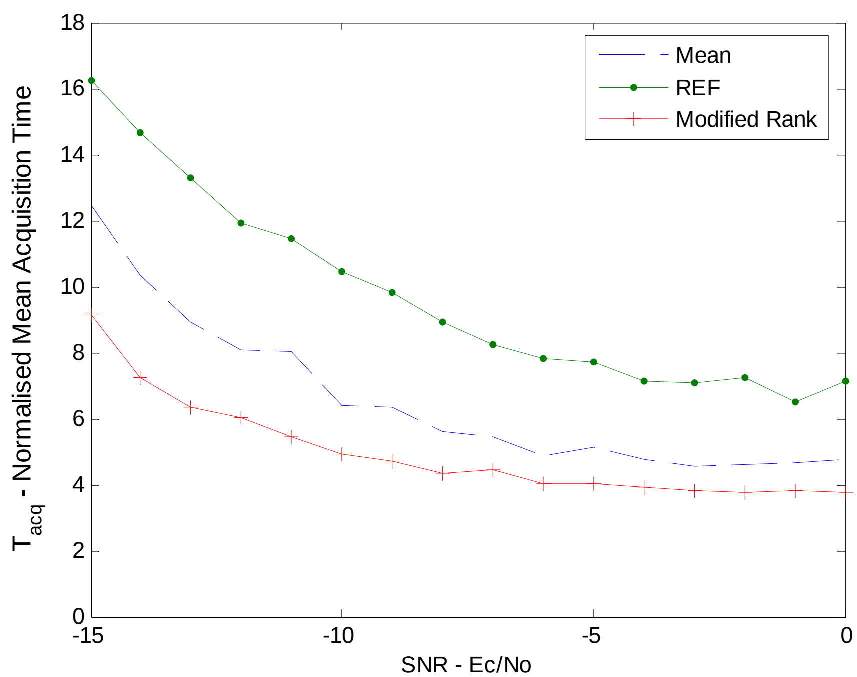


Figure 27 Mean Acquisition Time (single path)

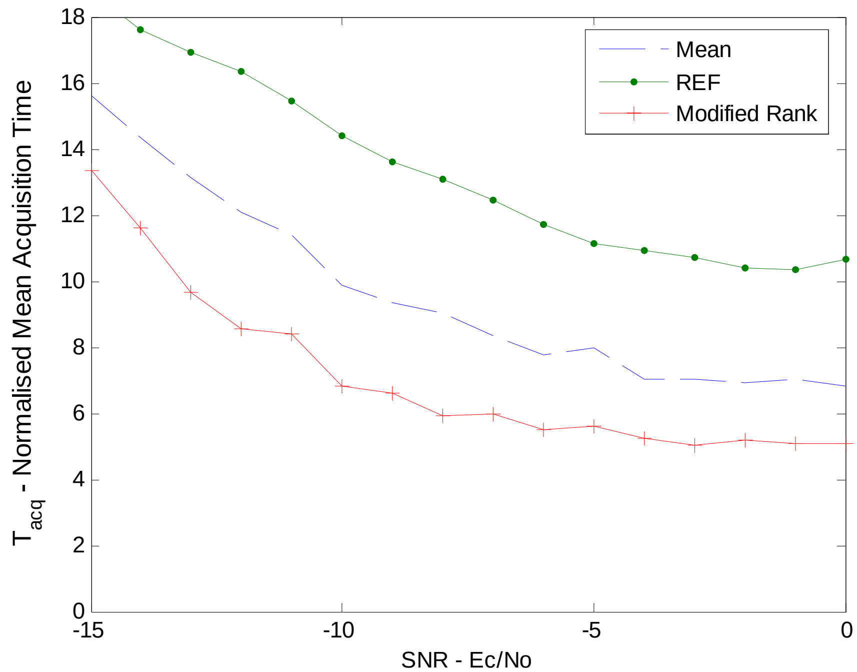


Figure 28 Mean Acquisition Time (3-paths)

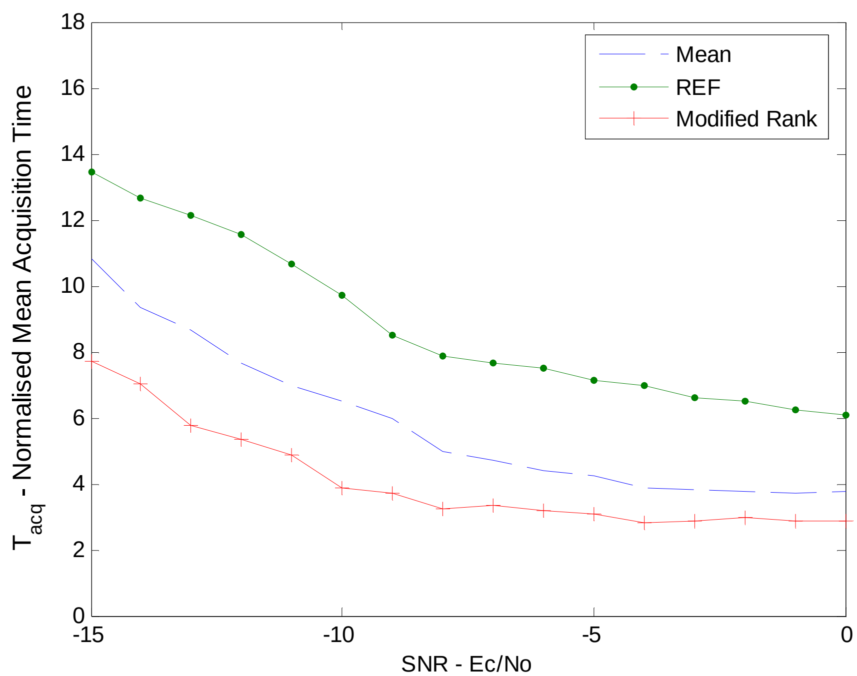


Figure 29 Mean Acquisition Time (half chip sampling & single path)

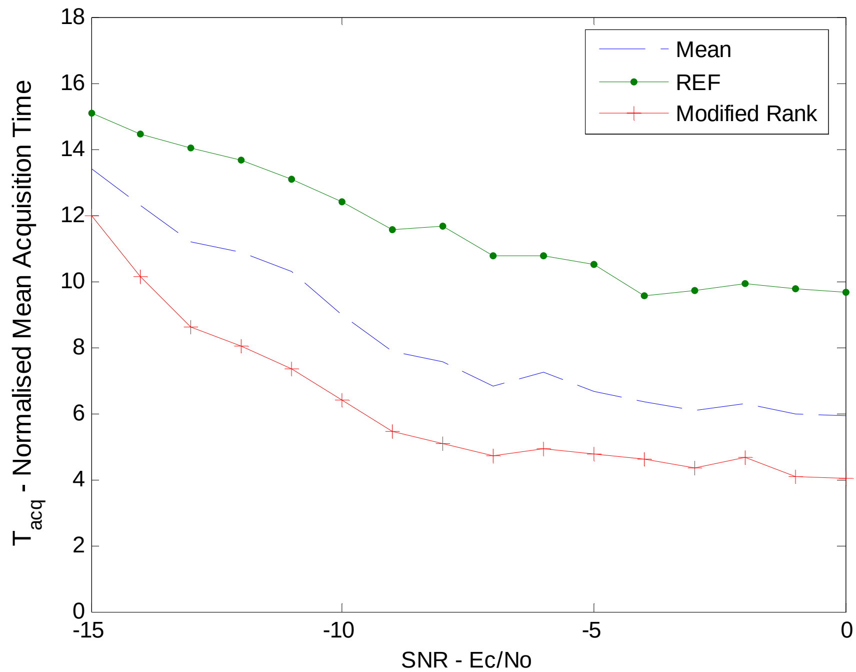


Figure 30 Mean Acquisition Time (3-path & half chip sampling)

4.3.2 FADING CHANNEL

Table 6 Simulation Parameters

N_S	1
N_P	2
Filter Length	64
Search Space	512
Post Integration length	320
Pf	10^{-4}
Pulse shape	Half Sine
Number of users	8

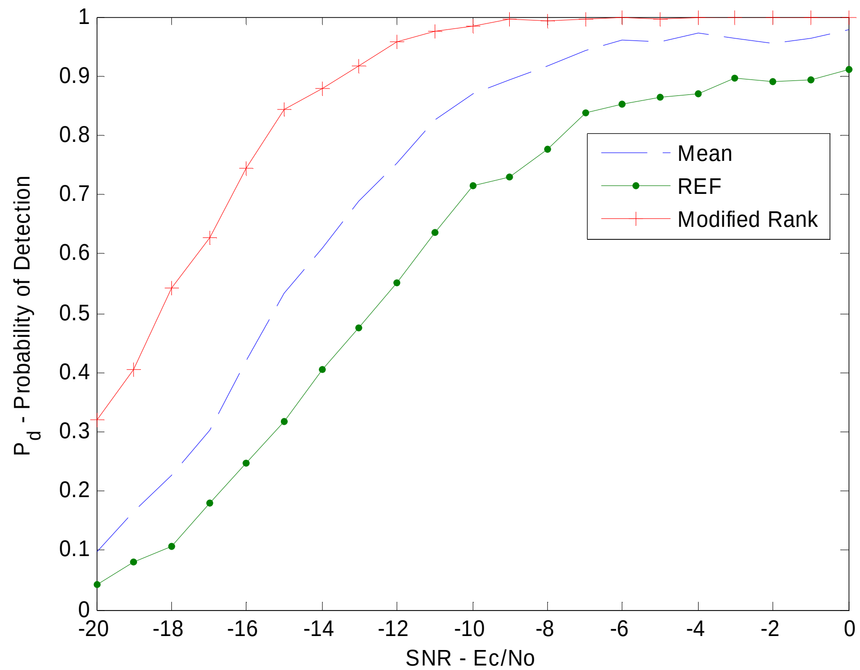


Figure 31 Probability of Detection (single path)

