Short-delay multipath mitigation for indoor positioning applications

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Short-Delay Multipath Mitigation for Indoor Positioning Applications

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A thesis submitted to the Nanyang Technological University
in fulfillment of the requirement for the degree of
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Summary

The works presented in this thesis mainly focus on mitigating short-delay multipath signals to enhance the ranging accuracy for Global Positioning System. The background behind this research is that nowadays GPS is widely used in urban areas and is being extended to indoor environments, where multipath signals are present. The multipath signals received by receiver antenna can degrade the C/A code ranging accuracy, and thus have significant effect on the estimation of user position. The existing techniques are able to mitigate the long-delay multipath signals quite well, but are less efficient in mitigating short-delay multipath signals which are numerous in urban areas and indoor environments.

GPS fundamentals closely related to multipath mitigation, e.g., signal acquisition and tracking, are presented in this thesis. Multipath related issues including its concept, properties, effects on the GPS signals, as well as the existing multipath mitigation techniques are studied extensively.

Multipath mitigation based on the partial early edge of the C/A code autocorrelation function is proposed in this thesis. It is based on the observation that multipath signals do not affect the leading portion of the early edge of the direct C/A code autocorrelation function. Multipaths pass through longer paths than the direct signal and hence have later arrival time. This fact has been well known and was mentioned in some other works, but here it is used to mitigate multipath signals, and the ranging errors under this technique are investigated analytically which is one of the contributions in this thesis. The 32 C/A codes adopted in GPS are studied and classified into 7 groups according to the shape of the first side lobe of the autocorrelation function. Mathematical analysis and simulations in ideal conditions indicate that the tracking errors caused by multipaths from these 7 groups are different, and for 25 C/A codes out of the total

32, the proposed technique is helpful to reduce the range error and has an advantage over the 0.1-chip-spacing narrow correlator, especially for the short-delay multipath. The range errors performance for the other 7 C/A codes is similar to existing technique. A measurement method for collecting GPS signal is presented in this work and the above technique is then tested based on the collected signal from a GPS simulator.

The virtual multipath based method for mitigating short delay multipath signals is an innovative technique proposed in this thesis. In this technique, the concept of virtual multipath is firstly defined and used to represent the overall effect of all the real multipath signals received by a GPS receiver. It is found that the peak location of the autocorrelation function of the direct signal can be represented by the peak of the virtual line-of-sight correlation function which is the resultant function after the virtual multipath correlation function is subtracted from the received composite correlation function. Mathematical analysis is also made in this thesis to prove the above idea based on the ideal C/A code and its triangular autocorrelation function. However, the peak of the autocorrelation function is actually rounded due to the finite bandwidth filtering of GPS receiver. To address this problem, a method called learning is proposed to reduce the effect of the rounded peak. Ideal simulations based on nine deterministic multipath signals as well as the simulation based on real GPS signal indicate that this technique is able to achieve great improvements on the range errors caused by multipath when compared with 0.1-chip-spacing narrow correlator and strobe correlator techniques.

Asymmetric phase only correlation (APOC) is another technique proposed in this thesis and its characteristics are presented through ideal simulations in Matlab as well as mathematical analysis on FFT. The performance of APOC on ranging is verified through practical measurements in the Positioning and Wireless Technology Center (PWTC), NTU. The results indicate that APOC has narrow correlation peaks and thus has better ability to separate short delay multipath

signals from the direct signal. Once the first arrival signal or the direct signal is detected, the multipath signals can be mitigated. Binary phase only correlation (BA-POC) is the resultant algorithm after the APOC is further improved by using two-bit level quantization to sample the received signal. Measurements on BA-POC are also conducted in indoor office environment, and high ranging accuracy is obtained. The limitation of BA-POC is that it is only applicable when the line-of-sight signal is present.

This thesis also presents the implementation of a software GPS receiver and its test results. The implemented software receiver is made up of the conventional functional blocks, i.e., FFT based acquisition, tracking and navigation. Through the implementation of the software receiver, the GPS technology is better understood and the receiver is also helpful to other related research works and projects in PWTC.

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Chapter 1 Introduction

1.1 Motivations

Global Positioning System (GPS) is a great successful story, which was developed firstly for U.S. military to provide precise estimates of position, velocity and time. Civil use is its second objective, whose accuracy is not as good as that of military use. But what is surprising to the GPS designers is that the civil applications grew at an astonishing rate throughout the 1990s, which can be found in land transportation, civil aviation, maritime commerce, etc. The value of the worldwide GPS products is expected to grow to \$22 billion in 2008 [1], up from \$13 billion in 2003, according to the Allied Business Intelligence of Oyster Bay, New York. GPS is on its way to become a part of our daily lives as an essential element of the commercial and public infrastructure.

For years, GPS has served society remarkably well when the satellites are clearly visible. But in recent years, applications are being extended into areas where the satellites are not visible, e.g., in the indoor or urban canyon environments. There are two issues significantly affecting the positioning performance in such environments:

The first one is the signal strength or the receiver sensitivity. Sensitivity, defined as the minimum signal strength with which a GPS receiver is able to work [2], is a function of various elements. Too weak satellite signals will result in failure of positioning. Current techniques, mainly focusing on increasing the pre-detection integration interval [2-4] or aiding from wireless network [5-7], have achieved great improvements on the receiver's sensitivity. For instance, a minimum signal power of -183.5dBW was detected in the literature [4].

The second issue is multipath, which means that the reflected signals from

ground or walls can induce a significant positioning error and hence have a negative impact on the receiver's performance. The multipath issue has been noticed for many years and great improvements have been achieved. The typical techniques, e.g., Narrow Correlator [8], Multipath Estimation Delay Lock Loop (MEDLL) [9], Strobe Correlator [10], etc., have been widely used in commercial GPS receivers. However, all these techniques only have good performance on mitigating long-delay (larger than 0.1 CA code spacing or approximately 30 meters) multipath signals. While in indoor or urban environments, numerous short-delay (less than 30 meters) multipath signals are present, which result in poor positioning accuracy. Furthermore, Narrow Correlator and Strobe Correlator can work in the region around the peak of the Pseudorandom Noise (PRN) code correlation function, but they require wide or even full GPS signal bandwidth (20MHz for the L1 band), and hence larger noise and interference could be introduced into the receiver. The computation load of MEDLL results in high hardware cost, because its high performance on mitigating multipath signals is based on the principle of maximum likelihood and needs to compute multiple correlations concurrently, and thus lots of correlators are a 'must'.

The above discussions imply that researches on short-delay multipath mitigation are demanding, especially for some typical environments like indoor and urban canyon. Works on short-delay multipath mitigation [11, 12] were published recently. The idea in [11] relies on a conclusion that the signal amplitude measurement (reported as the C/N_0 by GPS receivers) is highly correlated with the multipath error in the code phase measurement. A scaling factor, depending on multipath environment, is then used to estimate the multipath caused range error from the signal amplitude, and then the estimated range error is used to correct the measured pseudorange; the work in [12] involves development of a statistical model which is able to capture the behavior of pseudorange error distributions observed in real data. Based on this model, location can be estimated

by maximizing likelihood or minimizing expected squared error.

The above two techniques offered good ideas on mitigating short-delay multipath signals, but their performances can be significantly affected by the multipath environment. In other words, for a specific environment, some prior knowledge must be known in order to mitigate multipath. Therefore, both of them will face difficulties when put into practice.

1.2 Objectives

GPS signal is on two frequency bands, but one is open to limited authorized users, and the other one is for civilian purpose and free worldwide. The researches in this thesis will be based on the civilian signal or the C/A code only.

As presented above, mitigation on the range error caused by the short-delay multipath signals which are dominant in indoor and urban canyons environments is the major challenge for Global Positioning System nowadays, and so far there is no promising solution. The researches conducted in this thesis are based on this background and with the following objectives:

- i) Understand the methodologies and the principles behind the existing typical multipath mitigation techniques in GPS and investigate whether they can be improved and applied into short-delay multipath mitigation;
- ii) Seek innovative ideas, compatible with the current GPS technology, to re-shape the distorted autocorrelation function of the C/A code, caused by the short-delay multipath signals, to reduce the pseudorange error;
- iii) In indoor environments, there could be no any GPS signal detected, and other ranging systems, e.g., Wireless LAN (WLAN) based ranging system, can be used as an alternative method for positioning, so investigations on the multipath mitigation techniques for this kind of

systems are to be involved in this thesis as well.

1.3 Major Contributions

The major contributions achieved in this thesis are listed below:

- i) A fine delay estimation technique based on the partial autocorrelation function is proposed. This technique is based on the assumption that multipath signals are later than the direct signal, and thus on the leading edge of the autocorrelation function of C/A code, there is a portion which is ideally not affected by multipath signals. This portion is investigated and used to estimate the pseudorange error caused by multipath. Originally, this idea was mentioned in [13] and was used as an empirical method to estimate the delay of the direct signal. While in this thesis, it is investigated analytically. In addition, the property of the first side lobes of the autocorrelation functions of all the 32 PRN codes are analyzed, because all the first side lobes have a significant impact on the multipath tracking errors. Matlab simulations indicate that the multipath tracking errors from 25 PRN codes out of the total 32 codes are better than the 0.1-chip-spacing Narrow Correlator, and the tracking errors from the rest of the 7 PRN codes are acceptable as well. Measurements based on the GS1010 GPS simulator were also conducted to further verify the performance of the proposed technique [Chapter 5].
- ii) A novel technique based on virtual multipath signal is proposed to mitigate short-delay multipath signals. In this technique, the concept of virtual multipath is proposed for the first time, which is constructed according to the maximum value of the composite correlation function obtained at the receiver side. Based on this virtual multipath signal, the

delay of the direct signal can be determined from the virtual line-of-sight correlation function, which is the result after subtracting the virtual multipath correlation function from the composite correlation function. To address the problem of the limited bandwidth of real GPS signal, a method named as learning is proposed in this technique. Matlab simulations based on nine Rayleigh faded short-delay multipath signals with/without direct signal show that the proposed technique can achieve an average range error within 10 meters. The technique is further evaluated based on real GPS signal, which also shows superior performance when compared with the 0.1-chip Narrow Correlator and Strobe Correlator [Chapter 6].

- iii) Phase only correlator is a popular technique in image processing and pattern recognition, but it is not suitable to process communication signals, mainly due to heavy noise. The modified version, called Asymmetric Phase Only Correlation (APOC), is proposed in this thesis and for the first time applied into the area of ranging. Simulations show that APOC is able to produce multiple narrow and sharp correlation peaks when multiple reflected signals are present. Measurements based on the 2.45GHz WLAN signal were conducted in indoor office environments and the results indicate that APOC has superior performance on mitigating multipath signals. The APOC is further improved for saving hardware cost, which is named as Binary Asymmetric Phase Only Correlation (BA-POC) and is applicable to the environments when the strong line-of-sight signal is present. Measurement indicates that BA-POC is able to achieve high ranging accuracy with range errors around 4 meters [Chapter 7].

The above three multipath mitigation techniques handle the multipath issue

without incorporating any assistance from the cellular communication links and are based on the assumption that the signal delay has been estimated up to the integer number of C/A code chips and hence only focus on the estimation of the fraction of the signal delay. The integer part of the signal delay normally can be determined as long as the GPS signal can be detected. The basic method of detecting indoor GPS signal is through performing coherent integration for a longer period of time, so in this thesis we assume that the transitions of the navigation data are known. Additionally, the performances on multipath mitigation of the three proposed techniques are achieved by utilizing the geometrical properties of the C/A code autocorrelation function, and therefore all of them are geometrically based rather than statistically based. Finally, the three proposed techniques are based on the C/A code autocorrelation function, no matter whether it is obtained by coherent or by noncoherent integration, so the three techniques proposed here fit the current GPS receiver technology quite well.

Besides the above researches on short-delay multipath mitigation, the implementation of a software GPS receiver (SGR) will also be introduced. Testing based on the real GPS signals indicates that the SGR can achieve a user position error within 10 meters in the East-West and the North-South direction.

1.4 Organizations of Thesis

This thesis consists of eight chapters organized as follows:

Chapter 1 presents an introduction to the motivations, objectives and the contributions of the research work.

