國 立 交 通 大 學

電信工程學系

碩 士 論 文

微弱訊號環境下之適應性衛星訊號擷取法

Adaptive GPS Acquisition Technique in Weak Signal Environment

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摘　要

衛星定位系統，在室外有著良好的定位能力。近年來，對於室內衛星訊號擷取的需求，與日遽增，於是在本論文中，我們提出了一種新的，混合式的訊號擷取方法，使得在微弱訊號的環境下，也能使接收成功率大幅度提升。此種方法結合了同步積分及差分同步積分，對於接收到的訊號，作了適度運算，藉由找出導航訊號中位元反轉的位置，適度地在同步積分及差分同步積分間切換，以期達到使用較少的接收訊號，就能完成訊號擷取的動作之效果。對於偵測結果的判定，我們使用了可調的判斷門檻，可以依照外在的訊號對雜訊比作適當地調整，以達到增加擷取成功率的效果。

Adaptive GPS Acquisition Technique in Weak Signal Environment

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Abstract

In recent years, there has been increasing demands in providing precise location estimation in weak signal environment (e.g. in urban area or inside a building). Conventional GPS acquisition techniques are considered to be adequate on positioning capability in outdoor environment. However, most of them are not satisfactory for applications with weak signals. It has been studied that a longer time of data integration is required in order to provide sufficient data for acquisition under weak signal circumstances. In this thesis, the detail about the techniques to acquire GPS signals in weak signal environment is introduced and an adaptive GPS acquisition technique is proposed, which adaptive adjusts the detection threshold and the length of data integration for signal acquisition. The location of the navigation bit-transition is perceived within the acquisition data in order to adaptively determine either the coherent or the differential coherent combining method should be used. The environmental SNR (Signal-to-Noise Ratio) is estimated and the acquisition detection threshold is adaptively adjustable according to it. The simulation results show that the proposed adaptive acquisition technique outperforms both the non-coherent combining and the differential-coherent combining algorithms. As the SNR value of the incoming signal decreases, the effectiveness of the proposed adaptive scheme can be observed.

誌　謝

終於將畢業論文完成了，碩士生涯也將到此結束。這兩年內，體會了很多，也學習到很多，都將成為我一輩子珍惜的回憶。謝謝指導教授方凱田教授，沒有他的用心，就不會有這篇看起來還頗有架勢的論文，我也不會覺得這麼充實，從老師身上學了很多，希望以後還有繼續向他學習的機會。謝謝口試委員吳文榕教授及王國禎教授，在口試時提供了許多寶貴的建議，使這篇論文能夠更加完整。

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Contents

1 Introduction **5**

2 Related Work **8**

2.1 GPS system 10

2.2 Acquisition Process 12

2.2.1 Time-domain Acquisition 13

2.2.2 Acquisition with FFT 15

2.2.3 Averaging Correlation 17

2.2.4 Modified C/A Code 19

2.3 A-GPS 20

2.4 Summing Over Multiple Blocks 23

2.4.1 Even-Odd Block Combining 24

2.4.2 Three Combining Schemes 24

3 The Proposed Adaptive Coherent Combining (ACC) Scheme **29**

3.1 Motivation of the ACC Scheme 29

3.2 Carrier to Noise Power Estimator 30

3.2.1 Theory of C/N0 Estimation 32

3.2.2 Introduction to PID 33

3.3 Adaptive Peak Detection Threshold (PDT) 34

3.4 Bit-Transition Detector 35

3.5 Procedures of the ACC Scheme 36

4 Performance evaluation **39**

4.1 Simulation Parameters 39

4.2 Fading Channel 39

4.3 Simulation Results 42

5 Conclusions **48**

List of Figures

2.1 A Fundamental GPS Receiver 12

2.2 Concept of the Two-dimensional C/A Code Search 13

2.3 Illustration of Acquisition with Periodic Correlation 17

2.4 The 5 Sequences of Average Steps 19

2.5 Assisted-GPS Concept 22

2.6 Acquisition with Narrow Range 22

2.7 Concept of Summing Over Multiple Blocks 24

2.8 Even Odd Integration 25

2.9 Receiver with Combining Scheme 26

3.1 The Effect of the Navigation Bit-Transition in the CC Scheme 31

3.2 The C/N0 Estimator 32

3.3 The Flow Chart of the Proposed ACC Scheme 35

4.1 The Peak Detection Threshold vs The SNR Value 40

4.2 The Total Integration Time vs the SNR Value (with 95% of Detection Probability) 44

4.3 The Detection Probability vs the SNR Value (with 50 ms of Total Integration Time) 44

4.4 The Total Integration Time vs the SNR Value with Center Frequency Uncertainty (with 95% of Detection Probability) 45

4.5 The Detection Probability vs the SNR Value with Center Frequency Uncertainty (with 50 ms of Total Integration Time) 45

4.6 The Estimation Curve of Open Loop Estimator and Close Loop Estimator in -30dB 46

4.7 The Estimation Curve of Open Loop Estimator and Close Loop Estimator in -35dB 46

4.8 The Estimation Curve of Open Loop Estimator and Close Loop Estimator in -39dB 47

4.9 The Detection Probability of Known SNR and Estimated SNR 47

Chapter 1

Introduction

Acquisition is the first and the most time-consuming process of the GPS receiver. There are many studies about how to improve the acquisition precision and to make it more time-efficient. Acquisition process contains a two-dimensional search, one is the C/A code phase and the other is the Doppler frequency shift caused by the relative movement of the satellites and the receiver. The acquisition process is usually implemented by taking correlation of the received signal and the locally generated signal. The acquisition process is to take correlation of these possible combinations of received signal and local generated code by taking each possible code phase and frequency shift into consideration, the one with a obviously highest correlation peak represents the correct estimate of received code phase and frequency shift. When the acquisition process is finished, the estimated code phase and frequency shift is passed to the tracking loop and the more precise prediction is executed. Serial search method are the most used and traditional acquisition strategy in hardware receiver implementation, but most of them are not time-effective. The general solution of hardware GPS receivers is to use several correlators to calculate the correlation in parallel. A new correlation method using FFT has been proposed in 1991, which enhances the correlation speed tremendously. But based on the hardware constraint on the FFT blocks, the acquisition using FFT is usually implemented in software simulation. Another solution to decrease the acquisition time is to adopt the Assisted GPS (A-GPS) method. The concept of the A-GPS method is to provide some information to the GPS receiver. The total uncertainty range of acquisition can be decreased and the acquisition time is reduce. The additional information is provided by a reference network that monitor the whole satellites constellations and calculate the predictions of ephemeris and Doppler frequency shift.

Besides the issue of increasing the computation capability, there has been increasing demands in providing precise location estimation in weak signal environment (e.g. in urban area or inside a building). Conventional GPS acquisition techniques are considered to be adequate on positioning capability in outdoor environment. However, most of them are not satisfactory for applications with weak signals. It has been studied that a longer time of data integration is required in order to provide sufficient data for acquisition under weak signal circumstances. When integrating the received signals for more than one periods, there can exist problems causing the degeneration of correlation results. The problem is resulted from the inherent characteristic of the GPS signals. The GPS signal is BPSK DSSS(Directed-Sequence Spread Spectrum) transmitted. Therefore, the signal is the product of data sequence and the spread code. The data sequence is 20 milliseconds per bit, and the spreading C/A code is 1 millisecond per period. If the integration time exceeds 20 milliseconds, there may be data bit transition within the integration. Therefore, even longer integration period can not improve the probability of finding the correct correlation peak. Several methods has been proposed to resolve this problem and make the multi-period integration useful in weak signal environment, including the Even-Odd blocks combining method, and the differential coherent combining scheme.

In this thesis, an adaptive GPS acquisition technique is proposed, which adaptive adjusts the detection threshold and the length of data integration for signal acquisition. In general, a GPS acquisition block uses a predefined detection threshold, which obeys the Neyman-Pearson criterion to obtain the maximized detection probability in the constraint false alarm rate. If we can estimate the environmental SNR and adaptively adjust the detection threshold in the acquisition process then we can significantly improve the sensitivity of GPS receivers. We modified an existed carrier to noise power estimation method in order to have better estimation result. The location of the navigation bit-transition is perceived within the acquisition data in order to adaptively determine either the coherent or the differential coherent combining method should be used. The bit-transition detector is an important component in our algorithm. With the knowledge of the occurrence of bit-transition, we can avoid the degeneration problem in the multi-period integration. The simulation results shows that the proposed adaptive acquisition technique outperforms both the non-coherent combining and the differential-coherent combining algorithms. As the SNR value of the incoming signal decreases, the effectiveness of the proposed adaptive scheme can be observed.

The remainder of this thesis is organized as follows. Chapter 2 introduces the related works, including the fundamental theory of the GPS system, the acquisition process and the techniques to enhance the acquisition capability in weak signal environments. The proposed Adaptive Coherent Combining (ACC) algorithm is presented in Chapter 3. Chapter 4 illustrates the performance evaluation of the proposed scheme in simulations. Chapter 5 draws the conclusion.

Chapter 2

Related Work

The Global Positioning System (GPS), a satellite-based locationing system, has been studied and effectively implemented in outdoor environment for many years [1] [2]. The signal acquisition process is the first required functional block within a GPS receiver. It detects the occurrence of the GPS signal after the incoming signals are received by the antenna. The acquisition process of a GPS signal involves a two-dimensional search problem. Both the phase of the C/A code and the Doppler offset of the received signal have to be determined simultaneously. The time domain serial search algorithms are studied to be the simplest schemes for acquisition, but most of them are considerably time-consuming. A Tone search detector is proposed in [3] [4] to effectively reduce the processing time for time domain serial search. Conventional acquisition schemes are mostly hardware-based approaches, which include both the time domain serial search [2] [3] and the frequency domain FFT methods [2] [5] - [8].

The FFT methods significantly speed up the acquisition process comparing with the time domain schemes. However, the limitation on the number of FFT components within the receiver hardware make it difficult to acquire further enhancement on the processing time. The software-based GPS receiver has been investigated in recent years [2] [9]. Due to the advancement on the processing capability from the DSPs and the embedded processors, the advantages of adopting the software-based acquisition approach is revealed, e.g. system modularity and flexible algorithm design.

There is a great amount of attention on exploring the techniques for location estimation in weak signal environment. It has been investigated in [10] that the attenuation and interference to the received GPS signal are severe in stringent circumstances. Different types of technique have been proposed to overcome the problem, which can be categorized into the Assisted-GPS (A-GPS) methods [11] - [14] and the combining schemes [2] [15] - [20] and others [21] [22] [23]. In order to accommodate the indoor acquisition environment, the simulation will be implemented in Rayleigh fading channel [24]- [30]. Several receiver design methods are proposed to improve the sensitivity of receivers in the multi-path environment [31] [32].

The A-GPS methods primarily utilize the messages coming from the cell-based systems to assist the acquisition of the received GPS signal. Alternatively, the multi-period combining schemes effectively acquire the GPS signal without the assistance from the external messages. The block acquisition scheme proposed in [15] alleviates the navigation bit-transition problem using the Coherent Combining (CC) method. However, a longer integration time is required for the acquisition process. An advanced acquisition method utilizing multi-period combining and can avoid the problem caused by navigation data bit-transition is introduced in [33]. The Differential Coherent Combining (DCC) scheme is proposed as in [16] - [18]. It compromises between the CC and the Non-Coherent Combining (NCC) methods with reduced squaring loss during the integration process.

In this thesis, an Adaptive Coherent Combining (ACC) scheme is proposed for GPS acquisition under weak signal environment. The main concept of the ACC scheme is to provide a mechanism to detect the position of the navigation bit-transition within the received GPS signal. The CC method is utilized in the first stage before the navigation bit-transition occurs; while the DCC scheme is adopted as the combining method after the bit-transition has been detected. The individual benefit of using either the CC or the DCC method can be maximized by exploiting the proposed ACC scheme. Moreover, the peak detection threshold within the ACC scheme is designed to be adjustable w.r.t. the levels of surrounding noise. We modified an existed carrier to noise estimator [34] [35] [36] to enable a faster convergence of the estimated value.

It will be shown in the simulation results that the proposed ACC scheme can provide both (*i*) a higher percentage of the detection probability and (*ii*) a smaller amount of the integration time, comparing with the DCC and the NCC methods.

The remainder of this chapter is organized as follows: Section 2.1 introduces the history and basic information of the GPS . Section 2.2 investigates the acquisition process of the GPS receiver, including time domain acquisition and frequency domain acquisition. Two major methods to enhance the sensitivity of the GPS receiver will be illustrated in Section 2.3 and Section 2.4. Section 2.3 illustrates the A-GPS and Section 2.4 introduces the multi-period integration method.