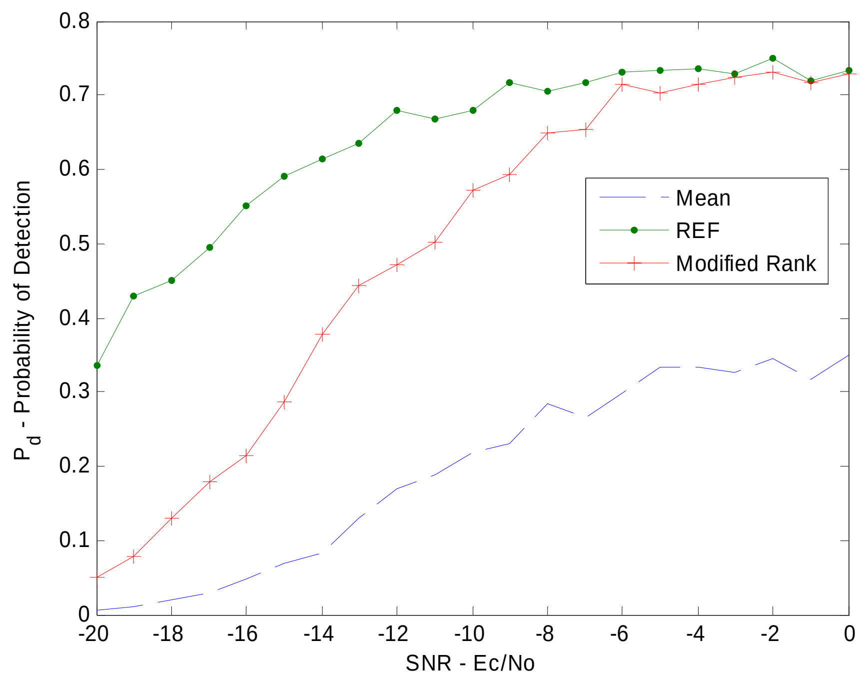


Figure 32 Probability of Detection (3-paths)

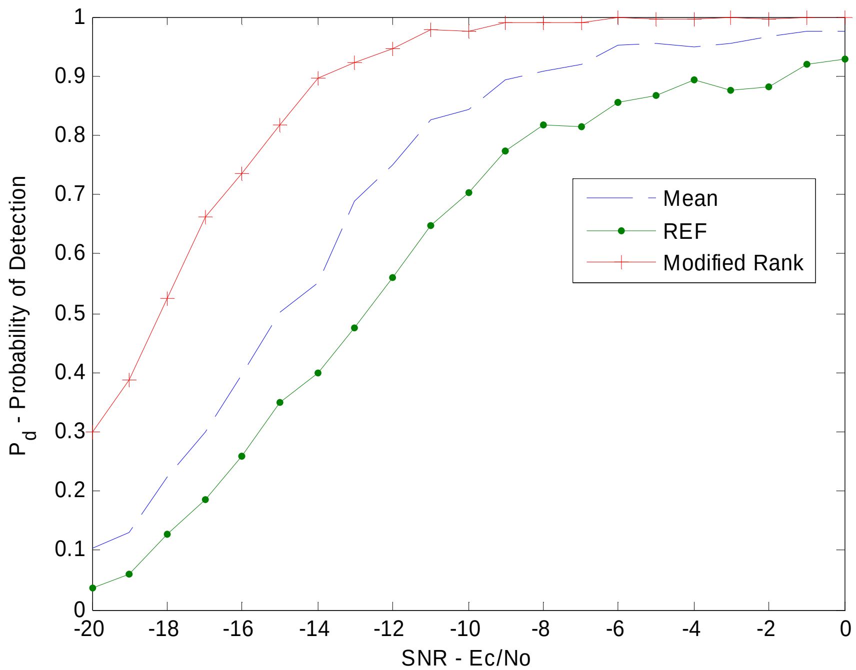


Figure 33 Probability of Detection (half chip sampling & single path)

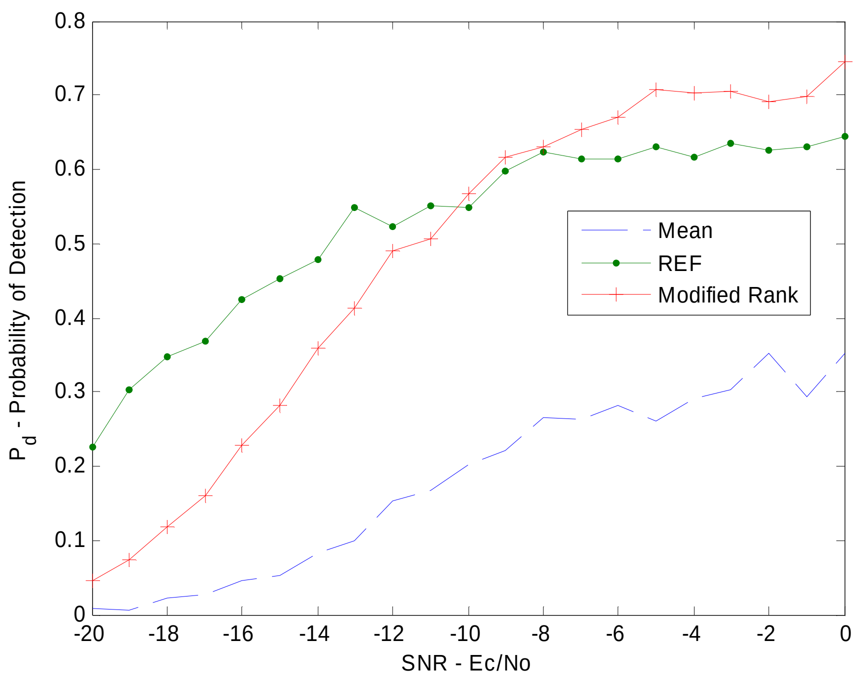


Figure 34 Probability of Detection (3-paths & half chip sampling))

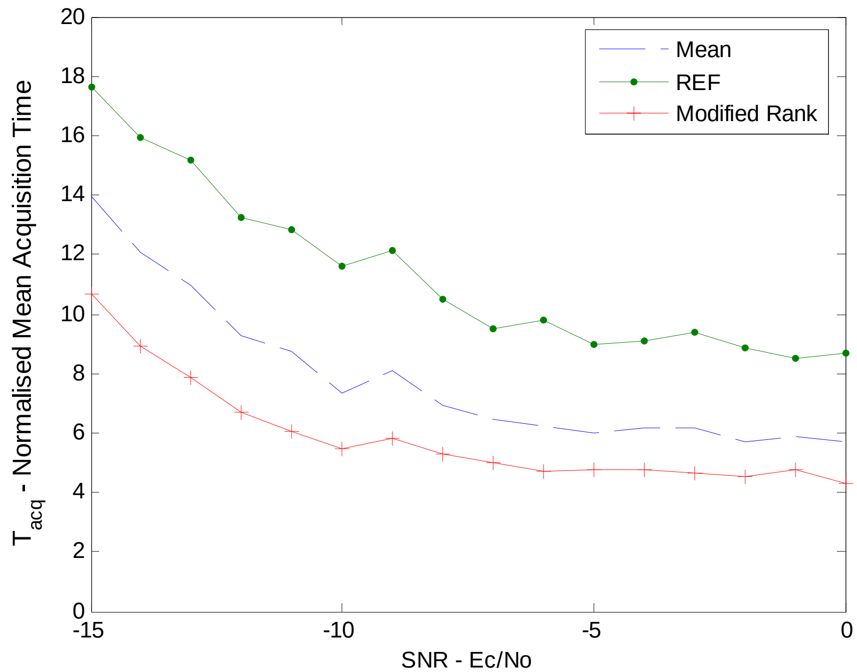


Figure 35 Mean Acquisition Time (single path)

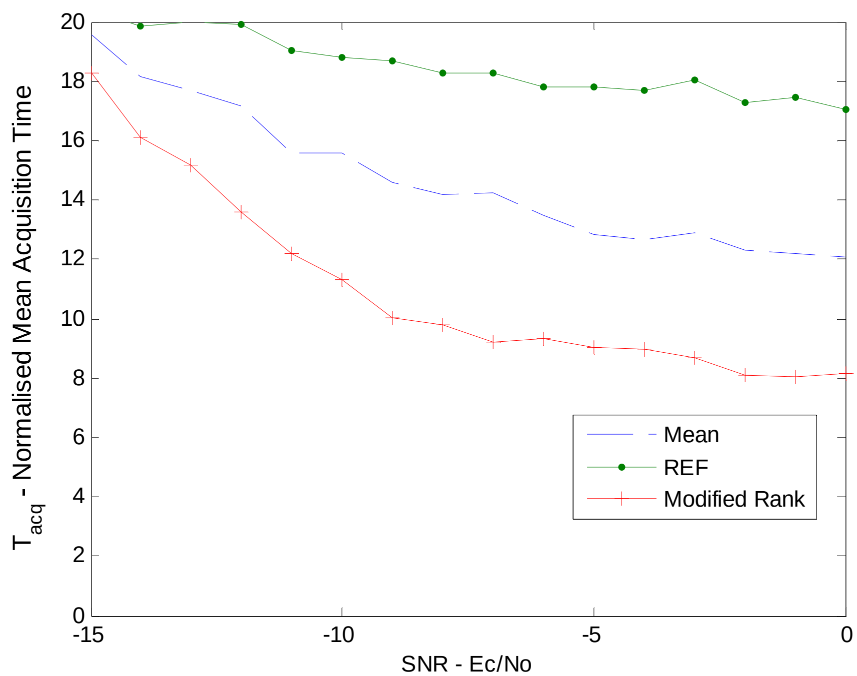


Figure 36 Mean Acquisition Time (3-paths)