To better understand the issue on multipath, some necessary fundamentals of GPS are given in Chapter 2. Multipath mainly has impact on the estimation of pseudorange, so only the knowledge closely related to pseudorange estimation,

e.g., signal acquisition and tracking, are presented in detail. In this chapter, a brief introduction to the GPS Modernization is also mentioned.

In Chapter 3, the literature review on multipath will be presented, including multipath characteristics, analysis on the multipath tracking errors, existing multipath mitigation techniques, etc.

The detailed implementation of the software GPS receiver is presented in the Chapter 4, which includes the receiver block diagram, the parameters adopted and the testing results based on both the GS1010 simulator signal and the real navigation satellite signals.

Chapters 5, 6 and 7 will present the major contributions mentioned in Section 1.3 respectively. The measurements and performance verifications of APOC and BA-POC are based on WLAN signal, but potentially they can be applied to process a Pseudolite [14] signal, which will be further mentioned in the Chapter 8.

Chapter 8 will conclude this thesis and present some ideas on the future works.

Chapter 2 Global Positioning System Fundamentals

This chapter will present the fundamentals of global positioning system, which are the basic knowledge to understand the remaining chapters in this thesis. The most challenging task of a GPS receiver is to estimate the distances between it and at least four satellites, i.e., pseudoranges, based on which the user position can be calculated, so in this chapter only the fundamentals related to the pseudorange estimation will be introduced in details.

2.1 GPS Overview

2.1.1 Space Segment [15]

Fig.2-1 GPS satellite constellation

The GPS Space Segment consists of 24 Navstar satellites in semi-synchronous (approximately 12-hour) orbits. Six orbital planes are arranged for all the 24 satellites, and each one contains four satellites. The inclination angle of each orbital plane is 60 degrees relative to the earth's equator. The average radius of the satellite orbit is 26,560 kilometers (or 20200 kilometers above the earth surface). Fig.2-1 shows the GPS satellite constellation.

The period for a GPS satellite to complete running one orbit is approximately 11 hours and 58 minutes, and the earth is rotating under the satellites, so a satellite will trace a track over the earth's surface which repeats every 23 hours and 56 minutes, and hence a stationary user on the surface of the earth will observe the same satellite in the sky once each day, but the satellite will rise and set four minutes earlier each day, due to the four minute difference between the rotational period of the earth and two orbital periods of a satellite. Under this arrangement, at any time and any location with a clear view of sky, a GPS receiver is able to observe four to eleven satellites.

The GPS satellite signals are transmitted in a form of Code Division Multiple Access (CDMA) technique, which contains carrier, Pseudorandom Noise (PRN) codes (i.e., ranging codes), and navigation data. There are two frequency bands for the carrier. One is Link 1 (L1) at 1575.42 MHz, and the other is Link 2 (L2) at 1227.6 MHz. There are two types of PRN codes in the satellite signals. One is a 1.023 MHz Coarse/Acquisition code (C/A code) on L1, and the other is a 10.23 MHz Precision code (P code) on both L1 and L2. Both C/A code and P code are used for ranging, but the difference between them, except the chip rate, is that C/A code is free and open to all users over the world, while the P code is encrypted and available only to authorized users, e.g., US military. When encrypted, the P code is known as Y code. 50 Hz navigation data are superimposed on both the P code and the C/A code. The contents of navigation data include satellite clock-bias, satellite ephemeris (precise orbital) data for the transmitting satellite, ionospheric signal

propagation correction data, and satellite almanac (coarse orbital) data for the entire constellation.

2.1.2 Control Segment [16]

The control segment primarily consists of a Master Control Station (MCS), at Falcon Air Force Base (AFB) in Colorado Springs, USA, plus Monitor Stations (MS) and Ground Antennas (GA) at various locations around the world. The monitor stations are located at Falcon AFB, Hawaii, Kwajalein, Diego Garcia Atoll, and Ascension. All monitor stations except Hawaii and Falcon AFB are also equipped with ground antennas. The Control Segment includes a Pre-launch Compatibility Station (PCS) located at Cape Canaveral, USA, and a back-up MCS capability.

The MCS is the central processing facility for the control segment and is responsible for monitoring and managing the satellite constellation. The functions of MCS include the control of satellite station-keeping maneuvers, reconfiguration of redundant satellite equipment, regularly updating the navigation messages transmitted by the satellites, and various other satellite health monitoring and maintenance activities. The monitor stations passively track all GPS satellites in view, collecting ranging data from each satellite. This information is transmitted to the MCS, where the satellite ephemeris and clock parameters are estimated and predicted. The MCS uses the ground antennas to periodically upload the ephemeris and clock data to each satellite for retransmission in the navigation message. Communications between the MCS and the MSs are typically accomplished via the U.S. Defense Satellite Communication System (DSCS).

2.1.3 User Segment

The user segment mainly refers to GPS receivers, which are dedicatedly designed hardwares to receive, decode, and process the GPS satellite signals. The most general receivers are stand-alone hand-held facilities, which are usually for personal use. Of course, GPS receivers can also be integrated with or embedded into other systems. Due to different applications, e.g., general navigation, accurate positioning, time transfer, surveying and satellite attitude estimation, the performance as well as the price of GPS receivers can have significant difference.

2.2 GPS Positioning Principle

The general goal of a GPS receiver is to obtain the user's position, which is usually described by longitude, latitude and altitude, by which the user is able to determine his or her location in a map.

The positioning principle behind GPS is based on Time of Arrival (TOA). A similar but simpler case is shown in Fig.2-2 [17], which is one-dimensional, i.e., the user can only move left or right along the x -axis. In order to determine the user's position on the x -axis, one needs to know two reference points and the corresponding distances from them to the user's position. As indicated in Fig.2-2, X_1 and X_2 are the two reference points, and d_1 and d_2 are the corresponding distance from them to the unknown user position respectively. The user position thus can be determined uniquely, because there is only one point on the x -axis, which is with a distance of d_1 to the reference X_1 and a distance of d_2 to the reference X_2 .

Fig.2-2 One-dimensional case

Similarly, Fig.2-3 represents the case for the two-dimensional coordinates [17]. In order to determine the user position, three reference points and three distances are needed, i.e., if the positions of the three reference points (X_1 , X_2 , X_3) and the corresponding distances (d_1 , d_2 , d_3) from them to the unknown user are known, the user position can be uniquely determined in a plane by finding the common intersection of the three circles.

Fig.2-3 Two-dimensional case

It can be derived that in a three-dimensional case, i.e., the real case for GPS, four reference points and four distances are needed in order to determine the user position. The equal-distance trace to a fixed point is a sphere in the three-dimensional case. Two spheres intersect to make a circle. This circle intersects the third sphere to produce two points, and by the fourth sphere, the user position can be determined from these two points.

The positioning methodology in GPS is similar to that in the three-dimensional case mentioned above. The positions of at least four satellites

can be computed by a receiver according to the demodulated navigation messages, which are transmitted by the navigation satellites. Simultaneously, the receiver is able to measure the corresponding distances from it to the four satellites, and hence the user position can be obtained.

However, the successful calculation of the user position through four reference points or satellites is based on the assumption that the four distances from the satellites to the GPS receiver are measured accurately. In fact, there is a constant unknown clock bias in the pseudorange measurement, and this constant is same to all the four pseudoranges. Under this condition, the drawn spheres, as mentioned above, will never intersect at a point. Therefore, in order to address this additional unknown constant and uniquely determine the user position, totally five satellites are needed theoretically.

While in practice, the GPS receiver is still able to calculate a user position based on only four satellites and the corresponding measured distances which are with the constant bias error. The reason is that although two possible solutions or positions are obtained when using four satellites, one of the solutions is close to the earth surface and the other one is far away from the earth. Users are always close to the earth surface, so the earth's center can be regarded as the fifth reference point, and thus four satellites are actually sufficient to calculate the user's position, even if the unknown clock bias exists.

The basic equations used for calculating the user position in GPS can be illustrated through the three-dimensional coordinates, as shown in Fig.2-4.

The four satellite positions, calculated from the navigation data, are (x_1, y_1, z_1) , (x_2, y_2, z_2) , (x_3, y_3, z_3) and (x_4, y_4, z_4) . The distances from these satellites to the user measured by a GPS receiver are d_1 , d_2 , d_3 and d_4 respectively. Let (x_u, y_u, z_u) be the position coordinates of the user, then the basic equations are given in equation (2.1).

Fig.2-4 The GPS three-dimensional coordinate

$$\begin{aligned}
 d_1 &= \sqrt{(x_1 - x_u)^2 + (y_1 - y_u)^2 + (z_1 - z_u)^2} + e \\
 d_2 &= \sqrt{(x_2 - x_u)^2 + (y_2 - y_u)^2 + (z_2 - z_u)^2} + e \\
 d_3 &= \sqrt{(x_3 - x_u)^2 + (y_3 - y_u)^2 + (z_3 - z_u)^2} + e \\
 d_4 &= \sqrt{(x_4 - x_u)^2 + (y_4 - y_u)^2 + (z_4 - z_u)^2} + e,
 \end{aligned} \tag{2.1}$$

where the variable e is the constant clock bias in the distance measurement. There are two solutions for the user position in the above equations. As has been mentioned earlier on, one solution will be far away from the earth, and the other one, which is the desired position, is close to the earth surface.

The only question that remains is how to measure the distance between a satellite and the receiver, i.e., the estimation of the pseudorange. The concept is rather simple and easy to be understood, which can be illustrated from the following basic distance equation [18],

$$d_i = c \cdot [T_{R_i} - T_{T_i}] \quad (\text{m})$$

where,

c : speed of light, i.e., 2.99792458×10^8 (m/sec);

T_{Ri} : signal receiving time of the i -th satellite;

T_{Ti} : signal transmitting time of the i -th satellite.

Therefore, the receiver's main task is to extract the transmitting and the receiving time from the received GPS signals, and then using the above equation, one can theoretically estimate the range between a satellite and the receiver.

2.3 GPS Signal

2.3.1 Time Domain Description of GPS signal

In order to cater to the accuracy requirements of GPS, the signals used must possess the necessary properties [19]:

- 1) Tolerance to signals from other GPS satellites sharing the same frequency band; i.e., multiple access capability.
- 2) Tolerance to some level of multipath interference – there are many potential sources of multipath reflection.
- 3) Tolerance to reasonable levels of unintentional or intentional interference, jamming, or spoofing by a signal designed to mimic a GPS signal.
- 4) Ability to provide ionosphere delay measurements

Basically, two types of signals are selected for GPS, i.e., the C/A code and the P code. The P code is not available to civilian users but the military, so only the C/A code will be discussed here.

The signal on L1 band contains both C/A code and P code, which are in quadrant phase of each other and can be written as [20]:

$$S_{L1} = A_p P(t) D(t) \cos(2\pi f_1 t + \Phi) + A_c C(t) D(t) \sin(2\pi f_1 t + \Phi) \quad (2.2)$$

where A_c is the amplitude of C/A code; $C(t)$, equal to +1 or -1, represents the chips of C/A code; $D(t)$, equal to +1 or -1, is the navigation data code. A_p and $P(t)$ have similar meanings except that they correspond to P code. Φ is the initial phase, and f_1 is the L1 frequency, i.e., 1575.42 MHz. P code is used only by military, so only the latter part in equation (2.2) is useful for civilian user. The carrier frequencies in all the 24 satellites are same, but each satellite transmits one unique C/A code. The concept of generating GPS signal is shown in Fig.2-5 [21].

Fig.2-5 Concept of generating GPS signal

2.3.2 Signal Power Level and Noise Estimation

The minimum power levels of the signals received on the earth are listed in Table 2.1 [20].

Table 2.1 Power Level of GPS Signals

	P code	C/A code
L1	-133dBm	-130dBm
L2	-136dBm	-136dBm*

* Presently not in L2 frequency

A typical diagram for estimating receiver noise is shown in Fig.2-6 [19]. First, the GPS signal hits the antenna, and then is passed to the bandpass filter and the low noise amplifier (LNA) through a transmission line.

Fig.2-6 C/N₀ computation

Where,

P_c = Received signal power at antenna output;

C = Received signal power at preamplifier input;

N_0 = Received noise density at preamplifier input.