2.1 GPS system

The GPS (Global Position System) program was first developed in Decenber 1973. The first satellite was launched in 1978. In August 1993, there are total 24 satellites in orbit and in December of the same year the initial operational capability has been established. In February 1994, the Federal Aviation Agency (FAA) announced that the GPS was ready for aviation use. There are total of 24 satellites of the GPS, which are divided into six orbits and each orbit has four satellites. Each orbit makes a 55-degree angle with the equator and the orbits are separated by 60 degrees from each other and therefore cover the complete 360 degrees. These designs make at least 4 satellites visible in the sky anytime in most of the locations. Several frequencies make up the GPS electromagnetic spectrum:

- L1 (1575.42 MHz): The L1 frequency band carries a coarse-acquisition (C/A) code as well as an encrypted precision P(Y) code.

- L2 (1227.60 MHz): The L2 frequency band usually carries the P(Y) code. The encryption keys required to directly use the P(Y) code are tightly controlled by the U.S. government and are generally provided only for military use. The keys are changed on a daily basis. In spite of not having the P(Y) code encryption key, several high-end GPS receiver manufacturers have developed techniques for utilizing this signal (in a round-about manner) to increase accuracy and remove error caused by the ionosphere. Recognizing the civilian need for increased accuracy, "modernized" IIR-M (IIR-14 (M) and later) satellites carry a civilian signal interleaved with an improved military signal on both the L1 and L2 frequencies.

- L3 (1381.05 MHz): The L3 frequency band carries the signal for the GPS constellation's alternative role of detecting missile/rocket launches (supplementing Defense Support Program satellites), nuclear detonations, and other high-energy infrared events.

- L4 (1841.40 MHz): The L4 frequency band is studied for additional ionospheric correction.

- L5 (1176.45 MHz): The L5 frequency band is proposed for use as a civilian safety-of-life (SoL) signal. This frequency falls into an internationally protected range for aeronautical navigation, promising little or no interference under all circumstances. The first Block IIF satellite that would provide this signal is set to be launched in 2008.

There are basically two types of PN (pseudo noise) code of the GPS signals: the C/A (coarse/acquisition) and the P (precision) codes. The P code is modified by a Y code, which is therefore referred to as the P(Y) code. The C/A code is for civilian used and the P(Y) code is primarily used by the military. The following discussion in the thesis will only consider the signals modified by the C/A code. The GPS signal is a bi-phase shift keying (BPSK) direct sequence spread spectrum (DSSS) signal. In general, a code division multiple access (CDMA) signal is a spread spectrum signal. All of the signals use the same center frequency in the system. The signals are modulated by a set of orthogonal codes. In GPS, there will be 24 orthogonal codes for each satellite individually. In order to acquire the desired signal, the spreading code of the signal must be used to correlate with the received signal. Since all the signals use the same center frequency, there is a possibility that the signals will interfere with each other. In order to avoid this situation, all the signals are transmitted with the same transmission power. The satellite transmits two kinds of information to the receiver: one is the ephemeris and the other is almanac. The ephemeris includes the information whether

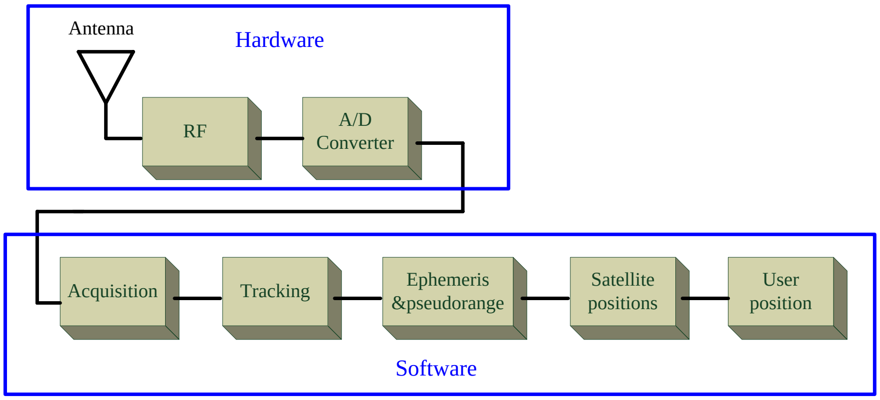

Figure 2.1: A Fundamental GPS Receiver

the satellite is healthy or not, and the date and time to inform the receiver to update. The almanac data tell the receiver about the orbits information and the positions of each satellites. A GPS receiver can receive signals from 4 to 11 satellites at the same time. There are usually two approaches to utilize these satellites, one is to use all the satellites to calculate the user position, and the other approach is to choose only four satellites to fix the position. Because additional information can be obtained, usually all the satellites will be used. After received the information from the satellites, the positions of satellites will be first calculated, and the user position can be calculated from the triangulation method.

2.2 Acquisition Process

A fundamental GPS receiver is shown in Fig. 2.1. After demodulation and digitizing, the base-band received signal is passed to the acquisition and tracking process to extract the navigation data. The signal acquisition process is the first required functional block within a GPS receiver. It detects the occurrence of the GPS signal after the incoming signals are received by the antenna. The received signal is despread by synchronizing with a local Pseudo Random Noise (PRN) code, i.e. the Coarse/Acquisition (C/A) code with 1023 chips on the L1 carrier. During the acquisition process, both the coarse code phase and the Doppler

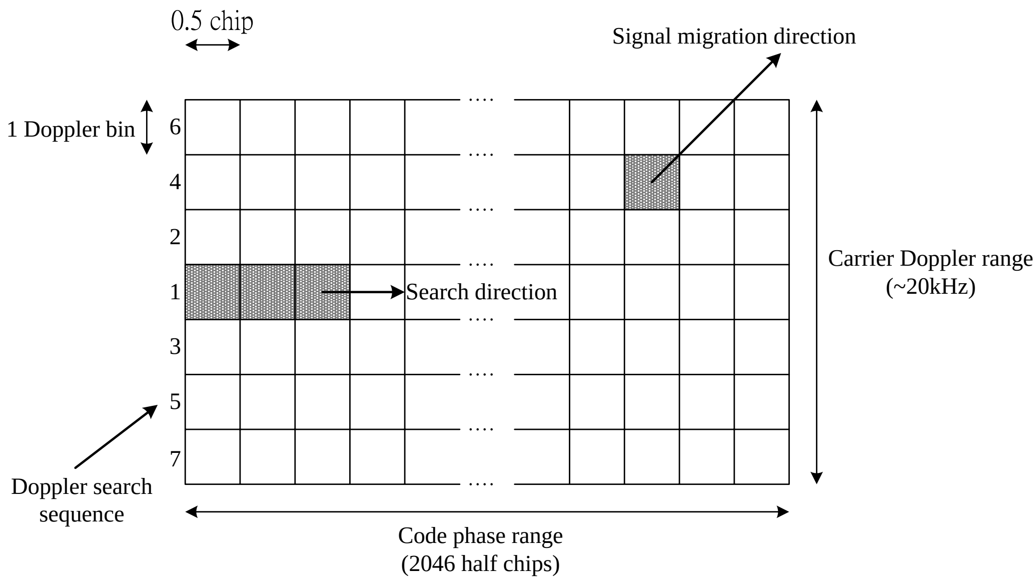

Figure 2.2: Concept of the Two-dimensional C/A Code Search

frequency shift can be estimated, and the results are delivered to the receiver's tracking loop for further processing. The acquisition process of a GPS signal involves a two-dimensional search problem. Both the phase of the C/A code and the Doppler offset of the received signal have to be determined simultaneously.

Conventional acquisition schemes are mostly hardware-based approaches, which include both the time domain serial search [2] [3] and the frequency domain FFT methods [2] [5] [6]. The time domain serial search algorithms are studied to be the simplest schemes for acquisition, but most of them are considerably time-consuming.

2.2.1 Time-domain Acquisition

It is assumed that the received GPS signal is digitized with a 5 MHz sampling rate, which results in 5000 samples of each C/A code with 1 ms of period in length. As mentioned in the previous section, the correlation between the received GPS signal and the local code involves a two-dimensional search process, as shown in Fig. 2.2. The x-direction of the two-dimensional search relates to the 5000 possible code shift after sampling. The frequency shift due to the

Doppler effect are considered in the y-direction of the two-dimensional search. The frequency range of the received signal with the Doppler shift is assumed as 1250 $\pm$ 10 kHz, where 1250 kHz corresponds to the center frequency. There are a total of 21 frequency components with the 1-kHz step size for the frequency range in consideration. Consequently, the two-dimensional search process can be represented as $M = \{m(i, j) \,|\, i, j \in \Re, 1 \le i \le p, 1 \le j \le q\}$, with $p = 5000$ and $q = 21$ as described above. The position with the correct code phase and frequency shift are denoted as $m(i_c, j_c)$, with $1 \le i_c \le p$ and $1 \le j_c \le q$. The received GPS signal $r(t)$ can be formulated as

$$r(t) = \sqrt{A} \cdot N(t)s(t - \tau)e^{j2\pi f_d t} + n(t) \tag{2.1}$$

where

1. A represents the power of the received GPS signal.
2. $N(t)$ denotes the time series of the navigation data, which may change its value every 20 ms.
3. τ is the unknown code phase of the received signal.
4. f_d denotes the Doppler shift.
5. $n(t)$ is the assumed AWGN with the power spectral density of N_0.

The correlation between the received GPS signal $r(t)$ and the locally replicated C/A code for a specific satellite $s(t)$ can be obtained as

$$y^{(n)}(i, j) = \int_{(n-1)T}^{(n)T} r(t)[s(t - \tau_i)e^{j2\pi f_j t}]^* \, dt \tag{2.2}$$

where

1. T represents the code period.
2. τ_i is the code shift of the i^{th} correlator.

3. f_j represents the adjustable frequency shift.

4. $y^{(n)}(i, j)$ denotes the $(i, j)^{th}$ correlation results with $s(t - \tau_i)$ at the n^{th} integration period, which corresponds to a matrix element $m(i, j)$ in the two-dimensional search process.

2.2.2 Acquisition with FFT

To overcome the shortcoming of time-domain acquisition, a FFT (Fast Fourier transform) acquisition method has been studied [2] [5] [6]. If an input signal passes through a LTI (linear time-invariant) system, the output signal can be calculated by the time domain convolution or the frequency domain Fourier transform. The convolution between the received signal and the impulse response of LTI system can be represented by

$$z(n) = \sum_{m=0}^{N-1} r(m)h(n-m) \tag{2.3}$$

Because the discrete operation is periodic, the time shift of $h(n-m)$ is periodic. Take the DFT (Discrete fourier transform) of the above equation, we can obtain:

$$Z(k) = \sum_{n=0}^{N-1}\sum_{m=0}^{N-1} r(m)h(n-m)e^{-j2\pi kn/N} \tag{2.4}$$

$$= \sum_{m=0}^{N-1} r(m)[\sum_{n=0}^{N-1} h(n-m)e^{-j2\pi(n-m)k/N}]e^{-j2\pi mk/N} \tag{2.5}$$

$$= H(k)\sum_{m=0}^{N-1} r(m)e^{-j2\pi mk/N} = R(k)H(k) \tag{2.6}$$

As shown in Fig. 2.3, the correlation process of GPS acquisition is not a linear correlation. Because the received signal is multiplied by a periodic PN (Pseudo noise) code which is periodic, it make the correlation a circular correlation. Similar to the circular convolution equation obtained above, the FFT equation of circular correlation can be formulated to be used in the acquisition process. A correlation between the digitized received signal $r(n)$ and the digitized locally generated signal $s(n)$ can be written as

$$y(n) = \sum_{m=0}^{N-1} r(m)s(n+m) \tag{2.7}$$

If we perform the DFT of $y(n)$, the result becomes

$$Y(k) = \sum_{n=0}^{N-1}\sum_{m=0}^{N-1} r(m)s(n+m)e^{-j2\pi kn/N} \tag{2.8}$$

$$= \sum_{m=0}^{N-1} r(m)[\sum_{n=0}^{N-1} s(n+m)e^{-j2\pi(n+m)k/N}]e^{j2\pi mk/N} \tag{2.9}$$

$$= S(k)\sum_{m=0}^{N-1} r(m)e^{j2\pi mk/N} = S(k)R^{-1}(k) \tag{2.10}$$

where R^{-1} represents the inverse DFT. The above equation can also be written as

$$Y(k) = \sum_{n=0}^{N-1}\sum_{m=0}^{N-1} r(n+m)s(m)e^{-j2\pi kn/N} \tag{2.11}$$

$$= S^{-1}(k)R(k) \tag{2.12}$$

If the s(n) is real, $s^*(n)$=$s(n)$, which is the complex conjugate. With this relation, the magnitude of $Y(k)$ can be represented as

$$|Y(k)| = |S^*(k)R(k)| = |S(k)R^*(k)| \tag{2.13}$$

This relationship can be used to find the correlation of the received signal and the locally generated signal. The above equation provides a periodic correlation and this is the desired procedure of GPS acquisition. With this equation, we can perform the acquisition process in frequency domain which is faster than acquisition in time domain. The steps of acquisition using FFT methods can be listed in the following:

1. Take the FFT of the 1 ms received data $r(n)$ and convert it into frequency domain as $R(k)$, where n and k range from 0 to 4999.