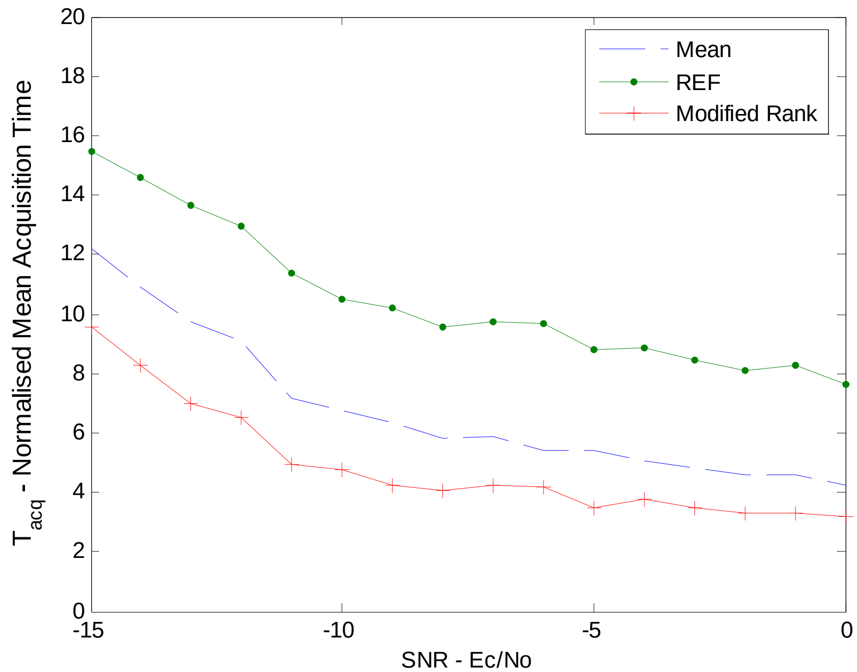


Figure 37 Mean Acquisition Time (half chip sampling)

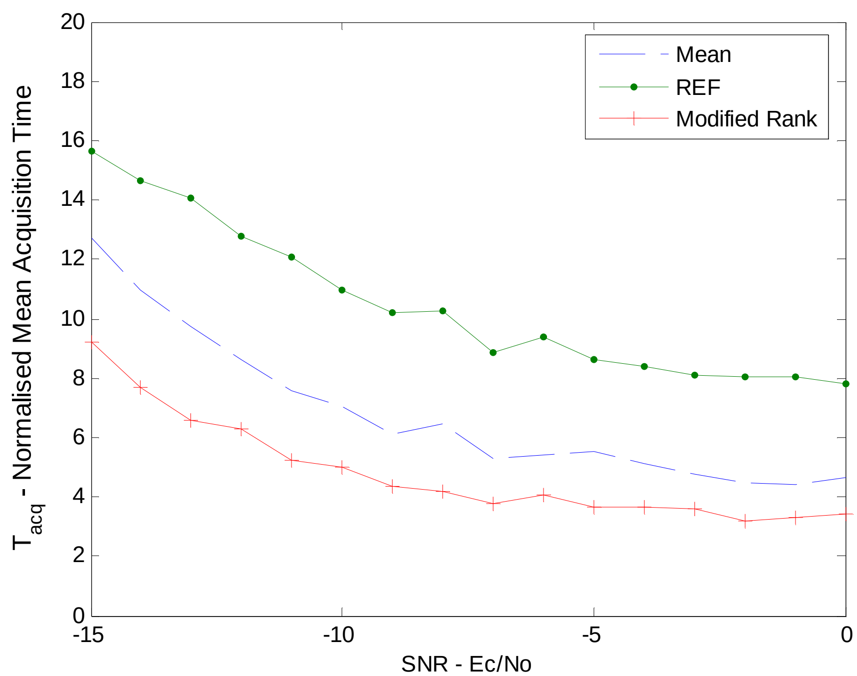


Figure 38 Mean Acquisition Time (3-path & half chip sampling)

4.4 REMARKS

The modified scheme is seen in the simulation results to outperform other schemes when the time to acquire the correct phase is considered. In the case were there is fading in a 3-path channel (Figure 33 and Figure 34), the Probability of detection of the Ref scheme is higher. However looking at the mean time to acquire, it has a higher mean time to acquire. This is due to a high false alarm rate. The Probability of detection is only meaningful when the Mean acquisition time is considered. For the same scenario, the time to acquire the correct phase for the modified scheme is lower. In summary the Modified scheme outperforms all the other schemes considered.

CHAPTER 5

DIGITAL SIGNAL PROCESSOR IMPLEMENTATION

Digital Signal Processor (DSP) is one of the key technologies that have made the mobile phone revolution possible. Recently, there has been a strong trend toward implementing more and more functionality of digital communication modems using flexible, programmable DSP platforms, especially in the early stages of system development and prototyping. DSPs are used for a wide range of applications, from communications and controls to speech and image processing.

A migration to a highly integrated, cost and power efficient final hardware realization (*Application Specific Integrated Circuit* - ASIC) can then follow at a stage once the design has been validated. This recent trend has been even more successful with the tremendous improvement in processing speed and capability of the recently developed DSP platforms[70-73].

In the case of CDMA, since the signalling schemes are spread-spectrum in nature (operating at a very high rate, the chip rate), the signal processing requirements are very demanding. However, with the constant improvement in DSP capability, it has recently become possible to implement all (or most) of the baseband CDMA modem functionalities using programmable DSPs[16].

Specifically for the thesis work we worked on Texas Instruments' DSK C6713-DSP based card [74] shown in Figure 52. Digitals signal processors such as the TMS320C67x family of processors are like fast special-purpose microprocessors with a specialized architecture and an instruction set is appropriate for signal processing[75].

5.1 DSK C6713 DESCRIPTION

The DSK package from spectrum Digital is a complete DSP system with hard and software support tools. The board includes a TI C6713 floating point DSP and a 32-bit codec.

The DSK board includes 16Mb of synchronous dynamic random access memory and 256Kb of flash memory. Four connections on the board provides audio input and output; MIN in, LINE in, LINE out and Headphone. There are four user dip switches that provide users with feedback control. The DSK operates at 225 Mhz.

The DSK supports a TMS320C6713 DSP which can operate at a clock frequency of up to 225 MHz. The DSP core is designed for high performance floating point operation. Beyond the DSP core, the C6713 integrates a number of on-chip resources that improve functionality and minimize hardware development complexity.

5.2 DSP TOOLS

Code composer Studio

Code composer studio is a fully Integrated DSP Development Environment. Code Composer StudioTM (CCStudio) Development Tools are a key element of the eXpressDSP Software and Development Tools strategy from Texas Instruments.

CCStudio delivers all of the host tools and runtime software support for your TMS320 DSP based real-time embedded application to market faster.

The fully integrated development environment includes real-time analysis capabilities, easy to use debugger, C/C++ Compiler, Assembler, linker, editor, visual project manager, simulators, XDS560 and XDS510 emulation drivers and DSP/BIOS support.

Embedded Target for TI TMS320 C6000

Embedded Target for the TI TMS320C6000 DSP Platform integrates Simulink® and MATLAB® with Texas Instruments eXpressDSP™ tools. The software collection lets you develop and validate digital signal processing designs from concept through code. The Embedded Target for TI C6000 DSP consists of the TI C6000 target that automates rapid prototyping on your C6000 hardware targets. The target uses C code generated by Real-Time Workshop® and your TI development tools to build an executable file for your targeted processor. The Real-Time Workshop build process loads the targeted machine code to your board and runs the executable file on the digital signal processor.

Link for Code Composer Studio

Link for Code Composer Studio Development Tools lets you use MATLAB functions to communicate with Code Composer Studio™ and with information stored in memory and registers on a target. With the links you can transfer information to and from Code Composer Studio and with the embedded objects you get information about data and functions stored in your signal processor memory and registers, as well as information about functions in the project being implemented on the DSP.

5.3 SYSTEM BLOCKS

5.3.1 TRANSMITTER

The CDMA transmitter system model shown in Figure 39 was implemented in MATLAB. The signal generated in MATLAB was then loaded unto the DSP which performed the acquisition.

5.3.1.1 USER DATA

The desired user is generating a series of one, this is the preamble sequence. All the other users are generating random data.

5.3.1.2 GOLD CODES

The user data are spread using Gold Codes. Gold sequences are PN sequences which are optimal in terms of reaching the Sidelnikov bound[76] for the maximum auto and cross-correlation peaks for binary sequences with a given periodicity. Gold codes are generated using a shift register.

In [77], Gold describes a method of generating large families of binary pseudorandom sequences and argues that they have consistently uniform correlation characteristics. These have subsequently been known as Gold sequences. They are formed by taking the modulo-2 sum of two preferred-pair m-sequences.