The received noise density (one-sided) is $N_0 = KT_{eq}$, where T_{eq} is the equivalent noise temperature and is defined as:

$T_{eq} = T_A/L + (L-1)T_0/L + T_R$ °K, where T_A is the antenna noise temperature; L is the power transmission loss between the antenna and the preamplifier. T_0 is ambient temperature of the transmission line and typically equals 290 °K; T_R is receiver noise temperature, which equals $T_0 \cdot (F-1)$ and here F is the receiver noise figure. All temperatures are in °K. The value of Boltzman's constant is $K = -228.6$ dBW/°K-Hz.

For instance, if the LNA receiver noise figure is 1.0 dB, then $F = 1.259$, $F-1 = 0.259$, and $T_R = 75.1$ °K. For a typical example of small line loss, $L = 1.1$ (0.4dB), and $T_A = 130$ °K, the equivalent noise temperature and noise density for this example are,

$$T_{eq} = 130^\circ \text{K} / 1.1 + (0.1 / 1.1) \cdot 290^\circ \text{K} + 75.1^\circ \text{K} = 23.42 \quad \text{dB}$$

$$N_0 = KT_{eq} = -205.2 \text{ dBW/Hz} .$$

The C/A signal power density (energy per chip) for the specified $P_s = -160 \text{ dBW}$ and $f_c = 1.023 \times 10^6 \text{ Hz}$ is -220.1 dBW/Hz . Compare this with the noise density of -205.2 dBW/Hz , the C/A signal spectrum even at its spectral peak, is 14.9 dB below the noise power spectral density. Therefore, the GPS signal cannot be directly observed from a spectrum analyzer. Even when the signal is amplified to a reasonable power level, the spectrum of the C/A code cannot be observed because the noise is stronger than the signal.

2.3.3 C/A Code Properties

The C/A code used in GPS is a kind of pseudorandom noise code, also known as Gold codes [22]. Its chip rate is 1.023 MHz with rectangular pulse shape, so in the frequency domain, the width of the spectrum main lobe is 2.046 MHz . The total code period contains $1,023$ chips and lasts for 1 ms , i.e., the C/A code is 1 ms long, so one C/A code chip is about 977.5 ns ($1/1.023 \text{ MHz}$).

The most important property of the C/A code is its autocorrelation, which has a high narrow peak. The cross correlation value between any two different C/A codes is very small. Thus, it is very suitable for signal acquisition. High autocorrelation peak and low cross correlation value also make it effective in acquiring a weak GPS signal which is accompanied by a strong interference signal, because it makes it possible that the autocorrelation peak value of the weak signal is larger than the cross correlation value of the strong signal.

The Gold codes are not orthogonal but near orthogonal, so its cross correlations are not zero but have small values. Table 2.2 lists all the possible cross correlation values of the Gold code [20].

One complete period of C/A code in GPS consists of 1023 chips, so $n = \text{even}$

= 10 and $P = 1023$. It is easy to calculate the cross correlation values of the C/A code based on the equations in table 2.2: $-65/1023$, $-1/1023$ and $63/1023$. This is further illustrated in the figures below where Fig.2 - 7 (a) shows the normalized autocorrelation function of the C/A code 1, and the cross correlation function between C/A 1 and C/A 2 are shown in Fig.2 - 7 (b).

Table 2.2 Cross Correlation of Gold Code

Code Period	Number of Shift Register Stages	Normalized Cross Correlation Level	Probability of Level
$P = 2^n - 1$	$n = \text{odd}$	$\begin{cases} \frac{2^{(n+1)/2} + 1}{P} \\ -\frac{1}{P} \\ \frac{2^{(n+1)/2} - 1}{P} \end{cases}$	<p>0.25</p> <p>0.5</p> <p>0.24</p>
$P = 2^n - 1$	$n = \text{even}$	$\begin{cases} \frac{2^{(n+2)/2} + 1}{P} \\ -\frac{1}{P} \\ \frac{2^{(n+2)/2} - 1}{P} \end{cases}$	<p>0.125</p> <p>0.75</p> <p>0.125</p>

(a) Autocorrelation of C/A code 1

(b) Cross correlation between C/A code 1 and C/A code 2

Fig.2-7 Auto and cross correlation of C/A code

In Fig.2-7 (a), the maximum value of the autocorrelation peak is 1 (normalized by 1023), and the rest of the correlation has three values, i.e., $63/1023$, $-1/1023$ and $-65/1023$. The cross correlation shown in Fig.2-7 (b) only has the three values, i.e., $63/1023$, $-1/1023$, $-65/1023$ and has no high peak.

Fig.2-7 indicates that if the received signal contains a certain C/A code, the correlation between the received signal and the local code will produce a high peak. While, if the received signal does not contain the C/A code, the correlation value is very small. This property is utilized by GPS receivers to identify different satellites, since each navigation satellite transmits a unique C/A code.

As has been introduced, the C/A code is modulated onto the carrier in a form of Bi-Phase Shift Keying (BPSK), which means that it changes the carrier phase between 0 and π at a rate of 1.023 MHz. The navigation data bit is also bi-phase code, whose frequency is 50 Hz, i.e., each navigation data bit is 20ms long. C/A

code and navigation code exist on L1 frequency simultaneously, and the beginning of a navigation data bit is synchronized with the beginning of the C/A code, so one navigation code bit contains exactly 20 C/A codes. If there is a navigation data bit transition at an epoch, the C/A codes in the following 20ms will be “contrary” to those codes 20ms prior to the epoch, which means that the phase of the C/A codes before the epoch and the phase of C/A codes after the epoch are different by $\pm\pi$.

2.4 GPS Signal Acquisition

In order to calculate the user position, a GPS receiver needs to obtain the signal delay and decode GPS navigation data, which is the main task of signal tracking. But in order to make the signal tracking loop start running, two important parameters must be passed to it. One parameter is the rough delay or rough beginning point of the C/A code, and the other is the rough carrier frequency of the received signal. The task of achieving these two parameters is accomplished by the block prior to the signal tracking, called signal acquisition.

In acquisition, Doppler frequency shift is the most important issue having significant impact on the acquisition, so in this section, it will be introduced first.

2.4.1 Doppler Frequency Shift

In the previous sections, the L1 carrier frequency of GPS was clearly stated as 1575.42 MHz. However, due to the motion of navigation satellites relative to a GPS receiver, the carrier frequency received at the receiver side is actually uncertain. This phenomenon is named as Doppler frequency shift [23].

Briefly, the Doppler frequency shift can be estimated through equation (2.3).

$$\Delta f_{\text{doppler}} = \frac{f_r \cdot v_d}{c} \quad (2.3)$$

where, f_r is the L1 frequency, i.e., 1575.42 MHz; c is the speed of light; v_d is the satellite velocity component toward the user who is on the earth. v_d can be illustrated through Fig.2-8 [23].

Fig.2-8 Concept of Doppler frequency shift

In Fig.2-8, r_e and r_s are the radius of the earth and the satellite orbit respectively; v_s is the velocity of navigation satellite, and v_d , called Doppler velocity, is the velocity component of v_s toward the user.

It can be calculated that the maximum Doppler velocity occurs when the satellite is at the horizon position, i.e., position A in Fig.2-8, and the value equals to 929 m/s; the minimum Doppler velocity occurs when the satellite is exactly overhead, i.e., position B in Fig.2-8, and the value is zero [23].

Thus, based on equation (2.3), the maximum Doppler frequency shift can be calculated.

$$\Delta f_{\max} = \frac{f_r \cdot v_d}{c} = \frac{1575.42 \times 10^6 \times 929}{3 \times 10^8} \approx 4.9 \text{ KHz}.$$

Therefore, for a stationary observer on the ground, the maximum Doppler frequency shift is around ± 5 KHz. For a high speed aircraft, the Doppler frequency shift can be up to ± 10 KHz when the aircraft moves toward the satellite at about 2,078 miles/hr. So, if a GPS receiver is designed for a low-speed vehicle, the frequency uncertainty or the Doppler frequency shift is ± 5 KHz around the L1 central frequency, and if the receiver is to be used with a high-speed aircraft, the receiver has to search the frequency in the range of ± 10 KHz around the L1 frequency to determine the exact carrier frequency.

Similarly, the Doppler frequency shift on the C/A code can be estimated as,

$$\Delta f'_{\max} = \frac{f_c \cdot v_h}{c} = \frac{1.023 \times 10^6 \times 929}{3 \times 10^8} \approx 3.2 \text{ Hz},$$

which is very small. Even for high-speed aircraft, it is only around 6.4 Hz.

The above discussion indicates that when the signal reaches the receiver's antenna, the exact frequency has to be determined in order to perform signal acquisition. In addition, the C/A code has a phase delay in the received signal, due to the distance between GPS satellites and receiver. These two aspects determine that the process of GPS signal acquisition is a two-dimensional search, i.e., the frequency search to determine the Doppler frequency shift and the code search to determine the code phase delay.

2.4.2 Time Domain Acquisition Methodology

There are numerous algorithms for GPS signal acquisition [24-28], but the most straightforward one is the sliding window based time-domain correlation method, which will be introduced in this section.

Fig.2-9 shows the normalized ideal autocorrelation function of a C/A code which contains 1023 chips, where the side lobes are neglected and T_c is one C/A

code chip length. It reveals that if the misalignment between two C/A codes is more than one chip, the correlation value will be approximately zero; if the two C/A codes are exactly aligned, i.e., the two codes are synchronized, the correlation will output the peak value of the autocorrelation function shown in Fig.2-9.

Fig.2-9 Ideal autocorrelation function of C/A code

As has been mentioned, the process of satellite signal acquisition is a two-dimensional search, i.e., frequency direction and code direction. This search process is illustrated in Fig.2-10, in which each small square represents a correlation operation, which dumps a single correlation value.

Fig.2-10 Acquisition based on two-dimensional search

The entire frequency range, i.e., $(f_c - \Delta f, f_c + \Delta f)$, is divided into a number of small frequency bins, and each frequency bin represents a possible carrier frequency. Here, f_c is the carrier central frequency, which is determined by the receiver RF front end and is usually at the Intermediate Frequency (IF) band, e.g., 4.309 MHz [29]. In each frequency bin, the receiver generates a local reference signal, i.e., a carrier modulated by a certain C/A code, and then attempts to align it with the incoming signal by sliding the reference signal in time and performing correlation. After searching is completed, the maximum correlation value from all the outputs is selected and compared with the pre-defined threshold. If it is higher than the threshold, the corresponding C/A code being used in the local reference signal will be reported found in the received signal. The frequency bin and the code phase corresponding to the maximum correlation value, e.g., the dark square at (f, n) in Fig.2-10, is the obtained coarse frequency and coarse delay of the C/A code. How accurate the frequency and the delay are is mainly determined by the step size used in the searching.

2.4.3 Step Size in Signal Acquisition [30]

Step size here refers to, in acquisition, how much one should slide the local C/A code each time in the code search and how much the frequency bin width is used in the frequency search.

In frequency search, using a narrower step for searching means to take more steps to cover the entire desired frequency range, i.e., ± 10 KHz around the central frequency, and hence will consume more time. A coarse step size can reduce computation load, but will result in a relatively poor sensitivity. This can be illustrated by the following equations.

$$S_{in}(t) = C(t - \tau_0) \cdot \cos(2\pi \cdot f_1 \cdot t) \quad (2.4)$$

$$S_{\text{local}}(t) = C(t - \tau) \cdot \cos(2\pi \cdot f_2 \cdot t) \quad (2.5)$$

Equation (2.4) represents the received signal (navigation data are neglected), where $C(t)$ is the C/A code; τ_0 is the signal delay; f_1 is the received carrier frequency. Equation (2.5) is the constructed local reference signal, where f_2 is the estimated incoming frequency and τ is a variable representing the delay of the local signal. To calculate correlation, $S_{\text{local}}(t)$ needs to be multiplied with $S_{\text{in}}(t)$, the result of which is then integrated, as shown in equation (2.6).

$$A(\tau) = \int_{T_1}^{T_2} C(t - \tau_0) \cdot C(t - \tau) \cdot \frac{1}{2} \cdot \cos[2\pi \cdot (f_1 - f_2)t] \cdot dt \quad (2.6)$$

In equation (2.6), the frequency component of $(f_1 + f_2)$ is neglected, because the integrator acts as a low pass filter. Ideally if f_2 exactly equals to f_1 , $A(\tau)$ will be the correlation value between the incoming and the local codes. However, if there is an offset between f_1 and f_2 , the correlation result will be changed. Let's say $\Delta f = f_1 - f_2 = 1$ KHz. For C/A code, the typical integration interval is 1 ms, during which the $\cos[2\pi(f_1 - f_2)t]$ signal will vary one cycle. It is known that the integration of a pure cosine waveform in one cycle is zero, so the final correlation value in equation (2.6) will be very small. In order to address this problem, the frequency step size must be reduced. It is chosen that the maximum variation of the signal in (2.6), which is with a carrier frequency of $(f_1 - f_2)$, should be 0.5 cycle during the integration interval. If the data record is 1ms, in order to keep the maximum variation within 0.5 cycles in 1ms, the frequency step size or the frequency bin width used in acquisition should be 1 KHz, because under this bin width, the maximum frequency difference between the input signal and the local signal is ± 500 Hz, and a 500Hz carrier can vary 0.5 cycles at most within 1ms. Similarly, if the data record is 10 ms, a searching frequency step size of 100 Hz will fulfill the requirement.