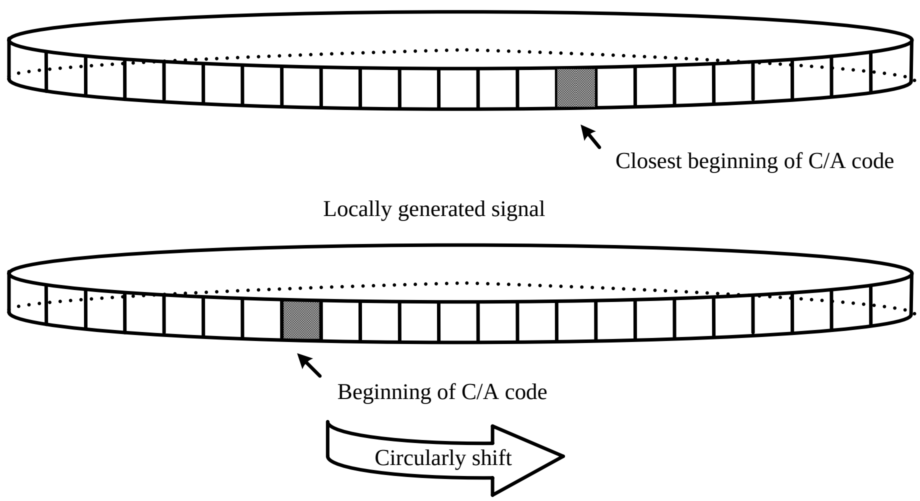

Figure 2.3: Illustration of Acquisition with Periodic Correlation

2. Take the complex conjugate of $R(k)$ and obtain $R^*(k)$.

3. Generate 21 local codes $s_{sl}(n)$ where $l = 1, 2, ..., 21$. $s_{sl} = C_s \cdot e^{j2\pi f_l t}$, where C_s is the C/A code of satellite s.

4. Take the FFT of $s_{sl}(n)$ to transform it to the frequency domain as $S_{sl}(k)$.

5. Multiply $R^*(k)$ and $S_{sl}(k)$ point by point and obtain the result as $Y_{sl}(k)$.

6. Take the inverse FFT of $Y_{sl}(k)$ to transform it into time domain as $y_{sl}(n)$.

7. Find the absolute value $|y_{sl}(n)|$. There are total of 105,000 $(21 \times 5,000)$ points of $|y_{sl}(n)|$.

8. The maximum of $|y_{sl}(n)|$ in the ith code phase and the jth frequency bin represents the beginning location of C/A code.

2.2.3 Averaging Correlation

The sampling rate of the GPS signal is 5 MHz, so every 1 ms received signal will be 5000 points, if we want to implement the acquisition process using FFT method, a 5000-point FFT correlator will be needed. But obviously it is difficult and not economic to build a 5000-point FFT correlator with today's technology. A new method has been proposed to improve this problem. In order to use cheaper and simpler FFT blocks, the new method called "averaging correlation" is introduced [6]. The C/A code rate is 1023 MHz and the GPS sampling rate is 5 MHz, therefore each C/A code chip will be sampled 4 to 5 times. If we make consecutive 4 or 5 points averaged into one chip, the received signal can be recovered back to 1023 chips per millisecond. Therefore only 1023 point FFT blocks are required, which are easier to be implemented. The averaged chip is similar to a chip of C/A code. Since the signal cannot be observed in the forms of square waves, there is not enough information to determine whether 4 or 5 points should be grouped together and averaged. Although this accurate position of chips cannot be exactly found, we can roughly estimate them using several time averaging searches. Among the first consecutive 5 samples, one of them must be the first sample of a chip since no chip contains more than 5 samples. If this sample is regarded as the beginning position of the first chip, the relative position of the 1023 chips and all the samples can be determined and recovered completely. Although the beginning point will not be the exact starting position in normal cases, it is a fairly good approximation. Thus the problem becomes to search for the beginning sample within the 5 ones at the beginning of the samples, as shown in Fig. 2.4 . Beginning with the correct position, the averaged sequence of chips is close to the original one and the correlation peak of this sequence and the local C/A code is higher than the others. The 5 sequences beginning with the first 5 samples respectively are tried and one of the 5 sequence would be chosen as the starting point of chips. The correlation of C/A code and these 5 sequences respectively would produce 5*1023 values. With the 5 averaged sequences and 1023 chip C/A code, the correlation computation needs to be done 5 times. But 5 times 1023-point FFT is still cheaper than 1 time 5000-point FFT, no mater in hardware implementation or software simulation. The simulation results in [6] shows that the averaging correlation can achieve the same acquisition successful rate with simpler and cheaper 1023-point FFT blocks.

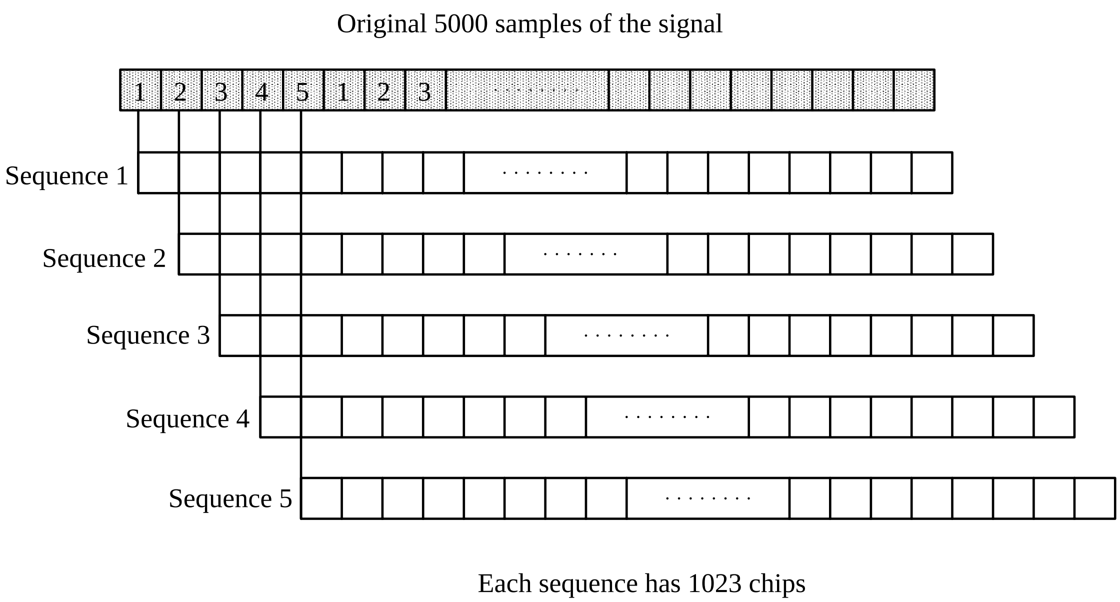

Figure 2.4: The 5 Sequences of Average Steps

2.2.4 Modified C/A Code

Following the averaging correlation, a modified correlation scheme has been introduced [7]. As mentioned in the previous subsection, 1023-point FFT blocks is utilized in the averaging correlation scheme. Since 1023 is not the order of 2, the complexity of FFT computation is increased. In order to overcome this inefficiency, some modifications are conducted. Since the period of the C/A code is 1023 chips, there will be 1023 points for each 1 millisecond received signal. A zero-padding-based solution was introduced in [8]. The signal energy loss on average was acceptable, but the necessary hardware and the total computation time are beyond a certain acceptable level which makes it useless. One possible solution was represented in [7] using available Xilinx's 1024-point FFT core. The new method can be implemented much easier than the 5000-point FFT based method and has a rather good performance. The size incompatibility of 1023 chips C/A code and the available 1024-point FFT core can be solved by changing the averaging rate, or the down-sampling rate from 1023 to 1024. So the 5000 samples will be down sampled to 1024 points. The same procedure will be done to the locally generated signal. Therefore, the locally generated signal will be first sampled by the

5 MHz sampling rate and down sampled to 1.024 MHz, which makes the 1024 points become 1 millisecond signal. The resulting averaging correlation will use 1024 averaged samples and 1024-point modified C/A code. The same as the 1023-point averaging correlation, the modified 1024-point correlation needs to be applied five times with five consecutive starting samples. Within the five correlation results, the one which has the largest peaks keeps the important information necessary for acquisition. Different from normal C/A code acquisition, the correlation peak of the modified C/A code is not based on equilateral triangle, but it is still good enough to estimate the code phase roughly for acquisition consideration. One thing need to be noticed is that each chip of the modified C/A code is 1/1024 millisecond instead of 1/1023 millisecond. Therefore, while estimating the code phase, this change should be taken into consideration.

2.3 A-GPS

Assisted GPS, or A-GPS, is a technology that uses an assistant server to shorten the time needed for determining a location using the GPS system. It is useful in urban areas, where the user is located in "urban canyons", under heavy tree cover, or even indoors. It has become more popular for location estimation and is commonly associated to Location Based Services (LBS) over cellular networks.

The development of these services is fuelled, in part, by the U.S. Federal Communications Commission's E911 mandate requiring the position of a cell phone to be available to emergency call dispatchers. General GPS receivers take several minutes to search and acquire the satellites especially in the cold start situation. Since the receiver does not have any information about the possible code phase and frequency shift. In the warm start situation, the acquisition speed will be faster because the receiver just need to search the possible range of the received signal. In regular GPS networks, there are only satellites and GPS receivers involved. In the A-GPS networks, the receiver, being limited in processing power and normally under less than ideal locations for position fixing, communicates with the assistance server that has high processing power and access to a reference network. Since the A-GPS receiver and the assistance server share tasks, the process is quicker and more efficient than regular GPS, albeit dependent on cellular coverage. The concept of the A-GPS system is shown in Fig. 2.5. Besides the time-consuming problem, general GPS receivers cannot acquire the satellites if the received signals are weaker than the outdoor minimum of about -130 dBm. Imagine that if we can provide some information to a GPS receiver while it performs acquisition process, the total search range can be reduced, as shown in Fig. 2.6. Therefore the required acquisition time will be reduced. The concept of the A-GPS [11] is to provide the GPS receiver some information about which frequency bins to search as mentioned above. The total searching range can be reduced and also enhance the GPS acquisition sensitivity. The aided information has traditionally been provided by distributing the broadcast satellite ephemeris, (or derivative information, such as expected Doppler frequencies) over an alternative data channel. This idea was first proposed by Taylor and Sennot (1981) [11], and has been applied in many variations for almost 25 years. However, the broadcast ephemeris is only valid for an average of three and a maximum of four hours. To provide a more precise information of satellites, infrastructures like reference network are needed. This network tracks the GPS satellites all the time. With this information, orbit models can be created that can be remained good for days in the future. These orbit models provides A-GPS aiding data that has usability time span greater than broadcast ephemeris. Ordinarily, a standard GPS device needs to have a clear line-of-sight to at least four GPS satellites before it can calculate its position. In addition, it requires enough processing power to transform the data streams from the satellites into a position. Using the A-GPS system, the cell tower will receive the signal from several satellites and perform the calculations. As a consequence the phone will relay the GPS signal that it receives to the tower. The A-GPS system has many advantages, including better accuracy, availability, coverage and shorter time-to-first-fix. The disadvantages of the A-GPS system includes: (*i*) network assistance increases signaling load; (*ii*) inter-operability between network and mobiles requires additional standards and delaying deployment.