General Properties of Gold Sequences

They are periodic with periodicity $N_C = 2^n - 1$

They have cross-correlation given as

$$A(\tau) = \begin{cases} 1 & \tau = 0 \\ \dfrac{1}{Nc}\left\{-1, -2^{\frac{n}{2}} - 1, 2^{\frac{n}{2}} - 1\right\} & \tau \neq 0 \end{cases} \tag{56}$$

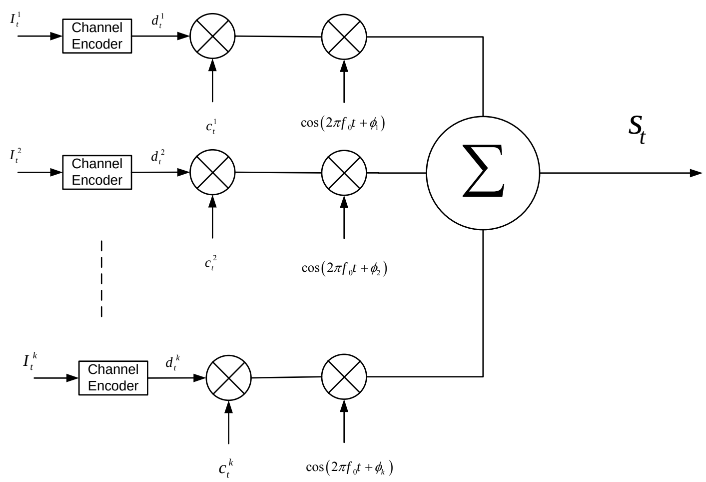


Figure 39 CDMA Transmitter

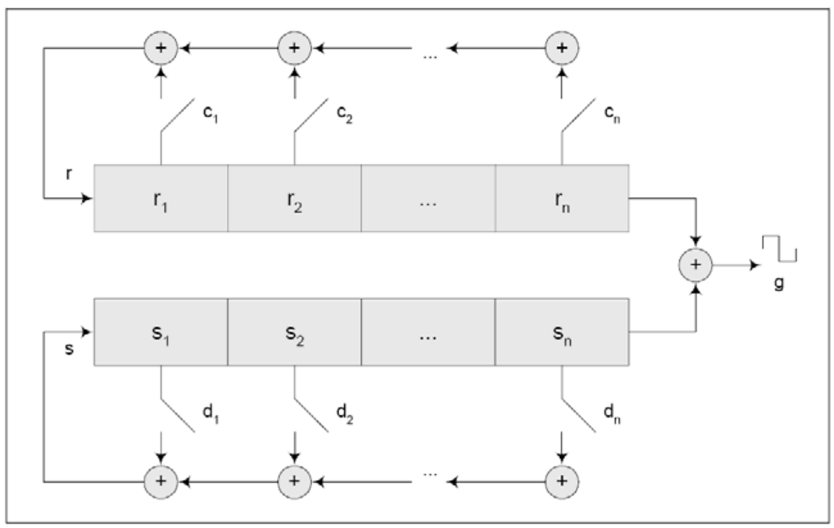


Figure 40 Gold Sequence Generator [5]

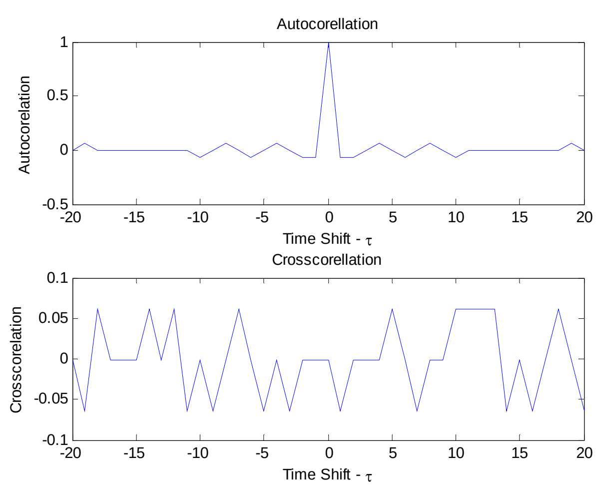


Figure 41 Autocorrelation and Cross correlation properties of Gold codes

And cross correlation given as

$$C(\tau)=\left\{\frac{1}{Nc}\left\{-1,-2^{\frac{n}{2}}-1,2^{\frac{n}{2}}-1\right\}\right. \quad \tau\neq 0 \tag{57}$$

A codelength of 511 was used. The generator polynomial is 001000010001.

Gold codes are not perfectly orthogonal, i.e. $C(\tau)\neq 0,\quad \tau\neq 0$ thus the cross correlation will have an effect on the acquisition system performance.

5.3.1.3 PULSE SHAPING FILTER

The chip waveform was spread using this filter. A design problem is the number of sample per pulse that is desired for a performance level. The number of sample per pulse affects the performance measures. More samples per chip give better performance results however at the expense of system speed and computation requirements. See Figure 42, Figure 43 and Table 7. Using a number of samples per pulse of 16 samples gives a good compromise between complexity and performance.

5.3.1.4 FADING CHANNEL SIMULATOR

A 3-path Rayleigh fading channel was used. The envelope of the Rayleigh was generated using the improved Jakes model [78].

5.3.2 RECEIVER

The CDMA receiver system model is shown in Figure 45. The receiver was implemented on the DSP.

5.3.2.1 RECEIVER FILTER

The matched filter matched to the spreading waveform transmitter filter is used to first filter the received signal.

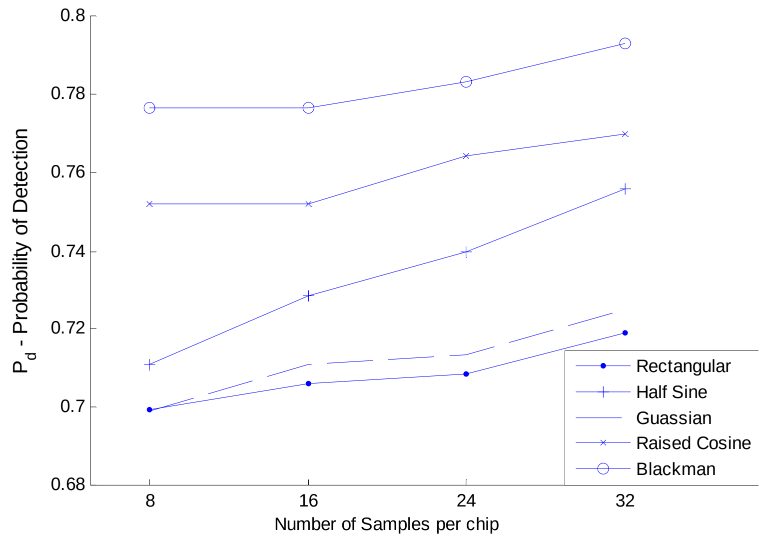


Figure 42 Performance with different sample per chip – Probability of Detection

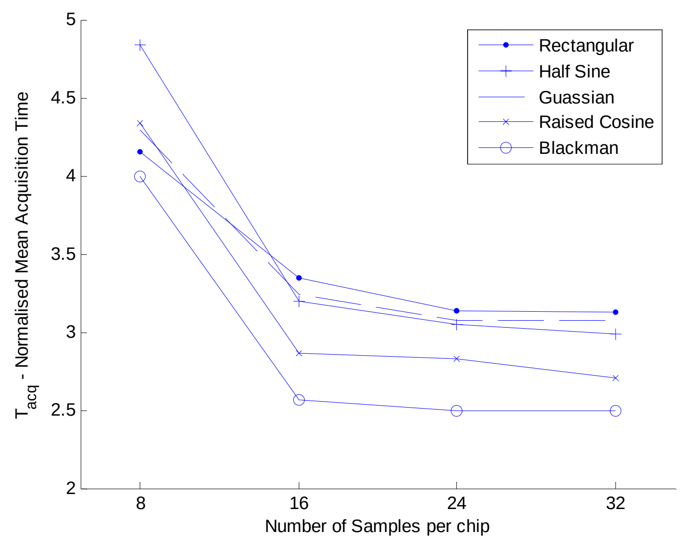


Figure 43 Performance with different sample per chip – Normalised Mean Acquisition Time

Table 7 Effect of Pulse samples on DSP performance

Number of Samples Per Chip	Number of CPU Cycles	Processing Time
8	1010109	4.4894ms
16	2754238	12.2410ms
24	5054684	22.4652ms
32	8330783	37.0257 ms

5.3.2.2 HYBRID SEARCH ALGORITHM

The simulation parameters of the system and the flowchart of for the algorithm is show in Table 8 and Figure 46 respectively.

5.4 REMARKS

The algorithm was implemented on the DSP. The number of sample per pulse affects the performance measures. More samples per chip give better performance results however at the expense of system speed and computation requirements.

A good compromise between complexity and performance is to sample at a rate of 16 samples per chip.

