CODE ACQUISITION TECHNIQUE USING ADAPTIVE FILTERS IN WIRELESS COMMUNICATION

BY

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This thesis, written by **MUHAMMAD HAMID** 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 fulfillment of the requirements for the degree of

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Dedicated to

My Parents

and

My

Grand Mother

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In the name of Allah, the Most Gracious and the Most Merciful

All praise and glory to Almighty Allah (SWT) who gave me courage and patience to carry out this work. Peace and blessing of Allah be upon His last Prophet Muhammad (Sallulaho-Alaihe-Wassalam) and all his Sahaba (Razi-Allaho-Anhum) who devoted their lives for the spread of Islam.

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Contents

List of Tables

List of Figures

THESIS ABSTRACT

Name: Muhammad Hamid

Title: Code Acquisition Technique using Adaptive Filters in Wireless Communication

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Direct-Sequence Spread Spectrum (DS/SS) systems offer several attractive features which motivates growing research on code synchronization (involves both Code Acquisition and Tracking) for **SS** *systems as result of the crucial role that it plays in operation of* **SS** *systems. This thesis introduces a modified adaptive filter structure for Code Acquisition which is different from the one proposed earlier.*
*An adaptive filter (AF) is used to carry the function of code acquisition. The taps of the filter contains the results of the sum of a block of chips. These contents are then compared with the similar summation of the locally generated sequence. The filter coefficients are adjusted towards their optimum value by using the Least-Mean-Square (***LMS***) algorithm and the filter tap weights are used to estimate the delay offset between the incoming* **SS** *signal and a locally generated* **PN** *sequence by minimizing the mean square error* **MSE** *between the received and the locally generated sequences. This new method first locates the region or the cell in which the correct phase lies and then the phase itself is acquired by zeroing onto the chip within the cell identified.*
The analysis of the system is done in AWGN environment for single user and theoretical and simulation results are presented which shows that the system performance is quite significant for low SNR conditions.
It is also shown that our modified scheme is robust to fading (frequency selective and flat rayleigh fading) for low SNR environment. The adaptive nature of the system is exploited to acquire and track the strongest received path.
Moreover, it has been shown that the proposed acquisition scheme significantly outperforms conventional Matched Filter based acquisition schemes in the presence of Multiple Access Interference (MAI).

Keywords: *CDMA, Adaptive Filter, Phase Synchronization, Code Acquisition, LMS, fading, Multiple Access Interference*

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خلاصة الرسالة

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أنظمة التوسيع الموجي ذو التتابع المباشر تقدم ميزات جذابة و متعددة تحث البحث المتنامي عن التوقيت الرمزي (يتظمن كلا من توافق الرمز و التتابع) لأنظمة التوسيع الطيفي كنتيجة للدور الهام الذي يلعبه في عمل أنظمة التوسيع الطيفي. هذه الرسالة تقدم تركيب مصفي تكيفي معدل للتوافق الرمزي بما هو مختلف عن المقترح سابقا. استخدم مصفي تكيفي وظيفة التوافق الرمزي. أجزاء المصفي تحتوي الناتج من مجموع عدد من الشرائح. هذه المحتويات تقارن بعد ذلك مع مجموع مشابه من متابعة مولدة محليا. معاملات المصفي تعدل بحسب القيمة الأفضل باستخدام طريقة المتوسط التربيعي الأقل LMS وأوزان أجزاء المصفي تستخدم لتقدير التأخر بين موجة SS القادمة و متابعة PN المولدة محليا. بتقليل متوسط الخطأ التربيعي بين المتابعة المستقبلة والمولدة محليا. هذه الطريقة الجديدة تعين أولا المنطقة أو الخلية حيث زاوية الموجة تقع ثم الزاوية نفسها يتحصل عليها بتصغير الشريحة خلال الخلية المتعرف عليها. تم تحليل النظام أولا في قنوات AWGN لمستخدم واحد و النتائج النظرية و نتائج المحاكاة معطاة حيث تبين أن أداء النظام متميز إلى حد كبير عند حالات SNR المنخفضة. أيضا بينا أن طريقتنا المعدلة جيدة للضعف الموجي (الثابت و الانتقائي) لقنوات SNR الضعيفة. الطبيعة المتكيفة للنظام استخدمت لتحصيل و متابعة أقوى موجة مستقبلة . اظافة إلى ذلك وضحنا أن طريقة التوافق المقترحة تفوق طرق التوافق العادية المرتكزة على المصفي المؤقت في وجود التداخل الموجي المتعدد MAI .

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Chapter 1

Introduction

1.1 Spread Spectrum System Concept

The traditional approach to digital communications is based on the idea of transmitter as much information as possible in as narrow a frequency bandwidth as possible. Therefore, a concept called *Narrowband Signal* (s_n) is used to yield narrowband systems. The most general concept of spread spectrum systems is presented in Fig 1.1 Formally, the operation of both the transmitter and receiver can be partitioned into two steps. In the first step, which we refer to as the primary modulation, the narrowband signal s_n is formed. In the second step, or secondary modulation, the operation $\epsilon(\ )$ is applied, resulting in the expansion of the signal spectrum to a very wide frequency band. This signal will be denoted as s_w.

At the receiver side, the first step is despreading, which is formally presented by

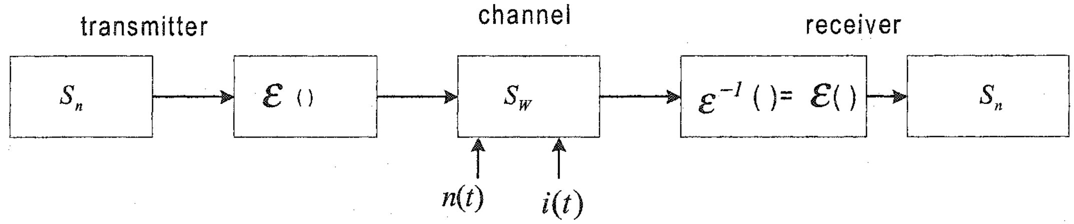

Figure 1.1: Spread Spectrum system Concept

the operation $\epsilon^{-1}(\ ) = \epsilon(\ )$. In other words, after despreading (which is identical to spreading process), the wideband signal s_w is converted back to the original form s_n and standard methods for narrowband signal demodulation are used.

In short, in a spread spectrum system, the transmission bandwidth used by the signal is usually much larger than the bandwidth of the information source; a pseudo-random code is used to spread the information signal power over a bandwidth much wider than the required minimum signal bandwidth. At the receiver, the same code is used to despread the received signal to its original information bandwidth. This is followed by data detection. This processing results in rejection of any interfering signals that may exist in the same transmission bandwidth.

1.1.1 Motivation

In this section, we will explain the origins of spread spectrum systems to highlight the reasons behind using spread spectrum systems in certain applications, among those are:

Antijamming Capabilities

Spread spectrum systems provide reliable communication in the presence of intensive jamming. If the spreading of the input signal is done at the transmitter side (i.e; resulting in wider band) and that the signal is received in the presence of a relatively narrowband and a much stronger jamming signal, then at the receiver side, the despreading process will convert the input signal into a sum of the narrowband useful and the wideband interfering signals. After narrowband filtering, only a very small portion of the interfering signal energy will pass the filter and remain as residual interference. Processing gain is the parameter that defines how much interfering signal is suppressed. The antijamming capability of spread spectrum system has been exploited in military communications.

Operation in Multipath Environment

As a result of multipath propagation, a transmitted signal will be received as a number of its mutually delayed replicas. Most of the time, these signal components will act as interference to each other and the net result will be the degradation of the system performance. We can separate these components and combine them synchronously into a larger signal vector that would provide good signal modulation conditions by using so-called Rake Receiver which works on the principle of spread spectrum.

Code Division Multiple Access

Since bandwidth is a limiting resource, spread spectrum systems employ a transmission bandwidth that is several orders of magnitude greater than the minimum required signal bandwidth (modulation scheme is called CDMA i.e; Code Division Multiple Access). While this system is very bandwidth inefficient for single user, the advantage of spread spectrum is that many users can simultaneously use the same bandwidth without significantly interfering with one another. In a multiple-user where *multiple access Interference* (MAI) is very dominant, spread spectrum system becomes bandwidth efficient.

Multiple Access Interference Rejection Capability

A very important advantage of spread spectrum system is its inherent Interference rejection capability. Since each user is assigned a unique **PN** code (described in next section), which is approximately orthogonal to the codes of other users, the receiver can separate each user based on their codes, even though they occupy the same spectrum at all times. This implies that, up to a certain number of users sharing the same frequency, interference occurring between spread spectrum signals using the same frequency is negligible.

1.2 Spreading Codes

The waveform used to spread and despread the data in spread spectrum systems is commonly known as spreading codes. These spreading codes are usually generated using linear feedback shift registers. For the spread spectrum system to operate efficiently, spreading codes are selected to have certain desirable properties [1]. For example, the autocorrelation function must be closer to a delta function in order to facilitate synchronization and Rake reception. The spreading codes must have low cross-correlation so that multiple access interference (MAI) should be insignificant. Ideally, the length of the code must be infinite which is obviously not possible, thus shorter periodic codes known as *pseudorandom codes (PN codes)* are sought. Specific PN codes include the maximal-length codes and Gold codes, among others. A summary of these codes and other related topics are as follows:

1.2.1 Maximal Length Sequences (m-sequences)

Maximal length sequences, are by definition, the largest periodic codes that can be generated by a given shift register or a delay element of a given length [2]. The feedback connections are generally represented by polynomials called *generator polynomials* like

$$f(X) = f_0 \oplus f_1 X \oplus \cdots \oplus f_{n-1} X^{n-1} \oplus f_n X^n \tag{1.1}$$

for linear feedback shift register (LFSR) of length n as shown in Fig. 1.2 where $\oplus$ denotes the modulo2 sum of feedback connections. In (1.1), for $0 < i < n$, f_i represents the connection from the ith unit of the shift register and may have a value of 0 or 1 showing the 'presence' or 'absence' of the feedback connection, respectively.

For a given shift register length, the feedback connections or $f(X)$ determine whether the sequence will be maximal length or not [2]. $f(X)$ must be a *primitive polynomial* of degree n to generate m-sequence with period $2^n - 1$. This means that all zero state is not permissible. A table of primitive polynomials of degree less than 35 is available in [3].

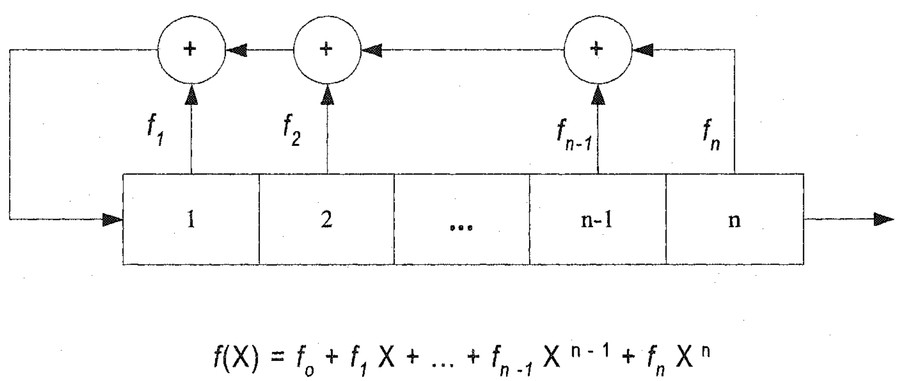

Figure 1.2: Linear Feedback Shift Register.

1.2.2 Gold Sequences

The cross-correlation between the pairs of the family of m-sequences varies from pair to pair. It is possible to find some pairs of m-sequences that have three valued cross correlation such sequences are known as Gold sequences named after Gold [4]. Such pairs of m-sequences are called *preferred sequences.* A list of preferred sequences of degree 18 or less can be found in [5]. Gold codes are constructed by the modulo-2 addition of the output of two LFSRs whose generator polynomials of these LFSRs form a pair of preferred $m-$sequences. The complete family of Gold codes for this generator is obtained using different initial loads of the shift register [6].

1.3 Code Synchronization

Code synchronization means that the time delay between the incoming code and the locally generated code is made to be very small compared to the time duration of a code symbol, which is known as a chip interval. Furthermore, Code synchronization must be maintained by the system for the duration of communication. The time delay, which the receiver must estimate to synchronize, is a random parameter which is usually assumed to be uniformly distributed over the time uncertainty region. Several sources contribute to the randomness of the time delay estimation process including initial random phase offset between the transmitters and receivers codes, random channel variations, interference effects, AWGN, unstable clock generators, Doppler variations, and more importantly the distance dependent time delays.

Code synchronization is usually developed over two steps, namely acquisition and tracking ([7]-[8]). "Acquisition is the process by which the two codes are coarsely aligned such that their phase offset is within a chip duration, while tracking is used to minimize the delay offset within a certain fraction of a chip and to maintain the time alignment of the two codes."

1.3.1 Motivation

Due to the noise-like correlation characteristic of the PN code, the despreading code must be perfectly synchronized to the code in the incoming signal in order to ensure complete recovery of the information data.

If code synchronization is not established accurately then not only that none of the claimed advantages of spread spectrum, such as interference rejection, multiple access, multipath diversity, etc., are obtained, but also reliable exchange of information even in the absence of receiver noise becomes very difficult if not impossible. Therefore, fast code synchronization is desirable to reduce the overhead time needed in accessing the system; this is particularly important in simplex applications and packet-switching data networks. In the proposed CDMA digital cellular systems, code synchronization has a critical role in the operation of such systems since the efficiency of the access channels may also suffer which may result in capacity loss.

Several performance measures could be used to assess the efficiency of a code synchronization scheme. For instance, performance of acquisition systems is usually measured by the statistics of the time elapsed prior to acquisition. In addition, depending on the intended application, the mean acquisition time or the probability of achieving acquisition within a specific time limit may be used to compare different acquisition schemes. Another factor which need to be considered is the hardware complexity of the system since some schemes achieve fast acquisition by expanding on the hardware requirements. For tracking systems, the mean time to lose lock and the variance of the estimation error are usually important.

1.4 Thesis Organization

The Thesis is organized as follows. Chapter 2 presents an overview of the classification of the Code synchronization schemes proposed earlier and the previous research in the area of code synchronization of **SS** signals. Different techniques are described as well as a general description of the most recent adaptive filter based code synchronization technique is also presented and detail description is given to Tarhuni-Sheikh proposed adaptive filter (AF) based algorithm. Finally the motivation behind the Modified adaptive filter (MAF)-based structure proposed in this thesis is described.

In Chapter 3, a detailed description of the proposed acquisition system is presented. The complete analysis for the proposed acquisition system in AWGN environment is given both in cell identification stage and phase acquisition stage. A brief description on the selection of LMS step size parameter is presented. The acquisition process is modelled as a Markov Process and the transition probability matrix is presented for the proposed system. Simulation and Analytical results for the Mean Acquisition Time are shown to agree quite well for the proposed system.

In Chapter 4, acquisition performance of the proposed system is studied under pedestrian fading conditions. Simulation results for the Mean Acquisition Time for the proposed system is presented both for frequency selective (for both Exponential MDP and Uniform MDP) and frequency non-selective (flat) Rayleigh fading environment.

In Chapter 5, the proposed acquisition system is investigated in the presence of Multiple Access Interference (MAI). The analysis of probability of cell acquisition failure for the cell identification stage is presented and simulation and analytical results are shown to agree quite well for four users case. Moreover the effect of adaptation time on acquisition failure is also presented in the presence of increasing number of users and it has been shown that the acquisition based capacity for the proposed system is far better than conventional matched filter scheme. Simulation results for the mean acquisition time performance is also presented for four users to see the degradation in performance in the presence of MAI.

Chapter 6 presents some concluding remarks and future work to follow this research.

Chapter 2

Synchronization of DS/SS Systems

Spread spectrum systems have received considerable attention in military applications due to their attractive features such as anti-jam capability, low probability of intercept, and accurate ranging. Recently, it was recommended to use SS techniques in commercial systems such as digital cellular radio systems where it is expected to provide larger capacity improvement than that obtained using TDMA systems [9]. Also, SS is a prime candidate for use in PCS and indoor wireless systems [10] [11] [12]. All SS applications require code synchronization prior to data detection. There have been extensive research activities in developing fast, reliable, and simple synchronization schemes for SS systems.

Recent research in synchronization techniques for spread spectrum (SS) systems is motivated by the growing interest in using SS for commercial digital cellular mobile radio systems as well as in many wireless data networks ([9]-[13]). The development of reliable and fast synchronization techniques for SS signals has received considerable attention in spread spectrum communication systems research ([7] [14] and references therein).

2.1 Classification of Code Acquisition Techniques

Code acquisition schemes can be classified as active serial search, passive matched filters, sequential estimation methods, and the adaptive filter based synchronization scheme which will be considered separately.

2.1.1 Serial Search System

In active serial search techniques [15] as depicted in Fig 2.1 the time uncertainty region is quantized with a fraction of a chip interval which are called as cells (i.e; each chip is called a cell). A locally generated PN code, with its phase set at the start of the quantized uncertainty region, is correlated with the incoming signal. The correlator output is compared to a threshold after a fixed examination period which is generally much longer than the chip duration. If the threshold is not exceeded, the local code phase is advanced to the next cell and the process is repeated until the threshold is crossed. Once the correct cell is detected, the tracking circuit is initiated either directly or after a verification mode which is used to reduce the effects of false alarm errors (wrong indication of acquisition).

Serial search schemes are generally simple and reliable under small signal-to-noise ratio conditions because they makeup the processing gain in the acquisition process, but their acquisition time performance is relatively poor especially if the uncertainty region is large. Their performance could be improved by increasing the search rate of the system through parallel or hybrid searching strategies. In certain situations, where the probability distribution of the delay offset is not uniform, other searching strategies may improve the performance.

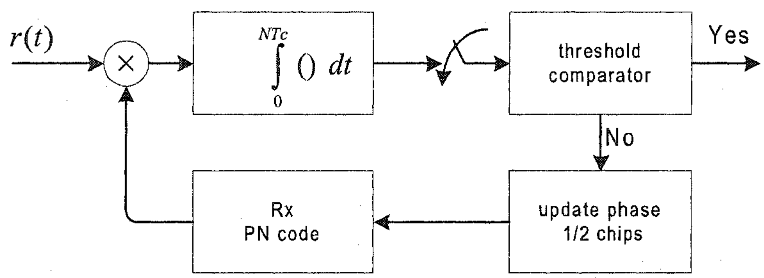

Figure 2.1: Serial Search System with 1/2-chip step size

2.1.2 Parallel Search systems

In parallel acquisition schemes [16] as depicted in Fig 2.2, all the cells (each chip corresponds to one cell) are tested simultaneously and the cell that results in a maximum correlation is assumed tentatively as the correct phase offset. Further verification of this cell is used to minimize false alarm errors. Parallel systems have fast acquisition time performance but their hardware requirements are usually prohibitively complex, particularly if the uncertainty region is large.

On the other hand, a hybrid acquisition system tries to compromise the acquisition speed with hardware complexity. This is achieved by dividing the uncertainty region into disjoint intervals each with more than one cell. The system starts searching the intervals sequentially while the cells within each interval are tested concurrently.

Passive acquisition systems provide fast acquisition performance by using matched filters. The receiver does not generate a replica of the PN code but it uses a filter with an impulse response matched to the PN code. The output of the matched filter (MF) will have a peak at the correct time instant which corresponds to the delay offset between the two codes. Nevertheless, this technique is only used for systems with short codes due to the implementation complexity encountered in designing matched filters with long periods. Some variation of the MF technique may be used as an acquisition aiding subsystem in which a synchronization preamble is appended to the spreading code. A filter matched to the preamble is used to detect the code starting time.

2.1.3 Sequential Estimation Method

Acquisition by sequential estimation [17] is based on estimating the present state of the incoming PN code. The local code generator is initially loaded with this estimate and a decision is made to check for acquisition after a correlation period. If

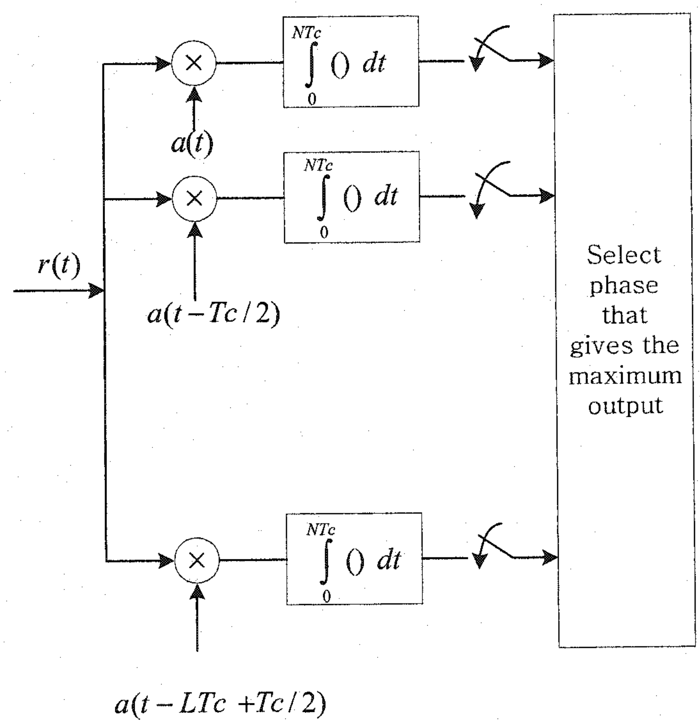

Figure 2.2: Parallel Acquisition System

acquisition is not detected, the estimation process is repeated. Sequential estimation methods provide fast acquisition in moderate to high signal-to-noise ratio conditions, however, for low signal-to- noise ratios the estimation process is unreliable and the acquisition time can be very high. Also this scheme is extremely vulnerable to interfering signals since it will try to acquire them. It is therefore not useful for multiple-access applications where other users share the same channel at the same time.