The following items list several important steps of the operational principle of an A-GPS receiver:

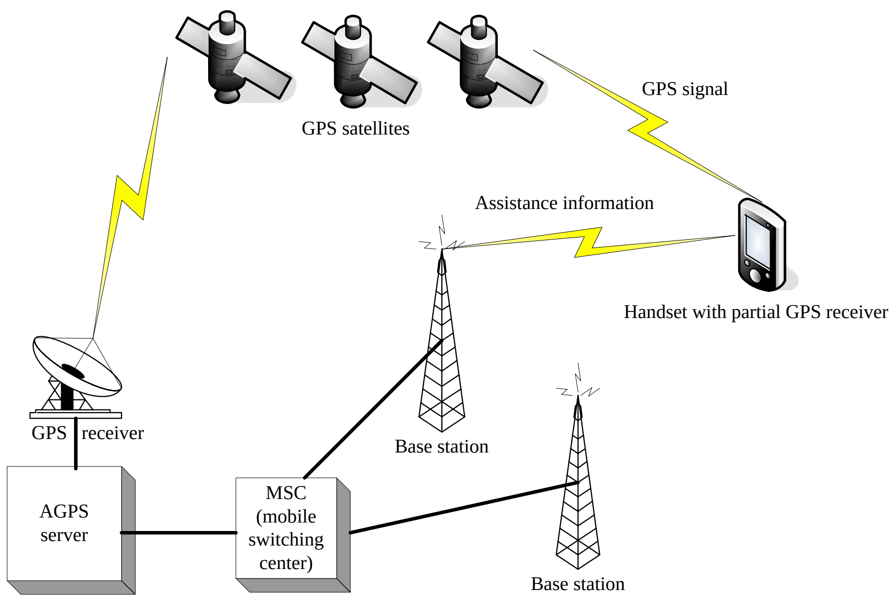

Figure 2.5: Assisted-GPS Concept

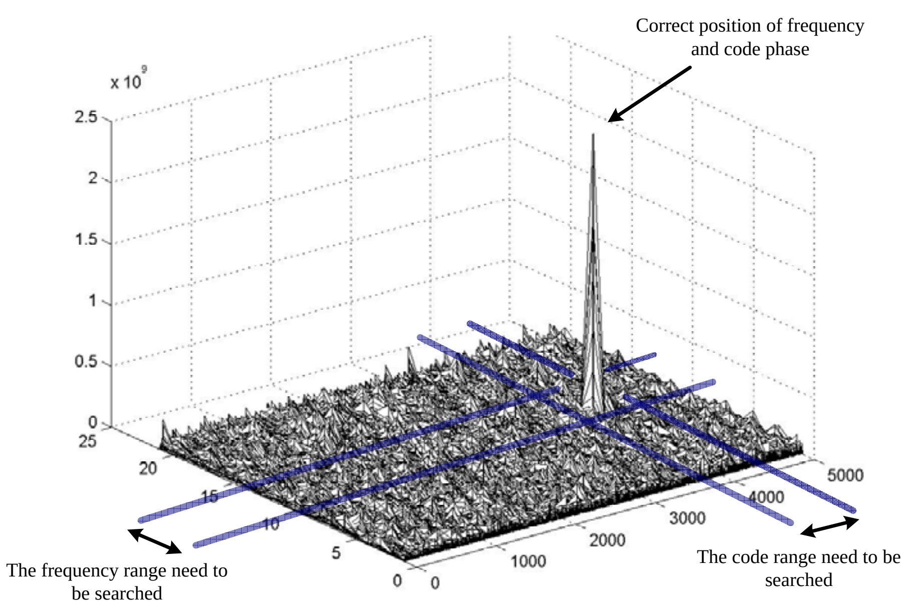

Figure 2.6: Acquisition with Narrow Range

1. The A-GPS handset transmits the position of the corresponding base station to the A-GPS server.

2. The A-GPS server transmits the related assistance information (e.g. almanac data, the elevation angle) according to the receiver's position.

3. The handset starts to receive the GPS signal according to the assistance information to reduce the TTFT(time-to-first-fit).

4. The handset demodulates the received signals and calculates the pseudo range of each satellite, and transmits these information to the A-GPS server.

5. The A-GPS server completes the calculation according the pseudo range information transmitted from the handset, and estimates the location of the handset.

2.4 Summing Over Multiple Blocks

In outdoor environment, it is sufficient to integrate the received signal with the local code for just one code period, which is 1 ms and obtain a good acquisition result. However, weak signals will require integration more than one period to achieve a successful acquisition. Since it can suppress the noise level in the received signal and the desired signal will become visible. As shown in Fig. 2.7, ideally we would like to integrate over an arbitrary number of code period as need. Unfortunately, it is unfeasible because of the natural characteristic of the received signal. The received signal is produced by the transmitted data bits, which is 20 milliseconds per bit, multiplied by the C/A code, which is 1 millisecond per chip. So integration over multiple code period may have problem encountering the polarity changes every 20 milliseconds. In the multiple period integration, the problem of signal polarity is always an important issue to be considered. Several methods has been proposed to overcome this problem, and they are briefly summarized in the following.

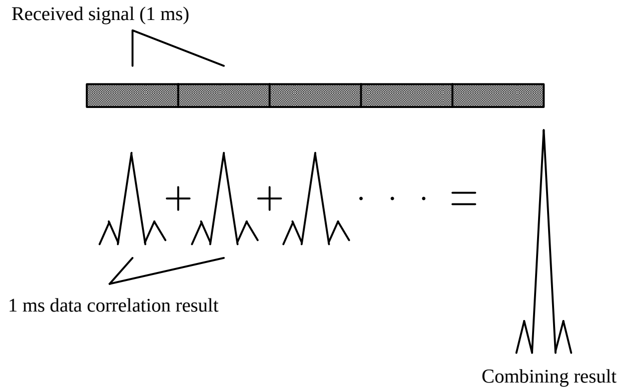

Figure 2.7: Concept of Summing Over Multiple Blocks

2.4.1 Even-Odd Block Combining

One of the summing method has been proposed in [33]. For every 20-millisecond received signal, there can be data bit transition. By summing over two consecutive 10-millisecond received data blocks, we are assured that at least one of them will be free of data bit transition as shown in Fig. 2.8. From the figure, we can easily conclude that only the even blocks or the odd blocks will suffer from bit transition problem, and the other will be free from bit transition. With this method, we can integrate an arbitrary number of code period by summing the correlation results individually for the even code blocks and the odd code blocks, and choose the one with higher correlation peaks as the final correlation result. This method avoids the attenuation effect caused by the data bit transition when multiple-period integration is required.

2.4.2 Three Combining Schemes

For acquisition of the GPS signal in a higher Signal-to-Noise Ratio (SNR) circumstance (e.g. outdoors or rural area), the 1-ms correlation is generally satisfactory for the detection of the code phase and the frequency shift. However, multi-period combining (i.e. integration for more than 1 ms) will be required for detecting the peak value from the correlation results under weak signal environment (i.e. with comparably low SNR situation). The combining

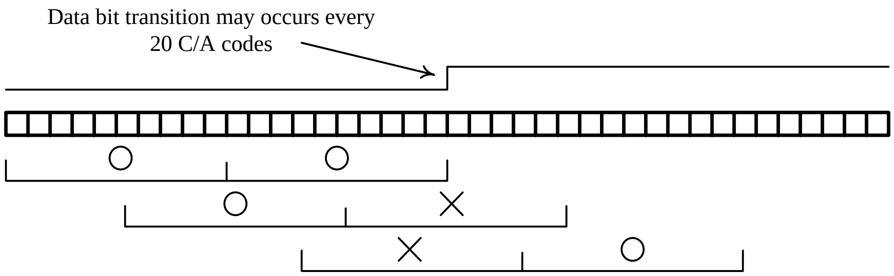

Figure 2.8: Even Odd Integration

methods can effectively suppress the noises and increase the successful rate for the correlation peak detection. The sensitivity of the receiver can therefore be amplified. The difference of a general GPS receiver and a receiver with combining method is, a general receiver compare the correlation peaks with the detection threshold after 1-ms signal correlation, and a combining receiver first calculate the combining result from the received signal and then compare it with the detection threshold. A receiver with the combining scheme is as shown in Fig. 2.9.

There are three different types of combining methods [17] [18], including the Coherent Combining (CC), the Non-Coherent Combining (NCC), and the Differential-Coherent Combining (DCC) schemes. After N ms of integration period, the correlation peaks (i.e. $P_{CC}(N)$, $P_{NCC}(N)$, $P_{DCC}(N)$) that result from these three combining methods are listed as follows:

$$P_{CC}^{N} = \max_{m(i,j)} \left\{ \left| \sum_{n=1}^{N} y^{(n)}(i,\, j) \right|^2 \right\} \tag{2.14}$$

$$P_{NCC}^{N} = \max_{m(i,j)} \left\{ \sum_{n=1}^{N} \left| y^{(n)}(i,\, j) \right|^2 \right\} \tag{2.15}$$

$$P_{DCC}^{N} = \max_{m(i,j)} \left\{ \left| \sum_{n=1}^{N} [y^{(n-1)}(i,\, j)]^* y^{(n)}(i,\, j) \right| \right\} \tag{2.16}$$

where

$$y^{(n)}(i,\, j) = \int_{(n-1)T}^{(n)T} r(t)[s(t-\tau_i)e^{j2\pi f_j t}]^*\, dt \tag{2.17}$$

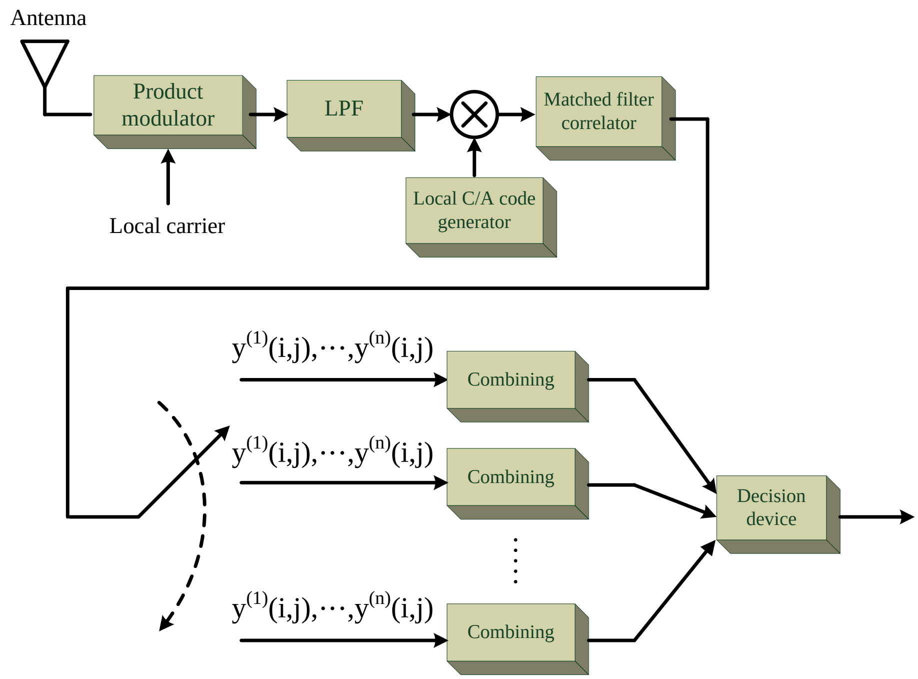

Figure 2.9: Receiver with Combining Scheme

and

1. T represents the code period.
2. τ_i is the code shift of the i^{th} correlator.
3. f_j represents the adjustable frequency shift.
4. $y^{(n)}(i, j)$ denotes the $(i, j)^{th}$ correlation results with $s(t - \tau_i)$ at the n^{th} integration period, which corresponds to a matrix element $m(i, j)$ in the two-dimensional search process.

It is noted that the three methods relate to obtaining the maximum value within the two-dimensional search space $m(i, j)$ after a particular type of combining scheme has been adopted.

There are tradeoffs between the three combining schemes. The CC method can provide the highest successful rate for peak detection without the squaring loss (as can be seen from

(2.14)) in the assumption that there is no bit-transition happening in the received signal. However it is not the case in real world. The CC method suffers from the problem due to the navigation bit-transition, which may occur every 20 ms of the received GPS data. The bit-transition deteriorates the results obtained from the CC method, which make it unfeasible to be adopted for the correlation peak detection with a longer period of integration time. The NCC method takes the absolute value of each 1-ms correlation result before summation, which alleviates the problem occurred from the navigation bit-transition. However, it suffers from the excessive amount of squaring loss as the SNR value decreases [18]. The results make the NCC scheme impractical to be utilized in weak signal environment.

The non-coherent integration is a technique integrating both the in-phase correlation result and the quadrature phase correlation result as shown in (2.15). Therefore it is not necessary to know the navigation message bit transition. That is, the non-coherent integration is not influenced by sign inversion of the navigation message bit during the integration, and an allowable carrier frequency error is not related to the number of the noncoherent integration but to the integration time of correlation. The disadvantage for the noncoherent integration is that it induces the squaring loss for weak GPS signals [18]. Particularly the squaring loss is the dominant factor among the acquisition losses of assisted GPS dealing with weak GPS signals. The squaring loss is defined as the ratio of the SNR before the non-coherent integration for the SNR after the non-coherent integration as:

$$L_{sq} = \frac{\alpha_c}{\alpha_{nc}} \tag{2.18}$$

$$\alpha_c = \sqrt{2 f_c T_i}\,\frac{\sqrt{2P_s}}{\sigma_n} \tag{2.19}$$

$$\alpha_{nc} = 2\frac{\Gamma(1/2+1)F(\frac{-1}{2};1;\frac{-\alpha_c^2}{2}) - \sqrt{\pi}}{\sqrt{4-\pi}} \tag{2.20}$$

where $\Gamma(\cdot)$ is the gamma function, $F(\cdot)$ is the confluent hypergeometric function, α_c is the SNR before the non-coherent integration, and α_{nc} is the SNR after the non-coherent integration.