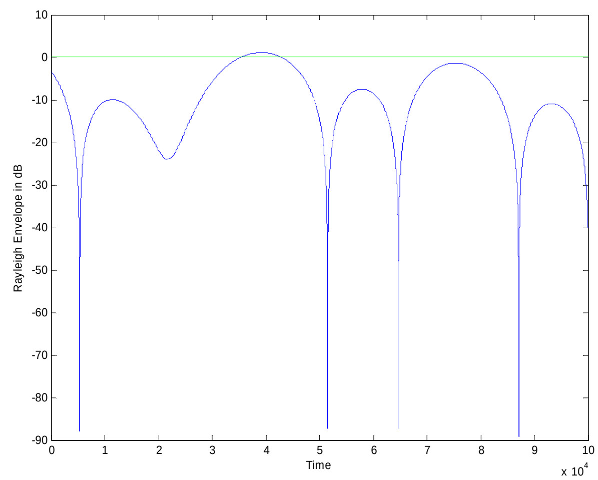


Figure 44 Rayleigh Fading Envelope

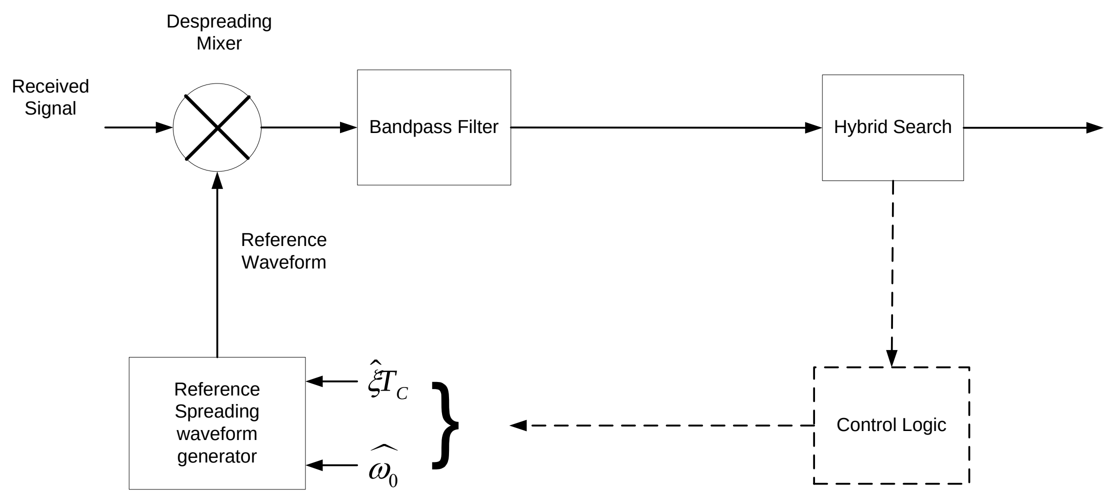
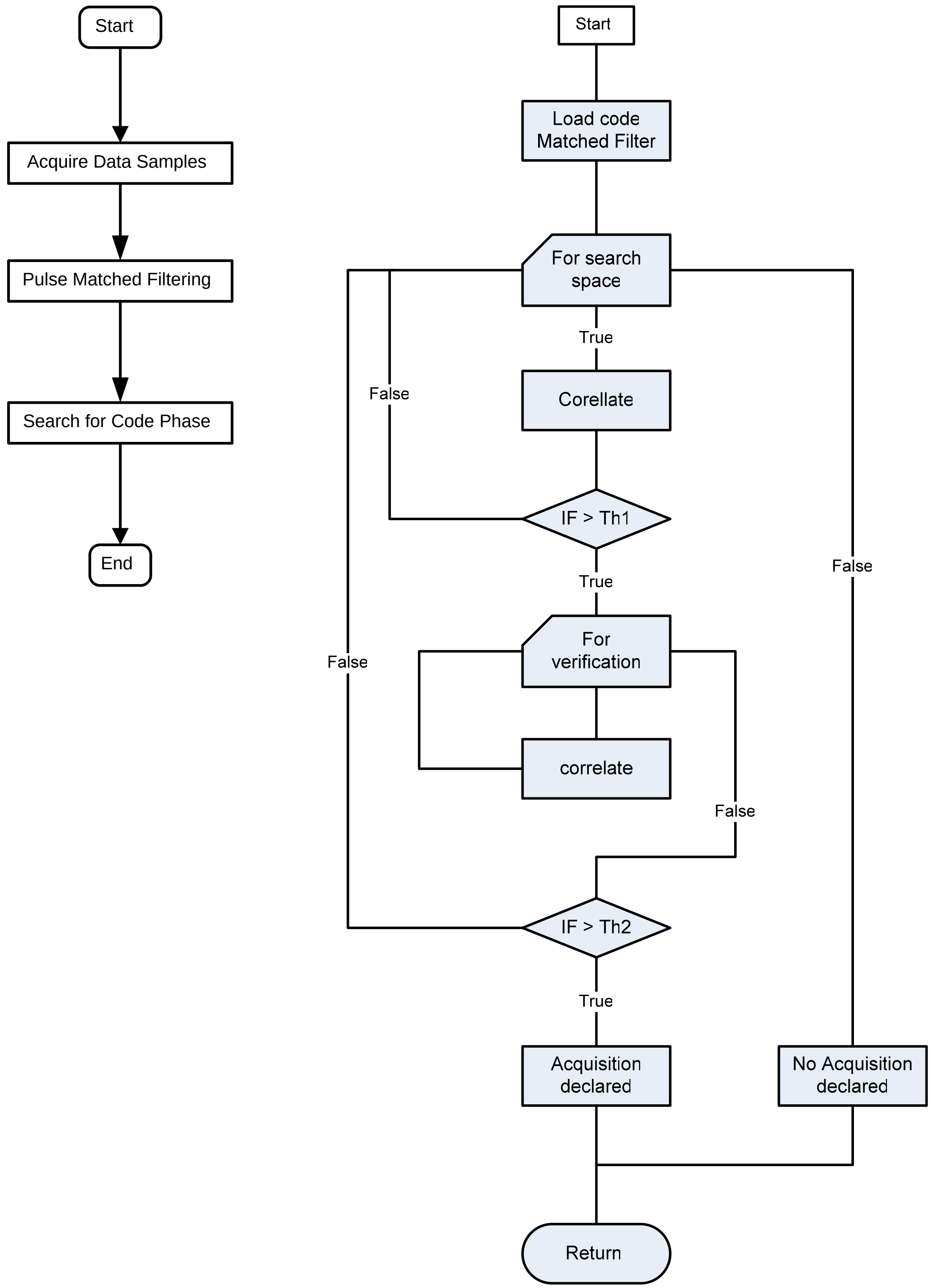


Figure 45 CDMA receiver

Table 8 DSP Implementation parameters

Number of Samples per Symbol	16
Sampling after matched filtering	Half the chip time
Codelength	511
Code Matched filter Length	64
Verification	640 sampling periods



(a) (b)

Figure 46 (a) Flowchart for Receiver & (b) Flow chart for search algorithm

CHAPTER 6

THESIS CONCLUSIONS

6.1 THESIS SUMMARY

The research goal of the thesis is to study the Hybrid Search scheme which is considered as one of the best candidates to fulfil the objectives of the acquisition subsystem. Our paramount objective was to investigate the effect of multiple access interference on the synchronisation performance of the search scheme.

The Improved Gaussian MAI approximation was applied in the analysis of the noncoherent Hybrid search receivers instead of the traditional Gaussian approximation. Modifications were made to the analysis of the noncoherent Hybrid search schemes to accommodate a more accurate multiple access interference model. Numerical results show differences in the performance of the system for both models. Results show that the standard Gaussian approximation gives results that are optimistic. To ensure proper planning for capacity the improved Gaussian approximation should be used.

The effect of different chip pulse shaping on the synchronisation performance was investigated. Blackman pulse gave the best results for both performance measures. The Blackman pulse performed best when compared with other pulse for both performance measures. In order of performance we have, Blackman, Raised Cosine, Half sine, truncated Gaussian and the worst being the Rectangular Pulse.

A modified Adaptive threshold setting scheme was proposed for the hybrid noncoherent receivers. Results show that the modified scheme outperforms other threshold schemes. The algorithm is also easy to implement.

The Hybrid search scheme and the threshold setting schemes were implemented on a TI C6713 DSK.

6.2 FUTURE WORK

The improved Gaussian approximation should also be applied in the modelling of fading channels. In our work we have considered Time limited pulses, investigation should be made into the effect of bandlimited pulses on the acquisition schemes.

We have assumed that the uncertainty period is uniformly distributed in our work. Future work should consider situations were there is apriori knowledge of where the correct cell is located.

The acquisition subsystem was implemented on the DSP. A goal for future work should be to implement the whole system on a DSP and hardware implementation on FPGA & ASIC. Other acquisition schemes for instance the one proposed in [32], should also be implemented in te DSP.

The algorithm was implemented using C and Assembly programming on the Code Composer Studio. Integration of Simulink and Code Composer studio for rapid prototyping on the DSP is a promising and simpler approach to software radio design. A future work should consider this young technology. This technology will provide a testbed for analysis of communication systems.

APPENDIX

Appendix - A   Pulses Examples

Rectangular Pulse

The time limited Rectangular Pulse Characteristics is given as

$$x(t)=\begin{cases}1, & \dfrac{-T_c}{2}\le t<\dfrac{T_c}{2}\\ 0, & otherwise\end{cases} \tag{58}$$

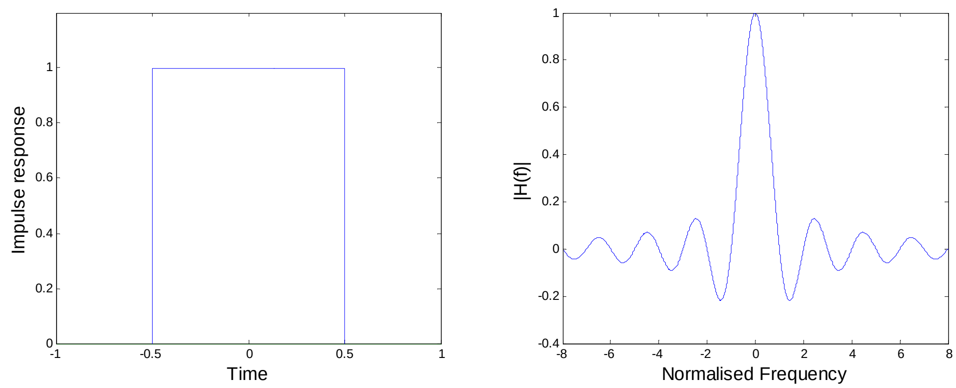


Figure 47   Rectangular Pulse

Half-Sine Pulse

The Half Sine Pulse Characteristics is given as

$$x(t)=\begin{cases}\sqrt{2}\sin\left(\pi t/T_C\right), & 0\le t<T_c\\ 0, & otherwise\end{cases} \tag{59}$$