2.2 Classification of Code Tracking techniques

For tracking of PN signals a delay-locked loop (DLL) is usually employed. Two identical codes with a relative delay offset of one chip or less are independently correlated with the incoming signal. The difference of the correlators outputs is used as a control signal to adjust the clock rate of the local codes to keep the error signal within the pull-in range of the DLL. Another tracking loop is known as the tau-dither loop (TDL) where a single correlator is used. A low frequency square wave is used to dither the timing of the local code generator which results in amplitude modulation of the correlator output. The amplitude shift is used as an error signal to drive the clock generator of the local PN code.

2.3 Related Research on Synchronization of SS Signals

One of the early work on code acquisition using a sliding correlator was conducted by Sage [15]. Both coherent and noncoherent serial search acquisition systems were studied in AWGN channels. Expressions for the false alarm and detection probabilities were derived, however no analysis for the acquisition time was presented.

Holmes and Chen in [18] analyzed a single dwell serial search acquisition system with noncoherent detection. The time uncertainty region was divided into cells which were searched serially and the step size for the search update was assumed to be a fraction of a chip. A flow graph technique was used to model the acquisition process as a Markov chain which was simplified to obtain an acquisition time generating function. The mean and variance of the acquisition time were analyzed as a function of the false alarm and detection probabilities. Doppler effects were also considered in an approximate analysis. A double dwell time system, in which a short time correlation was initially used to increase the cell search rate followed by a longer correlation time to verify the more probable cells, was introduced to reduce the acquisition time. Their technique could be altered to suit other systems however, a considerable amount of algebraic manipulations will be needed.

In [19], Hopkins considered the complete problem of code synchronization. He used Markov chain theory to derive the transition matrix that defines the acquisition and tracking processes. Frequency offset errors due to Doppler were found to either increase or decrease the search rate depending on the sign of the Doppler offset and the direction of local code update (advancing or retarding phase). Also, it was shown that the mean acquisition time and the mean time to lose lock could be improved by using different dwell times and different thresholds for the acquisition and tracking modes, respectively.

Another measure of performance for the single dwell serial search acquisition system was introduced by DiCarlo and Weber [20]. The probability of successful synchronization was derived as a function of the detection probability, false alarm probability, detector integration time, and false alarm penalty time. The acquisition time generating function was arranged in such a way to permit identification of the probability of successful acquisition terms. The false alarm penalty time was found to have a strong effect on performance optimization.

In [21] DiCarlo and Weber considered a multiple dwell serial search system. Variable examination intervals were used to allow fast dismissal of incorrect cells. This was accomplished by testing a cell for a number of dwells of an increasing period provided that all the previous tests have indicated synchronization. The mean and variance of the acquisition time were derived. It was shown that significant improvement over a single dwell system is gained by using a double dwell system. A diminishing gain is noticed as the number of dwells were increased.

Polydoros and Weber [22] developed a simple yet generalized analysis for fixed dwell serial search code acquisition systems. Their scheme provides a simpler way of analysis than that introduced in [18]. A flow graph representation of the acquisition process based on symbolic notation was presented to obtain the generating function. Thus, by proper definition of the flow graph branches according to the system in hand, it is possible to use the developed expressions, with little or no modifications, to obtain different performance measures of the acquisition system. Polydoros and Weber then applied their analysis to a noncoherent matched filter code acquisition scheme [23]. A high rejection rate for the incorrect cells was obtained (a fraction of a chip interval), while a verification mode based on majority voting was used to minimize the false alarm rate. The mean acquisition time was found to depend on the matched filter length. It was also concluded that an optimum threshold value which minimizes the mean acquisition time exists for each value of signal-to- noise ratio.

Due to the relatively large acquisition time of serial search acquisition systems, Milstein et al. [24] proposed a parallel processing acquisition scheme that uses multiple surface acoustic wave (SAW) convolvers. The spreading sequence is divided into several subsequences which are used as references for the convolvers. The search rate is reduced by a factor proportional to the number of chips per subsequence times the number of convolvers used. It was shown that acquisition is possible within a period twice the number of chips in a subsequence. However, a large amount of hardware is required.

Su [25] discussed several parallel search acquisition algorithms that use matched filters. A parallel search with fixed dwell time detector, based on the maximum likelihood principle, is used. Another variation of the above algorithm that compares the decision statistic with a threshold prior to starting verification is also presented. Moreover, parallel algorithms with Walds sequential detection were analyzed.

Sourour and Gupta [16] [26] proposed an acquisition technique that uses a bank of noncoherent matched filters to improve the performance under fading channels. The code period was divided into disjoint subsequences and two filters were matched to each subsequence. First, under AWGN channel, the mean acquisition time was derived along with the false alarm and detection probabilities. Next, the performance over non-selective, slow Rayleigh fading channels was obtained by averaging the AWGN channel results over the fading distribution. Finally, for fast fading in which the channel changes within the length of a matched filter period, the performance measures were re-evaluated. It is interesting to note that, under non-fading and slowly fading conditions with small signal-to- noise ratio, it was found that it is better to increase the filter lengths and decrease the number of parallel matched filters. In contrast, under fast fading, more parallelism is better. Also, it was noted that in general fading degrades code acquisition systems performance significantly (about 9 to 10 dB in SNR).

As a result of the increasing interest in using CDMA systems for commercial applications, several papers that address the acquisition problem in the presence of multiple access interference were presented. Rick and Milstein [27] studied the reverse link acquisition performance for a DS-CDMA system. A noncoherent parallel acquisition scheme was investigated for a chip synchronous CDMA system with random sequences and perfect power control conditions. Also, no Doppler, fading, or data was assumed to be present in the received signal. The false alarm probability was used to assess the performance as a function of the observation interval and the number of active users in the system. It was concluded that the acquisition system performance may be the limiting factor for the CDMA system capacity rather than the bit-error rate (BER) performance. This is similar to the conclusion obtained by Madhow and Pursley [28] using asymptotic analysis for a matched filter system.

Recently, some parallel acquisition schemes were suggested ([29]- [30]). In [29], a differentially coherent system was shown to yield 4 to 5 dB improvement over noncoherent detection. Optimal and suboptimal estimators based on sequential probability ratio test (SPRT) were derived in [31] and [32]. In [30], a partial code correlation and majority decision rule scheme was proposed to operate under fast fading channels.

Between the two extremes, serial and parallel schemes, other methods based on hybrid strategies were studied by Baum and Veeravalli [33]. These systems reduce the hardware complexity incurred by parallel systems while keeping the acquisition time performance at an acceptable level. Several cells were examined for acquisition simultaneously and SPRT was used to increase the rejection rate of the incorrect intervals. The mean acquisition time was derived as a function of the false alarm and detection probabilities which were evaluated by computer simulation.

Rapid acquisition by sequential estimation (RASE) was originally developed by Ward [17]. It is based on estimating, from the received signal, the initial loading of the shift register used to generate the local PN code. The system then correlates the two codes for a period of time to check if the codes are synchronized or not. If not, a new estimate is made and the process is repeated. This technique provides fast acquisition for moderate input signal-to-noise ratio, however, the performance degrades significantly under interfering conditions. Thus, it is not suitable for multiple access applications.

Several variations of the RASE scheme were introduced [34] [35]. These schemes try to improve the initial phase estimation by using error correcting logic which results in a significant increase in hardware. In [36], a recursive-aided RASE (RARASE) technique was proposed that discards initial phase estimates. This scheme appear to suffer from errors without attempting to verify those errors. This improves the system performance without increasing the hardware required. However, the basic limitations of sequential estimation methods are still the same, i.e., susceptibility to interference.

Based on the previous survey of SS acquisition techniques, it is observed that most of the schemes, serial or parallel, rely either on correlation principles or sequential estimation.

Recently, some work on the hybrid acquisition system was carried out by Tarhuni-Sheikh using adaptive filters to search for the correct phase of a PN sequence[37] (we will consider it in detail in the next section). Even though the acquisition time experienced by this new type of acquisition circuit is slightly better than those of the more traditional types, adaptive filters have major advantage of being able to act as the tracking circuit too[38]. Furthermore, it was shown that the system performs well in fading and frequency selective environments [39].

The capacity of CDMA systems is essentially limited by the performance of the acquisition subsystem; i.e; the maximum number of users that can achieve signal acquisition with a certain degree of reliability. This capacity can be improved significantly by using acquisition strategies that minimize the effect of Multiple Access Interference (MAI) in Multiuser environment, such as adaptive filter structure proposed in [40] and [41] (based on the same idea proposed in [37]) and it has been shown that the adaptive filter acquisition system can support the maximum number of users as that obtained by the BER criterion while matched filter based acquisition schemes can limit the capacity significantly.

2.4 Application of Adaptive Filters in SS Systems

An FIR adaptive filter is basically a filter with a time varying impulse response [42]. The time variation is obtained by adjusting the filter taps in a certain way. Adaptive filters have been widely used in different areas due to their ability to adjust to their operation environment.

In SS systems, adaptive filters were used to suppress narrow-band interference in DS/SS techniques. An adaptive filter is used to estimate the narrow-band interference [43] [44] [45], which is highly correlated, from the received signal. The interference estimate is subtracted from the received signal which is then applied to the correlator for data detection. The same idea could be interpreted as using a filter with zeros at the interfering signals frequencies. These systems improve the performance if the processing gain is small to provide sufficient protection against the strong narrow-band interference.

A Near-Far multi-user receiver for DS-CDMA systems which uses an FIR adaptive filter has been suggested in [46]. It has been shown that, with or without power control, the adaptive filter receiver provides larger system capacity than conventional matched filter receivers. Furthermore, the adaptive receiver does not need any information regarding the spreading sequences of other users and their received powers. A similar system was also described in [47].

In [48], an FIR adaptive filter is used for code acquisition. The adaptive filter has a number of taps equals to the code length and the input to the filter is sampled at the chip rate while the output is sampled at the bit rate. A maximum likelihood decision about the delay is obtained from the tap-weight vector after convergence. The system has good acquisition performance; however, under severe interference conditions, the adaptation time required by the system before making a reliable decision about the delay estimate is very long which may increase the acquisition time performance. Also, the system requires that users access the system one at a time.

The structure proposed by Tarhuni-Sheikh (whose detail is in Sec 2.5) is different from that suggested in [48] in several aspects. Firstly, a smaller filter length is used in this study as opposed to the filter in [48] which has a number of taps equals to the code length. Secondly, the decision strategies used to detect acquisition and to extract timing information are different. Finally, the proposed scheme uses the same filter for code tracking as well while the system in [48] is used for code acquisition only.

2.4.1 Adaptive Filter Based Code Synchronization Scheme.

In this thesis, we will investigate SS code synchronization using an adaptive filter in a transversal configuration. The new adaptive filter based scheme has an improved acquisition time performance over serial search techniques; it also has less hardware complexity than other hybrid systems. Moreover, the same filter can also be used for code tracking (not under focus in this thesis) and as a result the SS system will have a simpler structure than other conventional SS systems which use a separate code tracking loop. A simplified block diagram for the adaptive filter based synchronization system is shown in Fig 2.3 (first proposed in [37]). The adaptive filter

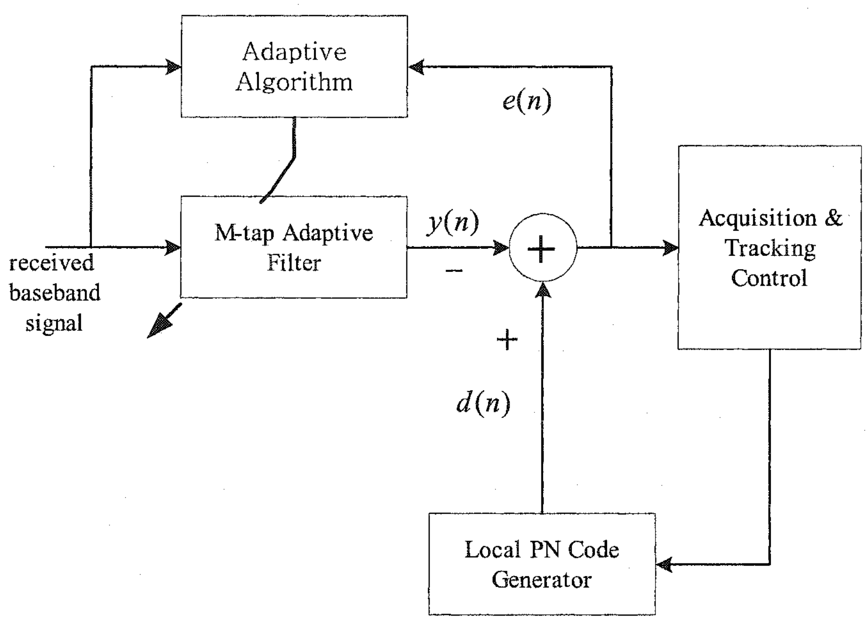

Figure 2.3: A Simplified Block Diagram of the Adaptive Filter based Code Synchronization System

has M taps which are adjusted to minimize the mean-square error (MSE) between the received SS signal and a locally generated replica of the spreading PN code. Initially, the adaptive filter operates in acquisition mode in which the time uncertainty region is divided into intervals each of M chips duration. For each interval, the filter coefficients are adapted toward their optimum value (in the MSE sense) and, after an adaptation period, convergence in the MSE is checked. If convergence is detected, the delay offset is estimated from the tap weights, otherwise the next interval is tested and the adaptation process is repeated.

Once acquisition is declared, the system switches to tracking mode in which the received signal is interpolated to increase the time resolution of the adaptive filter. The local code phase is continuously adjusted to maintain the peak of the filter coefficients at the center tap while the MSE is maintained below a preset threshold value. If tracking has been initiated due to a false alarm error, the tracking algorithm will sense the increase in the MSE and restart the acquisition process.

2.5 Tarhuni-Sheikh (T-S) adaptive filter based Code Acquisition Algorithm

In this section, we will give a brief overview of Adaptive Filter based hybrid code acquisition scheme proposed in [37] .The adaptive acquisition system block diagram is shown in Figure 2.4. The system consists of a filter with M adaptive taps, an adaptation algorithm to adjust the filter coefficients iteratively, a pseudo-noise (PN) code generator, and other blocks to check for acquisition. The received **SS** signal, $r(t)$ is

$$r(t) = \sqrt{2P}i(t-\tau)a(t-\tau)cos(2\pi f_c t + \varphi) + n_p(t) \tag{2.1}$$

where P is the power of the received signal, $i(t)$ is the information data, $a(t)$ is the PN spreading signal, f_c and φ are the frequency and random phase of the carrier signal, respectively, and τ is the random time delay that must be estimated by the synchronizer. It is assumed that τ is uniformly distributed over the spreading code length. The receivers thermal noise, $n_p(t)$, is modelled as an additive white Gaussian noise (**AWGN**). The information data is

$$i(t) = \sum_i i_j \psi_{T_b}(t - jT_b) \tag{2.2}$$

where $\{i_k\}$ is a sequence of ± 1 with equal probability, and ψ_{T_b} is a shaping rectangular pulse defined as

$$\psi_{T_b} = \begin{cases} 1 & ; 0 \leq t \leq T_b \\ 0 & ; elsewhere \end{cases} \tag{2.3}$$

where T_b is the symbol duration. The spreading waveform, $a(t)$, is given by

$$a(t) = \sum_i a(i)\psi_{T_c}(t - iT_c) \tag{2.4}$$

where $\{a\}$ is spreading sequence such that

$$\{a\} = \{...., a(-2), a(-1), a(0), a(1), a(2), ...\} \tag{2.5}$$

with $a(i) \in \{-1, 1\}$, T_c is the chip duration, and the sequence is assumed to be periodic with period L, i.e; $a(i \pm L) = a(i)$, for all i where L is the length of spreading sequence. Also, it is assumed that the system's processing gain is equal to the code length, i.e; $T_b/T_c = L$. It has been assumed that the receiver is perfectly synchronized with the RF carrier phase.

Let us now assume, without loss of generality, that the received signal power is normalized to unity. The equivalent baseband signal, $x(t)$ is then given by

$$x(t) = b(t - \tau)a(t - \tau) + n(t) \tag{2.6}$$

where $n(t)$ is a zero mean **AWGN** with two sided power spectral density $N_0/2$. The baseband signal is passed through a Chip Matched Filter (**CMF**), whose output is sampled at a rate of f_s such that

$$f_s = 1/T_c \tag{2.7}$$

The samples of the **CMF** output, $x(n)$ are applied to the FIR filter whose output at the nth time instant is given by

$$y(n) = \mathbf{w}^T(n)\mathbf{x}(n) \tag{2.8}$$

where $\mathbf{w}$(n) is the tap-weight vector of the filter coefficients at the nth time instant represented as

$$\mathbf{w}(n) = [w_0(n) \;\; w_1(n)....w_{M-1}(n)]^T \tag{2.9}$$

with the subscripts $0, 1, ..., M-1$ denote the tap number. Also, $\mathbf{x}(n)$ is the filter input vector at the nth instant given by

$$\mathbf{x}(n) = [x(n) \;\; x(n-1)....x(n-M+1)]^T \tag{2.10}$$

which consists of the present sample and past $M-1$ samples of the received signal. The superscript T denotes vector transposition operation.

At every sampling instant, a vector of the received signal samples is used to produce an output, $y(n)$, that represents an estimate of the desired signal, $d(n)$. The desired signal is chosen as the locally generated despreading sequence at a certain phase, i.e., $d(n) = a(n - \hat{\tau})$, where $\hat{\tau}$ can assume any integer value between 0 and $(L-1)T_c$. Initially, the phase of the local sequence is set at the beginning of the time uncertainty region, i.e., $\hat{\tau} = 0$. The error signal, $e(n)$, is defined as the difference

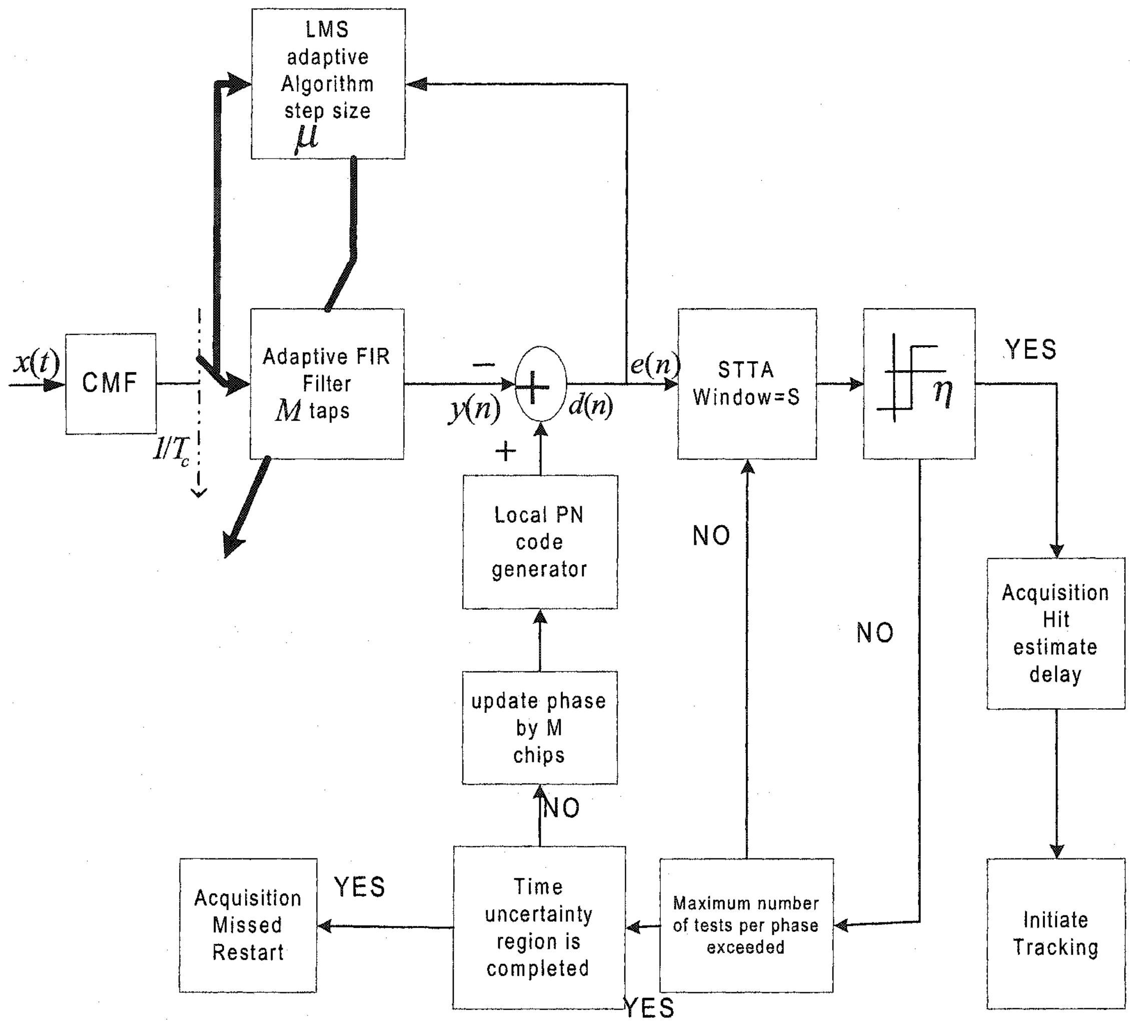

Figure 2.4: Block Diagram of the Tarhuni-Sheikh Code Acquisition System

between the filter output and the desired signal,

$$e(n) = d(n) - y(n) \tag{2.11}$$

As time progresses, the filter coefficients are adjusted according to a certain algorithm and the filter output is produced using the updated tap weights. This process continues for a period of time which has been assumed to be long enough for the coefficients to converge near their optimum value. After the end of this adaptation period, a decision period is started in which the error signal is checked for convergence towards the minimum mean-squared error (**MMSE**) obtained by this filter with coefficient vector set to the optimum Wiener value.