For code search, recall that the correlation property in Fig.2 - 9 show that if the code offset between two C/A codes is more than one chip, the correlation

value will be approximately zero. Therefore, for code search, the maximum step size is one PRN code chip. Finer step size will let the replica code approach the incoming code more accurately, but with a price of high computation load. So it is similar to that in frequency search, i.e., there is a tradeoff between the search accuracy and computation load.

After the accomplishment of signal acquisition, the carrier frequency and the beginning point of the C/A code in the received signal is roughly determined. These two values are then passed to the tracking loop, which is in charge of further fine search and signal demodulation.

2.5 GPS Signal Tracking

The major task of tracking, including carrier tracking and code tracking, is to demodulate the navigation data from the received signal. Carrier tracking is usually implemented through a Phase Lock Loop (PLL), which adjusts the carrier phase of the replica signal to follow that of the incoming signal. The PRN code of the GPS signal is conventionally tracked by a Delay Lock Loop (DLL), which keeps aligning the local code with the incoming code.

2.5.1 Delay Lock Loop (DLL) [31]

Typically, the signal acquisition only determines the beginning of the C/A code in units of integer number of code chips, and there is still a code delay ambiguity less than one chip. One chip of C/A code corresponds to approximately 300 meters in GPS, so it will be unable to provide satisfactory accuracy of user position if this delay ambiguity is not determined. To calculate the user position more accurately, the beginning point of C/A code usually is re-determined by the

DLL in a standard GPS receiver.

A straightforward idea of code tracking may be to shift the local code by a fine step size each time, calculating the correlation between it and the received code, and then obtain the accurate code delay when the maximum correlation value is found. However, two questions exist in this method. The first is that the maximum correlation value may not correspond to the actual beginning point of the C/A code, due to noise or interference. The second is that the relationship between a correlation value and the code delay is not one-to-one, caused by the symmetry of the code autocorrelation function, so it is ambiguous as to which direction the local code should be shifted.

Fig.2-11 Early and late correlation function

Fig.2-12 Discrimination function

Fortunately, a delay lock loop provides a good solution to the above two problems. The idea behind the code tracking is quite simple. The property of C/A code autocorrelation will be used again, just as that in acquisition. In the literature, the local replica code is also widely called prompt code, and its ideal autocorrelation function was already shown in Fig.2-9. DLL uses one early correlator and one late correlator to assist to align the prompt code with the received code. The early correlator is obtained by shifting the prompt code $d/2$ earlier, and the late code is obtained by shifting the prompt code $d/2$ later. The early and late autocorrelation function is shown in Fig.2-11 respectively, where 'd', generally named as early-late spacing, conventionally equals to one C/A code chip length. If we define $E(\tau)$ as the early correlation function and $L(\tau)$ as the late correlation function, then the result of $E(\tau)-L(\tau)$ can be obtained as that in Fig.2-12. The linear segment from peak P_1 to peak P_2 in Fig.2-12 is widely named discrimination function.

It is seen that the discrimination function represents a one-to-one relationship between the code phase offset and the early-minus-late correlation value. Its characteristics are monotone and linear. The zero point of the discrimination function means that the code offset between the received C/A code and the local prompt code is zero, in other words, the local prompt code is exactly aligned with the received C/A code when the discrimination function outputs zero.

The operation process of the delay lock loop can be simply described as: generate a local prompt code, an early and a late code, calculate the correlation value between the received code and the early-minus-late code, and then use this correlation value to generate a correction information, which is used to shift the local prompt code earlier or later so that the early-minus-late correlation value converges to zero step by step. When the discrimination function reaches zero, the correction value is also zero, so the loop enters a stable state, which is called

‘locked’. The total code phase that the prompt code has been shifted is the fine time delay of the received signal. The derivation of the correction information is a strict mathematic process; for details please refer to [31].

A generic DLL architecture is shown in Fig.2-13 [31], which follows the principle that has been mentioned above.

Fig.2-13 Coherent DLL architecture

2.5.2 Phase Lock Loop

Fig.2-14 Generic tracking loop architecture

The generic tracking loop [32] diagram is depicted in Fig.2-14, which is to make the phase of the reference signal follow the phase of the input $x(n)$. The input signal is run through a discriminator to produce the error value $\epsilon(n)$. The discriminator is represented as a simple subtraction with a gain of k_p . The output

is low-pass filtered to reduce noise. The resulting value is used by the numerically-controlled oscillator (NCO), with a gain of k_v , to produce a new replica signal for the next iteration of the loop. We want the loop output, $y(n)$, to be as close to zero as possible.

Fig.2-15 Phase lock loop in GPS

The phase lock loop block diagram widely adopted in GPS receivers is shown in Fig.2-15 [33], which is a combination of the generic tracking loop and I/Q channel demodulation. Given that the input signal $x(n)$, we strip the code and carrier, leaving navigation data as the output $y(n)$. The complex signal $I(n) + jQ(n)$ is passed through a discriminator and filtered to determine the carrier tracking error $\varepsilon_f(n)$, which is then used by the NCO to generate a new reference signal.

Fig.2-16 Costas tracking loop

The characteristics of a PLL are determined by the filters and the phase discriminator adopted. In GPS, an ‘integrate and dump’ filter is typically used in the I-arm and Q-arm respectively, and the typical integration interval is 1ms, which means that the filter outputs one single value every one millisecond. The task of discriminator is to determine the phase difference between the input carrier and the local carrier in one integration interval. There are several types of discriminators available [18], and the most popular one is the arctangent discriminator. The low pass filter after the discriminator, called loop filter, is usually a one-order or two-order filter with a bandwidth that is suitable for GPS [34]. The architecture described here is shown in Fig.2-16, widely known as Costas Phase Lock Loop. In the locked mode, the demodulated data bits will be present on its in-phase arm.

The $\text{atan}(Q/I)$ discriminator is suboptimal, but good at high and low SNR. The most important advantage of $\text{atan}(Q/I)$ is its ability on handling the 180° phase shifts caused by navigation data bit transition and continuing tracking, which makes it very suitable for demodulating the GPS signal. Its major disadvantage is the higher computational burden when compared with other discriminators [18].

2.6 A Conventional GPS Receiver Architecture [35]

Fig.2-17 A generic GPS receiver architecture

To organize all the individual function blocks introduced in this chapter together, this section will present an overall description on the architecture of a generic GPS receiver, which is shown in Fig.2-17 [35].

A) Antenna and Pre-amplifier

The antenna is normally a right-hand circularly polarized to match the incoming GPS signals, and the pattern is essentially hemispherical in most applications.

As stated, the minimum received satellite signal can be about -160 dBW, so the pre-amplifier is usually a Low Noise Amplifier (LNA), which determines the noise figure of the entire receiver. A typical LNA has a gain on the order of 25-40 dB and a noise figure less than 2dB, but the other components and other associated losses usually result in a total noise figure of 3-4 dB.

B) Front End

What are usually included in this part are filters and down converters. The GPS signal must pass through several filters to reduce the unwanted noise. The center frequency of these band pass filters is either at the L1 frequency or the intermediate frequency (IF) generated from the RF down converter. The filter centered on the L1 frequency typically has a pass band of 2 MHz – 20 MHz, by which the out-of band noise is removed. At each stage of the down converting process, a filter is also needed to wipe off the corresponding noise added by the hardware.

The RF down converter down converts the L1 frequency to lower intermediate frequencies in one, two or multiple stages, determined by designers and applications. Adequate image suppression and general band pass filters can be used in the case of multistage down conversion, and the final IF can be placed to near base band, e.g., 4 MHz. The final IF obtained from a single-stage down

conversion is usually higher, e.g., 30 – 100 MHz.

C) A/D Conversion

By A/D converter, the analog signal is sampled and quantized. Most low-cost commercial receivers employ 1-bit sampling in narrow bandwidths, i.e., 2 MHz. High-end receivers typically use from 1.5-bit to 3-bit sampling in bandwidths ranging from 2 – 20 MHz.

D) Baseband Signal Processing

Baseband signal processing is related closely to signal acquisition and tracking, the methodology of which has been illustrated in the earlier sections. Usually, a GPS receiver uses dedicated designed hardware, e.g., numerically controlled oscillators, correlators, Integrate and Dump accumulators, filters and DSP to accomplish the measurements, acquisition and signal tracking. Integration intervals are different for acquisition and tracking. The integration prior to data bit synchronization, called pre-detection integration, is typically less than the data bit period. This is done to avoid correlation loss caused by the navigation data bit transitions. The integration after data bit synchronization, called post detection integration, is done typically over a single data bit period. The outputs from the baseband signal processing include pseudorange, delta-range, carrier phase, and navigation data.

E) Navigation Processing

In this block, the demodulated navigation data will be processed, including sub-frame matching, parity check, polarity check, reading navigation data etc., based on which the satellite positions are calculated. Satellite positions together with the corresponding pseudoranges are used to calculate user position in the typical format of longitude, latitude and altitude.

2.7 Theoretical Accuracy Limit for C/A Code Ranging

This section will briefly introduce the theoretical limit on the C/A code pseudoranging [36], which provides a potential reference to evaluate the performance of a positioning technique in GPS. If the ranging accuracy limit has been almost reached, it will be un-meaningful to put more costly efforts into further research. Inversely, if there is still a big gap between the pseudorange accuracy of the current GPS technology and the theoretical limit, the designers still need to work hard to improve the performance of the GPS receiver.

Assume that the GPS signal has been down converted to baseband, i.e., the signal model is,

$$Z(t) = c(t - \tau) + n(t) \quad (2.7)$$

where $c(t)$ is the baseband C/A code, and $n(t)$ is zero-mean Gaussian noise in a frequency range of $[-\omega, \omega]$ Hz, and zero otherwise. Furthermore, the frequency spectrum of the C/A code is $[-\omega_c, \omega_c]$, which is included in $[-\omega, \omega]$. Navigation data are not included in (2.7), because they can be obtained from the outputs of a GPS receiver, and in turn can be multiplied with the raw GPS signal to remove the navigation bits.

In a certain time interval, say $[0, T]$, the discrete version of $Z(t)$ in (2.7) can be obtained after performing sampling on $Z(t)$, i.e.,

$$Z_i(t) = c_i(t - \tau) + n_i(t) \quad i = 0, 1, 2, \dots, N \quad (2.8)$$

Based on the equation (2.8) and the principle of maximum likelihood, the theoretical ranging accuracy limit can be derived as:

$$\sigma = \frac{(\sqrt{2}/4)}{\sqrt{K} \cdot \sqrt{P/N_0} \cdot \sqrt{W_c} \cdot \sqrt{T}} \quad (2.9)$$

where T is the integral interval or the time interval in which the signal is observed, K is a constant and equals 512×10^3 , P is the power level of the C/A code and equals to $c^2(t)$, N_0 is the flat power spectral density of the noise in the frequency range of $[-\omega, \omega]$ Hz, and ω_c is the single-sided width of the C/A code frequency spectrum. The σ in (2.9) is in units of seconds, and can be converted to meters by multiplying by the speed of light, i.e., 3×10^8 meters/sec.