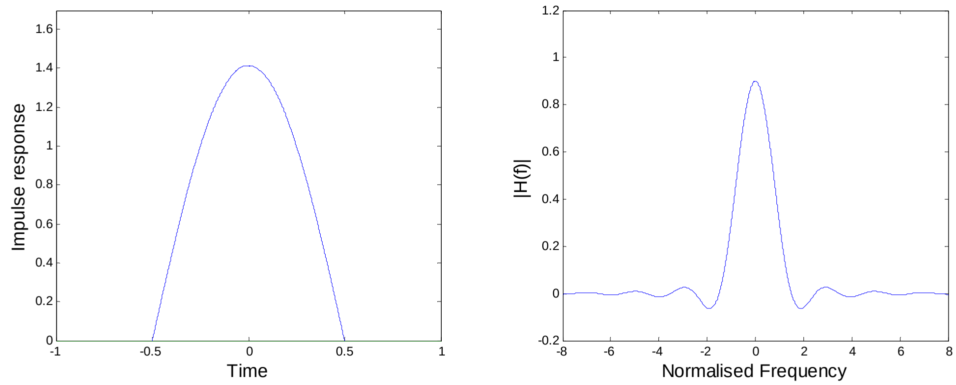


Figure 48 Half Sine Pulse

Gaussian Pulse

The Gaussian Pulse Characteristics is given as[47]

$$x(t)=\begin{cases}1 - \left|\frac{2t}{T_c}\right|, & 0 \le t < T_c \\ 0, & otherwise\end{cases} \tag{60}$$

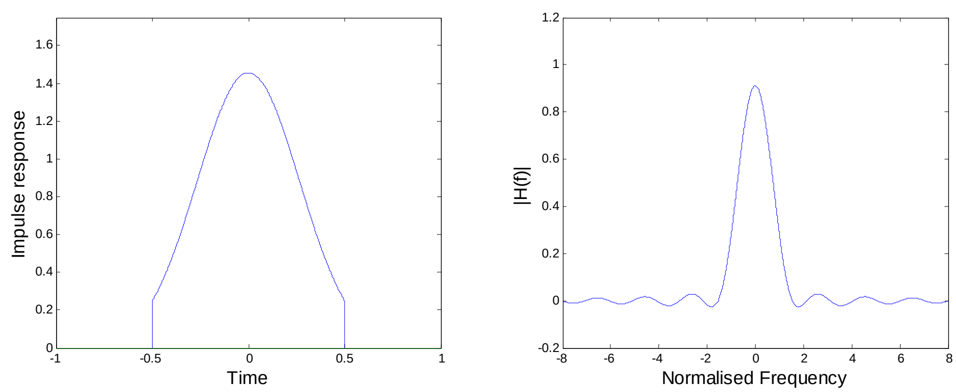
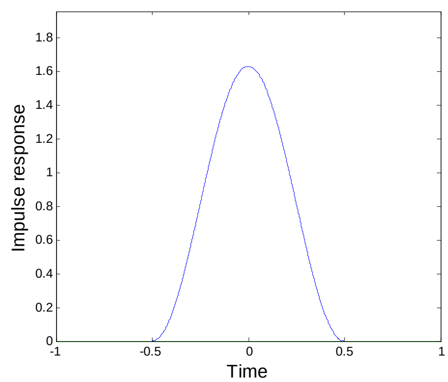


Figure 49 Gaussian Pulse Sine Pulse

Raised Cosine Pulse

The Time Domain Raised Cosine Pulse Characteristics is given as

$$x(t)=\begin{cases}\sqrt{2/3}\left[1-\cos\left(\dfrac{2\pi t}{T_c}\right)\right], & 0\le t<T_c \\ 0, & otherwise\end{cases} \quad (61)$$



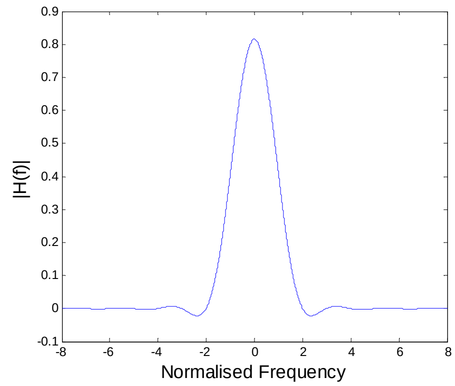


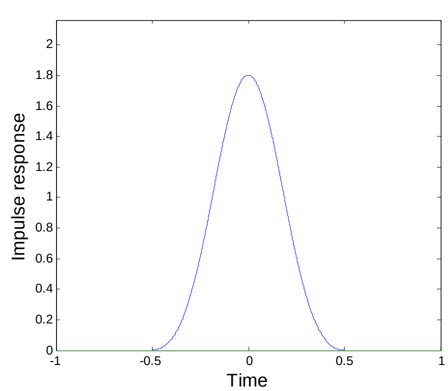
Figure 50 Raised Cosine Pulse

Blackman Pulse

The Blackman Pulse Characteristics is given as

$$x(t)=\begin{cases}c\left[k_1-k_2\cos\left(\dfrac{2\pi t}{T_C}\right)+k_3\cos\left(\dfrac{4\pi t}{T_C}\right)\right], & 0\le t<T_c \\ 0, & otherwise\end{cases} \quad (62)$$

$$where\ c^2=\left(k_1^2+k_2^2/2+k_3^2/2\right)^{-1},\quad k_1=0.42,\quad k_2=0.5,\quad k_3=0.08$$



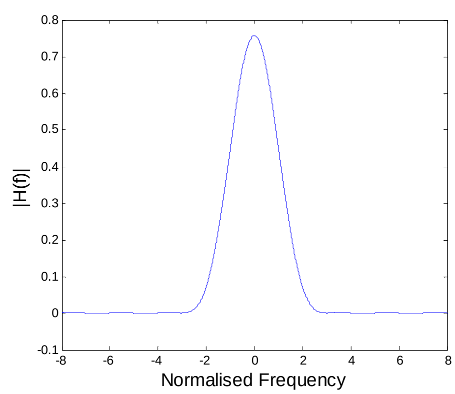


Figure 51 Blackman Pulse

Appendix - B Texas Instruments C6713 DSK

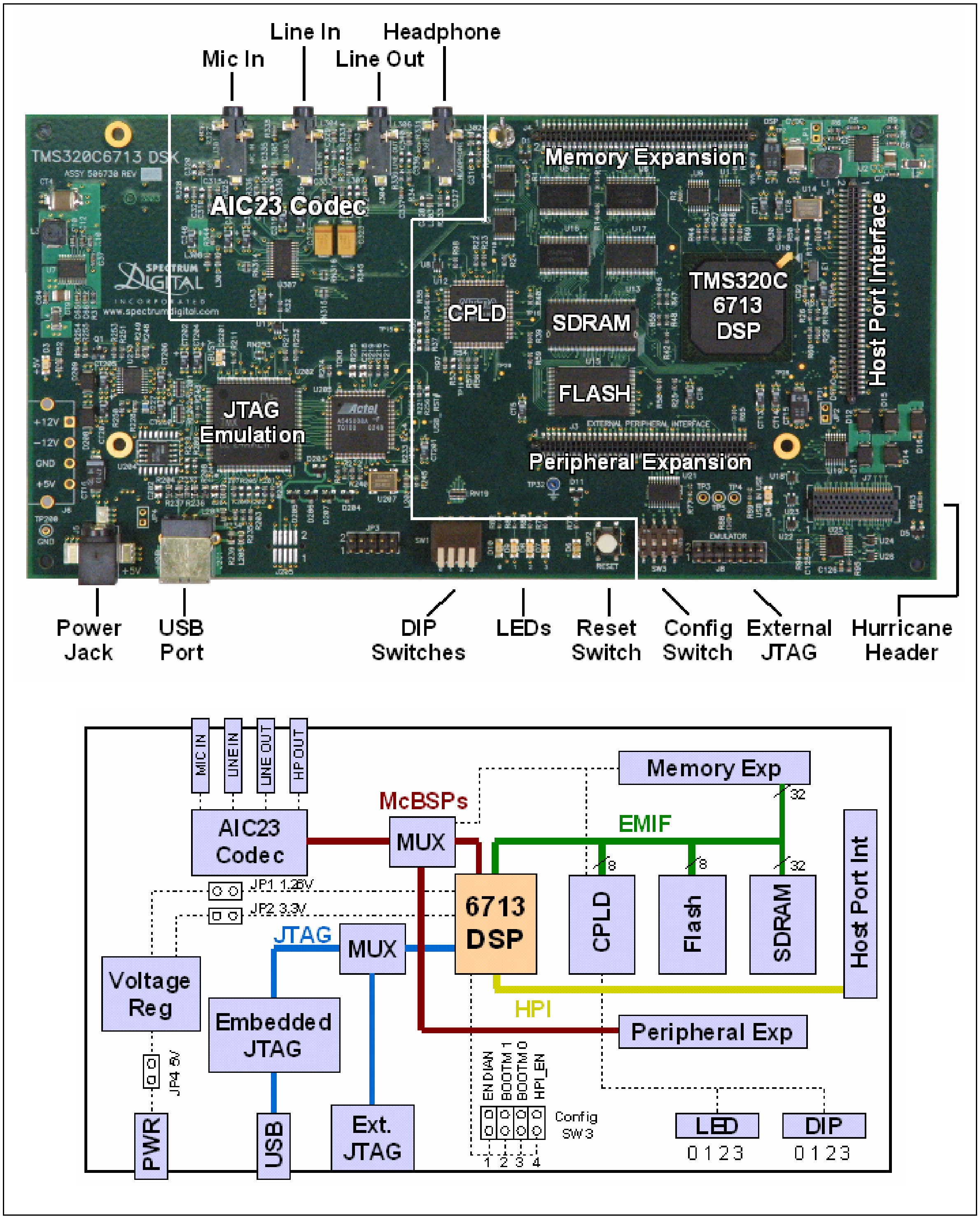


Figure 52 C6713-DSP Board

Features of the C6713

Features of the processor include:

- VLIW Core – VLIW is a processor architecture that allows many instructions to be issued (8 on the 6713 DSP) in a single clock while still allowing for very high clock rates.
- 192Kbytes Internal Memory – high speed internal memory for maximum performance.
- 64Kbytes L2 Cache/RAM – four 16Kbyte blocks of internal RAM that can be configured as RAM or cache.
- 4Kb Program/Data Caches – separate caches for program code and data.
- On-chip PLL – generates processor clock rate from slower external clock reference.
- 2 Timers – generates periodic timer events as a function of the processor clock. Used by DSP/BIOS to create time slices for multitasking.
- EDMA Controller – Enhanced DMA controller allows high speed data transfers without intervention from the DSP.
- 2 McBSPs – Multichannel buffered serial ports. Each McBSP can be used for high speed serial data transmission with external devices or reprogrammed as general purpose I/Os. 2 McASPs – Multichannel audio serial ports. Used for multi-channel and professional audio applications.
- 2 I2C Interfaces – Inter-Integrated Circuit Bus. An I2C bus is a serial bus format that can support several standard devices per bus.
- EMIF – External Memory Interface. A 32-bit bus on which external memories and other devices can be connected. The EMIF can interface to both synchronous and asynchronous memories.

REFERENCES

[1] J. Iinatti and P. Leppanen, "On the mean acquisition times of DS receiver code synchronization using matched filters and median filters in AWGN-channel and continuous/pulsed tone jamming," presented at Military Communications Conference, 1993. MILCOM '93. Conference record. 'Communications on the Move'., IEEE, 1993.

[2] L. B. Milstein and M. K. Simon, "Spread Spectrum Communications," in *The Communications Handbook*, J. D. Gibson, Ed., 2nd ed: CRC Press, 2002.

[3] J. Proakis, *Digital Communication*, 4th ed: McGraw-Hill, 2000.

[4] T. S. Rappaport, *Wireless Communications Principles and Practice*, 2nd ed: Prentice Hall, 2002.

[5] R. L. Peterson, R. E. Ziemer, and D. E. Borth, *Introduction to Spread Spectrum Communications*: Prentice Hall, 1995.

[6] G. L. Stuber, *Mobile Communication*, 2nd ed: Kluwer Academic Publishers, 2001.

[7] S. Glisic and B. Vucetic, *Spread Spectrum CDMA system for Wireless Communications*: Artech House Inc., 1997.

[8] P. Hopkins, "A Unified Analysis of Pseudonoise Synchronization by Envelope Correlation," *Communications, IEEE Transactions on*, vol. 25, pp. 770-778, 1977.

[9] A. Polydoros and C. Weber, "A Unified Approach to Serial Search Spread-Spectrum Code Acquisition--Part I: General Theory," *Communications, IEEE Transactions on*, vol. 32, pp. 542-549, 1984.

[10] C. Gumacos, "Analysis of an Optimum Sync Search Procedure," *Communications, IEEE Transactions on*, vol. 11, pp. 89-99, 1963.

[11] M. Katz, "Code Acquisition In Advanced CDMA Networks," in *Department of Electrical and Information Engineering*. Oulu: University of Oulu, 2002.

[12] D. DiCarlo and C. Weber, "Statistical Performance of Single Dwell Serial Synchronization Systems," *Communications, IEEE Transactions on*, vol. 28, pp. 1382-1388, 1980.

[13] J. Holmes and C. Chen, "Acquisition Time Performance of PN Spread-Spectrum Systems," *Communications, IEEE Transactions on*, vol. 25, pp. 778-784, 1977.

[14] A. J. Viterbi, *CDMA Principles of Spread Spectrum Communication*: Addison-Wesley Publishing Company, 1995.

[15] S. Glisic, "Code Acquisition," in *Wideband Wireless Digital Communication*, A. F. Molisch, Ed.: Prentice Hall, 2001.

[16] A. Polydoros and S. Glisic, "Code synchronization: a review of principles and techniques," presented at Spread Spectrum Techniques and Applications, 1994. IEEE ISSSTA '94., IEEE Third International Symposium on, 1994.

[17] M. H. Zarrabizadeh and E. S. Sousa, "A differentially coherent PN code acquisition receiver for CDMA systems," *Communications, IEEE Transactions on*, vol. 45, pp. 1456-1465, 1997.

[18] S. Rappaport and D. Grieco, "Spread-spectrum signal acquisition: Methods and technology," *Communications Magazine, IEEE*, vol. 22, pp. 6-21, 1984.

[19] B. G. Lee and B.-H. Kim, *Scrambling Techniques for CDMA Communications*. Boston: Kluwer Academic Publishers, 2001.

[20] L. Milstein, J. Gevargiz, and P. Das, "Rapid Acquisition for Direct Sequence Spread-Spectrum Communications Using Parallel SAW Convolvers," *Communications, IEEE Transactions on*, vol. 33, pp. 593-600, 1985.

[21] M. K. Simon, J. K. Omura, R. A. Scholtz, and B. K. Levitt, *Spread spectrum communications handbook*, Revised ed: McGraw-Hill, Inc., 1994.

[22] A. Polydoros and M. Simon, "Genralized Serial Search Code Acquisition: The Equivalent Circular State Diagram Approach," *Communications, IEEE Transactions on*, vol. 32, pp. 1260-1268, 1984.

[23] E. Sourour and S. C. Gupta, "Direct sequence spread spectrum parallel acquisition in a fading mobile channel," presented at Vehicular Technology Conference, 1989 IEEE 39th, 1989.

[24] G. Sage, "Serial Synchronization of Pseudonoise Systems," *Communications, IEEE Transactions on*, vol. 12, pp. 123-127, 1964.

[25] A. Polydoros and C. Weber, "A Unified Approach to Serial Search Spread-Spectrum Code Acquisition--Part II: A Matched-Filter Receiver," *Communications, IEEE Transactions on*, vol. 32, pp. 550-560, 1984.

[26] V. M. Jovanovic, "Analysis of strategies for serial-search spread-spectrum code acquisition-direct approach," *Communications, IEEE Transactions on*, vol. 36, pp. 1208-1220, 1988.

[27] A. Weinberg, "Generalized Analysis for the Evaluation of Search Strategy Effects on PN Acquisition Performance," *Communications, IEEE Transactions on*, vol. 31, pp. 37-49, 1983.

[28] W. Zhuang, "Noncoherent hybrid parallel PN code acquisition for CDMA mobile communications," *Vehicular Technology, IEEE Transactions on*, vol. 45, pp. 643-656, 1996.

[29] B.-J. Kang, H.-R. Park, and Y. Han, "Hybrid acquisition in DS/CDMA forward link," presented at Vehicular Technology Conference, 1997 IEEE 47th, 1997.

[30] C. Baum and V. Veeravalli, "Hybrid acquisition schemes for direct sequence CDMA systems," presented at Communications, 1994. ICC 94, SUPERCOMM/ICC '94, Conference Record, Serving Humanity Through Communications. IEEE International Conference on, 1994.

[31] E. A. Homier and R. A. Scholtz, "Rapid acquisition of ultra-wideband signals in the dense multipath channel," presented at Ultra Wideband Systems and Technologies, 2002. Digest of Papers. 2002 IEEE Conference on, 2002.

[32] M. G. El-Tarhuni and A. U. Sheikh, "Adaptive synchronization for spread spectrum systems," presented at Vehicular Technology Conference, 1996. 'Mobile Technology for the Human Race'., IEEE 46th, 1996.

[33] R. F. Smith and S. L. Miller, "Acquisition performance of an adaptive receiver for DS-CDMA," *Communications, IEEE Transactions on*, vol. 47, pp. 1416-1424, 1999.

[34] P. K. P. Cheung and P. B. Rapajic, "CMA-based code acquisition scheme for DS-CDMA systems," *Communications, IEEE Transactions on*, vol. 48, pp. 852-862, 2000.

[35] L. B. Milstein, "Interference suppression to aid acquisition in direct-sequence spread-spectrum communications," *Communications, IEEE Transactions on*, vol. 36, pp. 1200-1207, 1988.

[36] L. D. Gottesman and L. B. Milstein, "On the use of interference suppression to enhance acquisition performance in a CDMA overlay scenario," presented at Military Communications Conference, 1995. MILCOM '95, Conference Record, IEEE, 1995.

[37] Z.-L. Shi, P. F. Driessen, and W. Du, "PN code acquisition for asynchronous CDMA communications based on interference cancellation," presented at Communications, Computers and Signal Processing, 1993., IEEE Pacific Rim Conference on, 1993.

[38] R. Cameron and B. Woerner, "Synchronization of CDMA systems employing interference cancellation," presented at Vehicular Technology Conference, 1996. 'Mobile Technology for the Human Race'., IEEE 46th, 1996.