If convergence is detected, i.e., the MSE is less than a preset threshold η ,acquisition is declared, and the time delay between the received SS signal and the local despreading sequence will be within the time span of the adaptive FIR filter during the acquisition phase of M chips. The exact value of the delay is estimated from the weight distribution of the filter coefficients, $w(n)$, which is peaked at the tap with the correct delay estimate. On the other hand, if no convergence in the MSE is detected, the phase of the locally generated sequence, $d(n)$, is adjusted by M chips, the tap weights assume their initial values, and the test for convergence is repeated.

The time uncertainty region is divided into q disjoint intervals or cells, such that

$$q = \left\lceil \frac{L}{M} \right\rceil \tag{2.12}$$

where $\lceil z \rceil$ is the smallest integer greater than or equal to z. There are, at least, $q-1$ intervals each of width M chips, while the last interval has a width of M chips or less. The system tests the cells for acquisition one at a time. The time required to test a cell equals to the adaptation time plus the decision time. Convergence in the MSE is then used to decide either to accept the cell currently under test as the acquisition cell or to advance the local code to test the next cell.

The FIR filter coefficients are adapted toward their optimum value, in the **MMSE** sense, during every test for acquisition. The LMS adaptation algorithm is used to update the coefficients. The algorithm adaptation step size, μ, is chosen to guarantee both convergence in the MSE and convergence in the filter coefficients to the optimum wiener value.

2.6 Problem Definition

The Acquisition time experienced by the system proposed by Tarhuni and Sheikh is only slightly better than the other traditional types. Particularly, acquisition time performance of the system is not good for the low signal-to-noise-ratio (SNR) conditions because the decision parameters for phase acquisition are adjusted according to the estimated signal-noise-power. So for low SNR conditions, the performance of the system will be badly affected as the average squared error will not be less than the preset threshold after convergence period and false alarm errors will be very significant for that system. The performance of the system will be even worse for the longer spreading sequences in low SNR environment, as the system may not detect the phase even when it is searching the correct cell. It then needs to step through all the cells again which will considerably increase the acquisition time.

We propose in this thesis a modified adaptive filter (AF) based Code Acquisition scheme to acquire the correct phase delay of spreading sequence speedily, without having to do a sequential search. Our system differs from the T-S algorithm in two main aspects. Firstly the code acquisition in the modified system is done in two stages. In the first stage called the cell identification stage, the system first locate the neighbourhood i.e; the cell (in our system, cell concept is different from T-S algorithm and it is a collection of chips rather than a single chip) where the chip phase delay can be found. Then the system advances to the phase acquisition stage to locate the chip phase delay itself i.e; to locate the chip, called the phase identification stage. Secondly, in our system, the decision criteria is different from the T-S algorithm i.e; a record of maximum tap weights will indicate the location of the cell (in cell identification stage) and the location of the phase (in phase identification stage) where phase delay can be found. The advantage of this decision strategy is that the system is able to detect the tap weight corresponding to the correct cell or chip as maximum tap weight even for low SNR conditions.

It has been shown that the Acquisition time performance of the modified structure is far better than T-S structure for low SNR environment. The modified structure also falls in the category of Hybrid acquisition schemes.

In the next chapter, we will explain the proposed modified structure for code acquisition problem in detail.

Chapter 3

Proposed Adaptive Filter Based DS/SS Code Acquisition System in AWGN Environment

3.1 Introduction

In this chapter, I will describe in detail our proposed modified adaptive filter based code acquisition system structure and the performance of the system will be presented in AWGN environment.

3.2 Acquisition System Description

The proposed code acquisition system in Fig 3.1 is based on an Adaptive FIR filter with M coefficients (length of adaptive filter), or tap-weights, which are adjusted using the Least-Mean-Square (LMS) Algorithm to minimize the mean-squared-error (MSE) between the filter output and the desired reference signal. The baseband signal at the receiver consists of the received PN maximal length spreading code sequence $a(i) = \pm 1$ (generated by the shift register) at a certain delay and additive white gaussian noise (AWGN). Suppose no information data is transmitted. The delay is assumed to be over the code length. This baseband signal is sampled at chip rate to form the samples. With no loss of generality, we shall take the local spreading sequence at zero phase, and sampled at chip rate T_c to form the samples $a(n)$.

The first task of the system is to find the neighbourhood in which the correct phase delay is located, before it searches the phase itself. In the cell identification stage, the desired sequence into the adaptive filter is formed by adding together a fixed number c, called a cell, of these samples. In our system, the cell concept is different as explained earlier for other scheme. Here cell is a taken as a collection (sum) of chips rather than a single chip. The number c is chosen so that $cM = L$, where M is the adaptive filter length and L is the length of the spreading sequence. The same number c of the received samples are also added to form the input to the

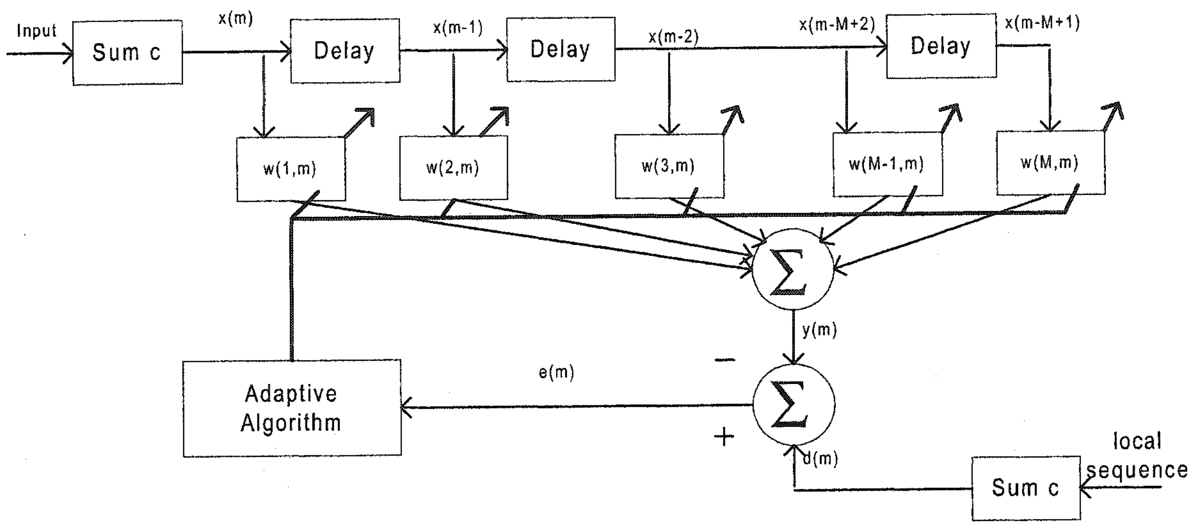

Figure 3.1: Schematic of an Adaptive Filter for the proposed Code Acquisition System

adaptive filter. Note that there will be a delay in multiples of T_c involved between the received sequence relative to the local spreading sequence as received sequence can start from anywhere within the code length. The adaptive filter input vector consists of the present and past $M-1$ samples (a total of M samples) of the sums of the received chips. The adaptive filter is clocked at cT_c to produce the output of the filter and error signal is computed as the difference between the desired sequence and the output of the filter. The LMS Algorithm (brief detail of LMS Algorithm is given in Appendix A) is used to search for minimum of the mean-squared error by updating the filter taps. Following convergence of the tap weights, the maximum of the tap weights is identified. As will be seen later, a record of the maximum tap weight will indicate the location of the cell where the chip phase delay can be found i.e; the location where the maximum of the chip overlap will occur.

To see how the above system works, suppose we have a system of $M = 3$ taps, and a spreading sequence of $L = 12$ chips is used. The cell size is therefore $c = 4$. Suppose the system starts with the local sequence at chip1, and the received sequence at chip 8. After 3 clock ticks of the adaptive filters, i.e., $12T_c$, the 3 taps will contain the sums

Tap 1	Tap 2	Tap3
$a_7 + a_6 + a_5 + a_4$	$a_3 + a_2 + a_1 + a_{12}$	$a_{11} + a_{10} + a_9 + a_8$

The summer of the local sequence will produce the value $a_{12} + a_{11} + a_{10} + a_9$. This sum has 3 chips which match those in Tap 3, and 1 that match in Tap 2. It can be easily seen that at the next clock tick of the adaptive filter, i.e., at $16T_c$, the summer of the local sequence will still produce a value which has three chips matching in Tap 3 and 1 which match that in Tap 2. Thus the tap weight corresponding to the Tap 3 will show maximum peak.

Once one of the cells has been identified as having the correct phase, the system will go into the phase acquisition mode i.e; to acquire the phase delay itself. It does this by changing the clock-rate of the adaptive filter to T_c, and loading the subsequence that starts $\lceil \frac{M}{2} \rceil$ from the beginning of the identified cell. So suppose if there are 11 Taps, and each cell has 8 chips, with the second cell being the identified cell, then the subsequence to be loaded is $a_3, a_4, \ldots a_{13}$.

A fixed number of samples are taken again for the phase acquisition stage, and the tap where the majority of the peaks occur will indicate the position of the delay phase.

3.3 Analysis of the Acquisition System

The received spread signal is given by

$$r(t) = \sqrt{2P}i(t - T_c)a(t - \tau)cos(2\pi f_c t + \phi) + n(t) \tag{3.1}$$

where P is the received power, $i(t)$ is the information data, T_c is the chip duration, $a(t)$ is the PN maximal length spreading code sequence, τ is the random time delay between the received signal and the local time reference that must be estimated by the synchronizer, f_c and ϕ are the carrier frequency and phase respectively, and $n(t)$ is Additive White Gaussian Noise. It is assumed that time delay τ is uniformly distributed over the spreading code length.

The spreading signal is given by

$$a(t) = \sum a(i) \prod_{Tc}(t-T_c) \tag{3.2}$$

where $a(i) = \pm 1$ generated by the shift register for a period of 127 with a linear shift register of length 7 having preferred pair of [7,3,2,1]which means that preferred polynomial is $1 + X + X^2 + X^3 + X^7$, and $\Pi_{T_c}(t)$ is a rectangular shaping pulse with unit amplitude and duration T_c.

To simplify the analysis, we shall make some assumptions:

1. We assume that no information data is transmitted during acquisition. This situation occurs when the system allows an acquisition period prior to commencing data transmission or if a separate pilot signal is available for acquisition.

2. The received power is normalized to unity.

3. The receiver has already been perfectly synchronized with RF carrier frequency and phase.

4. The spreading sequence is independent of the samples of AWGN, that is $E\{a(i)n(j)\} = 0 \quad ; all \;\; i, j.$

5. It is also assumed that the system is chip synchronous, i.e; τ is an integer value over the code period LT_c i.e; $\tau \in [0, 1, 2, ..., L-1]$. This assumption has been made to simplify the analysis.

The third assumption that the receiver is perfectly synchronized with RF carrier frequency and phase has been widely used in the literature, as in [27] [32] [33] [31]. Even though from a practical point of view, this might be a valid assumption to make, it simplifies the analysis-a reason why it is widely used in literature. The subject of adaptive equalization for non-coherent and differentially coherent detection did not receive much attention in the literature because of the analytical complexity introduced by the nonlinear detector. This nonlinearity preludes the commonly used adaptive algorithms to adapt the filter taps such as the simple LMS algorithm. Recently, equalization techniques for non-coherent receivers have been proposed in [49]. It has been shown that, if the receiver neglects the carrier phase completely, the equalizer coefficients will adjust themselves to compensate for the carrier phase and carrier phase does not have to be recovered provided that the phase variations are slow enough to allow the **LMS** algorithm to track channel variations. It has also been proposed to use a combination of a linear equalizer with a modified LMS algorithm and a differentially coherent receiver to cancel the the phase error. In short, although a coherent system is difficult to implement in practice, it enables a tractable analysis for the proposed system. Also, further research to develop non-coherent or differentially coherent adaptive system is needed.

With these assumptions, the received baseband signal at the receiver can be represented by just

$$x(t) = a(t - \tau) + n(t) \tag{3.3}$$

This baseband signal is sampled at the chip rate to form the samples $x(n)$. The local spreading sequence, sampled at T_c forms the samples $a(n)$. Since the desired sequence into the adaptive filter is formed by adding c chips called cell, of local spreading sequence samples, so

$$d(m) = \sum_{i=1}^{c} a(mc+i) \tag{3.4}$$

where $cM = L$, the length of the spreading sequence. Note that c can also be chosen so that $cM \approx L$ if $cM \neq L$. Similarly the same number c of received samples are also added to form input to adaptive filter, so

$$x(m) = \sum_{i=1}^{c} a(mc+k+i) + n(mc+k+i) \tag{3.5}$$

where k is the random delay of the received spreading sequence in T_c relative to the local spreading sequence.

This adaptive filter is clocked at every cT_c , so that the output of the filter is given by the convolution sum

$$y(m) = \sum_{j=0}^{M-1} w_j(m) x_k(m-j)$$

$$y(m) = \mathbf{w}^T(m)\mathbf{x_k}(m) \tag{3.6}$$

where $\mathbf{w}(m) = [w_0(m) \quad w_1(m) \quad \cdots \quad w_{M-1}(m)]^T$ is the tap weight vector at mcT_c, and $\mathbf{x_k}(m) = [x_k(m) \quad x_k(m-1) \quad \cdots \quad x_k(m-M+1)]^T$ is the filter input vector. The superscript T denotes Transposition operation.

The techniques that are adopted for code acquisition can use **LMS** (brief detail of LMS is in Appendix A) or **RLS** algorithm to adapt filter taps. The advantage of LMS over RLS is its simplicity of implementation and suitability for this problem. In general, the LMS algorithm exhibits more robust acquisition behaviour than the standard RLS algorithm. Because of these features, LMS algorithm is usually selected for the acquisition systems.

So the LMS Algorithm is used here also to search for the minimum mean-squared error by updating the filter taps according to [38]

$$\mathbf{w}(m+1) = \mathbf{w}(m) + \mu e(m)\mathbf{x}(m) \tag{3.7}$$

where

$$e(m) = d(m) - y(m) \tag{3.8}$$

is the error signal between the desired sequence and the filter output, and μ is the step size.

3.3.1 Cell Identification Analysis

According to [42], the optimum tap weight vector of the FIR filter is given by the Wiener Hopf equation

$$W_{opt} = \mathbf{R}^{-1}\mathbf{p} \tag{3.9}$$

where $\mathbf{R}$ is the autocorrelation matrix of the $\mathbf{x}$(m) and $\mathbf{p}$ is the cross correlation vector of $\mathbf{x}$(m) and $d(m)$. Now

$$\mathbf{R} = E\{\mathbf{x}(m)\mathbf{x}^{\mathrm{T}}(m)\} \tag{3.10}$$

which is $M \times M$ matrix with elements

$$r_{pq} = E\{x(m-p)x(m-q)\} \tag{3.11}$$

where p and $q = \{0, 1, ..., M-1\}$

Expanding out $x(m-p)x(m-q)$

$$\sum_i [a((m-p)c+k+i)+n((m-p)c+k+i)] \sum_j [a((m-q)c+k+j)+n((m-q)c+k+j)] \tag{3.12}$$

Assuming that the chips in the spreading sequence are uncorrelated, and also independent of the noise, we get $r_{pp} = c(1+\sigma^2)$, and $r_{pq} = 0$ if $p \neq q$, where σ^2 is the variance of the white noise. Therefore

$$\mathbf{R} = c(1+\sigma^2)\mathbf{I} \tag{3.13}$$

where $\mathbf{I}$ is the identity matrix.

The cross correlation vector between $\mathbf{x}$(m) and $d(m)$ is $\mathbf{p} = E\{\mathbf{x}(m)d(m)\}$ which is an $M \times 1$ vector with elements $p(q)$ where $q = \{0, 1, ..., M-1\}$. Expanding out

$$p(q) = E\{\sum_i [a((m-q)c+k+i) + n((m-q)c+k+i)] \sum_j a(mc+j)\} \quad (3.14)$$

Let us denote the larger number of matching chips between $\mathbf{x}(m)$ and $d(m)$ as v. Since the chips from only two cells of $\mathbf{x}(m)$ can match with those in $d(m)$, the other match involve $(c-v)$ chips.

Thus we have using the same assumptions above,

$$p(q) = \begin{cases} 0 & if \ \ a(n)'s \ \text{ do not match at all} \\ v & if \ \ a(n)'s \ \text{ match in v places with superior match} \\ (c-v) & if \ \ a(n)'s \ \text{ match in (c-v) places with inferior match} \end{cases} \quad (3.15)$$

It is shown in [42] that if $0 < \mu < 2/\lambda_{max}$ then the tap weights will converge to the optimum tap weight vector given by Eq 3.9. Using Eqs 3.13 and 3.15 in 3.9, we find that after convergence, the expected tap weights at location q is

$$E\{w(q)\} = \begin{cases} 0 & if \ \ a(n)'s \ \ do \ \ not \ \ match \ \ at \ \ all; \\ v/c(1+\sigma^2) & for \ \ cell \ \ with \ \ superior \ \ match \\ (c-v)/c(1+\sigma^2) & for \ \ cell \ \ with \ \ inferior \ \ match \end{cases}$$

It is shown in [50] that if $\mu\lambda_i/2 << 1$, where $\lambda_i's$ are the eigenvalues of the autocorrelation matrix $\mathbf{R}$, then the adaptive filter tap weights are the gaussian random variables with covariance matrix

$$\mathbf{C} \approx \frac{\mu}{2} J_{min} \mathbf{I} \tag{3.16}$$

where J_{min} is the minimum mean squared error (MMSE) given by [38]

$$J_{min} = \sigma_d^2 - \mathbf{p}^T \mathbf{R}^{-1} \mathbf{p} \tag{3.17}$$

with $\sigma_d^2 = c$, the variance of the desired sequence. In our case, the MMSE can be worked out to be

$$J_{min} = \frac{c^2 - 2v^2 SNR_c + 2cv SNR_c}{c(1 + SNR_c)} \tag{3.18}$$

where $SNR_c = 1/\sigma^2$, the signal to noise ratio chip.

It can be seen from Eq 3.16 that the tap weights are uncorrelated gaussian random variables with variance equal to $\frac{\mu}{2} J_{min}$. The mean of the tap weights after convergence is, from Eq 3.16, nonzero at only 2 locations. It is maximum at the cell location which has the largest match between the incoming sequence and the local sequence, and has a smaller value at an adjacent location where there is also an inferior match of chips. By examining the cell indicated by the maximum tap weight, the correct phase can be found. We find the maximum tap weight has a mean of

$$m_\alpha = w_{\alpha,opt} = \frac{SNR_c}{c(1 + SNR_c)} v \tag{3.19}$$

and next to it is a tap which has a mean weight of

$$m_\beta = w_{\beta,opt} = \frac{SNR_c}{c(1+SNR_c)}(c-v) \tag{3.20}$$

since $\frac{1}{1+\sigma^2} = \frac{SNR_c}{c(1+SNR_c)}$. All other tap weights have zero means. The variance of all tap weights is $\sigma_w^2 = \mu J_{min}/2$.

The probability that all the other tap weights are less than the one located at the correct position is

$$P_c = \int_{-\infty}^{\infty} p(w_0 < w_\alpha, ...w_\beta < w_\alpha, ..., w_{M-1} < w_\alpha | w_\alpha) f(w_\alpha) dw_\alpha \tag{3.21}$$

where w_α is the AF coefficient that corresponds to the cell that has the correct delay offset. Here exact joint PDF was difficult to drive, however a very good Gaussian approximation has been reported in [50], which we also applied in our case to write the expressions.

Since all zero mean $w_j's$ are independent Gaussian random variables.

$$p(w_j < w_\alpha | w_\alpha) = \int_{-\infty}^{w_\alpha} \frac{1}{\sqrt{2\pi}\sigma_w} e^{-\frac{w_j^2}{2\sigma_w^2}} dw_j = 1 - Q(\frac{w_\alpha}{\sigma_w}) \tag{3.22}$$

where $Q(t) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt$.