Therefore, the squaring loss is given by

$$L_{sq} = \frac{\alpha_c\sqrt{4-\pi}}{2[\Gamma(1/2+1)F(\frac{-1}{2};1;\frac{-\alpha_c^2}{2})-\sqrt{\pi}]} \tag{2.21}$$

And the squaring loss has properties as follows:

$$\frac{d}{d\alpha_c}L_{sq} < 0 \tag{2.22}$$

$$L_{sq}|_{\alpha_c=\sqrt{10}} = 1 \tag{2.23}$$

From (2.23), it is explained that (2.21) is a monotonic decreasing function and the non-coherent integration induces the squaring loss when the Signal-to-Noise Ratio (SNR) is below $\sqrt{10}$. If the SNR before the non-coherent integration is below $\sqrt{10}$, the squaring loss exists. But if the SNR before the noncoherent integration is above $\sqrt{10}$, the squaring loss does not exist.

The DCC method (as in (2.16)) multiplies the correlation results from the two adjacent 1-ms of data. It can be observed that the DCC scheme lessens the bit-transition problem with decreased squaring loss in comparison with the NCC method. It has been shown in [17] [18] that the DCC scheme is superior to the NCC method with the existences of the frequency offset and fading in the received GPS signal.

Chapter 3

The Proposed Adaptive Coherent Combining (ACC) Scheme

This chapter is organized as follows: Section 3.1 introduces the motivation of the proposed ACC method. Section 3.2 investigates the carrier to noise power estimator which will be needed in the proposed ACC scheme. The reasons of adopting the adaptive peak detection threshold will be introduced in Section 3.3. Section 3.4 investigates the design principle of bit-transition detector, which is an important component of the ACC scheme. And the detail procedures of the ACC scheme will be illustrated in Section 3.5.

3.1 Motivation of the ACC Scheme

As described in the previous section, the CC scheme should provide the highest successful detection rate comparing with the other two methods. However, the combining results are severely degraded by the navigation bit-transition problem. Fig. 3.1 shows the combined peak values (P_{CC}^{N}) as in (2.14) vs the total combined length (for $N = 1$ to 35 ms) from the CC scheme. The three lines correspond to the SNR values equal to 0, -15, and -30 dB. The navigation bit-transitions are assumed to locate at both 10 ms and 30 ms in order to observe their effects to the CC scheme. It can be seen that the P_{CC}^{N} value is increased as the combined length N is raised up to 10 ms as shown in Fig. 3.1. Due to the navigation bit-transition, the

P_{CC}^N value is decreased after 10 ms, which can be observed from (2.14) by summing negative values. The convex shape of the P_{CC}^N value between 10 ms and 30 ms results from taking the absolute value of the real and the imaginary parts of P_{CC}^N, which are essentially both negative values due to the navigation bit-transition. It is also noted that the combined peak values obtained from the NCC (P_{NCC}^N) and the DCC (P_{DCC}^N) schemes should both result in increasing functions.

The results obtained from Fig. 3.1 can facilitate to propose a new combining scheme by observing the position of the navigation bit-transition. The two combined peaks (i.e. P_{CC}^{10} and P_{CC}^{30}) acquired from the CC scheme denote the existence of the bit-transition. The combined peak value (P_{ACC}^N) of the proposed ACC scheme can be designed to adaptively switch between the CC and the DCC schemes. As illustrated in Fig. 3.1, P_{ACC}^N is assigned as P_{CC}^N before the navigation bit-transition (i.e. N = 1 to 10 ms); while $P_{ACC}^N = P_{CC}^{10} + P_{DCC}^{N-10}$ after the bit-transition has occurred (i.e. N = 21 to 20 ms). By using the adaptive mechanism, the proposed ACC scheme can preserve the benefits from both the CC and the DCC methods.

It is also important to mention that the location of the navigation bit-transition does not necessarily relate to the peak value with correct code phase and frequency shift. Owing to both the noise effects and the spreading of the correlation peak from the bit-transition, the corresponding $m(i, j)$ that makes the value of P_{CC}^{10} is not always equivalent to the matrix entry (i.e. $m(i_c, j_c)$) that possesses the correct code phase and frequency shift.

3.2 Carrier to Noise Power Estimator

In general, a GPS acquisition block uses a predefined detection threshold, which obeys the Neyman-Pearson criterion to obtain the maximized detection probability in the constraint false alarm rate. With the knowledge of environment carrier to noise power ration, the acquisition block can calculate a more precise detection threshold rather than using predefined threshold, to significantly improve the sensitivity of GPS receivers. The carrier to noise estimation is a joint estimation of the two parameters at the same time. Based on the joint estimation theory of [34], the carrier to noise power estimation can be represent in the

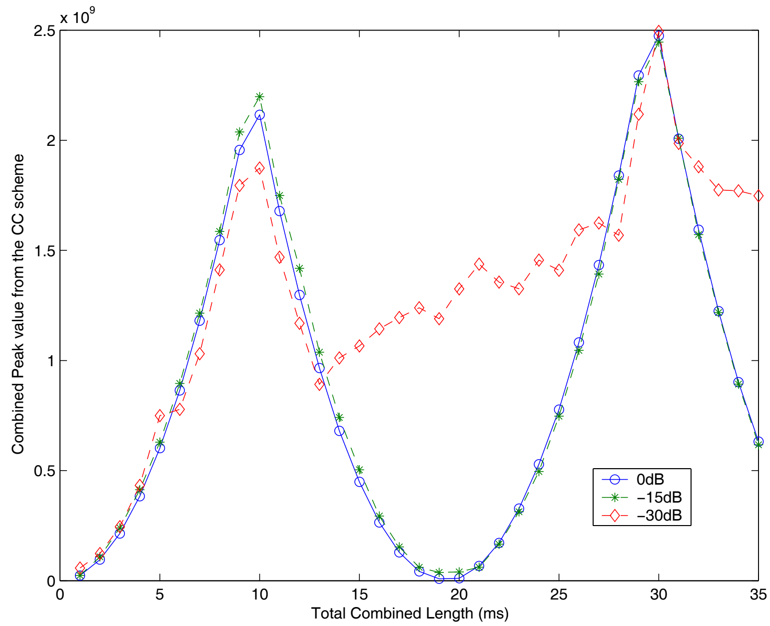

Figure 3.1: The Effect of the Navigation Bit-Transition in the CC Scheme

following [35]:

$$\frac{\hat{C}}{N_0} = \frac{\hat{P}F}{\hat{N}T_i} = \frac{F/T_i\sqrt{2(\sum_{\mu=0}^{M-1}|s_\mu|^2)^2 - \sum_{\mu=0}^{M-1}|s_\mu|^4}}{\sum_{\mu=0}^{M-1}|s_\mu|^2 - \sqrt{2(\sum_{\mu=0}^{M-1}|s_\mu|^2)^2 - \sum_{\mu=0}^{M-1}|s_\mu|^4}} \tag{3.1}$$

Where

- $\hat{P}$ denotes the estimated received signal power.
- $\hat{N}$ denotes the estimated noise power.
- s_μ represents the independent envelope samples of the received signal.
- F denotes the receiver noise figure and can be appropriately adjusted in simulation.

And the signal to noise power ratio is estimated after despreading and coherent integration over a period T_i. The estimation is from the amplitude moments, which leads to relatively simple calculations and fast convergence. The simulation results shows that the carrier to noise estimation performs well in static SNR environment. To improve the converging speed, we

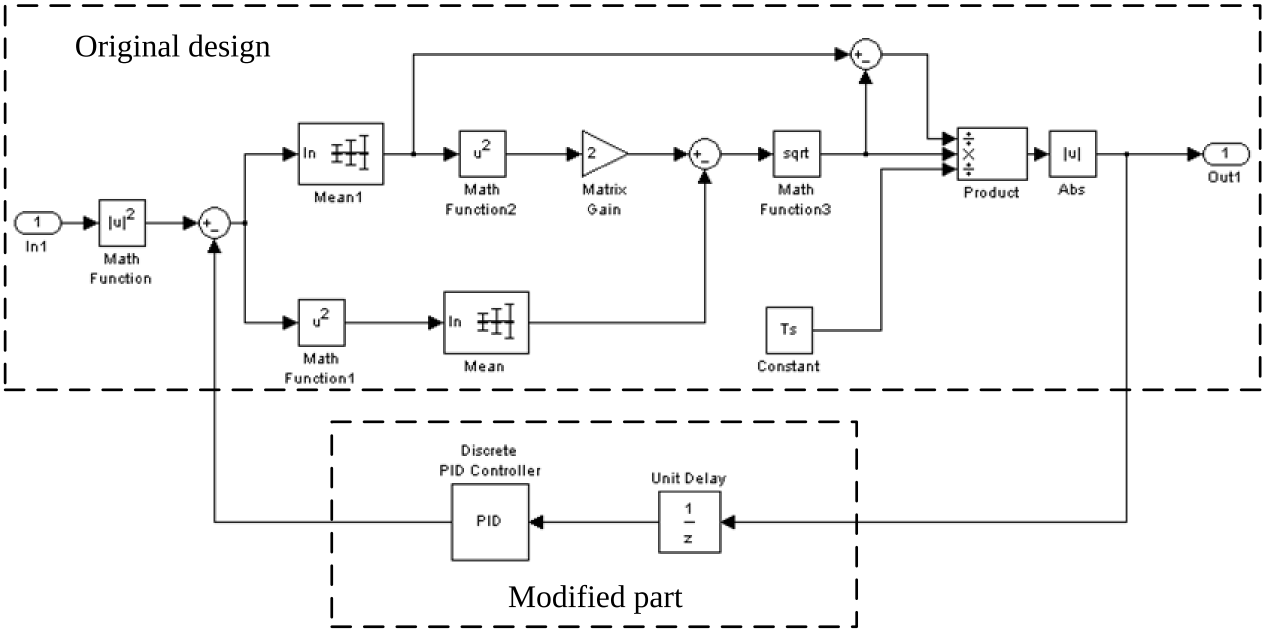

Figure 3.2: The C/N0 Estimator

add a PID (proportional-integral-derivative) feedback loop in the original estimation circuit, as shown in Fig. 3.2 and the simulation results shows that the new circuit outperforms the original circuit in speed of convergence and estimation accuracy. The theory of the C/N0 estimation and PID will be illustrated below.

3.2.1 Theory of C/N0 Estimation

The theory of the carrier to noise power estimator is based on the statistic method studied in [34]. The problem considered is to estimate two parameters, P and N from M independent envelope samples, s_1, s_2, ... ,s_{M-1}, which are denoted as S. The second moment and the forth moment are as follows [36]:

$$E[S^2] = P + N \tag{3.2}$$

$$E[S^4] = P^2 + 4PN + 2\sigma^2 \tag{3.3}$$

Denote $E[S^2]$ and $E[S^4]$ as $\bar{S^2}$ and $\bar{S^4}$ and the estimates of P and N as $\hat{P}$ and $\hat{N}$, the following equation can be obtained from (3.2) and (3.3) :

$$\frac{(\bar{S^2})^2}{\bar{S^4}} = \frac{P^2 + 2PN + N^2}{P^2 + 4PN + 2N^2} \tag{3.4}$$

Therefore, $\hat{P}$ and $\hat{N}$ can be computed as:

$$\hat{P} = [2(\bar{S^2})^2 - \bar{S^4}]^{1/2} \tag{3.5}$$

$$\hat{N} = \bar{S^2} - \hat{P} \tag{3.6}$$

3.2.2 Introduction to PID

A PID controller is a common feedback loop component in control systems. The PID loop adds positive corrections, removing error from the process's input. PID is named after its three correcting calculations, which all add to and adjust the controlled quantity. These three parameters are illustrated as following:

1. Proportional - To handle the present, the error is multiplied by a (negative) constant P (for "proportional"), and added to (subtracting error from) the controlled quantity. P is only valid in the band over which a controller's output is proportional to the error of the system.

2. Integral - To handle the past, the error is integrated (added up) over a period of time, and then multiplied by a (negative) constant I (making an average), and added to the controlled quantity. I averages the measured error to find the process output's average error from the setpoint. A simple proportional system oscillates, moving back and forth around the setpoint, because there's nothing to remove the error when it overshoots. By adding a negative proportion of (i.e. subtracting part of) the average error from the process input, the average difference between the process output and the setpoint is always being reduced. Therefore, eventually, a well-tuned PID loop's process output will settle down at the setpoint.

3. Derivative - To handle the future, the first derivative (the slope of the error) over time is calculated, and multiplied by another (negative) constant D, and also added to (subtracting error from) the controlled quantity. The derivative term controls the response to a change in the system. The larger the derivative term, the more rapidly the controller responds to changes in the process's output. Its D term is the reason a PID loop is also called a "Predictive Controller." The D term is a good thing to reduce when trying to dampen a controller's response to short term changes. Practical controllers for slow processes can even do without D.