Fig.2-18 plots the accuracy limit based on equation (2.9), where the full bandwidth of the C/A code, i.e., $\omega_c = 10$ MHz is assumed, and P/N_0 is assumed to be twice C/N_0 (because noise is split into I-arm and Q-arm, while the useful C/A codes only appear at the I-arm of the demodulator).

For a typical C/N_0 , e.g., 45 dB-Hz and an observation interval of one second, Fig.2-18 reveals that the error limit is about 16.6 cm. [36] also points out that if only considering noise as the error source, modern receivers, e.g., Narrow Correlator, are nearing the theoretical accuracy limits. But in real-life, there are other error sources degrading the ranging accuracy, among which multipath is the most significant one and is to be discussed in the following chapters.

Fig.2-18 Range error limits versus observation time

2.8 GPS Modernization

For many years, the current GPS technology has been recognized as the future of navigation for many peaceful applications, including those for various modes of transportation. The civil aviation community has been one of the main benefactors of GPS due to its flexibility and worldwide applicability. As a direct result of these benefits to the civil community, U.S. Vice President Al Gore announced a GPS modernization effort in January 1999 to extend the capabilities of GPS even further than those currently enabled by the existing GPS constellation. In this section, a brief introduction to the GPS modernization will be presented.

2.8.1 GPS Modernization Overview

One of the main features of the modernization effort is that two new navigation signals will be available for civil use. These signals will complement the existing civilian service broadcast at 1575.42 MHz (L1). The first of these two new signals will be a C/A code located at 1227.60 MHz (L2), and will be available for general use in non-safety critical applications. This will be available beginning with the initial GPS Block IIF satellite scheduled for launch in 2007 [37].

The other signal, located at 1176.45 MHz (L5), which falls in a band protected worldwide for aeronautical radio navigation, and therefore will be protected for safety-of-life applications. Additionally, it will not cause any interference to existing systems. Therefore, with no modification of existing systems, the addition of L5 will make GPS a more robust radio navigation service for many aviation applications, as well as all ground-based users (maritime, railways,

surface, shipping, etc.).

At the current GPS satellite replenishment rate, all three civil signals (L1-C/A, L2-C/A, and L5) will be available for initial operational capability by 2010 and for full operational capability by approximately 2013.

2.8.2 L5 Signal Structure

Fig.2-19 Block diagram of GPS L5 signal generator

The L5 signal is at 1176.45 MHz, and its PRN code has 10230 chips with a chip rate of 10.23 MHz (repetition period is 1ms). The minimum received signal power is expected to be -124 dBm. Fig.2-19 [38] shows the L5 signal generator consisting of a data channel (I5) and a data-free (Q5) channel.

The I5 channel in Fig.2 - 19 indicates that it is constructed by the PRN code modulated by the navigation data at 100 symbols per second and a 10ms-long Neuman-Hofman (NH) sequence (0000110101). To obtain the 100 sps data symbols, the 50bps data are convolutionally encoded with rate 1/2 and constraint length of 7, and the 50bps data are the original navigation data from a satellite

(276 bits per 6 seconds) plus CRC (24 bits per 6 seconds).

The Q5 channel is relatively simple, and it consists of 20 periods of PRN codes modulated by the NH20 code (00000100110101001110). Similar to that in the I5 channel, each chip of the NH20 code is 1ms long and is represented by one cycle of the PRN code (10230 chips).

Both the baseband I5 and Q5 data are then modulated to the 1176.45 MHz carrier using the Quadratic-Phase Shift Keying (QPSK) method, and transmitted via the antenna.

2.8.3 Benefits of L5

L5 will provide significant benefits above and beyond the capabilities of the current GPS constellation, including the precision approached navigation worldwide, increased availability of precise navigation operations in certain areas of the world, and interference mitigation.

For the current civilian applications of GPS, people frequently suffer from the lack of the GPS signals, especially in urban canyons and indoor environments. In the future, GPS receivers will be more robust to perform positioning, because the addition of L5 should significantly increase the ability in providing aviation and other applications with high-quality signals.

The L5 is also a complement to the Wide Area Augmentation System (WAAS) of US. The current receivers in the US usually obtain the ionospheric corrections through WAAS. When L5 is available, avionics will have one more option by using dual frequency, i.e., L1 and L5, to correct the ionospheric delay.

Similarly, in the event of interference on one of the L1 and L5 channels, the avionics will be able to conduct precise positioning using the other frequency. Therefore, the addition of L5 will have a profound impact on both the new and existing applications of GPS for navigation.

Summary

This chapter introduced the fundamental knowledge to understand the GPS technology. A GPS receiver is composed of many function blocks, but only the signal acquisition and the signal tracking, which significantly affect the estimation of pseudoranges, were introduced in details, because they, especially the tracking, can be affected by multipath, which is the main topic to be discussed in the following chapters.

Chapter 3 Multipath Overview

The fundamentals of GPS have been presented in the previous chapter, and as mentioned, for precise positioning, GPS receivers face a challenge in most applications, i.e., multipath, which significantly affects the accuracy of the obtained user position. This chapter will focus on the issue of multipath and the typical existing techniques for multipath mitigation.

3.1 Introduction to Multipath in GPS

3.1.1 Concept of Multipath

Multipath refers to the signals which arrive at a receiver antenna via multiple paths, due to reflection or diffraction. An example of multipath signals caused by the ground and a building wall is shown in Fig.3-1 [39, 40].

Fig.3-1 Multipath signals sensed by receiving antenna

3.1.2 Multipath Signal Characteristics

Some important characteristics of multipath are listed below [41, 42]:

- i) Multipath signals will always arrive after the direct path signal, because they travel a longer propagation path.
- ii) A multipath signal usually is weaker than the direct path signal, because some signal power will be lost due to the reflection. It could be stronger if the direct path signal is hindered in some way.
- iii) If the excess delay of the multipath signal is less than two PRN code chips, the locally generated receiver signal will partially correlate with it. If the delay is greater than two chips, the correlation power will be negligible.
- iv) The multipath signals are additive, i.e., the received composite signal is constructed by adding the multipath signals onto the direct signal.

3.1.3 Multipath Channel Description

The time-variant impulse response in equation (3.1) denotes the multipath propagation channel [43, 44]:

$$h(\tau, t) = \sum_{m=0}^N a_m(t) e^{-j2\pi f_1 T_m(t)} \delta(\tau - T_m(t)) \quad (3.1)$$

where $a_m(t)$ represents the time-varying attenuation factor of the m -th propagation path; $T_m(t)$ is the time varying propagation delay of the m -th path; f_1 is the L1 carrier frequency, and $\delta(t)$ represents the Delta function.

The GPS signals are delayed by the propagation time T_m before reaching the receiver. T_m is also associated with a frequency shift resulting from the Doppler Effect due to the motion of satellite and receiver. In the presence of N multipath signals, the signal transmitted from one satellite and presented at the receiver

signal processing stage input is given by:

$$S(t) = \text{Re} \left\{ \sum_{m=0}^N a_m D(t - T_m) p(t - T_m) e^{j(\omega \cdot t + \phi_m)} \right\} \quad (3.2)$$

where:

- the subscript m accounts for the presence of multipath propagation;
- a_m is the possible time-varying attenuation factors of the m -th component resulting from the reflection. For multipath signal, a_m is typically Rayleigh distributed;
- $D(t)$ denotes the navigation message broadcasted by the satellite and $p(t)$ is its spread-spectrum C/A code;
- T_m is the total time-varying propagation delay of each signal component. In fact, we have $T_m = \tau_m + kT_c$ (k is an integer). T_c denotes one chip period of the C/A code and τ_m is the raw measurement computed by the code tracking loop;
- ω is the received radian carrier frequency including the Doppler frequency shift;
- Φ_m is the time-dependent phase of the m -th component of the received signal. It is a function of the propagation delay T_m , of the initial signal phase θ_m and of the change of phase due to the reflection θ_r , that is, $\phi_m = -\omega \cdot T_m + \theta_m + \theta_r$.

As we are only interested in the measurement of navigation errors, the influence of the 50bps data message on the tracking processes will be neglected in the following [45, 46]. Furthermore, only the case of slow fading is considered, i.e., $B_F \ll B_L$. B_F represents the fading bandwidth and B_L is the tracking loop bandwidth [47]. The following discussions in this chapter will be based on equation (3.3) [48].

$$S_r = \sum_{m=0}^N a_m \cdot p(t - T_m) \cdot \cos[\omega \cdot t + \theta_m] \quad (3.3)$$

3.2 Analysis on the Multipath Tracking Error

3.2.1 Multipath Influence on Carrier Phase

(a)

(b)

Fig.3-2 Multipath influence on carrier phase

Fig.3-2 shows the effect of multipath on the carrier phase, where in (a) the phasor $\mathbf{D}$ represents the direct signal, $\mathbf{M}$ is the multipath phasor, and θ_m is the phase of multipath relative to the direct signal. The received signal will be the phasor sum of $\mathbf{D}$ and $\mathbf{M}$, i.e., $\mathbf{R}$, and its phase differs from that of the direct one by θ , which is defined as the carrier phase multipath error [35].

In order to determine an upper bound of the phase multipath error, it is assumed that the relative phase-rate between the direct and the multipath signal is zero, and as a result, the time dependence relating the phasors to the actual

time-domain signals, i.e., the $e^{j\omega t}$ in equation (3.3), can be ignored. Assume that the magnitudes of the direct and multipath phasors shown in Fig.3-2(a) are given by

$$\begin{aligned} D &= R(\tau) \\ M &= \alpha \cdot R(\tau - d) \end{aligned} \quad (3.4)$$

where $R(\tau)$ is the C/A code autocorrelation function, d is the relative delay of the multipath, and α is the multipath-to-direct ratio (M/D).

The multipath phasor can be decomposed into two components as shown in Fig.3-2(b), and it can be easily derived that

$$\theta = \arctan\left(\frac{\mathbf{M}_Q}{\mathbf{D} + \mathbf{M}_I}\right), \quad (3.5)$$

where $\mathbf{M}_I$ and $\mathbf{M}_Q$ are the horizontal component and the vertical component of the multipath phasor $\mathbf{M}$ respectively, i.e.,

$$\begin{aligned} \mathbf{M}_Q &= \mathbf{M} \sin(\theta_m) \\ \mathbf{M}_I &= \mathbf{M} \cos(\theta_m) \end{aligned} \quad (3.6)$$

Based on (3.4) and (3.6), (3.5) can be written as

$$\theta = \arctan\left\{\frac{\alpha \cdot R(\tau - d) \sin(\theta_m)}{R(\tau) + \alpha R(\tau - d) \cos(\theta_m)}\right\} \quad (3.7)$$

It can be easily shown that the maximum value of (3.7) is $\pi/2$. For the GPS L1 frequency, this amount equals to approximately 4.8 cm. The phase error typically is less than 1cm [35], which is very small and hence is negligible. The techniques on mitigating carrier phase multipath error can be found in [49-51].

3.2.2 Multipath Influence on Code Delay

The ideal autocorrelation function of the C/A code is triangular and symmetric. However, due to the presence of multipath, it will be distorted, i.e., skewed and asymmetric. A simple example is shown in Fig.3-3, where only a single multipath signal exists and the autocorrelation function of the direct PRN code is assumed

to be triangular and the side lobes are zero.

Fig.3-3 Distorted correlation function

The principle of a Delay Lock Loop has been discussed in Section 2.5.1, and the discrimination function of the DLL was plotted in Fig.2-12, which was obtained by subtracting the late correlation function from the early one as shown in Fig.2-11.

When a multipath signal is present, the prompt, early and late correlation functions will be replaced by the distorted composite correlation function in Fig.3-3, and the resultant discrimination function under multipath is depicted in Fig.3-4.

Fig.3-4 Discrimination function under multipath

Comparing Fig.3-4 with Fig.2-12 will see that when the distorted discrimination function outputs zero, the chip offset between the received code and the local reference code is nonzero. In other words, the DLL will lock at the point of A in Fig.3-4. This nonzero chip offset is known as the range error caused by multipath.

An analytical analysis on the tracking error will be presented next, based on a single multipath signal.

Fig.3-5 is the coherent delay lock loop architecture, in which $S(t)$ represents the received signal.