And $p(w_\beta < w_\alpha | w_\alpha) = 1 - Q(\frac{w_\alpha - m_\beta}{\sigma_w})$.

where w_β is the AF coefficient that corresponds to the cell that has the inferior match (adjacent to the tap having correct delay offset). Thus

$$P_c = \int_{-\infty}^{\infty} [1 - Q(\frac{w_\alpha}{\sigma_w})]^{M-2} [1 - Q(\frac{w_\alpha - m_\beta}{\sigma_w})] \frac{1}{\sqrt{2\pi}\sigma_w} e^{-\frac{(w_\alpha - m_\alpha)^2}{2\sigma_w^2}} dw_\alpha \qquad (3.23)$$

This probability may be small if we look at only one tap weight vector. To get a better estimate of the location of the correct cell, it is necessary to do a number of trials.

Before we do that, we need to work out the probability that the peak occurs at the wrong taps. The probability that the peak occurs at the tap which has the inferior match, adjacent to the correct one, is

$$P_{fa} = \int_{-\infty}^{\infty} p(w_0 < w_\beta, ...w_\alpha < w_\beta, ..., w_{M-1} < w_\beta | w_\beta) f(w_\beta) dw_\beta \qquad (3.24)$$

Using the same arguments above,

$$P_{fa} = \int_{-\infty}^{\infty} [1 - Q(\frac{w_\beta}{\sigma_w})]^{M-2} [1 - Q(\frac{w_\beta - m_\alpha}{\sigma_w})] \frac{1}{\sqrt{2\pi}\sigma_w} e^{-\frac{(w_\beta - m_\beta)^2}{2\sigma_w^2}} dw_\beta \qquad (3.25)$$

where w_β is the AF coefficient that corresponds to the inferior match of chips. And the probability that a peak occurs at any other of the wrong taps, assuming that all other taps are equally probable, is

$$P_f = \frac{1}{M-2}(1 - P_c - P_{fa}) \qquad (3.26)$$

If we take only one sample of the tap weights, and identify the cell having the correct phase as the one where the peak is located, we get a very low probability of correct identification when the signal-to-noise ratio is low. Instead, we shall take $N = 2n + 1$, an odd integer of samples, and consider the cell as having the correct phase only if the majority of the peaks occur at the tap corresponding to the position where that particular cell lies.

Since the decisions are independent, the probability that the cell is correctly identified is given by the cumulative binomial distribution as

$$P_{cc} = \sum_{i=1}^{n+1} \binom{N}{n+i} P_c^{n+i}(1 - P_c)^{N-n-i} \tag{3.27}$$

Note that the above is true only if the tap weights of the different samples are independent, which is not strictly true, especially when μ is small. However, if the samples are taken far enough apart, the tap weights are approximately independent.

The probability that the adjacent cell with the inferior match is wrongly identified as the one containing the correct phase is

$$P_{wca} = \sum_{i=1}^{n+1} \binom{N}{n+i} P_{fa}^{n+i}(1 - P_{fa})^{N-n-i} \tag{3.28}$$

Similarly the probability that any of the other cells is wrongly is identified

$$P_{wco} = \sum_{i=1}^{n+1} \binom{N}{n+i} P_f^{n+i} (1 - P_f)^{N-n-i} \tag{3.29}$$

Hence the probability that a wrong cell identification is made is

$$P_{wc} = P_{wca} + P_{wco} \tag{3.30}$$

Finally, the probability that no other cell can be identified as having the correct phase, and a new set of samples have to be taken, is

$$P_{nc} = 1 - P_{cc} - P_{wc} \tag{3.31}$$

3.3.2 Phase Acquisition

Once one of the cells has been identified as having the correct phase, the system will go into the phase acquisition mode. It does this by loading the subsequence that starts $\lceil \frac{M}{2} \rceil$ from the beginning of the identified cell.

According to [37], the peak appearing at the tap which corresponds to the correct phase delay has a mean of

$$w_p = \frac{SNR_c}{SNR_c + 1} \tag{3.32}$$

All the other taps have a mean of zero.

The variance of the taps is

$$\sigma_{wc}^2 \cong \begin{cases} \frac{\mu}{2}\frac{1}{SNR_c+1} & if \;\; phase \;\; lies \;\; in \;\; cell \\ \frac{\mu}{2} & otherwise \end{cases}$$

Should the cell be identified correctly, the probability that the maximum peak is located at the right tap is

$$P_{rc} = \int_{-\infty}^{\infty} [1 - Q(\frac{w_\alpha}{\sigma_{wc}})]^{M-1} \frac{1}{\sqrt{2\pi}\sigma_{wc}} e^{-\frac{(w_\alpha - w_p)^2}{2\sigma_{wc}^2}} dw_\alpha \tag{3.33}$$

The probability of the peak being located at a wrong tap in this case is

$$P_{wcl} = \frac{1}{M-1}(1 - P_{rc}) \tag{3.34}$$

If the cell has been identified wrongly,i.e., the correct phase does not lie within the cell being searched at all, then a peak will occur randomly among the taps, with probability

$$P_{wc2} = \frac{1}{M} \tag{3.35}$$

To cut down on the incidence of false alarms, which takes some time to be cleared, the system will again sample the peaks an odd number of times and make a decision only if more than half of the peaks are located at the same tap. Thus if the number of samples is taken to be $K = 2m + 1$, then the probability of acquiring the right chip phase delay, if the correct cell is identified, is

$$P_{ca} = \sum_{i=1}^{m} \binom{K}{m+i} P_{rc}^{m+i} (1 - P_{rc})^{K-m-i} \tag{3.36}$$

The probability of false alarm is

$$P_{fa1} = \sum_{i=1}^{m} \binom{K}{m+i} P_{wc1}^{m+i} (1 - P_{wc1})^{K-m-i} \tag{3.37}$$

The probability of no acquisition is

$$P_{na1} = 1 - P_{ca} - P_{fa1} \tag{3.38}$$

If the correct cell has not been located, then the probability of false alarm is

$$P_{fa2} = \sum_{i=1}^{m} \binom{K}{m+i} P_{wc2}^{m+i} (1 - P_{wc2})^{K-m-i} \tag{3.39}$$

And the probability of no acquisition is just

$$P_{na2} = 1 - P_{fa2} \tag{3.40}$$

3.4 Selection of LMS Step Size Parameter

The steady state mean-squared error produced by the LMS algorithm is determined by the step size parameter used to adjust the filter taps. A large step size results in a fast initial convergence and consequently a small transition period to reach steady state; however a larger steady state MSE is observed. The value of the adaptation step size must be carefully selected such that the limits of stability of the LMS algorithm, as described in Appendix A, are not violated. Provided that these limits are satisfied, the step size parameter has a large influence on the probability density function of the squared error.

Fig 3.2 shows the probability that the correct cell is identified as a function of the **LMS** step size μ for SNR_c of -5dB. The correct cell identification probability deteriorates as μ increases as shown in fig 3.2. However, for small values of μ, the correct cell identification probability can be made to approach to unity. The probability that the wrong cell is identified starts to increase with increasing μ because it is dominated by the false alarm error probability that the peak occurs at the tap which has the inferior match.

An important conclusion from these observations is that the step size parameter must be carefully selected to ensure good system performance. One possibility is to optimize the step size parameter such that the correct cell identification probability is maximized under the constraint that the false alarm probability does not exceed a

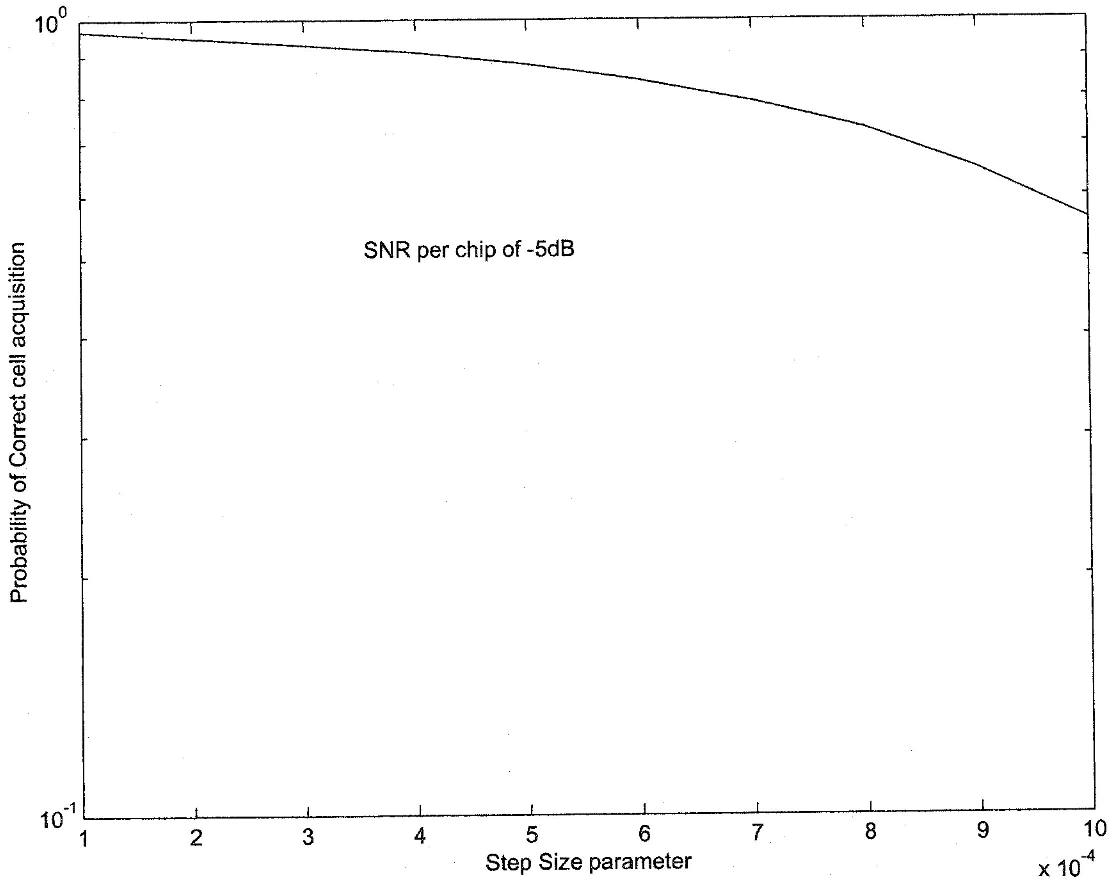

Figure 3.2: Probability of correct cell identification versus Step Size parameter

certain value. However, this criterion may not lead to the minimum mean acquisition time performance. For example, a small value of μ may be needed to meet design specifications which may increase the acquisition time due to the increased transition period required for the LMS algorithm to reach steady state.

3.5 System Performance Analysis

To reduce the inordinate amount of time the system spends in trying to identify the cell when the signal to noise ratio is low and v (superior match of chips) is small, we shall have the system advance the local sequence by $k = \lfloor \frac{c}{2} + 1 \rfloor$ chips every time it is unsuccessful in identifying the cell. This extra step does not affect the performance of the system at all when the signal to noise ratio is high, but when it is low, will shorten the cell identification process substantially.

The mean time to acquisition for the proposed system, can be obtained by modeling the acquisition process as a Markov chain process with one absorbing state.Markov chain process is a Markov random process [1] that has an initial state, a finite or countable set of possible states, and associated transition probabilities. The Markov chain is defined to be the set of states (outcomes) for a particular sequence of experiments. The state transition diagram for non-absorbing states (transient states) is shown in Fig 3.3

The average time in T_c that the system stays in the cell identification (CI) states (states 1 and 5), is K_{CI}. The time the system stays in both the Phase Acquisition (PA) states (states 2 and 6) and the wrong phase acquisition (WPA) states (states 3 and 7) is K_A chip times. The time the system stays in the False Alarm (FA)

[1]Markov Random Process is a random process where the occurrence of an event is dependent only upon the immediately preceding event and is said to be in one of the set of mutually exclusive states.

States (states 4 and 8) is the penalty time for identifying the delay phase wrongly. We shall take it as K_p chip times. The ACQ state is the absorbing state, and the system will eventually enter this state when the correct phase delay is required. Transition probability matrix [2] for Markov chain process of our system is

$$\mathbf{P} = \begin{pmatrix} 0 & P_{cc}(v) & P_{wc}(v) & 0 & P_{nc}(v) & 0 & 0 & 0 & | & P_{ca} \\ 0 & 0 & 0 & P_{fa1} & P_{na1} & 0 & 0 & 0 & | & 0 \\ 0 & 0 & 0 & P_{fa2} & P_{na2} & 0 & 0 & 0 & | & 0 \\ 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & | & 0 \\ P_{nc(v+k)} & 0 & 0 & 0 & 0 & P_{cc}(v+k) & P_{wc}(v+k) & 0 & | & 0 \\ P_{na1} & 0 & 0 & 0 & 0 & 0 & 0 & P_{fa1} & | & P_{ca} \\ P_{na2} & 0 & 0 & 0 & 0 & 0 & 0 & P_{fa2} & | & 0 \\ 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & | & 0 \\ - & - & - & - & - & - & - & - & - & - \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & | & 1 \end{pmatrix}$$

Here we have shown that the transition probability matrix can be broken into four parts. i.e; $\mathbf{P} = \begin{bmatrix} \mathbf{T} & \triangle \\ \psi & \mathbf{I} \end{bmatrix}$ where $\mathbf{T}$ is the transition probability matrix for the Non-absorbing (transient) states, $\triangle$ is the transition matrix from the non-absorbing

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2

A transition probability Matrix is a square matrix with zero or positive elements less than or equal to one such that the sum of elements in each row is unity. A row vector is called a probability vector whose element values are greater than or equal to zero and less than or equal to one and whose elements sum to one.

states to absorbing states, $\mathbf{I}$ is the identity matrix while ψ is the zero matrix. The average number of times the system stays in the non absorbing states can be found [51] from the matrix $[\mathbf{I}-\mathbf{T}]^{-1}$, where the matrix $\mathbf{T}$ here is

$$\mathbf{T} = \begin{pmatrix} 0 & P_{cc}(v) & P_{wc}(v) & 0 & P_{nc}(v) & 0 & 0 & 0 \\ 0 & 0 & 0 & P_{fa1} & P_{na1} & 0 & 0 & 0 \\ 0 & 0 & 0 & P_{fa2} & P_{na2} & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 \\ P_{nc(v+k)} & 0 & 0 & 0 & 0 & P_{cc}(v+k) & P_{wc}(v+k) & 0 \\ P_{na1} & 0 & 0 & 0 & 0 & 0 & 0 & P_{fa1} \\ P_{na2} & 0 & 0 & 0 & 0 & 0 & 0 & P_{fa2} \\ 1 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \end{pmatrix}$$

The top row entries of the matrix $[\mathbf{I}-\mathbf{T}]^{-1}$gives the average number of times the system stays in the various states before going into the absorbing state **ACQ** when it starts off in state 1.

3.6 Results

Numerical Results are all derived from a system using a 127 chips maximal length sequence.Fourteen tap Adaptive Filter is used and so fourteen cells are used, with each cell containing 9 chips, except for the last cell, which contains 10. For the cell

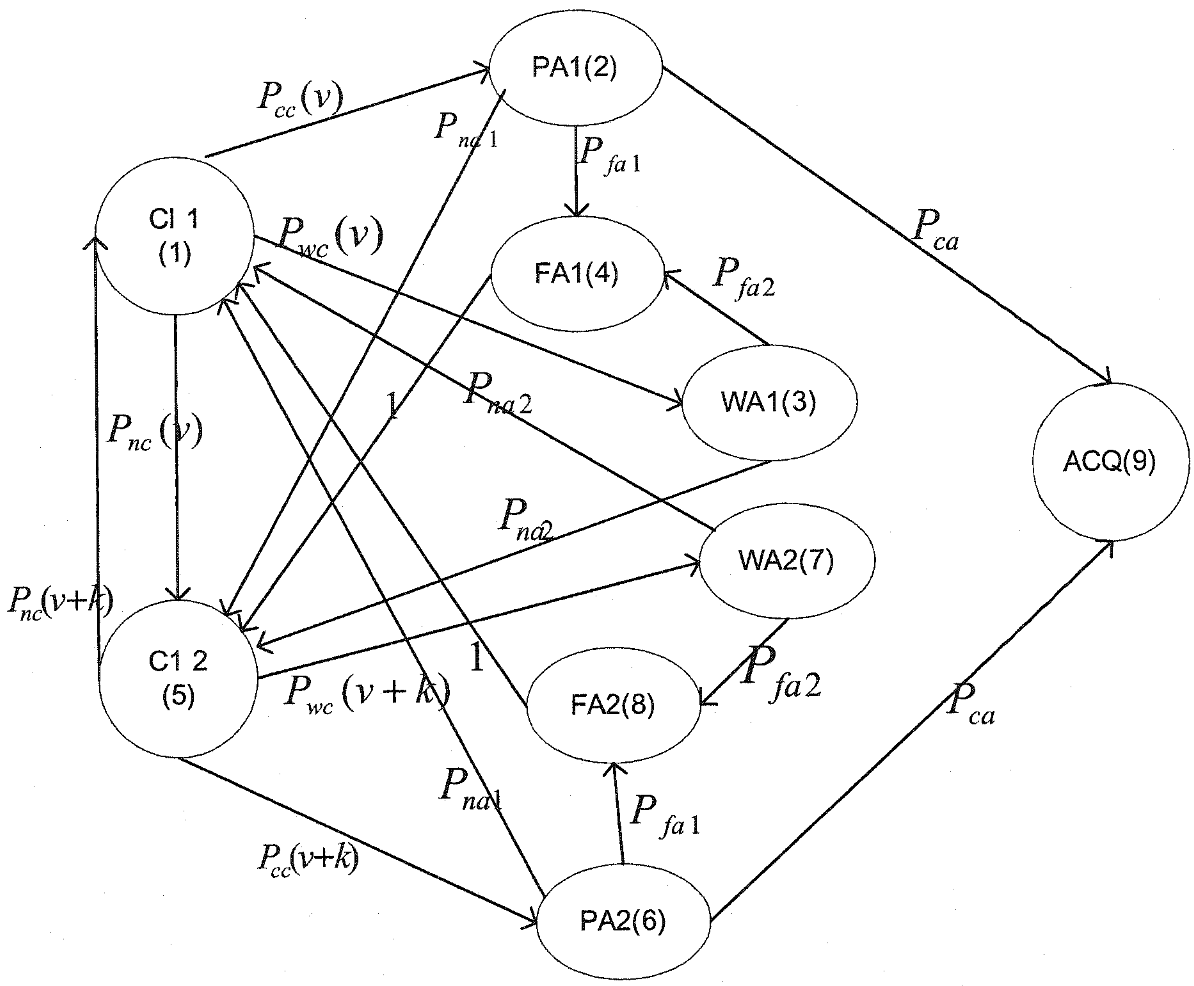

Figure 3.3: Transition diagram for the Proposed System

6 JUL 2000

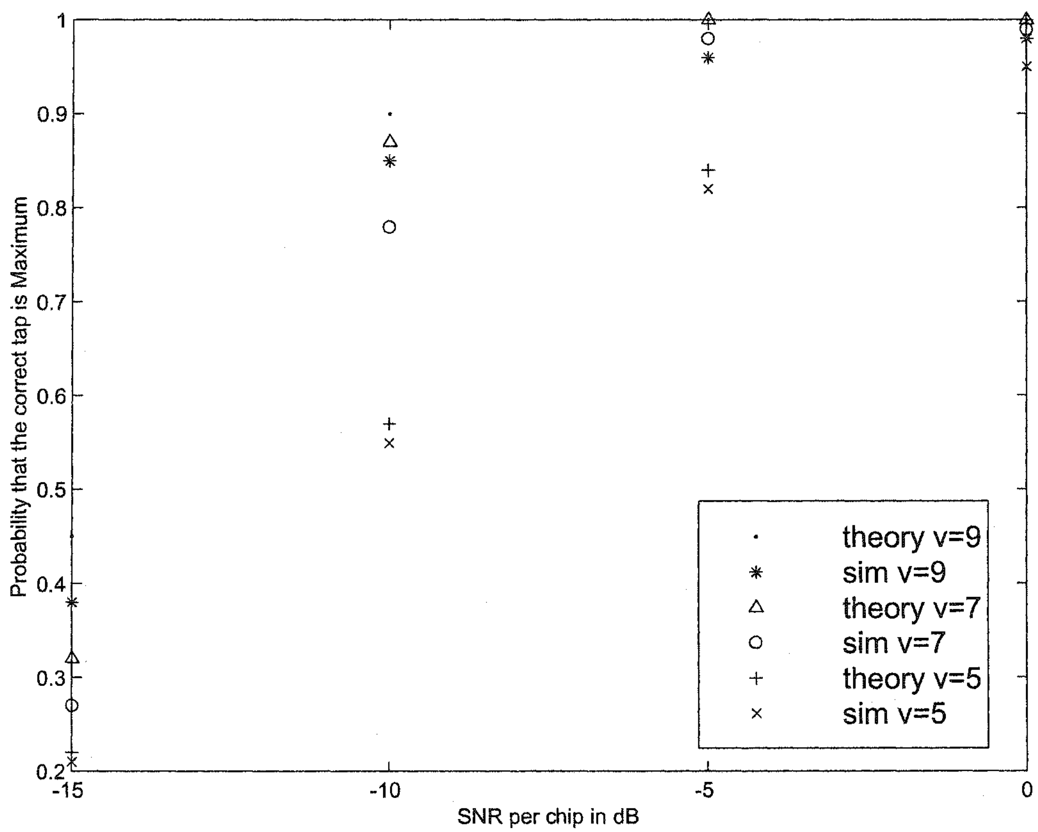

Figure 3.4: Graph for P_c against SNR_c for a system using a spreading sequence of 127 chips. Both Theoretical and Simulation results for chip matches of 9, 7 and 5 are shown

identification stage, a value of 0.0001 is used for μ. Fig 3.4 shows P_c, the probability of identifying the cell correctly using just a single sample, for SNR_c ranging from $0dBto-15dB$, and $v's$ varying from 5, the minimum, through 7 to 9, the maximum. We have run the system for 1000 experiments. Considering the short length of the spreading sequence used, the theoretical and simulation results agree quite well.