By tuning these three parameters, we can improve the desired property of the original open loop system.

3.3 Adaptive Peak Detection Threshold (PDT)

Due to the low SNR value in weak signal environment, it is necessary to provide an adjustable peak detection threshold to increase the probability of successful rate. The Peak Detection Threshold ($PDT(SNR)$) is designed to be a SNR dependent parameter. In the acquisition process, the SNR value associated with the input signal can be obtained by the C/N0 estimator. The SNR value can be calculated by utilizing the initial sets of the received GPS data (e.g. the first 1 ms of the received data). The pre-determined relationship between the PDTs and the SNR values will be illustrated in the performance evaluation section.

The ratio between the peak value and the second to peak value obtained from the combining process are utilized for peak detection. If the ratio is greater than the threshold $PDT(SNR)$, it is concluded that the the peak corresponding to the correct code phase and the frequency shift (i.e. $m(i_c, j_c)$) is obtained. More details of the proposed ACC scheme will be described in the Section III.D.

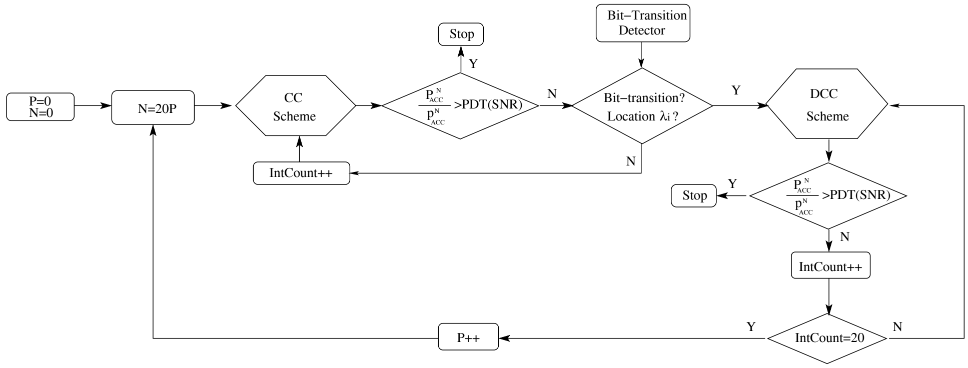

Figure 3.3: The Flow Chart of the Proposed ACC Scheme

3.4 Bit-Transition Detector

The bit-transition detector is the key component in the proposed ACC scheme. Every 20-sets of the correlated data (i.e. every 20-ms of data) are considered within the detector. As show in Fig. 3.1, the combined peak values (P_{CC}^{N}) vs the total combined length (for $N = 1$ to 35 ms) from the CC scheme will reveal the information of bit-transition. Therefore we can take it as the indication. In the detector, we use the concept of correlation coefficient. The correlation coefficient is an indicator of similarity of two values. It is a measure of how well the predicted values from a forecast model "fit" with the real-life data. The correlation coefficient is a number between 0 and 1. If there is no relationship between the predicted values and the actual values the correlation coefficient is 0 or very low (the predicted values are no better than random numbers). As the strength of the relationship between the predicted values and actual values increases, so does the correlation coefficient. A perfect fit gives a coefficient of 1.0. Thus the higher the correlation coefficient the better. From Fig. 3.1, we can see that the CC peaks will have a obviously slope discontinuity if there is a bit transition, therefore we can use this character as the detection feature. Following lists the steps of the bit-transition detector.

1. Design 20 models to model the CC shapes with bit-transition locates from 1 ms to 20 ms and call them "Model 1" to "Model 20" respectively.

2. Calculate the correlation coefficient of the CC results of received signal and the 20

models.

3. Choose the CC results whose correlation coefficient with each model that is great than 0.9.

4. Take the average of those chosen CC results.

5. Those with the highest average CC results represents the accurate bit transition position of the received signal.

For example, it is assumed that a bit-transition happens in the 5 ms of the received signal. If we take the average of the CC results whose correlation coefficient with the "Model 5" is greater than 0.9, the average will be the highest.

It is also found that the surrounding noises may deteriorate the detection of the navigation bit-transition with low SNR values. The successful rate for detecting the bit-transition goes down as the SNR value decreases. In the proposed ACC scheme, it is observed viable to partition the entire search space (i.e. $M_{p\times q}$) into several equally divided subspaces. $\gamma(i,\, j)$ corresponds to a subspace of the original search space $m(i,\, j)$, i.e. $\gamma(i,\, j) \in m(i,\, j)$.

3.5 Procedures of the ACC Scheme

The procedures of the proposed ACC scheme is shown in Fig. 3.3. As mentioned in the subsection 3.1, the proposed scheme adaptively adjust itself between the CC and the DCC methods, depending on the occurrence of the navigation bit-transition. The processes of the scheme are described as follows.

The SNR value of the received GPS signal is estimated after acquiring the first 1 ms of data. It is utilized to determine which $PDT(SNR)$ and $BTT(SNR)$ values should be selected. The bit-transition detector is enabled after the first 20-ms of the correlated data are accumulated. The bit-transition criterion as illustrated in the subsection 3.4 is exploited to detect the existence of a bit-transition within this 20-ms of data. As illustrated in Fig. 3.3, the position of the bit-transition will be fed into the executing loop of the ACC scheme.

At the beginning of the detection loop, the CC scheme is executed, i.e. $P^N_{ACC} = P^N_{CC}$. The ratio between the peak value (i.e. $P^N_{ACC}(\xi, L)$) and the second to peak value (represented as $p^N_{ACC}(\xi, L)$) in the ξ^{th} partition is utilized for peak detection. If $P^N_{ACC}(\xi, L)/p^N_{ACC}(\xi, L) > PDT(SNR)$, the ACC scheme terminates and the information of the coarse code phase and the frequency shift are delivered to the tracking system of the receiver. If the condition is not satisfied, the detection loop will continue to execute. The information supplied from the bit-transition detector will be used to determine if the ACC scheme should be switched from the CC to the DCC method. If there is no bit-transition occurred, the CC scheme will be utilized for the next 1-ms of the correlated data. Until the the bit-transition has been detected within the 20-ms period (e.g. located at the λ_1^{th} ms), the ACC scheme is adjusted into the DCC method for the remaining time of integration, i.e. $P^N_{ACC} = P^{\lambda_1}_{CC} + P^{N-\lambda_1}_{DCC}$. The ratio $P^N_{ACC}(\xi, L)/p^N_{ACC}(\xi, L)$ is continued to be compared with the $PDT(SNR)$ value for the remaining periods of time as shown in Fig. 3.3. If the peak ratio does not exceed the $PDT(SNR)$ value for the entire 20-ms of data, the ACC scheme will be adjusted back to the CC scheme. The process will be iteratively conducted until the peak ratio is greater than the $PDT(SNR)$ value. The combining peak P^N_{ACC} can therefore be obtained as

$$P^N_{ACC} = \sum_{i=1}^{n-1} \left[P^{\lambda_i}_{CC} + P^{20-\lambda_i}_{DCC} \right] + P^{\lambda_n}_{CC} + P^{N-\lambda_n}_{DCC} \tag{3.7}$$

where λ_i corresponds to the location of the detected bit-transition at the i^{th} 20-ms period. It is also noted that $\lambda_j = 20$ indicates no bit-transition occurred at that particular j^{th} 20-ms period. Unlike most of the existing combining algorithms with fixed integration time, the proposed ACC scheme provides an adjustable integration time based on the detection criterion. The effectiveness of the proposed ACC scheme will be evaluated and compared with the other combining methods in the next section.

The important steps of the ACC scheme are listed in the following:

1. The C/N0 estimator estimates the C/N0 using the 1 ms received signal.

2. The detection threshold are decided depends on the estimated C/N0.

3. The bit-transition detector activates after receiving 20 ms signals, and detects if there's bit transition within the 20 ms signal.

4. The acquisition process starts with the CC scheme, and switches to DCC scheme according the information provided by bit-transition detector.

5. The acquisition process will continue until the correlation peak of ACC exceed the PDT.

Chapter 4

Performance evaluation

This chapter is organized as follows: Section 4.1 illustrates settings of simulation parameters. Section 4.2 introduces the Rayleigh fading [24]- [26] channel that will be used in the simulation. And the simulation results will be investigated in Section 4.3.

4.1 Simulation Parameters

In the simulations, The parameters are selected based on the weak signal environment [21], where the SNR values are chosen between -33 and -39 dB. The power of the received GPS signal is assumed as $A = 30$ dBm; while the Doppler frequency shift f_d is set at 1251 kHz. Fig. 4.1 shows the peak detection thresholds ($PDT(SNR)$) vs the SNR values by averaging the experience data for 200 runs. In order to provide fair performance comparison, the $PDT(SNR)$ values for both the DCC and the NCC methods are also obtained. These adjustable $PDT(SNR)$ thresholds will be utilized in the simulations for performance evaluation.

4.2 Fading Channel

In this thesis, the communication channel will be modelled as the Rayleigh fading channel and the property of it will be illustrated as following. Rayleigh fading [24]- [26] is a statistical model for the effect of a propagation environment on a radio signal. It assumes that the power

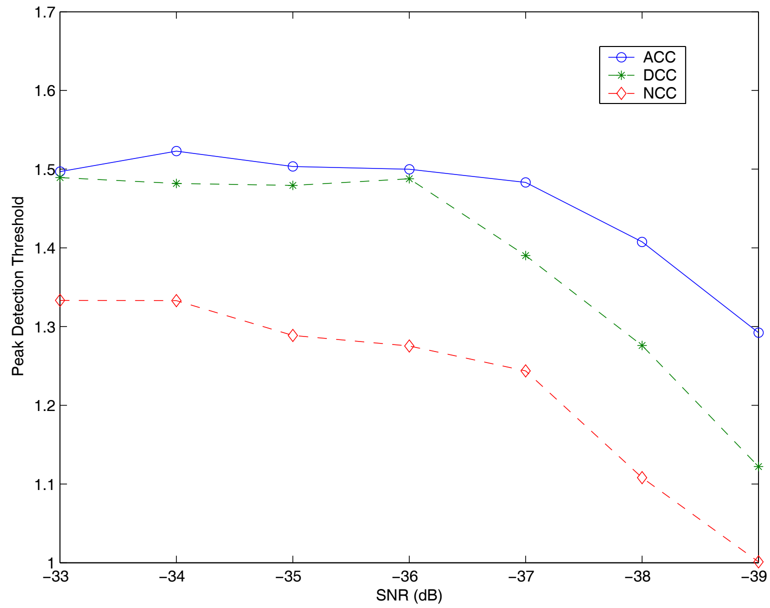

Figure 4.1: The Peak Detection Threshold vs The SNR Value

of a signal that has passed through a communications channel will vary randomly, according to a Rayleigh distribution. It is a reasonable model for tropospheric and ionospheric signal propagation as well as the effect of heavily built-up urban environments on radio signals. Rayleigh fading is most applicable when there is no line of sight between the transmitter and receiver. If there is a line of sight, Rician fading [37] [38] is more applicable.

Rayleigh fading is a reasonable model when there are many objects in the environment that scatter the radio signal before it arrives at the receiver. The central limit theorem holds that, if there is sufficiently much scatter, the channel impulse response will be well-modelled as a Gaussian process irrespective of the distribution of the individual components. The power, or envelope, of the channel response will therefore be Rayleigh distributed. Calling this random variable R, it will have a probability density function:

$$p_R(r) = \frac{r}{\sigma} e^{-r^2/2\sigma}, r \geq 0 \tag{4.1}$$

where $\sigma = E[R^2]$. Rayleigh fading is exhibited by the assumption that the real and imaginary parts of the response are modelled by independent and identically distributed (i. i. d.)

zero-mean Gaussian processes so that the amplitude of the response is the sum of two such processes.