Fig.3-5 Generic coherent delay lock loop architecture.

First, assume that there is no multipath signal, and the LOS signal (neglect the navigation data) can be expressed as,

$$S(t) = A_0 \cdot P(t - \tau_0) \cdot \cos(\omega t - \varphi_0) \quad (3.8)$$

where A_0 is the LOS signal amplitude; τ_0 and φ_0 are the initial time delay and phase; ω is the signal frequency in radian/s.

The outputs of the four correlation processes in Fig.3 - 5 can easily be derived as,

$$\begin{aligned} I_P(\Delta\tau, \Delta\varphi) &= A_0 \cdot R(\Delta\tau) \cdot \cos(\Delta\varphi) \\ Q_P(\Delta\tau, \Delta\varphi) &= A_0 \cdot R(\Delta\tau) \cdot \sin(\Delta\varphi) \\ I_D(\Delta\tau, \Delta\varphi) &= A_0 [R(\Delta\tau + d \cdot T_c / 2) - R(\Delta\tau - d \cdot T_c / 2)] \cdot \cos(\Delta\varphi) \\ Q_D(\Delta\tau, \Delta\varphi) &= A_0 [R(\Delta\tau + d \cdot T_c / 2) - R(\Delta\tau - d \cdot T_c / 2)] \cdot \sin(\Delta\varphi) \end{aligned} \quad (3.9)$$

where $\Delta\tau = \tau - \tau_0$ is the time difference between the local ranging code and the received ranging code; $\Delta\varphi = \varphi - \varphi_0$ is the phase difference between the local carrier and the received carrier; $R(\Delta\tau)$ is the autocorrelation function value corresponding to the lag value of $\Delta\tau$.

Conventionally, the code tracking loop will keep adjusting the local signal to make the output of I_D in equation (3.9) equal to zero, and then the delay of the local signal is the delay of the received signal. From equation (3.9), it can be seen that this is true, i.e., when $I_D = 0$, $\Delta\tau = 0$ ($\Delta\varphi$ will be tracked by the phase lock loop and equals to zero ideally).

When a single multipath signal is present, the received signal will be given by,

$$S(t) = A_0 \cdot P(t - \tau_0) \cdot \cos(\omega t - \varphi_0) + A_1 \cdot P(t - \tau_1) \cdot \cos(\omega t - \varphi_1) \quad (3.10)$$

where A_1 , τ_1 and φ_1 are the amplitude, delay and initial phase of the multipath signal respectively.

Similarly, the outputs below will be produced by the correlators in Fig.3-5.

$$\begin{aligned}
 I_P &= A_0 R(\Delta\tau) \cos(\Delta\varphi) + A_1 R(\Delta\tau - \delta_m) \cos(\Delta\varphi - \theta_m) \\
 Q_P &= A_0 R(\Delta\tau) \sin(\Delta\varphi) + A_1 R(\Delta\tau - \delta_m) \sin(\Delta\varphi - \theta_m) \\
 I_D &= A_0 [R(\Delta\tau + dT_c / 2) - R(\Delta\tau - dT_c / 2)] \cos(\Delta\varphi) \\
 &\quad + A_1 [R(\Delta\tau - \delta_m + dT_c / 2) - R(\Delta\tau - \delta_m - dT_c / 2)] \cos(\Delta\varphi - \theta_m) \\
 Q_D &= A_0 [R(\Delta\tau + dT_c / 2) - R(\Delta\tau - dT_c / 2)] \sin(\Delta\varphi) \\
 &\quad + A_1 [R(\Delta\tau - \delta_m + dT_c / 2) - R(\Delta\tau - \delta_m - dT_c / 2)] \sin(\Delta\varphi - \theta_m)
 \end{aligned} \tag{3.11}$$

where $\delta_m = \tau_1 - \tau_0$ and $\theta_m = \varphi_1 - \varphi_0$ are the delay difference and the phase difference between the multipath and the direct signal.

Again, the tracking loop will adjust the local signal to make the I_D zero. However, the situation here is different from that in (3.9) due to the presence of the multipath signal. In equation (3.11), I_D equal to zero does not mean that $\Delta\tau$ is zero, and this $\Delta\tau$ is widely named as the tracking error or range error caused by multipath.

For the conventional DLL, the magnitude of the range error is dominantly affected by the number of multipath signals, relative multipath amplitude, relative multipath delay as well as early-late spacing. For a single weaker multipath signal and an early-late-spacing of one chip, the maximum error is one-half PRN code chip, and as the C/A code as concerned, it is equal to 147 meters or so [35], which is very large. Pseudorange error of more than 100 meters was measured and reported in [52]. Compared with the multipath caused phase error presented in the section of 3.2.1, the code error is more severe, and this is the reason why most multipath mitigation techniques focus on the mitigation of code error.

3.3 Multipath Mitigation Techniques

Multipath is the dominant error source in GPS [53], and its effects on the

pseudorandom noise code ranging have been studied since the early 1970s. The relationships between the multipath delay and the PRN code tracking error are derived in [54], and in [55] the effects of the side lobes of the autocorrelation function on the multipath errors are discussed. A code tracking loop based on the extended Kalman filter is studied in [56]. The Wavelet algorithm based multipath mitigation techniques are discussed in [57-59]. And in [60, 61], the performances of several typical multipath mitigation techniques are compared.

As illustrated in Fig.3-4, in a standard GPS receiver, the accuracy of the pseudorange measurement greatly depends on the PRN code tracking loop, so in this section, techniques focusing on improving the performance of the PRN code tracking loop will be reviewed and the analysis will be mainly concentrated on the coherent DLL.

3.3.1 Narrow Correlator

As mentioned in Section 2.5, in a standard GPS receiver, there are two correlators: one is the early correlator, and the other is the late correlator. As a compromise between the acquisition time and the noise performance as well as because of the practical considerations on implementation, the conventional receivers usually set the early-late spacing equal to one PRN code chip length. However, some researches [8, 62, 63] reveal that smaller spacing has an advantage on reducing the PRN code tracking error. The simulations of tracking errors corresponding to the narrow and the standard early-late spacing are presented in [8].

The relationship between the early-late spacing and the maximum tracking errors caused by a single multipath is given analytically in [64, 65], from which one can clearly see the effect of the early-late spacing on the tracking errors.

Under the conditions of $d \leq T_c$ and $-a_0 < a_m \leq a_0$, the tracking error expressions are,

$$\begin{aligned}
 \tau_e &= \frac{a_m \tau_1}{a_0 + a_m} & 0 \leq \tau_1 < \frac{(a_0 + a_m)d}{2a_0} = \alpha \\
 \tau_e &= a_m d / 2a_0 & \frac{(a_0 + a_m)d}{2a_0} \leq \tau_1 \leq T_c - d(1 - \frac{a_0 + a_m}{2a_0}) = \beta \\
 \tau_e &= \frac{a_m}{2a_0 - a_m} (T_c + d/2 - \tau_1) & T_c - d(1 - \frac{a_0 + a_m}{2a_0}) < \tau_1 \leq T_c + d/2 \\
 \tau_e &= 0 & \tau_1 > T_c + d/2
 \end{aligned} \tag{3.12}$$

where τ_e is the tracking error; d is the early-late spacing; a_0 is the amplitude of the direct signal; a_1 , τ_1 is the amplitude and the delay of the first multipath respectively; a_m equals to the positive or the negative amplitude of the first multipath, i.e., $a_m = +a_1$ or $a_m = -a_1$, corresponding to the in-phase and out-of-phase multipath signal respectively; T_c is one PRN code chip length.

Fig.3-6 Tracking error in terms of multipath delay

Fig.3-6 depicts the tracking error as a function of the multipath delay τ_1 . The values of $\alpha +$ and $\beta +$ equal to the values of α and β in Equation (3.12) when $a_m = +a_1$, and $\alpha -$ and $\beta -$ are the values of α and β when $a_m = -a_1$. It is obvious that a straightforward way to reduce the maximum multipath errors is to reduce the early-late spacing d , because those maxima are proportional to d , which can be seen from equation (3.12).

Fig.3-6 corresponds to the scenario when only a single multipath signal is

present. If multiple reflections exist, it will be difficult to depict a similar curve as Fig.3-6, but can be concluded that the maximum tracking errors are still proportional to d , because multipath signals are additive at the receiver antenna.

The narrow correlator is not robust in mitigating short-delay multipath signals, which are with relative delays less than 0.1 C/A code chips. Furthermore, a narrow correlator requires wide filter bandwidth in the receiver's RF front end, because limited filter bandwidth can cause the peak of the autocorrelation function to be rounded, which is not desired in the narrow correlator technique. Compared to the standard DLL, the narrow correlator technique is easier to unlock, because the effective tracking range depends on the early-late spacing. For instance, if the early-late spacing is 0.1 chips, the maximum tracking error will be 0.1 chips, while in a conventional DLL, the tracking range is one chip. Therefore, the narrow correlator will require a smaller step size in the code delay search during signal acquisition.

3.3.2 E1&E2 Tracker

From Fig.3 - 3, it can be observed that the linearity of the early edge of the composite correlation function is better than that of the late edge due to the fact that multipath signals are later than the direct signal. This property is used in [66] to mitigate multipath, and the resultant technique is called E1&E2 tracker.

Fig.3-7 Illustration on E1&E2 tracker

The idea of E1&E2 tracker is quite straightforward. Instead of using an early and a late correlator, as in the conventional tracking loop, two early correlators, called E1 and E2 respectively shown in Fig.3-7, are used to only track the early edge of the composite correlation function. With the moving of E1 and E2 on the early edge, the loop will enter the 'locked' state and the delay of the direct signal is then determined when the ratio of E1/E2 equals a prior-defined value. By setting the spacing between E1 and E2 to one chip and the ratio of E1/E2 to unity, the E1&E2 tracker will be exactly equivalent to the conventional delay lock loop, as the E1' and E2' shown in Fig.3-7.

However, there are two challenges with the E1&E2 tracker. First, the optimum ratio of E1/E2 cannot be obtained easily, because it depends on the spacing between E1 and E2 as well as the measurement environments. Although a prior-defined ratio of 0.95 for the E1&E2 spacing of 0.125 chips is mentioned in the published paper [66], it is more like an empirical conclusion. Second, moving the E1 and E2 earlier will be helpful to mitigate multipath, but the correlation value of E1 and E2 will be affected by the heavy noise.

3.3.3 MEDLL Technique

Decreasing the early-late spacing is a step in the right direction to reduce tracking errors caused by multipath, but the range errors can still be considerable. Multipath Estimating Delay Lock Loop (MEDLL) [49, 67] is another technique, which simultaneously estimates the parameters of the line-of-sight plus multipath signals. By MEDLL, one can obtain the estimation of amplitude, delay and phase of the direct signal and the multipath signals. Among these parameters, delay and phase of the direct signal are the most important.

As has been introduced, the signal received by the antenna of a

spread-spectrum receiver can be written as [67]:

$$r(t) = \text{Re}\{x(t)\} = \text{Re}\left\{\sum_{i=0}^M a_i p(t - \tau_i) \exp[j(\omega t + \theta_i)] + n(t)\right\} \quad (3.13)$$

where $x(t)$ is the transmitted signal; $p(t)$ is the spread-spectrum code; $n(t)$ is white Gaussian noise. The variables of interest are the line-of-sight signal delay τ_0 and phase θ_0 .

The mathematical principle behind the MEDLL is the Maximum Likelihood, which states that the estimate of a certain parameter with the smallest mean square error is the estimate that maximizes the conditional Probability Density Function (PDF) of $r(t)$ [68].

Taking the logarithm of this PDF, the function that has to be maximized becomes:

$$L[r(t)] = -\int_0^T [r(t) - s(t)]^2 dt \quad (3.14)$$

where $s(t) = \sum_{i=0}^M \hat{a}_i p(t - \hat{\tau}_i) \cos(\omega t + \hat{\theta}_i)$ is the estimate of the line-of-sight plus multipath signals.