As P_c, the probability of correctly identifying the cell using just a single sample is quite low when signal to noise ratio is quite low, the system should take several samples of the peak to reduce the probability of false identification. For our simulations, we take 9 samples, at 127 chip sequence intervals, after settling down time (adaptation time) of eight 127 chip sequences. Note that the adaptation time is optimized experimentally such that it gives the optimum acquisition time performance. Thus K_{CI} comes out to be 2023 chips. Only if 5 or more samples have the peak at the same location will we declare that the cell has been correctly identified. This process reduces the wrong cell identification probability, and time is saved in not moving to the phase acquisition stage needlessly. The number of samples was chosen as 9 and not more because choosing to take more samples, even though can reduce the probability of wrong cell identification even further, will also cut down the probability of correct cell identification.

For the phase acquisition stage, the value of μ used is 0.0004. This bigger value is used to save settling down time, as the correct peak is more easily detected at this stage. Here, 7 samples are taken, at 70 chip intervals, after a settling down time of one full chip sequence of 127 chips. Thus K_A is 547 chips. Only if 4 or more out of the 7 samples have the peak located at the same location will the system declare that the correct phase is acquired. The penalty time for the false acquisition is taken to be five 127-chip sequences. In that amount of time, the system should be able to detect that the phase has not been acquired, and move back to the cell identification stage. Thus K_p is 635 chips.

Fig 3.5 shows the average chip intervals needed to acquire the phase of a 127-chip spreading sequence using our proposed system. The mean chip interval for acquisition of the system proposed by Tarhuni-Sheikh (description of the algorithm is given in Sec 2.5) using an adaptive filter with 16 taps, described in [37], is also shown. All the parameters for the T-S algorithm are the same as described in [37]. It is seen that for very low SNR_c of $-15dB$, our proposed system with only 14 taps works faster, needing only $14,200$ chips instead of $30,000$ chips on average to acquire the phase for low SNR_c condition. For high SNR_c, the mean time of acquisition remains constant over the range of -8 to $0dB$. This is due to the fact that the system is optimized for low signal to noise ratio environment, unlike the system described in [37], where signal power estimation is used, and the parameters of the phase acquisition are adjusted according to the estimated signal to noise power. If we use the signal to noise power too, the time required to acquire the phase can be decreased significantly under high SNR_c conditions. There is no need to do so, for as long as the system can acquire the phase of the spreading sequence prior to the

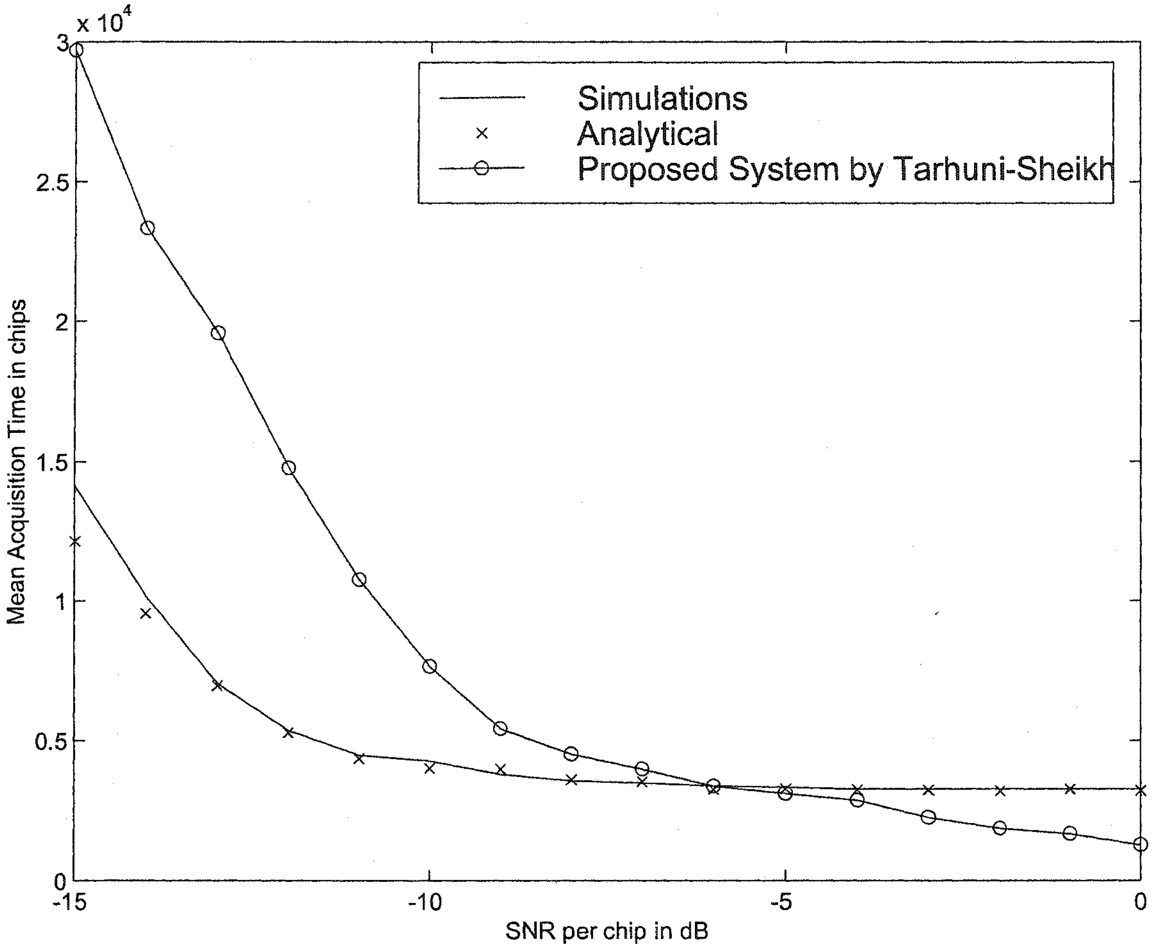

Figure 3.5: Graph of Mean Acquisition Time in chips against SNR_c in dB for system with 14 taps and using a 127 chip spreading sequence

end of the preamble preceding the message, it will operate correctly.

Chapter 4

Acquisition Performance Under Frequency Selective and Frequency Non-Selective Rayleigh Fading Channels

The aim of this chapter is to discuss the performance of the proposed code acquisition system under fading conditions usually encountered when the signal is transmitted over mobile radio channel.We present simulation results for the Mean Acquisition Time of the proposed acquisition system under frequency selective and frequency non-selective Rayleigh Fading Channels. For the frequency selective case, Exponential and Uniform Multipath delay profiles are used. Moreover in case of frequency selective fading, we also present simulation results when multipath components experiences different Doppler shifts.

4.1 The Mobile Radio Channel

Radio channel is the link between the transmitter and the receiver that carries information bearing signal in the form of electromagnetic waves. The radio channel is commonly characterized by scatterers (local to the receiver) and reflectors (local to the transmitter).

Fading experienced by a signal propagating through a mobile radio channel depends on the nature of the transmitted signal with respect to the characteristics of the channel. Two important factors that affect the type of fading are:

- **Multipath Propagation** — The presence of reflecting objects and scatterers in the environment give rise to multiple versions of the transmitted signal at the receiver. The multipath medium introduces time spread in the signal *i.e.* these replicas of the transmitted signal are displaced with respect to one another in time. *Delay spread or multipath spread* is used to quantify the signal distortion due to multipath propagation. The maximum time over which the received signal power is non-negligible is usually termed as the excess multipath delay.

- **Doppler Shift** — The time variations in the channel [1] are evidenced as a Doppler broadening and perhaps, in addition as a Doppler shift of a spectral line. *Doppler spread* is used to quantify the signal fading due to Doppler shift. This Doppler rate is proportional to the inverse of the coherence time of the channel. The coherence time is the maximum time over which the channel parameters are assumed to be significantly correlated.

Depending on the relation between the signal parameters (such as BW, symbol duration etc.) and the channel parameters (delay and Doppler spreads) different transmitted signals will undergo different types of fading. The time and frequency dispersion mechanisms in a mobile radio channel lead to four possible distinct effects, which manifest depending on the nature of the transmitted signal, the channel,the speed and orientation of the mobile. While multipath delay spread leads to *time dispersion* and *frequency selective fading*, Doppler spread leads to *frequency dispersion* and *time selective fading* [52].

4.1.1 Discrete-Time Channel Model

With the advent of statistical communication theory in late 1940's, the research in the characterization and modeling of fading channels gained considerable attention.

[1] These time variations may occur because of the physical characteristics of the medium or relative motion between base station and the mobile station or relative motion of the surroundings.

Over many years, a large number of experiments have been carried out to understand the nature of fading channels. Earlier work in this area includes contributions from Bello[53], Clarke[54] and Jakes [55].

Assuming low-pass equivalent model for the channel, the received signal $r(t)$ over a fading multipath channel can be represented by

$$r(t) = \int_{-\infty}^{\infty} h(\tau, t) s(t - \tau) d\tau \tag{4.1}$$

where $s(t)$ is the transmitted signal and $h(\tau, t)$ is the time-variant channel impulse response at delay τ and time instant t. In discrete form

$$r(n) = \sum_{i=-\infty}^{\infty} h(iT_c, n) s(n - iT_c) \tag{4.2}$$

where T_c is the chip duration and n represents the sampling index. Defining a compact notation for the time varying channel coefficients in the form $h_i(n) = h(iT_c, n)$, (4.2) can be written as

$$r(n) = \sum_{i=-\infty}^{\infty} h_i(n) s(n - iT_c) \tag{4.3}$$

The form of the received signal in (4.3) suggests that the impulse response of fading multipath channel can be modeled as a tapped delay line filter, modelled as a finite impulse response (FIR) filter, with tap spacing T_c and time varying tap coefficients

$h_i(n)$. For all practical purposes one can use truncated tapped delay line filter with $L = \frac{T_m}{T_c} + 1$ taps [56],i.e;

$$r(n) = \sum_{i=1}^{L} h_i(n)s(n - iT_c) \tag{4.4}$$

The time varying complex coefficients $h_i(n)$ are characterized as random processes because of the constantly changing physical characteristics of the media. The tap weights, $h_i(n)$, in Fig. 4.1 can be expressed as

$$h_i(n) = \sqrt{\rho_i}\, G_i(n) \tag{4.5}$$

where ρ_i is the strength of the ith path and $G_i(n)$ is the complex stochastic process specified by its mean square value and power spectrum density. The next sections discuss these two parameters in detail.

4.1.2 Fading Statistics

Fading describes the rapid fluctuation of the amplitude of radio signal when passing through radio channels over a short period of time or travel distance. The complex stochastic process $G_i(n)$ in (4.5) represents the fading and can be completely characterized by specifying the pdf of its amplitude and phase and the autocorrelation. The simplest process that can exhibit both time-selective and frequency-selective

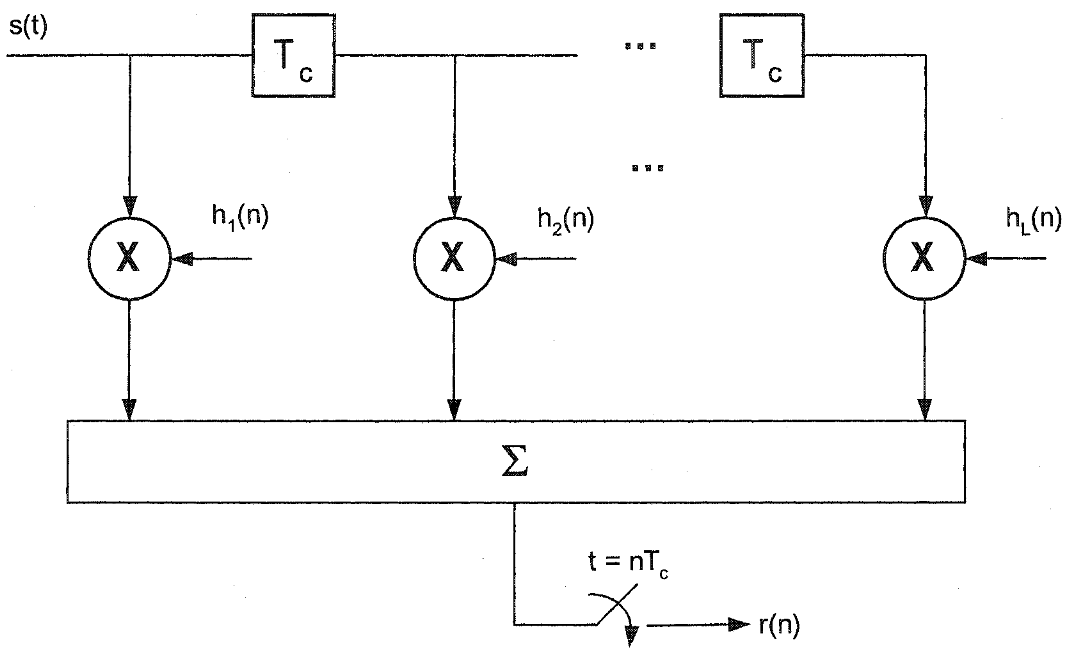

Figure 4.1: Tapped delay line model of fading channel

fading is *wide-sense stationary uncorrelated scattering (WSSUS)* process introduced by Bello [53]. The number of uncorrelated paths is sufficiently large so that the quadrature components of the fading process are Gaussian distributed according to the central limit theorem. The Gaussian WSSUS model fits well to land mobile radio channels and have been specified in the ITU's test environment [57]. In the absence of direct path, the Gaussian process has zero mean and the pdf of the envelope is Rayleigh [55, Chapter 1] given by

$$f_G(g) = \frac{g}{\sigma^2}\, e^{-g^2/2\sigma^2}, \quad g \geq 0 \tag{4.6}$$

where $\sigma^2 = E[GG^*]$ is the variance of the Gaussian process. The phase pdf has uniform distribution [54, 55, 58]

$$f_\Theta(\theta) = \frac{1}{2\pi}, \quad 0 \leq \theta \leq 2\pi \tag{4.7}$$

A typical and often-assumed shape for the power spectral density, also known as Doppler Spectrum, of the fading process for mobile radio channels is given by *Jakes' Spectrum* [54, 55, 59]

$$S(f) = \begin{cases} \frac{\sigma^2}{\pi f_m \sqrt{1-(f/f_m)^2}} & |f| \leq f_m \\ 0 & \text{elsewhere} \end{cases} \tag{4.8}$$

where f_m is the maximum Doppler frequency shift given by

$$f_m = \frac{v f_c}{C} \tag{4.9}$$

in which v, f_c and C are the vehicle speed, frequency of the carrier and speed of light respectively. The autocorrelation function of the fading process is given by [54, 55, 58]

$$\mathcal{R}(\tau) = \sigma^2 \ J_0(2\pi f_m \tau) \tag{4.10}$$

where $J_0(\cdot)$ is the zeroth order Bessel function of the first kind.

4.1.3 Power Delay Profile

To specify the tapped-delay line model of fading channels completely, the relative strength of the multi-paths, ρ_i, is also required as mentioned in (4.5). The relative strength of the paths is usually represented by *power delay profile (PDP)* of the channel which is dependent on the environment and is usually determined by experiments and measurements. To accurately evaluate the relative performance of candidate radio transmission technologies (RTT's), the test environment channel parameters or PDP has been specified by ITU [57].

The delay profile and Doppler spectrum for Pedestrian Channel A and Vehicular Channel A[2] is reproduced from [57] in Table 4.1 and 4.3. The term 'CLASSIC' for

[2]Channel A represents a low delay spread case

Doppler spectrum refers to Jakes' Spectrum in (4.8) and the fading statistics are assumed to be Rayleigh.

Table 4.1: Pedestrian Test Environment Tapped delay line parameters

Tap	Relative Delay [ns]	Average Power [dB]	Doppler Spectrum
1	0	0	CLASSIC
2	110	-9.7	CLASSIC
3	190	-19.2	CLASSIC
4	410	-22.8	CLASSIC

Table 4.2: Modified Pedestrian Test Environment Tapped delay line parameters

Tap	Relative Delay [ns]	Average Power [dB]	Doppler Spectrum
1	0	0	CLASSIC
2	260.4	-12.77	CLASSIC
3	512.8	-25.48	CLASSIC

Table 4.3: Vehicular Test Environment Tapped delay line parameters

Tap	Relative Delay [ns]	Average Power [dB]	Doppler Spectrum
1	0	0	CLASSIC
2	310	-1.0	CLASSIC
3	710	-9.0	CLASSIC
4	1090	-10.0	CLASSIC
5	1730	-15.0	CLASSIC
6	2510	-20.0	CLASSIC

Table 4.4: Modified Vehicular Test Environment Tapped delay line parameters

Tap	Relative Delay [ns]	Average Power [dB]	Doppler Spectrum
1	0	0	CLASSIC
2	260.4	-1.92	CLASSIC
3	512.8	-7.31	CLASSIC
4	781.2	-10.39	CLASSIC
5	1041.6	-10.89	CLASSIC
6	1302	-17.31	CLASSIC
7	1562.4	-19.48	CLASSIC
8	1823	-16.91	CLASSIC
9	2083	0	CLASSIC
10	2344	-24.43	CLASSIC
11	2604	-21.95	CLASSIC

However, these models cannot be used directly in the simulations because of the time resolution. In the simulations the sampling time is equal to one chip (T_c) thus the models must be modified so that the delay between two consecutive paths is a multiple of T_c. An interpolative-type method for this modification is given in [60] for which mathematical formulation for the modified pedestrian and vehicular channels and plots of modified channels proposed by ITU is also given in [61]. The modified delay profile for the Pedestrian Channel A and the Vehicular Channel A corresponding to 3.84 Mcps are listed in Table 4.2 and 4.4 respectively.

4.1.4 Simulation Model of Fading Channel

As discussed in Sec. 4.1.1 a fading channel can be represented by tapped delay line FIR filter with tap weights as given by (4.5). This section describes the method that will be used to simulate the filter or channel. Basically, we need colored Gaussian noise to realize fading channel statistics which may be Rayleigh, Rician or any other. These colored Gaussian processes can be generated either by *filtering* white Gaussian noise [54] or by *deterministic* methods [55, 59] or by *Monte Carlo* approach [62].

Filtering methods require large number of filter taps [63] to shape the spectrum of Gaussian processes as Jakes spectrum. A deterministic method to simulate mobile fading channels is based on Rice's sum of sinusoids [64, 65]. In this case, a colored Gaussian noise process is approximated by a finite sum of sinusoids with proper weights and frequencies. Jakes also presented a realization for the simulation of the fading channel model which generates real and imaginary parts of the channel tap coefficients as weighted sum of sinusoids. Jakes simulator has been widely used and extensively studied intermittently over the past three decades [66, 58, 59]. Recently, Pop and Beaulieu [67] have highlighted few shortcomings in the Jakes model. They proposed to include a random phase in the low frequency oscillators of the Jakes model. In this thesis, I have implemented this modified semi-deterministic Jakes model as fading channel simulator.

The real and the imaginary components of the channel tap are generated by [55]

$$\begin{aligned} g_I(t) &= 2\sum_{n=1}^{N_o} \cos\beta_n \cos(\omega_n t + \phi_n) + \sqrt{2}\cos\alpha\cos\omega_m t \\ g_Q(t) &= 2\sum_{n=1}^{N_o} \sin\beta_n \cos(\omega_n t + \phi_n) + \sqrt{2}\sin\alpha\cos\omega_m t \end{aligned} \tag{4.11}$$

$$\beta_n = \frac{n\pi}{N_o + 1} \qquad \omega_n = \omega_m \cos\left(\frac{2\pi n}{N}\right) \qquad N_o = \frac{1}{2}\left(\frac{N}{2} - 1\right) \tag{4.12}$$

where $t = kT_s$ and $\phi_1, \cdots, \phi_{N_o}$ are uniformly distributed random variables over $[0, 2\pi]$. For multipath uncorrelated scattering, I have implemented the technique proposed by Jakes [55] whose correct version is available in [58, page 80]. In this technique, the nth oscillator is given an additional phase shift $\gamma_{nl} + \beta_n$ with gains as before. For lth path the in-phase component of the fading can be written as:

$$g_l^I(t) = 2\sum_{n=1}^{N_o} \cos\beta_n \cos(\omega_n t + \theta_{nl} + \phi_n) + \sqrt{2}\cos\alpha\cos\omega_m t \tag{4.13}$$

$$\theta_{nl} = \gamma_{nl} + \beta_n \qquad \beta_n = \frac{n\pi}{N_o+1} \qquad \gamma_{nl} = \frac{2\pi(l-1)n}{N_o+1} \tag{4.14}$$

The quadrature phase component can be modified in the same manner. The normalized complex channel tap for lth path is

$$G_l(t) = g_l^I(t) + jg_l^Q(t) \tag{4.15}$$

which will be normalized such that $E[G_l G_l^*] = 1$. Thus, the complete channel tap for the model shown in Fig. 4.1 can be obtained by combining (4.5), (4.15) and (4.13) (and the mathematical expression for interpolation method used to modify the PDP proposed by ITU).