How rapidly the channel fades will be affected by how fast the receiver and/or transmitter are moving. Motion causes Doppler shift in the received signal components. The process of generating the Rayleigh fading channel can be found in [24] . Although we use the simulink in-build Rayleigh fading channel in our simulation, the detail process are listed below: Jakes' model: In his book, [24] Jakes popularised a model for Rayleigh fading based on summing sinusoids. Let the scatterers be uniformly distributed around a circle at angles α_n with k rays emerging from each scatterer. The Doppler shift on ray n is

$$f_n = f_d \cos \alpha_n \tag{4.2}$$

and, with M such scatterers, the Rayleigh fading of the kth waveform over time t can be modelled as:

$$R(t,k) = 2\sqrt{2}[\sum_{n=1}^{M}(\cos\beta_n + j\sin\beta_n)\cos(2\pi f_n t + \theta_{n,k}) + \frac{1}{\sqrt{2}}(\cos\alpha + j\sin\alpha)\cos 2\pi f_d t] \tag{4.3}$$

Here, α and the β_n and $\theta_{n,k}$ are model parameters with α usually set to zero, β_n chosen so that there is no cross-correlation between the real and imaginary parts of $R(t)$:

$$\beta_n = \frac{\pi n}{M+1} \tag{4.4}$$

and $\theta_{n,k}$ used to generate multiple waveforms. If a single-path channel is being modelled, so that there is only one waveform then θ_n can be zero. If a multipath, frequency-selective channel is being modelled so that multiple waveforms are needed, Jakes suggests that uncorrelated waveforms are given by:

$$\theta_{n,k} = \beta_n + \frac{2\pi(k-1)}{M+1} \tag{4.5}$$

In fact, it has been shown that the waveforms are correlated among themselves (they have non-zero cross-correlation) except in special circumstances [25]. The model is also deterministic (it has no random element to it once the parameters are chosen). A modified Jakes' model [26] chooses slightly different spacings for the scatterers and scales their waveforms using Walsh-Hadamard sequences to ensure zero cross-correlation. Setting

$$\alpha_n = \frac{\pi(n-0.5)}{M+1} \tag{4.6}$$

and

$$\beta_n = \frac{\pi n}{M} \tag{4.7}$$

results in the following model, usually termed the Dent model or the modified Jakes model:

$$R(t,k) = \sqrt{\frac{2}{M}} \sum_{n=1}^{M} A_k(n)(\cos\beta_n + j\sin\beta_n)\cos(2\pi f_n t + \theta_{n,k}) \tag{4.8}$$

The weighting functions $A_k(n)$ are the k^{th} Walsh-Hadamard sequence in n. Since these have zero cross-correlation by design, this model results in uncorrelated waveforms. The phases $\theta_{n,k}$ can be initialized randomly and have no effect on the correlation properties. The Jakes' model also popularised the Doppler spectrum associated with Rayleigh fading, and, as a result, this Doppler spectrum is often termed Jakes' spectrum.

In the simulation, the channel will be modelled using Rayleigh fading channel, with maximum Doppler shift 40 Hz, and the number of total multi-path is four, each with delay factor around $2 * 10^{-6}$ seconds and gain factor -4dB.

4.3 Simulation Results

The simulations are conducted to compare the proposed ACC algorithm with the DCC and the NCC schemes. Each technique is executed with 200 runs to acquire the averaged simulation results. Figs. 4.2 and 4.3 illustrate the performance comparison between these three schemes under different SNR values. Under the same 95% of detection probability (as shown in Fig. 4.2), the proposed ACC scheme results in comparably less integration time for signal acquisition, which is around 56 ms and 16 ms less than that from the NCC and the DCC schemes under SNR = -39 dB. As illustrated in Fig. 4.3, the detection probability obtained from the ACC scheme can achieve up to 68% at SNR = -39 dB (with 50 ms of integration time); while only 50% and 20% of the detection probabilities are obtained by using the DCC and the NCC scheme. To make the comparison more fair, we also compare the detection probability and the integration of DCC and NCC with the adaptive detection threshold. As shown in Fig. 4.2 and 4.3, the results of DCC and NCC with ADT(adaptive detection threshold) excel the ordinary DCC and NCC with pre-defined detection threshold.

It is also noted that the pure CC scheme is not utilized for comparison since it is comparably the worst scheme (which can not even achieve 10% of detection probability under various SNR values) due to the navigation bit-transition. The merits of using the ACC schemes can therefore be seen from the simulation results. To consider the more realistic situation, we add a un-deterministic term to the received center frequency, make it varies around the center frequency about 100 Hz for each 1-ms signals. We use the same criterion to compare the ACC with the NCC and DCC scheme. The simulation results can be shown in Fig. 4.4 and 4.5. The variations of the received center frequency decreases the probability of all three schemes, and increases the necessary integration length to achieve the 95% detection probability, but the advantage of ACC scheme is still obvious comparing with other two schemes.

We add a feedback loop in the ordinary C/N0 estimator and the simulation results are as shown in Fig. 4.6, 4.7, and 4.8. As we can see in these results, the modified C/N0 estimator outperforms the ordinary estimator in convergence speed and accuracy.

To investigate the performance of the C/N0 estimator in the whole system, the detection probabilities of the system with the SNR estimated and the SNR previously known are compared in Fig. 4.9.

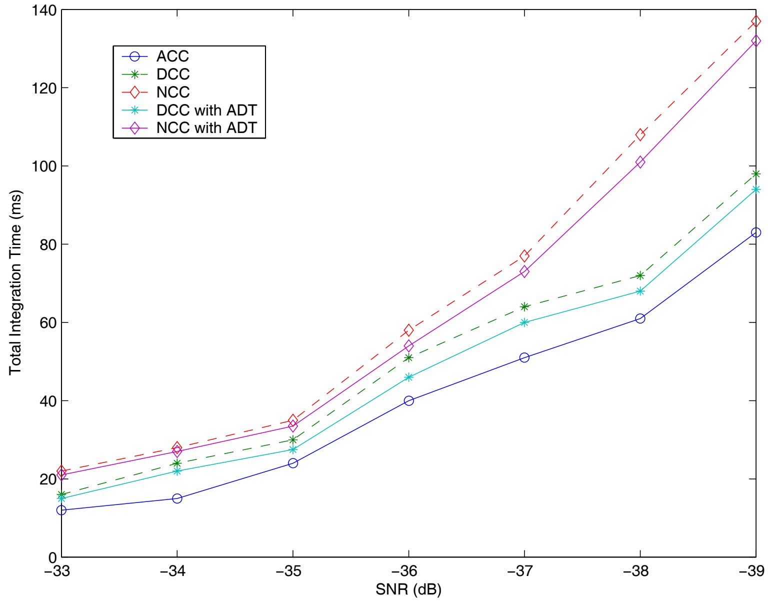

Figure 4.2: The Total Integration Time vs the SNR Value (with 95% of Detection Probability)

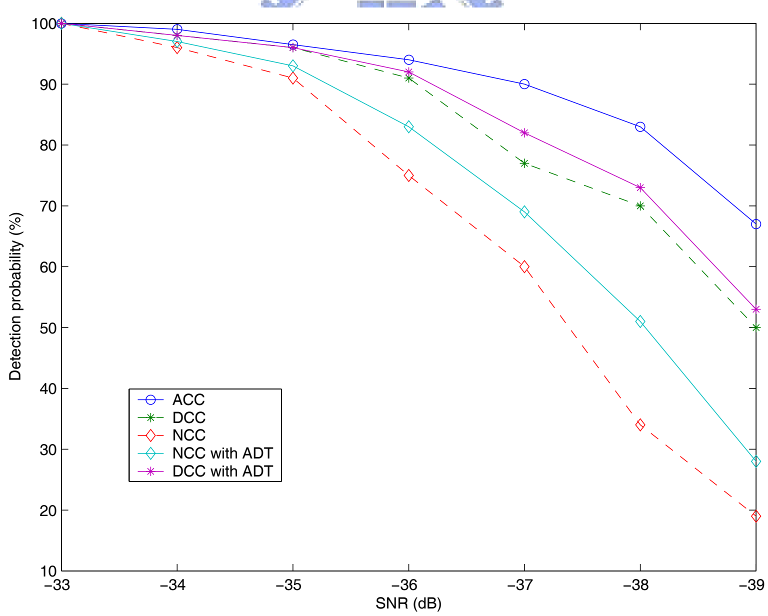

Figure 4.3: The Detection Probability vs the SNR Value (with 50 ms of Total Integration Time)

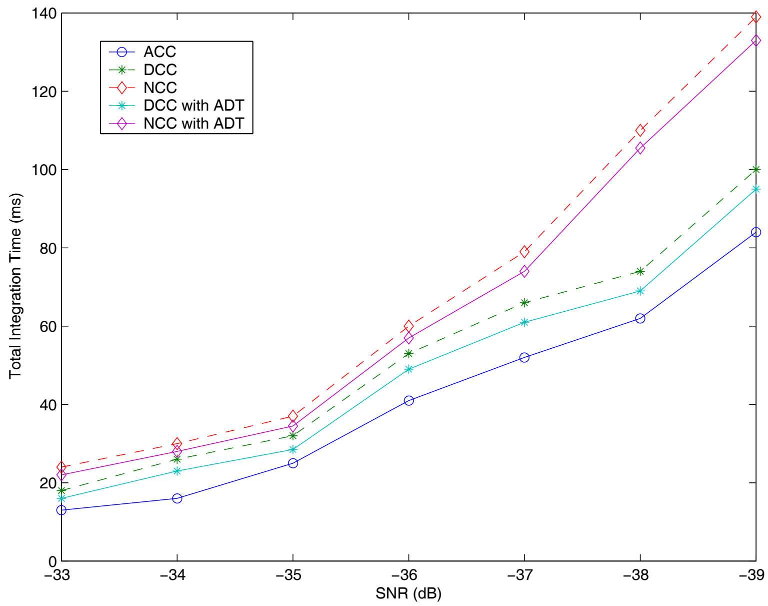

Figure 4.4: The Total Integration Time vs the SNR Value with Center Frequency Uncertainty (with 95% of Detection Probability)

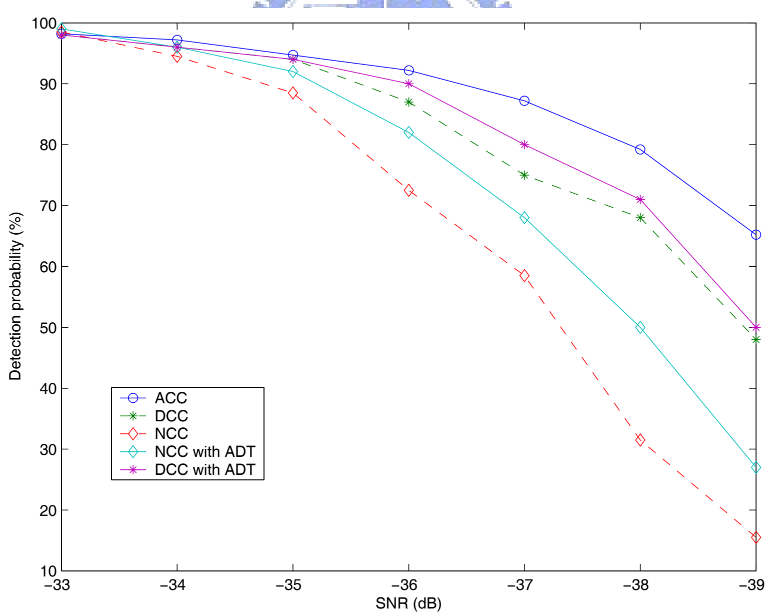

Figure 4.5: The Detection Probability vs the SNR Value with Center Frequency Uncertainty (with 50 ms of Total Integration Time)

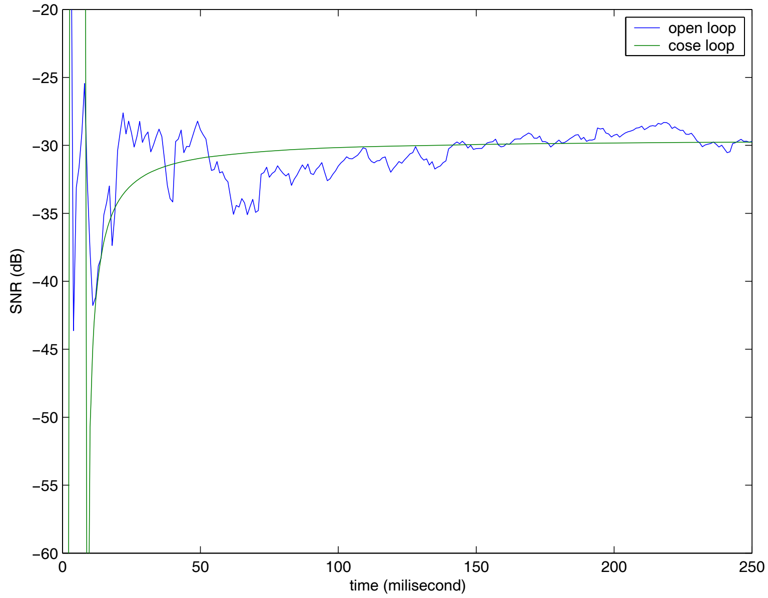

Figure 4.6: The Estimation Curve of Open Loop Estimator and Close Loop Estimator in -30dB