[39] U. Madhow, "MMSE interference suppression for timing acquisition and demodulation in direct-sequence CDMA systems," *Communications, IEEE Transactions on*, vol. 46, pp. 1065-1075, 1998.

[40] H.-C. Hwang and C.-H. Wei, "A new blind adaptive interference suppression scheme for acquisition and MMSE demodulation of DS/CDMA signals," *Vehicular Technology, IEEE Transactions on*, vol. 49, pp. 875-884, 2000.

[41] S. E. Bensley and B. Aazhang, "Maximum-likelihood synchronization of a single user for code-division multiple-access communication systems," *Communications, IEEE Transactions on*, vol. 46, pp. 392-399, 1998.

[42] J. B. Lee and S. K. Oh, "A code acquisition scheme with interference cancellation for DS-CDMA systems," presented at Vehicular Technology Conference, 2001. VTC 2001 Spring. IEEE VTS 53rd, 2001.

[43] S. E. Bensley and B. Aazhang, "Subspace-based channel estimation for code division multiple access communication systems," *Communications, IEEE Transactions on*, vol. 44, pp. 1009-1020, 1996.

[44] E. G. Strom, S. Parkvall, S. L. Miller, and B. E. Ottersten, "Propagation delay estimation in asynchronous direct-sequence code-division multiple access systems," *Communications, IEEE Transactions on*, vol. 44, pp. 84-93, 1996.

[45] U. Madhow, "Blind adaptive interference suppression for the near-far resistant acquisition and demodulation of direct-sequence CDMA signals," *Signal Processing, IEEE Transactions on [see also Acoustics, Speech, and Signal Processing, IEEE Transactions on]*, vol. 45, pp. 124-136, 1997.

[46] A. Mantravadi and V. V. Veeravalli, "Multiple-access interference-resistant acquisition for band-limited CDMA systems with random sequences," *Selected Areas in Communications, IEEE Journal on*, vol. 18, pp. 1203-1213, 2000.

[47] K. S. Zigangirov, *Theory of Code Division Multiple Access Communication*: Wiley-IEEE Press, 2004.

[48] S.-M. Pan, D. E. Dodds, and S. Kumar, "Acquisition time distribution for spread-spectrum receivers," *Selected Areas in Communications, IEEE Journal on*, vol. 8, pp. 800-808, 1990.

[49] H. Meyr and G. Polzer, "Performance Analysis for General PN-Spread-Spectrum Acquisition Techniques," *Communications, IEEE Transactions on [legacy, pre - 1988]*, vol. 31, pp. 1317-1319, 1983.

[50] M. D. Katz and S. Glisic, "Modeling of code acquisition process in CDMA networks-asynchronous systems," *Selected Areas in Communications, IEEE Journal on*, vol. 18, pp. 73-86, 2000.

[51] G. E. Corazza, "On the MAX/TC criterion for code acquisition and its application to DS-SSMA systems," *Communications, IEEE Transactions on*, vol. 44, pp. 1173-1182, 1996.

[52] R. R. Rick and L. B. Milstein, "Parallel acquisition in mobile DS-CDMA systems," *Communications, IEEE Transactions on*, vol. 45, pp. 1466-1476, 1997.

[53] J. Y. Kim, "Impact of power control error on PN acquisition performance for a code-division-multiple-access system," presented at Communications, Computers and Signal Processing, 1999 IEEE Pacific Rim Conference on, 1999.

[54] R. Anjaria and R. Wyrwas, "The effect of chip waveform on the performance of CDMA systems in multipath, fading, noisy channels," presented at Vehicular Technology Conference, 1992 IEEE 42nd, 1992.

[55] P. Xiao, E. Strom, P. Orten, and A. Svensson, "Pulse Shaping Considerations in DS-CDMA Systems with Long Spreading Sequences," presented at ISPACS'2002(IEEE International Symposium on Intelligent Signal Processing and Communication Systems), Kaohsiung,Taiwan, 2002.

[56] M. G. El-Tarhuni and M. A. Landolsi, "The effect of chip waveform shaping on the synchronization performance of DS-CDMA signals," presented at Vehicular Technology Conference, 1999 IEEE 49th, 1999.

[57] S. Thayaparan, T.-S. Ng, and J. Wang, "Half-sine and triangular despreading chip waveforms for coherent delay-locked tracking in DS/SS systems," *Communications, IEEE Transactions on*, vol. 48, pp. 1384-1391, 2000.

[58] C. Lee, K. Lee, Y. Choi, and K. Kim, "Analysis on detection performance of an energy detector for several chip waveforms in a DS/SS communication," presented at Military Communications Conference, 2001. MILCOM 2001. Communications for Network-Centric Operations: Creating the Information Force. IEEE, 2001.

[59] H. H. Nguyen and E. Shwedyk, "On error probabilities of DS-CDMA systems with arbitrary chip waveforms," *Communications Letters, IEEE*, vol. 5, pp. 78-80, 2001.

[60] J. M. Holtzman, "A simple, accurate method to calculate spread-spectrum multiple-access error probabilities," *Communications, IEEE Transactions on*, vol. 40, pp. 461-464, 1992.

[61] E. Siess and C. Weber, "Acquisition of Direct Sequence Signals with Modulation and Jamming," *Selected Areas in Communications, IEEE Journal on*, vol. 4, pp. 254-272, 1986.

[62] B. B. Ibrahim and A. H. Aghvami, "Direct sequence spread spectrum matched filter acquisition in frequency-selective Rayleigh fading channels," *Selected Areas in Communications, IEEE Journal on*, vol. 12, pp. 885-890, 1994.

[63] K. Choi, K. Cheun, and T. Jung, "Adaptive PN code acquisition using instantaneous power-scaled detection threshold under Rayleigh fading and pulsed Gaussian noise jamming," *Communications, IEEE Transactions on*, vol. 50, pp. 1232-1235, 2002.

[64] N. M. Stojanovic and V. M. Jovanovic, "New robust threshold setting method for frequency-hopping acquisition," *Electronics Letters*, vol. 28, pp. 815-817, 1992.

[65] J. Iinatti, "Robust method for threshold setting in a DS/DS-CDMA code acquisition," presented at Universal Personal Communications, 1996. Record., 1996 5th IEEE International Conference on, 1996.

[66] S. G. Glisic, "Automatic decision threshold level control (ADTLC) in direct-sequence spread-spectrum systems based on matched filtering," *Communications, IEEE Transactions on*, vol. 36, pp. 519-527, 1988.

[67] J. H. J. Iinatti, "On the threshold setting principles in code acquisition of DS-SS signals," *Selected Areas in Communications, IEEE Journal on*, vol. 18, pp. 62-72, 2000.

[68] J. Iinatti, "Comparison of two dwell code acquisition of DS signal using different threshold setting rules," presented at MILCOM 97 Proceedings, 1997.

[69] B. B. Ibrahim and A. H. Aghvami, "Direct sequence spread spectrum code acquisition in mobile fading channel using matched filter with reference filtering," presented at Global Telecommunications Conference, 1993, including a Communications Theory Mini-Conference. Technical Program Conference Record, IEEE in Houston. GLOBECOM '93., IEEE, 1993.

[70] P. M. Grant, "DSP chips for CDMA mobile communications," presented at Spread Spectrum Techniques and Applications Proceedings, 1996., IEEE 4th International Symposium on, 1996.

[71] D.-S. Lyu, J.-S. Kim, K.-M. Ha, J. H. Lee, J.-M. Kim, Y.-G. Jeong, J. S. Ha, and N.-J. Park, "Implementation of wideband CDMA modem using processors and programmable logic devices," presented at Circuits and Systems, 1997. ISCAS '97., Proceedings of 1997 IEEE International Symposium on, 1997.

[72] J.-W. Chung, J.-S. Kim, Y.-G. Jeong, and J.-S. Ha, "Implementation of code acquisition and code tracking loop for CDMA wireless local loop system," presented at Vehicular Technology Conference, 1998. VTC 98. 48th IEEE, 1998.

[73] P. R. Dent, "W-CDMA reception with a DSP based software radio," presented at 3G Mobile Communication Technologies, 2000. First International Conference on (IEE Conf. Publ. No. 471), 2000.

[74] "http://focus.ti.com/docs/prod/folders/print/tms320c6701.html."

[75] R. Chassaing, *Digital Signal Processing and Applications with the C6713 and C6416 DSK*: John Wiley & Sons, Inc., 2005.

[76] M. Stular and S. Tomazic, "Maximum periodic correlation of pseudo-random sequences in CDMA," presented at Electrotechnical Conference, 2000. MELECON 2000. 10th Mediterranean, 2000.

[77] R. Gold, "Optimal binary sequences for spread spectrum multiplexing (Corresp.)," *Information Theory, IEEE Transactions on*, vol. 13, pp. 619-621, 1967.

[78] M. F. Pop and N. C. Beaulieu, "Limitations of sum-of-sinusoids fading channel simulators," *Communications, IEEE Transactions on*, vol. 49, pp. 699-708, 2001.

VITA

Aregbesola Kassim Mayowa is a graduate student at department of Electrical Engineering, King Fahd University, Dhahran Saudi Arabia. He attained his bachelors of Science degree at Obafemi Awolowo University, Nigeria in 2001. He is currently conducting research on the synchronisation aspects of CDMA Spread Spectrum. Areas of interest is in Wireless Communication specifically Spread Spectrum techniques and Multiuser communication.

Prior to joining KFUPM, he has previously worked with Mobitel Nigeria Limited, deploying CDMA Wireless Local Loop systems. He also worked at Allied Technologies, Nigeria, as a Systems engineer and at Octagon Technologies as a Network Engineer.