In the general case of estimating M different signals, the equations for the i^{th} signal ($0 \leq i \leq M$) are:

$$\left\{ \begin{array}{l} \hat{\tau}_i = \max \left\{ \text{Re} \left[R_{xi} - \sum_{\substack{j=0 \\ j \neq i}}^M a_j R_{ij} \right] \right\} \\ \hat{a}_i = \text{Re} \left\{ R_{xi} - \sum_{\substack{j=0 \\ j \neq i}}^M a_j R_{ij} \right\} \\ \hat{\theta}_i = \arg \left(R_{xi} - \sum_{\substack{j=0 \\ j \neq i}}^M a_j R_{ij} \right) \end{array} \right. \quad (3.15)$$

where

$$R_{xi} = \int_0^T x(t) p(t - \hat{\tau}_i) \exp[-j(\omega t + \hat{\theta}_i)] dt ,$$

$$R_{ij} = \int_0^T p(t - \hat{\tau}_i) p(t - \hat{\tau}_j) \exp[-j(\hat{\theta}_i - \hat{\theta}_j)] dt.$$

What needs to be determined in equation (3.15) is the multipath number M . There are three methods to choose M . First, the maximum likelihood estimate of M is again the value that maximizes equation (3.14). Second, an alternative method is to calculate equation (3.15) for $M = 0, 1, \dots, M_{\max}$ and stop as soon as the following detection criterion is fulfilled [68].

$$\text{SRR} = \frac{\hat{a}_0^2}{2 \int_{\hat{\tau}_0 - D/2}^{\hat{\tau}_0 + D/2} \{ \text{Re}[R_{x0}(\tau, \hat{\theta}_0) - \sum_{i=0}^M \hat{a}_i R_{0i}(\tau, \hat{\tau}_i, \hat{\theta}_0, \hat{\theta}_0)] \}^2 d\tau} \geq K \quad (3.16)$$

In equation (3.16), D is the delay interval which contains the line-of-sight correlation peak. The Signal-to-Residual Ratio (SRR) is defined as the estimated signal power divided by the variance of the residuals after estimation of all parameters. If all signals are properly estimated, the SRR will be approximately equal to the Signal-to-Noise Ratio (SNR) after T seconds of correlation. Therefore, the threshold K should be chosen equal to the SNR value that yields the maximum allowable range errors. Third, the simplest strategy is to choose a fixed value for M , but a certain multipath error may be induced if M is improper.

Fig.3-8 Comparison of code range errors between MEDLL and DLL

The maximum range errors of MEDLL and DLL are given in Fig.3-8 [67], where T_c is one chip spacing; single multipath is present; the signal-to-multipath ratio equals 3 dB; the bandwidth of the spread-spectrum signals is 2 MHz. As can be seen, the MEDLL gives a great improvement on range error when compared with DLL.

MEDLL technique has two major disadvantages. The first one is that its computation burden is too high. As can be seen from the derivation of MEDLL in equation (3.15), a lot of parameters must be calculated concurrently. The second is that the number of multipath signals cannot be determined easily, which may result in a poor performance of MEDLL.

3.3.4 Strobe and Edge Correlator

A novel technique is proposed in [10, 51], the idea of which is to lower the multipath sensitivity by shortening the discrimination function pattern. In the Strobe technique, a linear combination of two narrow correlators is used, i.e., $2 \cdot \text{narrow}(d/2) - \text{narrow}(d)$, where $\text{narrow}(d)$ is the discrimination function pattern corresponding to an early-late spacing of d C/A chips, e.g., $d = 0.1$ for a narrow correlator.

Fig.3 - 9 Discrimination function pattern in strobe correlator

Fig.3 - 10 Range errors from different techniques

One typical discrimination function in the Strobe technique is depicted in Fig.3-9 [51], and the range errors in terms of relative multipath delay are shown in Fig.3-10 [10]. The discrimination function in Fig.3-9 reveals that its effective range is very narrow, and thus any multipath signal with a delay beyond this range will be mitigated ideally (the leftmost and the rightmost part in Fig.3-9 will not cause much trouble, because these two parts correspond to the multipath delay around one chip, which is not often observed). The tracking error plots in Fig.3-10 illustrate that the strobe correlator has superior performance on mitigating multipath compared with other techniques.

3.3.5 Multipath Mitigation by Reference Waveform Design

This technique proposes a specifically designed local reference waveform, and the correlation function between it and the received signal is much narrower when compared with the conventional correlation function, so the multipath signals can be separated from the direct signal, and hence the estimation of pseudorange can

be based on the direct signal only.

Fig.3-11 Optimal receiver for single-path signal

Fig.3-11 shows the baseband model for calculating the correlation function in a typical GPS receiver [69], where the left of the vertical dashed line in the top portion of Fig.3-11 indicates that the ideal C/A code is passed through the filter on a navigation satellite, $H(\omega)$, and the right of the vertical dashed line represents the processing blocks in the receiver. Before calculating a correlation, the ideal local reference C/A code is passed through a filter, which has the same response as the one on the satellite. The bottom plot in Fig.3-11 is the resulting correlation functions $R(\tau)$ of a direct plus a multipath signal, which indicate that the peak region of both correlation functions is rounded due to the filtering.

The contribution of this technique is to use $-R''(\tau)$, i.e., the negative of the second derivative of $R(\tau)$, instead of $R(\tau)$ to determine the delay of the direct signal. The benefits of it include: first, $-R''(\tau)$ has the same peak position as that of $R(\tau)$; second, the function $-R''(\tau)$ for a single signal (either direct or multipath) is very narrow, so the multipath signal can be separated from the direct

signal, as shown at the bottom of Fig.3-12.

Fig.3 - 12 The resultant improved correlation function

However, one problem is that the implementation of $-R''(\tau)$ is expensive in hardware and generally is not easy, because one has to obtain a complete correlation function in order to obtain the derivative of it. However, the relationship in equation (3.17) gives a good idea for implementation.

$$\begin{aligned} -\frac{d^2}{d\tau^2} R(\tau) &= -\frac{d^2}{d\tau^2} \int_0^T s(t) \cdot \tilde{c}(t-\tau) dt \\ &= -\int_0^T s(t) \cdot \tilde{c}''(t-\tau) dt \end{aligned} \quad (3.17)$$

The above equation reveals that the second derivative of $R(\tau)$ can be implemented through the second derivative of the local C/A code, which is the output of the local filter. Furthermore, the results of the derivative of the local C/A code can be stored in hardware and no need to calculate it each time.

The range error against relative multipath delay in meters for a single multipath signal is plotted in Fig.3-13, which indicates significant improvement on multipath mitigation, i.e., the maximum error is around 2.7 meters.

Fig.3-13 Range errors versus relative multipath delay

One of the disadvantages of this technique is that the effective SNR is reduced as the correlation function is made narrower. The other one is that when the relative delay of the multipath signal is very short, this technique may not be effective, in other words, the multipath signal may not be separated from the direct signal.

3.3.6 Multipath Mitigation Based Error Correction

The error correction based technique aims at estimating the multipath range error, and then subtracts the estimate of the error from the original pseudorange obtained by the normal techniques to achieve a more accurate pseudorange. In other words, the original pseudorange is corrected by the estimate of the error.

An efficient method, called Early Late Slope technique, is introduced in [41]. As has been introduced, in a conventional early-minus-late receiver, the received code will have equal 'distance' to the early correlator and the late correlator when

DLL is in its 'locked' state. But if multipath exists, the distance between the direct code and the early code is not equal to that between the direct code and the late code, due to the asymmetry of the autocorrelation function. The difference between these two distances represents the range error caused by multipath, which can be used to correct the estimate of the pseudorange.

Fig.3-14 Illustration on early late slope technique

A typical illustration to estimate the range error is depicted in Fig.3-14 [41]. In the figure, y_1 and y_2 are the measured value of the early and the late correlator respectively. The slope of the correlation function on the early edge of the peak is a_1 , and a_2 is the slope of the late edge of the peak. The spacing between the early and the late correlator is d in units of chips.

By the slope information, the following DLL discriminator in (3.18) can be derived to accurately estimate how much the early and the late correlator need to be moved again so that they are centered on the autocorrelation function peak of the direct signal.

$$T = \frac{2(y_1 - y_2) + d(a_1 + a_2)}{a_1 - a_2} \quad (3.18)$$

here T is the estimated tracking error caused by multipath. T equals to zero when the two correlators are positioned equidistant on each side of the peak. When T is non-zero, it can be used to feed back to the hardware to keep the local prompt code aligned with the received direct code.

Fig.3-15 Range error versus relative delay

Fig.3-15 shows the improved simulation result, because the Early Late Slope Technique can decrease the multipath caused range error by 30 to 70 percent for a single multipath signal with a relative delay greater than 0.1 chips. The positive error envelopes in Fig.3-15 are the responses to the in-phase multipath, while the negative ones correspond to the out-of phase multipath.

Similar ideas based on the pseudorange correction can be found in [70, 71], both of which have the same order of accuracy as that obtained in the Early Late Slope Technique.

3.3.7 Multipath Mitigation Based on SNR

In recent years, there has been a growing interest in applying signal-to-noise (SNR) to the field of multipath mitigation [72]. In this technique, the SNR is

defined as the correlation amplitude output by a GPS receiver, and the normalized SNR error is given by,

$$S_{error} = (SNR - A_0)/A_0$$

where A_0 is the amplitude of the line-of-sight or direct signal.

Fig.3-16 Relationship between SNR error, phase error and code delay error.

Fig.3-16 shows the relationship between SNR error, phase error and the code delay error, where ‘+’ is phase error in radians; ‘Δ’ is the normalized SNR error; the solid line represents code delay error caused by a single multipath; d in the figure denotes the early-late spacing.

From these simulations, it can be seen that the SNR error is in phase with the code delay error and in quadrature with the phase error, no matter what the correlator spacing is. Additionally, the normalized SNR error amplitude matches the phase error amplitude expressed in radians only for the narrow spacing case (the bottom portion in Fig.3-16).

Based on the above relationships, the SNR error can be applied as a part of multipath mitigation for phase and code delay respectively.

Fig.3-17 Phasor diagram of the direct, reflected and composite signals

Based on the phasor diagram in Fig.3-17 and when the multipath amplitude is quite small, the following relations can be yielded [73],

$$\Delta\varphi = \arctan\left(\frac{A_1 \sin \theta_m}{A_0 + A_1 \cos \theta_m}\right) \approx \frac{A_1}{A_0} \sin \theta_m$$

$$SNR = \sqrt{A_0^2 + A_1^2 + 2A_1A_0 \cos \theta_m} \approx A_0 + A_1 \cos \theta_m$$

And hence the relation between the SNR error and the phase error is,

$$\frac{SNR - A_0}{A_0} = \frac{A_1}{A_0} \cos \theta_m$$

The above relation indicates that the phase error can be estimated and corrected through the SNR error, while the SNR error can be obtained from a GPS receiver easily.

So far, the mitigation on the multipath caused code delay error by the SNR error has not been widely used, and it is only reported that the SNR error is combined with another technique using the day-to-day repeatability of the multipath on a permanent receiver [74]. With the assistance of the SNR error, the day-to-day repeatability based multipath mitigation technique is improved by 28%.

3.4 Effect of Filtering

The tracking errors presented in the previous sections are mostly based on the assumption that the received signal has infinite bandwidth. However, this is not true in practice, and there are two factors limiting the bandwidth of the C/A code, by which the pseudorange is estimated. First, the C/A codes transmitted from the navigation satellites are filtered by an onboard filter, the single-sided bandwidth of which is 10 MHz, and hence contains side lobes out to the 10^{th} [19]. Second, in order to reduce interference and noise, the RF front end of a GPS receiver usually contains several filters, which are distributed at different places. For low-end GPS receivers, the single-sided bandwidth of the filter is typically 1 MHz, which is very narrow. The effects of filtering on the performance of Early Minus Late tracking loop are discussed in [75-77]

The most significant effect of a band-limited filter on the C/A code is to make the peak of the C/A code autocorrelation function rounded, while the autocorrelation function of an infinite-bandwidth C/A code has a sharp peak. Fig.3-18 shows the comparison between these two cases, where a filter with single-sided bandwidth of 1 MHz is applied.