4.2 Simulation Results for the Mean Acquisition Time Under Fading Conditions

In our simulations, we consider both frequency selective and flat fading channels . In case of frequency selective case, both Uniform Multipath Delay Profile (UMDP) as well as Exponential Delay Profile (EDP) are studied. The fading is assumed to be Rayleigh distributed random process. Fading process is assumed such that the channel conditions remain unchanged for the symbol duration i.e. for the spreading code length. For each experiment, the delays were generated with a uniform distribution over the code length. In case of pedestrian channels, it has also been assumed that all paths are situated within the same cell, i.e. there is only one state that may lead to acquisition. Although this assumption does not affect the results significantly, it does simplify the simulation. But for the case of vehicular channel, this assumption can not be valid as we have 11 paths and cell size is 9 chips. The number of adaptive filter taps, M, is 14 and the LMS step size parameter value used is the same as used in case of AWGN Environment.

Fig 4.2 shows the absolute mean acquisition time in chips for frequency selective fading with Exponential MDP proposed by ITU, frequency selective with Uniform MDP and for Flat fading case and for T-S Algorithm (described in Sec 2.5) with Exponential MDP(ITU). The maximum Doppler rate is 10Hz. For case of T-S Algorithm, parameters taken are the same as described in [39]. Also the performance of AWGN is included to assess the degradation due to fading. From Fig 4.3, we see that, for a fixed acquisition time, flat fading degrades the performance by about 6dB for low SNR while frequency selective fading has a 4 to 5dB degradation depending upon the MDP. But the performance deteriorates more for high SNR conditions as the modified system is optimized for low SNR environment. we also compared the performance of our proposed algorithm with that of T-S Algorithm with the same channel conditions which we used for Exponential MDP case. We found that our proposed algorithm performs well for low SNR conditions as T-S algorithm degrades 10dB for Exponential MDP for low SNR conditions but this degradation is less for high SNR condition compared to our proposed algorithm. The reason is that in case of T-S algorithm, the parameters of the phase acquisition are adjusted according to the estimated signal to noise power. So for high SNR conditions, performance of T-S system is better than our proposed algorithm where the mean acquisition time remains constant for a wide range of SNR's. Moreover degradation in Mean Acquisition Time for our proposed system is far less than that reported in [16] for a parallel acquisition system for low SNR condition, which was also about 10-12 dB. Note that Uniform MDP is better than Exponential MDP due to larger inherent selection diversity. We note that the frequency selective environment is more favorable than flat fading since the system can exploit the inherent multipath diversity in acquiring the strongest available path. The ability of the proposed system to combat fading in general could be attributed to the adaptive nature of the system and the decision strategy which maximizes the detection probability. Although for data detection, a linear adaptive filter may not operate successfully in a fading environment [56], for code acquisition it provides good performance due to expected larger minimum MSE.

Fig 4.4 shows absolute mean acquisition time performance for the case when multipath components are subjected to Doppler rates of 8, 10 and 12Hz respectively.

Here we can see that Mean acquisition time performance is poor for the case when path with 12Hz has more average power and is better when path with 8Hz has high average power. So we conclude that for higher Doppler rates, the degradation in performance of Mean Acquisition time is greater for the proposed algorithm.

Fig 4.5 shows the absolute mean acquisition time performance of our proposed system for vehicular test environment with doppler frequency of 100Hz (for Exponential delay profile). In vehicular test environment, we have 11 paths and the cell size is 9 chips so all the paths can not be situated in the same cell. The distribution of the paths reveals that some of the stronger paths can be situated in the adjacent cell having inferior match which will increase the false alarm errors. Moreover fast fading effect (doppler frequency of 100Hz) is also compounded here and so the mean acquisition time performance is poorer than the case of pedestrian channel.

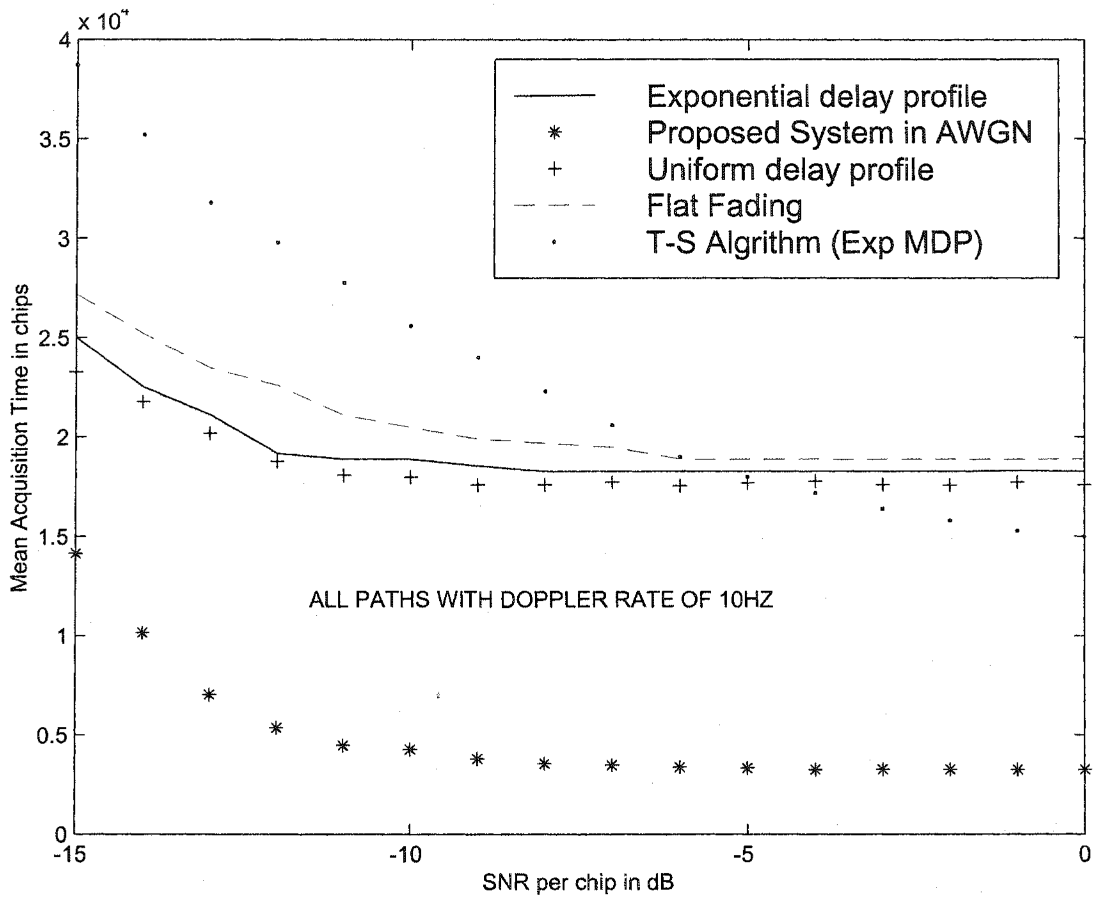

Figure 4.2: Mean Acquisition Time for Doppler rate 10Hz: AWGN, Exp MDP(ITU), Uniform MDP, Flat fading and T-S Algorithm (Exp MDP)

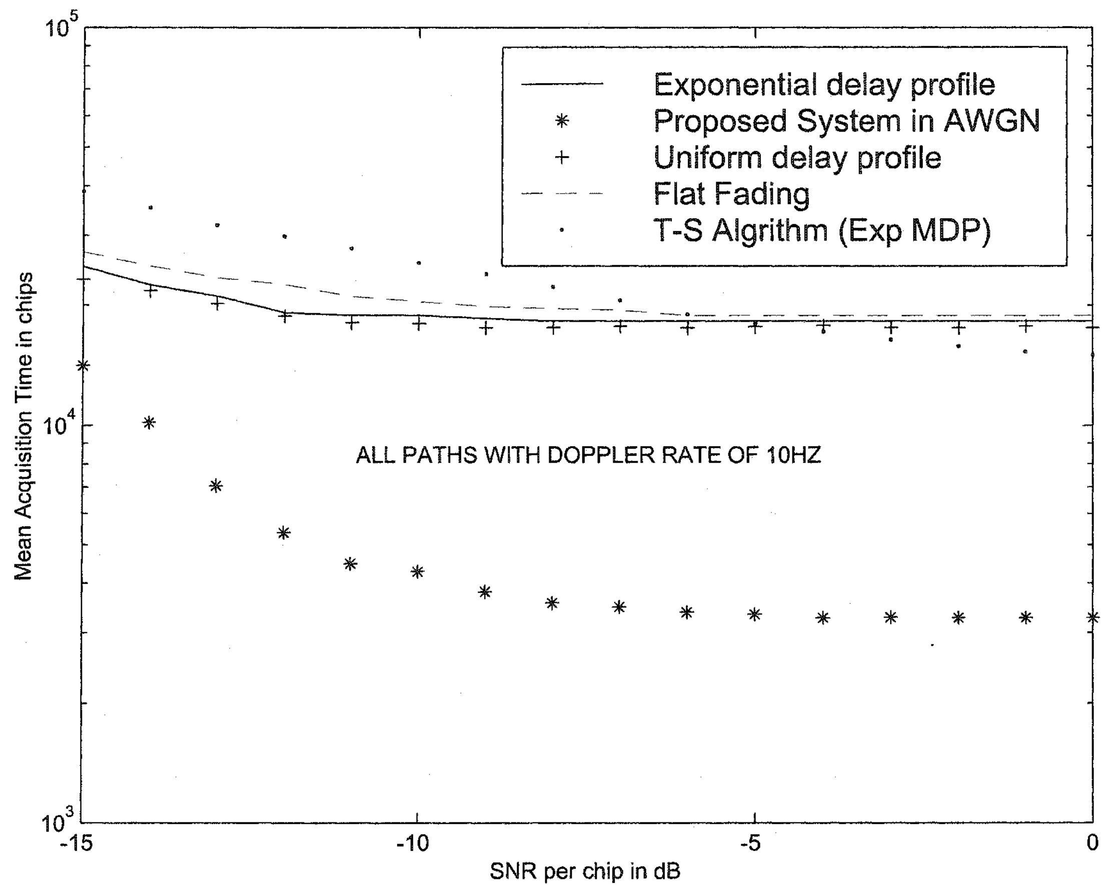

Figure 4.3: Mean Acquisition Time (on Log scale) for Doppler rate 10Hz: AWGN, Exponential MDP(ITU), Uniform MDP, Flat fading and Tarhuni's Algorithm (Exp MDP)

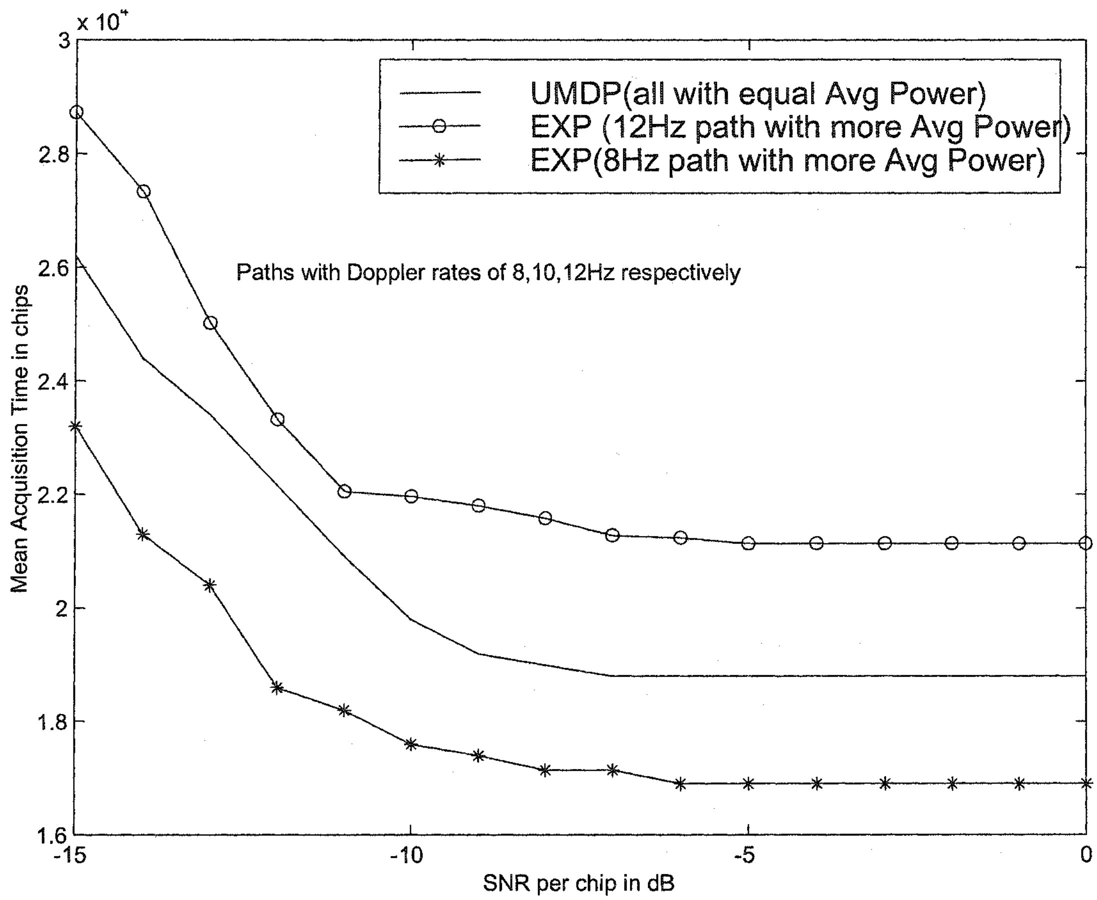

Figure 4.4: Mean Acquisition Time for varying Doppler shifts of 8,10 and 12Hz on multipath components of fading channels respectively (Exponential MDP by ITU and Uniform MDP)

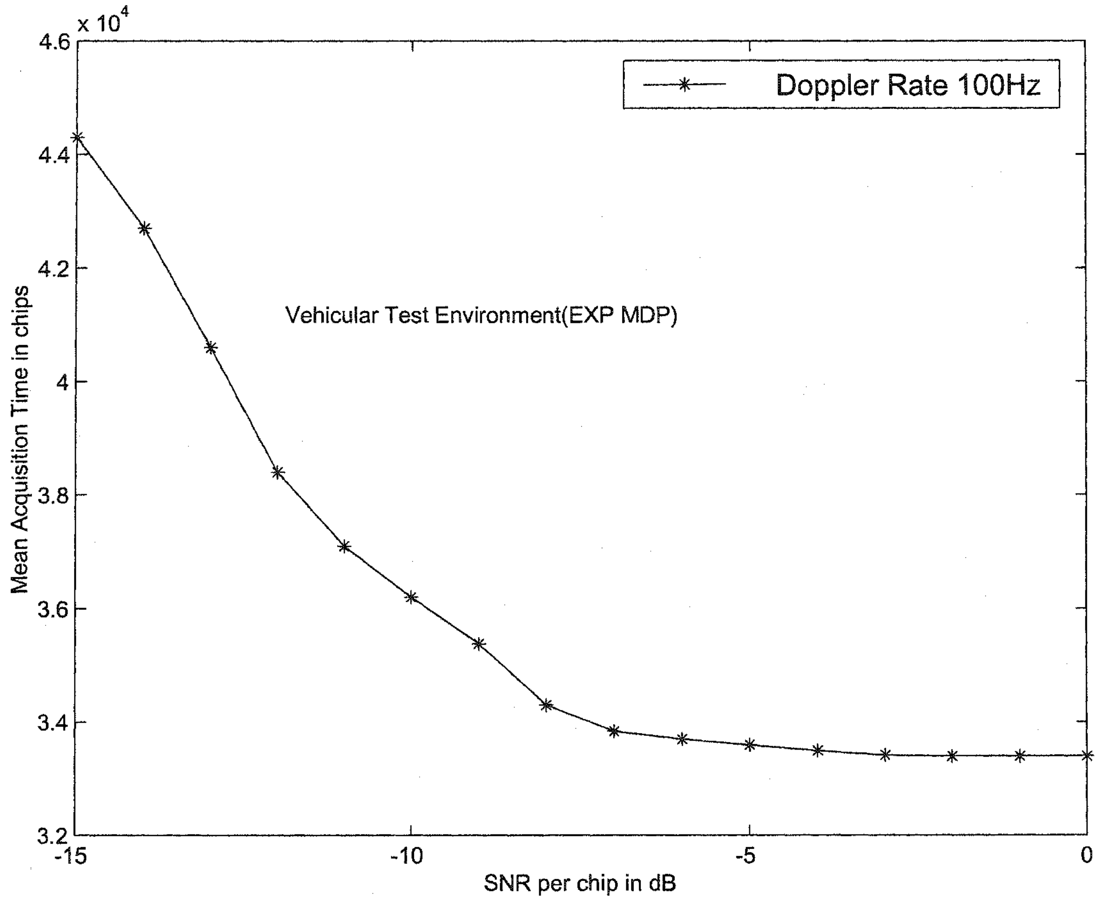

Figure 4.5: Mean Acquisition Time for Vehicular Test environment proposed by ITU with Doppler frequency of 100Hz (Exponential delay profile))

Chapter 5

Code Acquisition system performance in Multiuser Environment

5.1 Introduction

As explained in literature, the capacity of CDMA system is defined as the maximum number of simultaneous users that can achieve PN code acquisition with certain reliability. In CDMA systems, the Multiple Access Interference (MAI) can be dominant and the acquisition performance of both the simple correlator and the matched filter schemes is significantly degraded. Hence the capacity of such systems is essentially limited by the number of simultaneous users that can achieve and maintain code synchronization rather than by the number of users that can maintain a certain BER performance as explained in [68]. So using a more efficient code acquisition scheme can improve the capacity of such systems. This degradation is compounded when a near far problem exists. Therefore improving the acquisition system performance in CDMA systems by using more advanced near far resistant acquisition schemes would improve the actual capacity of the system to achieve the BER-based capacity. A Near far resistant PN code Acquisition system using MMSE equalizer was proposed in [69]. It was shown that the system can achieve the reliable acquisition in the presence of the strong MAI, but the adaptation time was very long. Moreover the complexity of the system is very high since the equalizer length should be the same as the code length.

Then a new hybrid scheme for Code Acquisition in MAI in [41] using adaptive filters was proposed which significantly outperforms the conventional matched filter based acquisition schemes. It improves the acquisition based capacity because it takes into consideration the presence of the Multiuser Interference while finding the optimum tap weight setting.

The goal of this chapter is to discuss the performance of our proposed PN code acquisition scheme (explained in Sec 3.2) which is also hybrid scheme employing the adaptive filter but implemented in a slightly different way than that explained in [41] and it has been shown that for the same SNR conditions, the modified system takes less time for code acquisition in MAI environment than that explained in [41] as well as it can achieve the BER based capacity.

5.2 System Model and System Description

The received baseband signal is given by

$$r(t) = \sum_{k=1}^{K} x_k(t - \tau_k) + n(t) \tag{5.1}$$

where K is the number of users, $n(t)$ is AWGN, τ_ki s the delay of the kth user, and $x_k(t)$ is the kth user DS/SS signal given by

$$x_k(t) = \sum_{n=-\infty}^{\infty} d_k(t - nT_b) \sum_{j=0}^{L-1} c_k(j) \prod_{T_c}{}_{(t-jT_c)} \tag{5.2}$$

where $\{d_k\}$ is the kth user information sequence with $d_k(m) = \pm 1$,$\{c_k\}$ is the kth user maximal length spreading sequence, L is the spreading code length and $\prod_{T_c}$ is a rectangular shaping pulse with unit amplitude and duration T_c.

To simplify the analysis , we shall make the same assumptions as we did in chapter 3 for single user case.

The received signal is sampled at the chip rate to produce the sequence

$$r(i) = \sum_{k=1}^{K} \sum_{n=-\infty}^{\infty} d_k(t - nT_b) \sum_{j=0}^{L-1} c_k(j) \prod(i - j - \gamma_k) + n(i) \tag{5.3}$$

where we have assumed that the delays are integer multiples of the chip duration, i.e. $\tau_k \in \{0, 1, 2...., L-1\}$. This assumption (chip synchronous) is merely to simplify the analysis and does not affect the results significantly. Notice that the dependence on the chip duration has been dropped for notational convenience.

Our objective is to estimate the delay of the desired user, τ_1, from the composite received signal, $\{r(i)\}$.

The same Adaptive Filter based DS/SS code acquisition system structure is used as in Fig 3.1. The first task of the system is to find the neighbourhood in which the correct phase delay of the desired user is located, before it zeroes it on the phase itself.