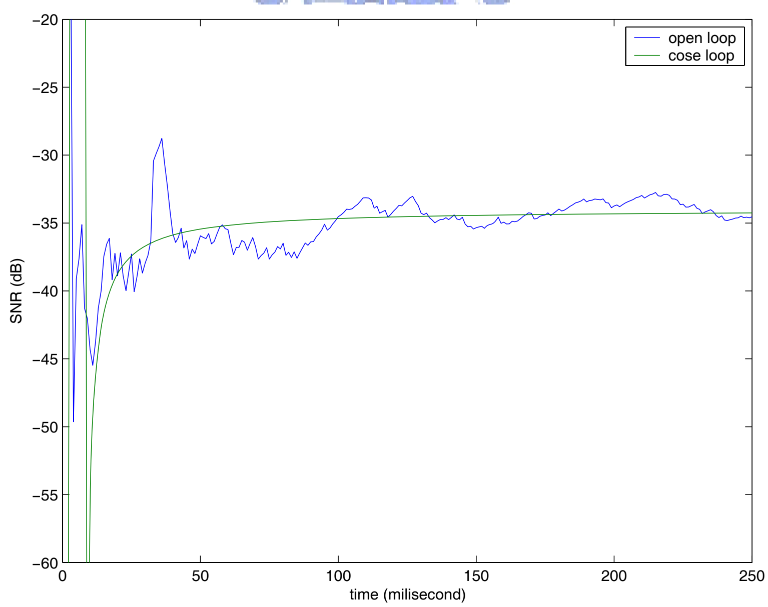

Figure 4.7: The Estimation Curve of Open Loop Estimator and Close Loop Estimator in -35dB

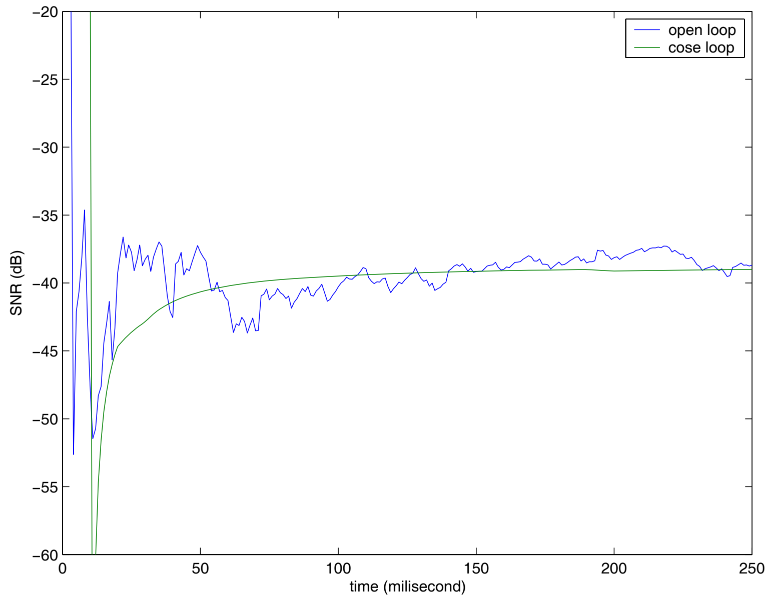

Figure 4.8: The Estimation Curve of Open Loop Estimator and Close Loop Estimator in -39dB

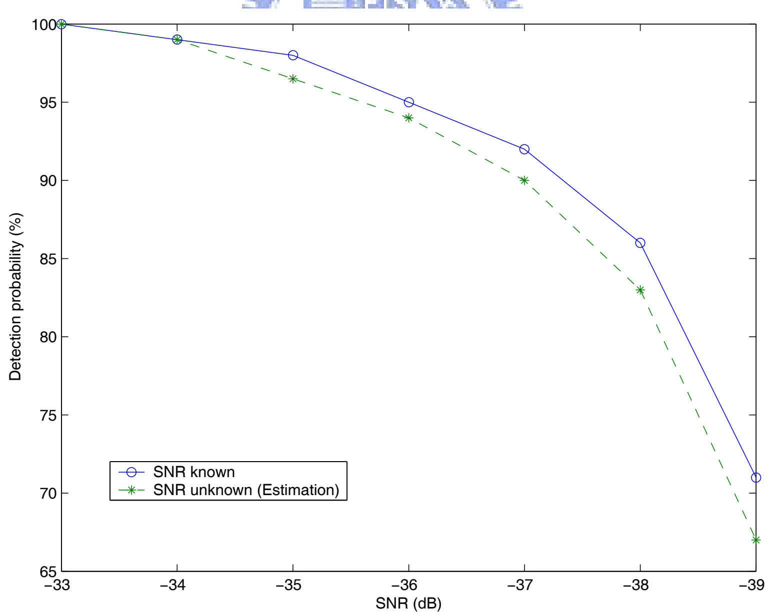

Figure 4.9: The Detection Probability of Known SNR and Estimated SNR

Chapter 5

Conclusions

In this thesis, the related techniques for the GPS receiver acquisition process has been investigated. The acquisition in time domain and frequency domain are introduces individually. The concept of A-GPS and the indoor acquisition methods, including three combining methods, even-odd multi-period integration method, are also introduced. An Adaptive Coherent Combining (ACC) acquisition technique is proposed. The ACC scheme is adaptively adjusted between the coherent or the differential coherent combining method based on the detection of the navigation bit-transition. The peak detection threshold is adaptively adjusted according to the environmental SNR estimated form the C/N0 estimator. The total integration time and the detection threshold are also adjustable based on the SNR values. As shown in the simulations, the proposed ACC scheme can provide better performance comparing with both the differential coherent and the non-coherent methods, especially under weak signal environment.

Bibliography

[1] M. S. Braasch and A. J. van Dierendonck, "GPS Receiver Achitectures and Measurements," *Proceedings of the IEEE*, vol.87, no.1, January 1999, pp. 48-64.

[2] James B. Y. Tsui, "Fundamentals of Global Positioning System Receivers: A Software Approach," 2nd Edition, *Wiley Series*.

[3] P. W. Ward, "GPS Receiver Serach Techniques," *Procedings of IEEE Position Location and Navigation Symposium*, April 1996, pp. 604-611.

[4] K. Krumvieda, P. Madhani, C. Cloman, E. Olson, J. Thomas, P. Axelrad and W. Kober, "A Complete IF Software GPS Receiver: A Tutorial about the Details," *ION GPS 2001*, Salt Lake City, UT, September 2001.

[5] D. J. R Van Nee and A. J. R. M. Coenen, "New Fast GPS Code-Acquisition Technique using FFT," *Electronics Letters*, vol.27, no.2, January 1991, pp. 158-160.

[6] J. A. Starzyk and Z. Zhu, "Averaging Correlation for C/A Code Acquisition and Tracking in Frequency Domain," *Proceedings of the IEEE Midwest Symposium on Circuits and Systems* , vol.2, August 2001, pp. 905-908.

[7] A. Alaqeeli, J. Starzyk, and F. V .Graas "Real-Time Acquisition And Tracking For GPS Receivers," *Proceedings of the 2003 International Symposium on Circuits and Systems*, vol.4, May 2003, pp. IV-500-IV-503.

[8] Zhu Z, "Averaging Correlation for Weak GPS Signal Processing, "*MS. Thesis*, Ohio University, April 2002.

[9] F. Daffara and P. Vinson, "Improved Search Algorithm for Fast Acquisition in a DSP-based GPS Receiver," *URSI International Symposium on Signals, Systems, and Electronics*, September 1998, pp. 310-314.

[10] T. Haddrell and A. R. Pratt, "Understanding The Indoor GPS Signal," *ION GPS 2001*, Salt Lake City, UT, September 2001, pp. 1487-1499.

[11] F. V. Diggelen, "Global Locate Indoor GPS Chipset and Services," *ION GPS 2001*, Salt Lake City, UT, September 2001.

[12] J. LaMance, "Low infrastructure Assisted GPS," *GPS World magazine*, November 2001.

[13] N. Harper, P. Nicholson, P. Mumford, and E. Poon, "Process for Improving GPS Acquisition Assistance Data and Server-side Location Determination for Cellular Networks," *The 2004 International symposium on GNSS/GPS*, Sydney, Australia, December 2004.

[14] D. Kubrak, C. Macabiau, and M. Monnerat, "Analysis of a Software-Based A-GPS Acquisition Performance Using Statistical Processes," *ION NTM 2005*, San Diego, CA, January 2005.

[15] M. L. Psiaki, "Block Acquisition of Weak GPS Signals in a Software Receiver," *ION GPS 2001*, Salt Lake City, UT, September 2001, pp. 2838-2850.

[16] O. -S. Shin and K. B. Lee, "Differential Cohorent Combining for Double-Dwell Code Acquisition in DS-CDMA Systems," *IEEE Transations on Communications*, vol.51, no.7, July 2003.

[17] H. Elders-Boll and U. Dettmar, "Efficient Differentially Coherent Code/Doppler Acquisition of Weak GPS Signals," *The 2004 International symposium on GNSS/GPS*, Sydney, Australia, December 2004.

[18] D. J. Cho, C. Park, and S. J. Lee, "An Assisted GPS Acquisition Method using L2 Civil Signal in Weak Signal Environment," *Journal of Global Positioning Systems*, vol.3, no.1-2, 2004, pp. 25-31.

[19] Y. K. Jeong, O. -S. Shin, and K. B. LEE, "Fast slot synchronization for intercell asynchronous DS-CDMA systems,"*IEEE Transactions on Wireless Communications*, vol.1, April 2002, pp. 353-360.

[20] Y. K. Jeong, K. B. LEE, and O. -S. Shin, "Differentially coherent combining for slot synchronization in intercell asynchronous DS/SS systems," *Procedings of the IEEE International Symposium on Personal, Indoor and Mobile Radio Communications*, September 2000, pp. 1405-1409.

[21] David M. Lin and James B. Y. Tsui, "A software GPS receiver for weak signals ," *IEEE MTT-S International Microwave Symposium Digest*, vol.3, 20-25, May 2001, pp. 2139-2142.

[22] David M. Lin and James B. Y. Tsui, "An Efficient Weak Signal Acquisition Algorithm for a Software GPS Receiver," *ION GPS2001*, Salt City, UT, September 2001.

[23] M. Yamada, Y. Kamio, and Y. Wada, "Acquisition of Direct-Sequence Spread-Spectrum Signal with Parallel Matched Filters," *Indoor and Mobile Radio Communications*, vol.2, September 2000, pp .1260-1264.

[24] W. C. Jakes, "Microwave Mobile Communications," *John Wiley and Sons Inc*, February 1975.

[25] S. V. Eckardstein and K. Isaksson, "Kanalmodeller for radiotransmission" (Channel models for radio transmission) "Royal Institute of Technology Master's thesis," Stockholm, Sweden, December 1991.

[26] P. Dent, G. E. Bottomley, and T. Croft, "Jakes Fading Model Revisited," *Electronics Letters*, June 1993, pp. 1162-1163.

[27] R. R. Rick and L. B. Milstein, "Optimal Decision Strategies for Acquisition of Spread-Spectrum Signals in Frequency-Selective Fading Channels," *IEEE Transactions on Communications*, vol.46, no.5, May 1998.

[28] A. M. Slonneger and D. V. Sarwate, ”Noncoherent parallel acquisition of PN sequences in direct sequence spread spectrum systems,” *ISSAT*, Yokohama, Japan, November 1992, pp. 31-34.

[29] Y. T. Su, ”Rapid code acquisition algorithms employing PN matched filters,” *IEEE Transations on Communications*, vol.36, June 1988, pp. 724-733.

[30] E. A. Sourour and S. C. Gupta, ”Direct sequence spread spectrum parallel acquisition in a fading mobile channel,” *IEEE Transations on Communications*, vol.38, July 1990, pp. 992-998.

[31] Chae-Hyun Lim, Hae-Sock Oh, Dong-Seog Han, and Myoung-Je Cho, ”Adaptive hybrid acquisition with antenna diversity in CDMA systems,” *Military Communications Conference*, vol.2, October 2001, pp. 1243-1247.

[32] B. B. Ibrahim and A. H. Aghvami, ”Direct sequence spread spectrum matched filter acquisition in frequency-selective Rayleigh fading channels,” *IEEE Journal On Selected Areas In Communications*, vol.12, no.5, June 1994, pp. 885-890.

[33] V. Przebinda, ”Advanced GPS Signal Acquisition,” May 2004.

[34] T. R. Benedict and T. T. SOONG, ”The Joint Estimation of Signal and Noise from the Sum Envelope,” *IEEE Transactions on Information Theory*, vol.13, no.3, July 1967.

[35] A. Schmid and A. Neubauer, ”Carrier to noise power estimation for enhanced sensitivity Galileo/GPS receivers,” *VTC 2005-Spring*, vol.4, June 2005, pp. 2629-2633.

[36] J. I. Marcum and P. Swerling, ”Studies of target detection by pulsed radar,” *IRE Transactions on Information Theory*, vol.6, April 1960.

[37] Y. Xie and Y. Fang, ”A General Statistical Channel Model for Mobile Satellite Systems,” *IEEE Transactions on Vehicular Technology*, vol.49, no.3, May 2000.

[38] M. Abramowitz and I. A. Stegun, ”Handbook of Mathematical Functions,” National Bureau of Standards, 1964, reprinted Dover Publications, 1965.