The desired sequence is formed by adding c, called a cell, of the spreading code of the desired user at a certain phase offset into the adaptive filter. Thus

$$u(m) = \sum_{i=1}^{c} c_1(mc + i) \tag{5.4}$$

The integer c is chosen so that $cM = L$, the length of the spreading sequence. Note that $c$can be chosen so that $cM \approx L$ if $cM \neq L$. The same number, c, of the received samples are $\mathbf{r_m}$ also added to form the input to the Adaptive Filter, so as described in chapter 3.

$$r(m) = \sum_{i=1}^{c} a(mc + k + i) + n(mc + k + i) \tag{5.5}$$

The adaptive filter is clocked at cT_c, so that the output of the adaptive filter is given by the convolution sum

$$\begin{aligned} y(m) &= \sum_{j=0}^{M-1} \mathbf{w}_j(m)\mathbf{x}(m-j) \\ &= \mathbf{w}^T(m)\mathbf{x}(m) \end{aligned} \tag{5.6}$$

where $\mathbf{w}(m) = [w_0(m) \;\; w_1(m)......w_{M-1}(m)]^T$ is the tap weight vector of the AF at mcT_c , and $\mathbf{r}(m) = [r(m) \;\; r(m-1)......r(m-M+1)]^T$ consists of the present and the past $M-1$ samples of the sums of the filter input vector.

An error signal is formed from the AF output and the reference signals such that

$$e(m) = u(m) - y(m) \tag{5.7}$$

which is used to adjust the filter taps according to the **LMS** algorithm

$$\mathbf{w}(m+1) = \mathbf{w}(m) + \mu e(m)\mathbf{r}(m) \tag{5.8}$$

where μ is the **LMS** algorithm step size which controls the convergence speed and the steady state mean square error. As we explained earlier also in chapter 3, following convergence of the tap weights, the maximum of the tap weights is identified. As will be seen later, a record of the maximum tap weights will indicate the location of the cell where the chip phase delay can be found.

Once the correct cell is located, the system will move on to acquire the phase delay itself. It does this by changing the clock rate of the adaptive filter to T_c, and loading the delay line elements with the subsequence that is half M chips ahead of the cell which has been pinpointed. A fixed number of samples are taken, and the tap where the majority of the peaks occur will indicate the position of the delay phase.

5.3 Probability of Cell Acquisition Failure

The optimum tap weight vector of the FIR adaptive filter is given by the Wiener Hopf Equation

$$\mathbf{w}_{opt} = \mathbf{R}^{-1}\mathbf{p} \tag{5.9}$$

where $\mathbf{R} = E\{\mathbf{r}(m)\mathbf{r}^T(m)\}$ is the $M \times M$ autocorrelation matrix of the AF input vector, and $\mathbf{p} = E\{\mathbf{r}(m)u(m)\}$ is the $M \times 1$ cross correlation vector between the AF input and the (desired user) spreading sequence.

The elements of $\mathbf{R}$ can be shown to be

$$r_{pq} = \begin{cases} c(1+\sigma^2) & \text{if } p = q \\ 0 & p \neq q \end{cases} \tag{5.10}$$

for $p, q = \{0, 1, 2, ..., M-1\}$.

The elements of the cross correlation vector between $u(m)$ and $r(m)$ is an $M \times 1$

vector with elements $p(q)$ where $q = \{0, 1, 2....., M-1\}$ and is

$$p(q) = \begin{cases} 0 & if \;\; a(n)'s \text{ do not match at all} \\ v & if \;\; a(n)'s \text{ match in v places with superior match} \\ (c-v) & if \;\; a(n)'s \text{ match in (c-v) places with inferior match} \end{cases} \tag{5.11}$$

Based on the previous observations, the optimum tap weight vector may be visualized as to have a peak at the tap corresponding to the correct delay offset i.e. corresponding to the cell where the delay offset is located. The decision strategy is to find the set of optimum coefficients $\mathbf{w}_{opt} = \{w_0 \;\; w_1 \;\; ..... \;\; w_{M-1}\}^T$, and then select the delay offset that corresponds to the maximum coefficient.

The performance of the proposed AF acquisition system can be measured by the probability of cell acquisition failure, P_{wc}. This occurs when the maximum of the optimum filter taps does not correspond to the correct delay estimate, i.e.

$$P_{wc} = 1 - P(w_\alpha \geq w_0, w_\alpha \geq w_1, ..., w_\alpha \geq w_{M-1}) \tag{5.12}$$

where w_α is the AF coefficient that corresponds to the cell having correct delay offset and $P(w_\alpha \geq w_0, w_\alpha \geq w_1, ..., w_\alpha \geq w_{M-1})$ is the joint probability density function (PDF). The exact PDF is difficult to drive, however a very good Gaussian approximation has been reported in [50]. The tap weight vector of the LMS adaptive filter has been shown to follow a gaussian random distribution with mean vector of

$$\underline{m} = \underline{w_{opt}} \tag{5.13}$$

and covariance matrix of

$$\mathbf{C} \approx \frac{\mu}{2} J_{min} \mathbf{I} \tag{5.14}$$

where $\mathbf{I}$ is an identity matrix, μ is the adaptation step size and J_{min} is the minimum mean square error (MMSE) given by

$$J_{min} = \sigma_d^2 - \mathbf{p}^T \mathbf{R}^{-1} \mathbf{p} \tag{5.15}$$

with $\sigma_d^2 = c$ is the power of the desired sequence. Equation (5.14) indicates that the tap weights are uncorrelated. The acquisition failure probability can be written as

$$P_F = 1 - \frac{1}{\sqrt{2\pi\sigma_w^2}} \int_{-\infty}^{+\infty} [1 - Q(\frac{w_c}{\sigma_w})]^{M-2} [1 - Q(\frac{w_c - m_\beta}{\sigma_w})] \exp(-\frac{(w_c - m_c)^2}{2\sigma_w^2}) dw_c \tag{5.16}$$

where $Q(x)$ is the Q function and m_c and σ_w^2 are the mean and variance of the correct tap which has the largest match between the incoming sequence and local sequence respectively whereas m_β is the adjacent cell which has an inferior match. We have also assumed that all the coefficients have the same variance σ_w^2 obtained from the diagonal elements of Eq 5.14. This is a reasonable assumption especially when the SNR per chip is small. we consider a CDMA system with code length of $L = 127$ chips. Fourteen cells are used again, with each cell containing 9 chips, except for the last that contains 10 chips. For the cell identification stage, a value of 0.0001 is used for μ. The adaptation period for the filter is selected as multiple of the code length.

Fig 5.1 shows P_F, the probability of identifying the cell wrongly for SNR_c ranging from 0 dB to -15 dB, using just a single sample for a short length spreading sequence of $L = 127$ in Multiple Access Interference with a maximum of 4 users. The decision criteria is the same as we take in case of single user. The adaptation period is taken to be of ten 127 chip sequences. The tap where the peak occur will indicate the position of the cell where the chip phase delay can be found. Here we have one desired user and three interfering users. The interfering users have their data as well while the desired user has been assumed to be without data sequence. The spreading sequence for the users have been generated with a linear shift register of length 7 and with different preferred pair combinations. For the desired user, the delays are generated with uniform distribution over the code length. We can see that both theoretical and simulation results agree quite well.

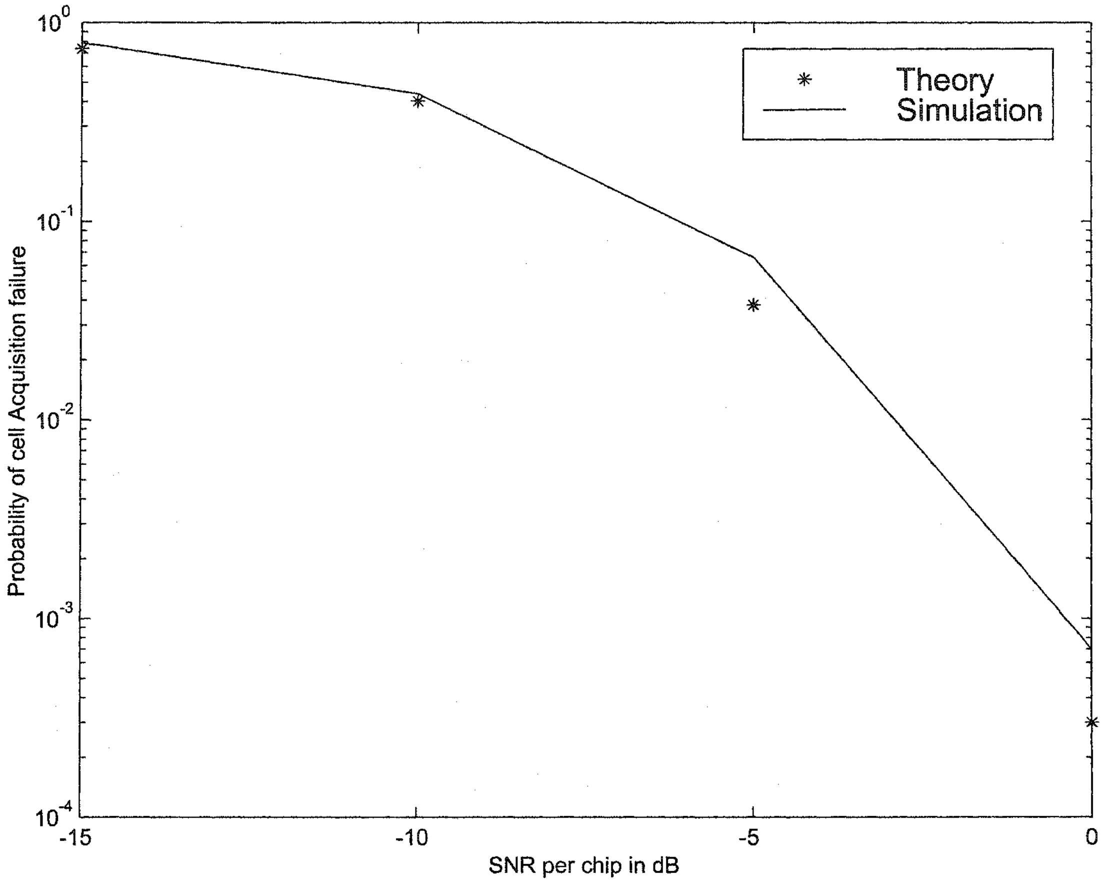

Figure 5.1: Theoretical and Simulation results for Probability of Cell Acquisition Failure against SNR_c for a system using spreading sequence of 127 chips.

5.4 Effect of Adaptation time on phase acquisition in Multiuser Environment

Fig 5.2 shows the effect of the **LMS** adaptation period on the acquisition failure probability with different number of users in a Multiple Access Interference environment for SNR_c of $-5\ dB$. Here we have generated spreading codes for users by using different initial loads of the shift register. The step size parameter is taken to be 0.0001 as in single user environment.

Simulation results are presented here for adaptation periods of 6, 8 and 12 code periods respectively for cell identification stage. For the phase identification stage, the adaptation period is taken to be one code length. The capacity of the system is significantly improved from 4 users with 90% success probability at 6 code periods adaptation time to nearly 40 users at 12 code periods. This is because the longer the adaptation period, the higher the probability that the AF coefficients converge to the optimum value. But a longer acquisition time is incurred because the AF should test all cells for a longer period and then phase itself. It has been reported in literature [41] that for 10% failure probability, Matched Filter can support upto 10 users at most which is significantly smaller than that supported by proposed **AF** scheme and the one proposed earlier.

The capacity of a **DS/CDMA** system based on bit-error-rate (BER) performance can be computed using a Gaussian approximation [52] and the results indicate that a CDMA system with perfect power control and processing gain of 127 can support up to 40 users with BER of 10^{-3} in AWGN. Hence Matched Filter (MF) system has a very limited capacity, while the previously proposed hybrid AF scheme as well as our modified scheme can achieve the capacity obtained by the BER criterion.

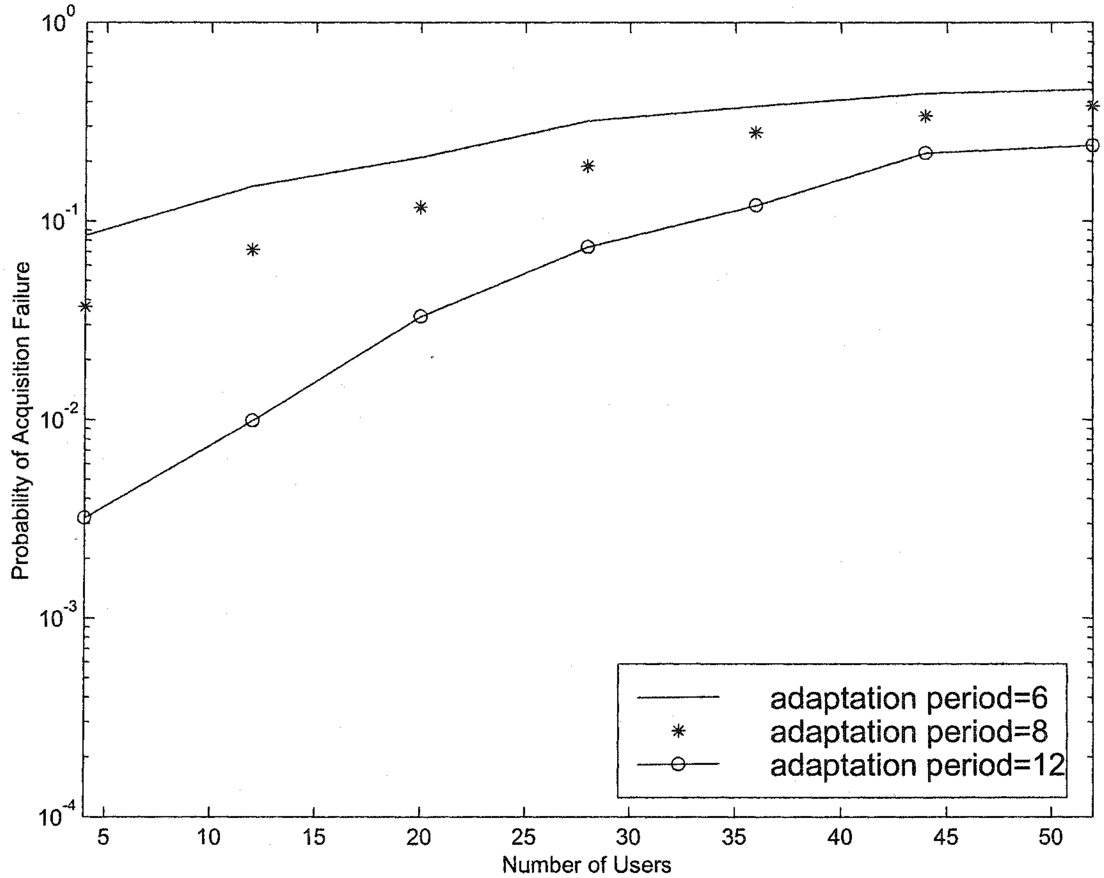

Figure 5.2: Probability of Acquisition Failure as a function of Number of users for different values of Adaptation periods with SNR_c of $-5\ dB$

5.5 Mean Acquisition Time Performance

Here for cell acquisition stage, to minimize the probability of false acquisition, we take 11 samples at 127-chip sequence intervals, after a settling down of twelve 127-chip sequences and only if six or more of samples have the peak at the same location will we declare that the cell has been correctly identified.

Once one of cells has been correctly identified as having correct phase of desired user, the system will go into the phase acquisition stage by loading the subsequence that starts $\lceil \frac{M}{2} \rceil$ from beginning of the identified cell. For phase acquisition stage, 9 samples are taken, at 127-chip intervals, after a settling down time of two chip sequence of 127-chips and if 5 or more samples have the peak located at the same location will the system declare that the correct phase is acquired.

Fig 5.3 shows the mean acquisition time in chips needed to acquire the phase of a 127 chip spreading sequence using our proposed system in case of Multiple Access Interference environment with a maximum of 4 users. Also the performance in case of single user (without MAI) is included to assess the degradation due to the Multiple Access Interference. We can see that for low SNR_c, degradation due to MAI is more than for high SNR_c case which means that code acquisition system is more affected by MAI for low SNR_c environment.

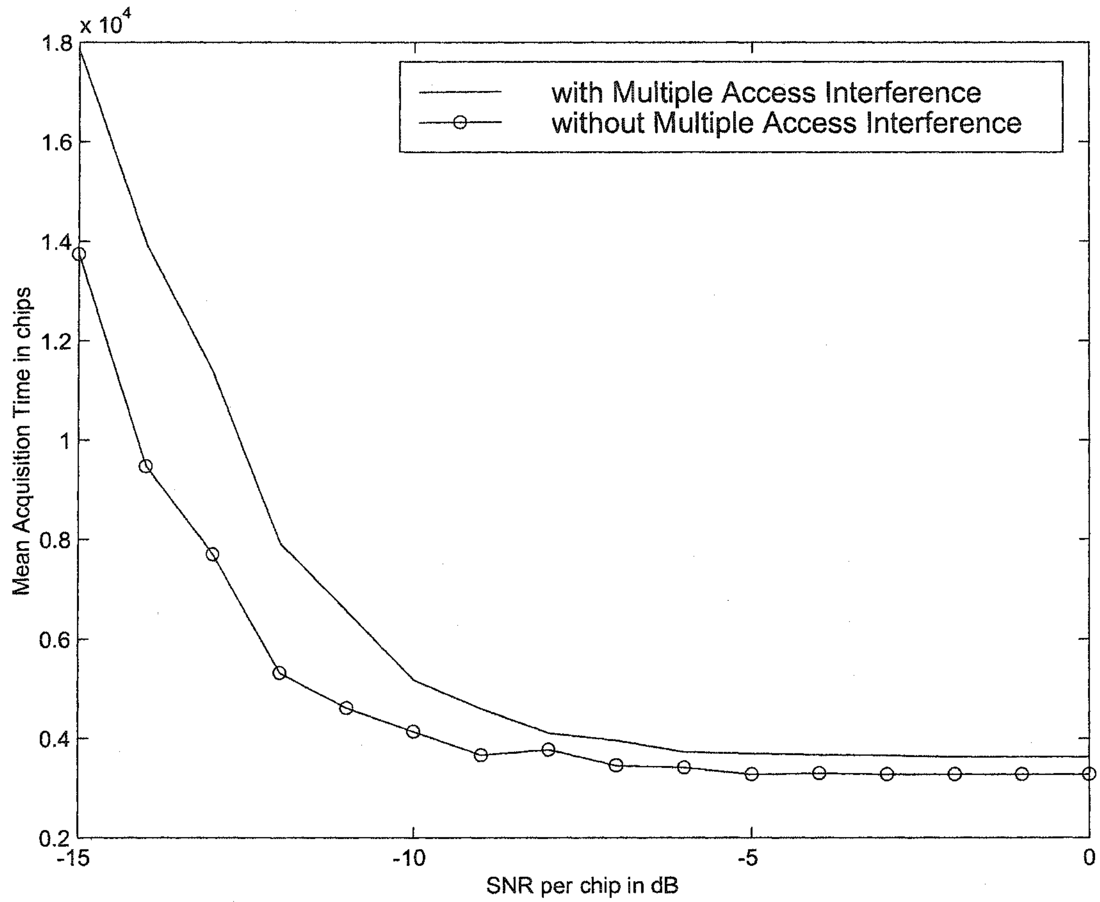

Figure 5.3: Graph of Mean Time to Acquisition in chip periods against SNR_c (signal-to-noise-ratio per chip in dB) in the presence of MAI (4 users))

5.6 Alternate Code Acquisition structure based on System Identification

Previously, we consider the adaptive filter based code acquisition system structure similar to that of Equalization problem. In this section, we will introduce the structure of the system based on system identification problem.

Fig 5.4 shows the adaptive filter based code acquisition system based on the system identification structure. The difference between the two structures is that in case of system identification based code acquisition structure, the desired user spreading sequence is chosen as the input to the adaptive filter, while the received samples are used as adaptive filter reference signal whereas in case of Equalization based structure, desired user spreading sequence is chosen as reference signal while the received samples are chosen as the input to the adaptive filter.

The rest of the system remains the same. Again, LMS algorithm is used to search for minimum of the mean square error and following convergence of tap-weights, maximum of tap-weights indicate the location of the cell where the chip phase delay lies. Then the algorithm switches to Phase acquisition stage to acquire the phase delay itself.

Fig 5.5 shows the comparison of the Mean acquisition time in chips for the two proposed structures for the code acquisition system. The performance of Equalization based structure is slightly better than the one based on system identification structure. The reason is that in case of system identification based structure, we have a noisy reference (i.e; the received samples are taken as reference) which slightly deteriorates the mean acquisition time performance of the system for low SNR conditions.

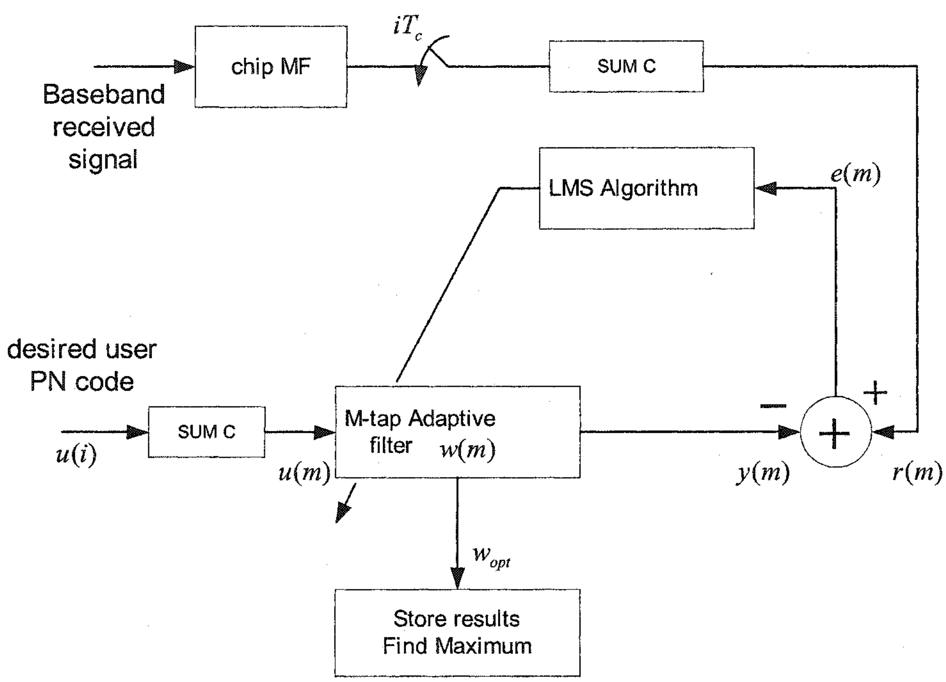

Figure 5.4: Block Diagram of System Identification based Adaptive Acquisition structure in Multiple Access Interference (MAI) Environment

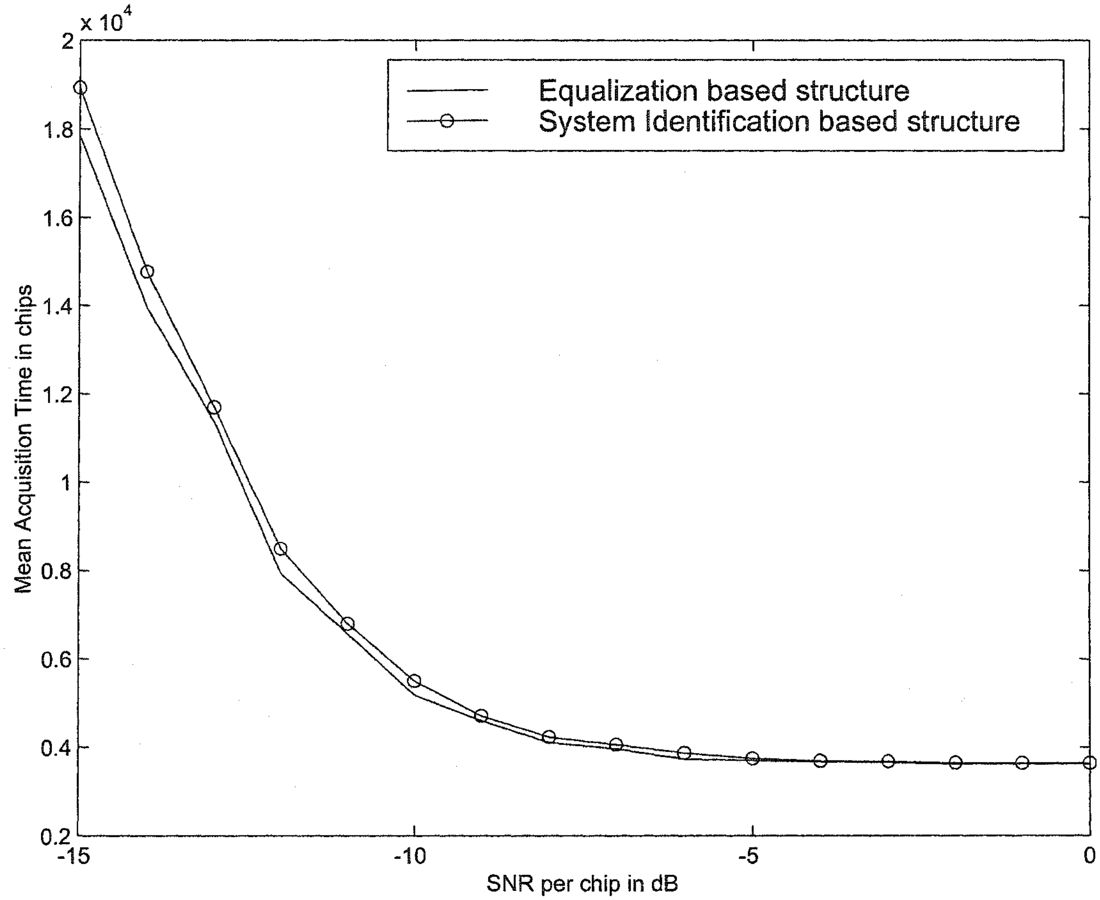

Figure 5.5: Comparison of Mean Acquisition Time in chips for the two structures in MAI Environment

Chapter 6

CONCLUSIONS AND FUTURE WORK

This chapter concludes the thesis by summarizing important observations, contributions and results. It also highlights some future avenues of research that can be originated from the work.

6.1 Conclusions

As we mentioned in earlier chapters that synchronization is essential in spread spectrum systems. Most of the previous work in the acquisition problem concentrate on correlation technique or sequential estimation methods. Recently, an adaptive filter was proposed to function as both acquisition and tracking circuit. This thesis presents the performance of the modified Hybrid Code Acquisition scheme using Adaptive Filters. The inferred conclusions from this research can be compiled in the following points:

- The acquisition mode of operation is described and analyzed in a single-user, AWGN channel model which is commonly used assumption for analyzing acquisition systems. Analytical expressions for both correct cell acquisition and phase acquisition probabilities as well as false alarm error probabilities are derived. Moreover the Mean Acquisition time is obtained by modelling the acquisition process as a Markov chain. It is shown to work better than another adaptive filter acquisition system [37] when signal-to-noise-ratio is low for a 127-chip sequence.

  The performance of the system should be even better for longer spreading sequences, since it needs not step through the cells sequentially searching for the correct phase. The time needed to find the right phase for a sequential system, such as the one in [37], may be shorter, when the right cell is searched, than that for the proposed system. However, sequential systems need to search, on average, half the total number of cells before it gets to the correct phase. For cases where the signal-to-noise ratio is low, the situation is worse, as the system may not detect the phase even when it is searching the correct cell. It therefore needs to step through all cells again, increasing the average time to achieve acquisition very much.

Besides taking a shorter time on average to achieve acquisition when SNR_c is low, the proposed system also requires much fewer computations to achieve synchronization when compared to the one in [37]. Instead of updating the tap-weights every single chip interval, which requires several operations, the proposed system updates the tap-weights once only every cc hip interval during cell identification.

- Simulations for the mean acquisition time performance for the modified hybrid code acquisition scheme based on AF are presented in Pedestrian test environment (taking both flat and frequency selective fading) with PDP (power delay profile) proposed by ITU for Exponential MDP case. It is concluded that the modified system performs better than the one proposed in [39] for low SNR_c conditions in Exponential Multipath delay profile (MDP). Moreover the system also exploits the inherent multipath diversity to acquire the strongest available path. Consequently, frequency selective fading produces lesser degradation than flat fading. We also note that since the gain is larger with Uniform MDP than Exponential MDP due to larger inherent diversity, so degradation is more in case of Exponential MDP than Uniform MDP. Finally we considered a more practical situation of taking varying doppler shifts on the Multipath components and concluded that mean acquisition time will be more if path with higher doppler rate will correspond to more average power than the path with smaller doppler rate.

- Finally the proposed system is analyzed in case of Multiple Access Interference environment. The capacity of the CDMA system is found to be improved significantly by using acquisition strategies such as the modified hybrid scheme based on adaptive filter structure, proposed in this thesis. The proposed system utilizes an LMS-based FIR adaptive filter to estimate the delay offset of the desired user from the tap-weight vector. Probability of cell acquisition failure is derived and simulation and analytical results are compared. It has been shown here that the adaptive filter acquisition system can support the same number of users as that obtained by the BER criterion and thus the degradation suffered by the matched filter based scheme can be avoided. Secondly the degradation due to MAI is less for the low SNR_c environment using the proposed system. Moreover the performance of the system will be more better for longer spreading sequence in the presence of MAI.

6.2 Future Work

This research work is ended up with some new directions in code synchronization for DS-CDMA systems which marks the vitality of the subject. The recommendations for future work can be summarized as:

- The capacity of the CDMA systems is limited by the Narrow-band interference environment which is very paramount in defense communications. The performance of the proposed code acquisition system can be investigated in this interference environment.

- In this thesis, the performance of the system is studied taking short length PN code sequence of period 127. We can also evaluate the performance considering long codes of period 255 or more and it is strong speculation to say that the proposed system will work more efficiently than the other acquisition systems for spreading sequence of long code period.

- The code tracking system can also be investigated in conjunction with the proposed code acquisition system and the parameters of mean hold-in time and mean penalty time can be investigated for the tracking mode of operation.

Appendix A

Review of Adaptive Filter Theory

A.1 Introduction

Adaptive filters have received considerable attention due to their applicability in many different areas such as echo cancelling, adaptive arrays, communication channels equalization, radar, and biomedical engineering. This is caused by their ability to operate in unknown environments and also to track the changes in the input signal characteristics. The theory of adaptive filters is well established [42] and only a brief overview is presented here.

A.2 Principles of Adaptive Filter

An adaptive filter is essentially a digital filter with adjustable coefficients or taps according to a certain criterion. The filter is completely described by the set of coefficients which determine the discrete impulse response of the filter or equivalently its transfer function. A block diagram of an adaptive filter is shown in Figure A.1. It consists of a digital filter with M taps and an adaptive algorithm to modify the coefficients of the filter. The filter is usually realized in a transversal or finite impulse response (**FIR**) structure. Other forms are possible such as the infinite impulse response (**IIR**) or the lattice structure, however, the FIR structure is commonly used because of its simplicity and guaranteed stability.

Two discrete-time signals, $x(n)$ and $d(n)$, are applied simultaneously to the adaptive filter. The signal $x(n)$ is the primary input to be processed by the filter to produce an output, $y(n)$, that is similar in some sense to the desired signal $d(n)$. The filter output is given by

$$y(n) = \sum_{k=0}^{M-1} w_k(n)x(n-k) = \mathbf{w}^T(n)\mathbf{x}(n) \tag{A.1}$$

where $\mathbf{w}(n)$ and $\mathbf{x}(n)$, the weight vector and input signal vector, respectively, are given by

$$\mathbf{w}(n) = \begin{bmatrix} w_0(n) \\ w_1(n) \\ \vdots \\ w_{M-1}(n) \end{bmatrix} \tag{A.2}$$

$$\mathbf{x}(n) = \begin{bmatrix} x(n) \\ x(n-1) \\ \vdots \\ x(n-M+1) \end{bmatrix} \tag{A.3}$$

The filter coefficients are adjusted to minimize some cost function of the error signal produced by the filter which is defined as

$$e(n) = d(n) - y(n) = d(n) - \mathbf{w}^T(n)\mathbf{x}(n) \tag{A.4}$$

One of the most popular cost functions is the mean-square error (**MSE**) criterion, defined by

$$J(n) = E\{e^2(n)\} \tag{A.5}$$

where $E\{.\}$ denotes the statistical expectation operation. The square of the error signal defined in A.4 is given by

$$e^2(n) = (d(n) - \mathbf{w}^T(n)\mathbf{x}(\mathbf{n}))(\mathbf{d}(\mathbf{n}) - \mathbf{w}^\mathbf{T}(\mathbf{n})\mathbf{x}(\mathbf{n}))^\mathbf{T}$$
$$= d^2(n) - 2d(n)\mathbf{x}^T(n)\mathbf{w}(n) + \mathbf{w}^T(n)\mathbf{x}(n)\mathbf{x}^T(n)\mathbf{w}(n) \tag{A.6}$$

Assuming that the input vector, $x(n)$, and the output, $y(n)$, are jointly stationary, the mean squared error (MSE) is given by

$$J(n) = E\{d^2(n)\} - 2E\{d(n)\mathbf{x}^T(n)\}\mathbf{w}(n) + \mathbf{w}^T(n)E\{\mathbf{x}(n)\mathbf{x}^T(n)\}\mathbf{w}(n)$$
$$= \sigma_d^2 - 2\mathbf{p}^T\mathbf{w}(n) + \mathbf{w}^T(n)\mathbf{R}\mathbf{w}(n) \tag{A.7}$$

where $\sigma_d^2 = E\{d^2(n)\}$ is the variance of the desired signal,

$$\mathbf{p} = E\{d(n)\mathbf{x}^T(n)\} \tag{A.8}$$

is the $\mathbf{M} \times 1$ cross-correlation vector and

$$\mathbf{R} = E\{\mathbf{x}(n)\mathbf{x}^T(n)\} \tag{A.9}$$

is the $\mathbf{M} \times \mathbf{M}$ autocorrelation matrix of the input vector. The **MSE** is, therefore, a quadratic function with a minimum J_{min} which is called the Minimum Mean-

Squared Error (**MMSE**). The gradient vector of the **MSE** with respect to $\mathbf{w}$(n) is

$$\nabla(J) = \frac{dJ}{d\nabla} = -2\mathbf{p} + 2\mathbf{R}\mathbf{w}(n)$$
$$= -2\mathbf{E}\{e(n)\mathbf{x}(n)\} \tag{A.10}$$

At the minimum point of **MSE**, the gradient is zero and the filter weight vector has its optimum value given by

$$\mathbf{w}_{opt} = \mathbf{R}^{-1}\mathbf{p} \tag{A.11}$$

The **MMSE** produced by this filter is

$$J_{min} = \sigma_d^2 - \mathbf{p}^T\mathbf{w}_{opt} \tag{A.12}$$

Equation (A.11) is known as the Wiener Hopf equation which requires the knowledge of both $\mathbf{R}$ and $\mathbf{p}$ a priori. It also involves the undesirable matrix inversion operation. Furthermore, for nonstationary signals both $\mathbf{R}$ and $\mathbf{p}$ are time varying and, therefore, $\mathbf{w}_{opt}$ is not fixed. Adaptive algorithms are used to adjust the filter taps to approximate the optimum value given by $\mathbf{w}_{opt}$ without having to compute $\mathbf{R}$ and $\mathbf{p}$ explicitly or perform the matrix inversion.

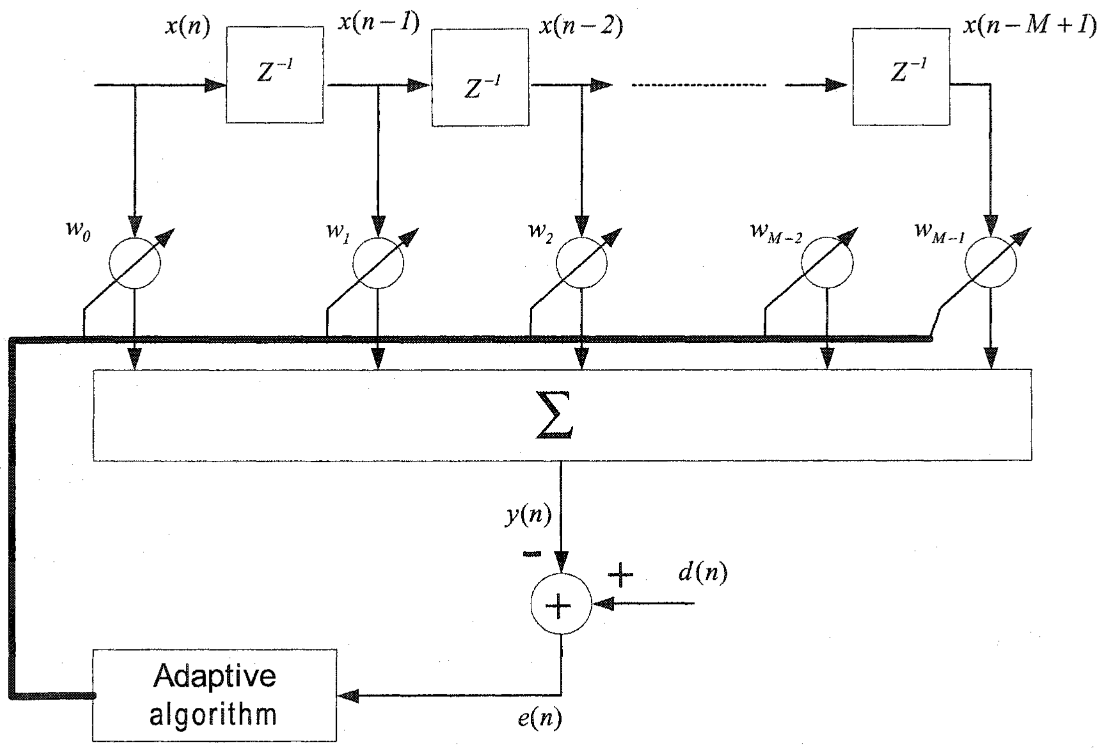

Figure A.1: Block Diagram of an Adaptive Filter

A.3 The LMS Adaptive Algorithm

Several techniques have been proposed to solve the problem of finding the minimum of the MSE performance surface. One of the oldest methods is known as the method of steepest descent which is used to adjust the filter coefficients recursively towards their optimum value. Starting from an arbitrary value, $\mathbf{w}(0)$, which could represent our best guess of the optimum solution, the algorithm computes the gradient vector given by (A.10) and changes the tap-weight vector in an opposite direction to the computed gradient. The coefficients update equation is given by

$$\mathbf{w}(n+1) = \mathbf{w}(n) - \frac{1}{2}\mu\nabla(\mathbf{J}) \tag{A.13}$$

where μ is a positive real-valued constant known as the step size parameter which controls the stability and rate of convergence. Equation (A.13) can be rewritten as

$$\mathbf{w}(n+1) = \mathbf{w}(n) + \mu\mathbf{E}\{e(n)\mathbf{x}(n)\} \tag{A.14}$$

The steepest descent algorithm still requires the knowledge of $\mathbf{R}$ and $\mathbf{p}$. However, by using an instantaneous estimate of $\mathbf{R}$ and $\mathbf{p}$, we obtain an estimate of the gradient vector given by

$$\nabla(J) = -2\mathbf{p} + 2\mathbf{R}\mathbf{w}(n) \tag{A.15}$$

where $P = d(n)\mathbf{x}^T(n)$ and $R = \mathbf{x}(n)\mathbf{x}^T(n)$. Incorporating (A.15) into (A.13), we obtain the simplest Least Mean Square (**LMS**) algorithm such that

$$\mathbf{w}(n+1) = \mathbf{w}(n) + \mu e(n)\mathbf{x}(n) \tag{A.16}$$

The LMS algorithm does not require the knowledge of $\mathbf{R}$ and $\mathbf{p}$ and , but only their instantaneous estimates. The coefficients obtained by the LMS algorithm are estimates of the optimum Wiener filter which improves gradually with time, but it will never reach the optimum value due to the noisy gradient used to adjust the coefficients.

There are two forms of convergence which need to be considered:

1. Convergence in the mean

$$\mathbf{E}\mathbf{w}(n) \longrightarrow \mathbf{w}_{opt} \qquad as \qquad n \longrightarrow \infty \tag{A.17}$$

2. Convergence in the mean square

$$\mathbf{E}\{e^2(n)\} < \infty \qquad as \qquad n \longrightarrow \infty \tag{A.18}$$

The step size, μ, must be selected to ensure both forms of convergence. The limits that must be satisfied by μ are

1. Convergence in the mean

$$0 < \mu < \frac{2}{\lambda_{max}} \tag{A.19}$$

2. Convergence in the mean square

(a)

$$0 < \mu < \frac{2}{\lambda_{max}} \tag{A.20}$$

(b)

$$\sum_{i=1}^{M} \frac{\mu\lambda_j}{2 - \mu\lambda_j} < 1 \tag{A.21}$$

where λ_j,$j = 1, 2, ..., M$, are the eigenvalues of the autocorrelation matrix $\mathbf{R}$.

The mean squared error produced by the **LMS** algorithm has the final value

$$J_{ss} = \frac{J_{min}}{\sum_{i=1}^{M} \frac{\mu\lambda_j}{2-\mu\lambda_j}} \tag{A.22}$$

which is always larger than the **MMSE**.

The learning curve is defined as the plot of the mean-squared error of the **LMS** algorithm with time. It shows how the filter performance is improved in the **MSE** sense as time progresses. It may be approximated by a single exponential with a time constant given by

$$\tau_{mse} \approx \frac{1}{2\mu\lambda_{av}} \tag{A.23}$$

where λ_{av} is the average eigenvalue of $\mathbf{R}$. The time to reach steady state by the exponential function is usually approximated as four times the time constant.

It should be noted that the step size has two conflicting effects. Reducing results in minimization of the steady state error which means a smaller difference between the steady state **MSE** and the **MMSE**. However, longer time is needed for the filter taps to converge to their optimum. The opposite is true, that is, increasing improves the convergence rate but the **MSE** misadjustment at steady state is large. Hence, careful attention has to be given to the selection of μ.

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