Fig.3-18 Comparison of autocorrelation function with/without filtering

Since the autocorrelation function is affected by the filtering, the tracking error caused by multipath signals will be definitely affected too. [77] points out that the maximum tracking error equals to the amplitude of the filtered multipath discrimination function divided by the slope of the filtered line-of-sight discrimination function, i.e.,

$$\tau_e = \lim_{\Delta \rightarrow 0} \frac{a_1 \left[R\left(\frac{T_c - d}{2}\right) - R\left(\frac{T_c + d}{2}\right) \right]}{2a_0 \left[R\left(\frac{d - \Delta}{2}\right) - R\left(\frac{d + \Delta}{2}\right) \right] / \Delta} \quad (3.19)$$

where $R(\bullet)$ represents the correlation function; a_0 and a_1 are the amplitude of the line-of-sight and the multipath signal respectively; T_c is one PRN code chip period, and d is the early-late spacing.

If no filtering is applied, the maximum tracking error in (3.19) equals to $da_1/(2a_0)$. If the signal is passed through a band-limited filter with a double-sided bandwidth of $2k/T_c$ MHz, the tracking error normalized by $T_c a_1/2a_0$ versus the early-late spacing d is plotted in Fig.3-19, where the dotted line corresponds to the case when there is no filtering.

Fig.3 - 19 Multipath tracking error versus early-late spacing

Fig.3-19 reveals that with no filtering, the tracking error linearly decreases

with the decrease of the early-late spacing. However, if the signal is filtered, the multipath signal caused tracking error will reach a minimum value after the early-late spacing is reduced to a certain value for a certain filtering bandwidth, and hence continuing to decrease the early-late spacing will not be helpful to reduce the tracking error. In conclusion, for a single multipath signal with a small multipath-to-direct ratio (e.g., -10 dB), there is no improvement on the range error by using an early-late spacing smaller than $1/BW$ [77], where BW is the single-sided signal bandwidth in Hertz.

Summary

This chapter discussed the issues on multipath, especially the techniques which have achieved great improvement on pseudorange accuracy by mitigating multipath signals, and most of them have been applied into commercial GPS receivers. However, a significant pitfall of the existing techniques is that they are not robust for mitigating short-delay multipath signals, which have relative delays less than 0.1 PRN code chips. With more and more GPS applications extended to urban and indoor environments where numerous short-delay multipath signals exist, new techniques need to be explored to achieve a satisfactory positioning accuracy. In the Chapter 5, Chapter 6 and Chapter 7, three proposals on mitigating short-delay multipath signals will be introduced.

Chapter 4 Software GPS Receiver Implementation

Software GPS Receiver (SGR) has become popular in recent years, because it has many advantages when compared with hardware receivers [101]. The software GPS receiver can be modified or updated much easier, and hence is more flexible to adopt new signal processing algorithms or to receive new navigation signals (e.g., Galileo, GLONASS, etc.); it can be used by companies to evaluate their navigation products prior to the hardware implementation and can be adopted by education institutes to conduct research on Global Navigation Satellite System (GNSS); it has the minimum hardware cost and hence is more economical. The goal of some software GPS receivers [33, 102, 103] mainly focuses on testing algorithms only and does not have too much of a requirement on running speed, while recently some institutes have announced their real-time software GPS receivers [104-109].

As an important work of this thesis, a software GPS receiver is implemented, and this chapter will present its main function blocks as well as the testing results conducted in the Positioning and Wireless Technology Centre. At the moment it is in Matlab format, and can be further modified by using other programming language, e.g., C/C++, to make it run faster.

4.1 Architecture of the Software GPS Receiver

Fig.4-1 Block diagram of the software GPS receiver

The generic GPS receiver architecture is shown in Fig.2-17, and Fig.4-1 gives the basic block diagram of the implemented SGR.

Although it is widely named “software” receiver, the entire receiver actually consists of hardware modules and software modules. The former, including RF front end and digitizer, is mainly for conducting RF signal down conversion and data collection, and the function of the latter part is to do signal post-processing to obtain user position, based on the raw data which are saved in PC hard disk.

4. 2 Receiver Hardware Modules

4.2.1 RF Front End

Fig. 4-2 RF front end with GP2010 chip

As that discussed in Section 2.6, a typical RF front end usually consists of one, two or multiple stages of mixers to down convert the RF signal to IF. The front end used for our SGR is with a GP2010 chip from Zarlink Inc [29], as shown in Fig. 4-2.

The front end contains three stages of mixers, and the output IF signal is at

4.309 MHz, which is band limited by a Surface Acoustic Wave (SAW) filter with 1 dB bandwidth of 1.9 MHz.

4.2.2 Digitizer

The analog IF GPS signal from the RF front end is digitized by a PCI interface data acquisition board, National Instrument (NI) 5112, as shown in Fig. 4-3, which is installed in a Dell Optiplex GX280 desktop. The NI-5112 board can be controlled through Labview, and Table 4.1 gives the key configurations of the Stream to Disk VI in Labview for collecting GPS signal.

Fig. 4-3 NI-5112 PCI digitizer

Table 4.1 Configurations for NI-5112

	Resource Name	Channel	Min Sample Rate	Max Points Per Fetch	Vertical Range	Vertical Coupling
Settings	DAQ::1	0	20 MS/sec	5 M	0.6 V	AC

By the above set-up and configurations, the NI-5112 digitizer is able to capture

the input signal under a sampling rate of 20 MSamples/sec and save it as a binary file in PC hard disk in real time. To calculate user position by GPS signal, at least a 30-second signal must be captured.

4.3 Receiver Software Modules

These modules include C/A code generator, GPS signal acquisition, tracking and user position calculation. In a standard GPS receiver, all these blocks are conventionally implemented by hardware, but here all of them will be replaced by software, and hence is named software GPS receiver.

4.3.1 C/A Code Generator

Fig. 4-4 C/A code generator

Table 4.2 Code phase assignments

Satellite ID Number	GPS PRN Signal Number	$a \oplus b$	Code Delay Chips	First 10 Chips C/A (Octal)
1	1	$2 \oplus 6$	5	1440
2	2	$3 \oplus 7$	6	1620
3	3	$4 \oplus 8$	7	1710
4	4	$5 \oplus 9$	8	1744
5	5	$1 \oplus 9$	17	1133
6	6	$2 \oplus 10$	18	1455
7	7	$1 \oplus 8$	139	1131
8	8	$2 \oplus 9$	140	1454
9	9	$3 \oplus 10$	141	1626
10	10	$2 \oplus 3$	251	1504
11	11	$3 \oplus 4$	252	1642
12	12	$5 \oplus 6$	254	1750
13	13	$6 \oplus 7$	255	1764
14	14	$7 \oplus 8$	256	1772
15	15	$8 \oplus 9$	257	1775
16	16	$9 \oplus 10$	258	1776
17	17	$1 \oplus 4$	469	1156
18	18	$2 \oplus 5$	470	1467
19	19	$3 \oplus 6$	471	1633
20	20	$4 \oplus 7$	472	1715
21	21	$5 \oplus 8$	473	1746
22	22	$6 \oplus 9$	474	1763
23	23	$1 \oplus 3$	509	1063
24	24	$4 \oplus 6$	512	1706
25	25	$5 \oplus 7$	513	1743
26	26	$6 \oplus 8$	514	1761
27	27	$7 \oplus 9$	515	1770
28	28	$8 \oplus 10$	516	1774
29	29	$1 \oplus 6$	859	1127
30	30	$2 \oplus 7$	860	1453
31	31	$3 \oplus 8$	861	1625
32	32	$4 \oplus 9$	862	1712
**	33	$5 \oplus 10$	863	1745
**	34 [*]	$4 \oplus 10$	950	1713
**	35	$1 \oplus 7$	947	1134
**	36	$2 \oplus 8$	948	1456
**	37 [*]	$4 \oplus 10$	950	1713

* 34 and 37 have the same C/A code.

** GPS satellites do not transmit these codes; they are reserved for other use.

The C/A code generator provides the local reference code for signal acquisition. As introduced in Section 2.3.3, the C/A codes in GPS belong to the family of Gold Codes, and each C/A code has 1023 chips with a cycle length of one millisecond.

The C/A code generator architecture is shown in Fig.4-4, which is a combination of two Gold Code generators, i.e., G1 and G2. The modulo-2 addition of these two Gold Code generators' output is the C/A code. There are 32 C/A codes in total, which are determined by the different combinations of 'a' and 'b' based on Table 4.2 [110].

4.3.2 GPS Signal Acquisition

As mentioned in Section 2.4 in Chapter 2, signal acquisition is a two dimensional search, i.e., code search and frequency search, and its aim is to find the high correlation peak.

In the SGR, a Fast Fourier Transform (FFT) based correlation algorithm is adopted to acquire GPS signal, and its block diagram is shown in Fig.4-5.

Fig. 4-5 FFT-based correlation algorithm

If the received signal is matched with the local reference signal in both carrier

frequency and code phase, the output will be a high correlation peak to indicate that a certain satellite signal is acquired.

In order to make the receiver be able to judge whether the acquisition results contain a high correlation peak or not, a detection criterion or threshold must be applied. Typically, a threshold is based upon an acceptable false detection probability that a noisy measurement, which does not contain the GPS signal, will appear to match the local replica. Based on [103], the threshold used in the software GPS receiver is,

$$V_t = \sigma_n \cdot \sqrt{-2 \cdot \ln(P_{fa})}$$

where,

V_t : threshold

σ_n : root mean square (rms) noise power.

P_{fa} : predefined false detection probability.

σ_n can be simply estimated from the acquisition output, i.e., $G_i = \sqrt{I_i^2 + Q_i^2}$, and the estimation can be expressed as [103]:

$$\hat{\sigma}_n = \sqrt{\frac{1}{2N} \sum_i^N G_i^2}$$

where N is the total number of output correlation values.

4.3.3 GPS Signal Tracking

Generally, the GPS signal can be tracked by a delay lock loop and a phase lock loop. However, sometimes the tracking loop may fail in tracking the signal even if the signal does exist. One of the reasons is that the offset between the coarse frequency obtained from acquisition and the real signal frequency is beyond the tracking range of the phase lock loop (e.g., 500 Hz for 1ms integration). In the implemented SGR, to enhance the reliability and robustness of

the tracking loop, a Frequency Lock Loop (FLL) is implemented to refine the coarse carrier frequency. Once the FLL is locked, the tracking is handled by the phase lock loop (typically Costas loop), because the phase lock loop is more precise than the frequency lock loop. The principles of delay lock loop and phase lock loop was presented in section 2.5 of Chapter 2, so in this section only the FLL will be presented in details.

Fig.4-6 gives the illustration on the interoperation between the DLL and the carrier tracking loop, which is implemented in the SGR.

Fig. 4-6 The overall tracking loop implemented in SGR

The multiplier before the DLL is to strip the intermediate frequency carrier in the incoming signal, and ideally only the C/A code (the navigation data bit can be neglected for the moment, because it is 20 ms long) is remaining, which is then processed by the DLL to estimate the delay of the code.

The architecture of DLL can be referenced to Fig.2-13. In the SGR, the signal under processing is discrete, so the low pass filter can be replaced by an accumulator, which integrates the input signal and dumps a single value typically every millisecond.

The discriminator function of the DLL, as discussed in section 2.5.1, is formed by a conventional normalized early-minus-late correlator with an early-late-spacing of one C/A code chip, i.e.,

$$\text{Discrimination} = \frac{\text{Early} - \text{Late}}{\text{Early} + \text{Late}}.$$

Similar to that in DLL, the input to the carrier tracking loop is the carrier only (again, the navigation data bit is neglected), because the C/A code is removed by the multiplier before the carrier tracking loop in Fig.4-6.

Fig. 4-7 Interoperation between FLL and PLL

Fig. 4-8 Architecture of FLL

Fig.4-7 gives the illustration on the transition from FLL to PLL, where CW is the discrete signal after the C/A code is wiped off, and f_0 is the signal frequency

refined by FLL. The FLL is set to un-locked by default, so it will run prior to the PLL and is based on the coarse frequency directly obtained from acquisition. Once the FLL is locked, tracking is passed to the PLL which is based on the refined frequency.

A frequency lock loop is to estimate the signal frequency, unlike the phase lock loop aiming to estimate the signal phase variation. The idea of FLL is quite easy to understand. It uses the phase difference between two time epochs to compute the corresponding frequency difference, i.e.,

$$f_n = \frac{\theta_n - \theta_{n-1}}{2\pi \cdot \Delta T}$$

where θ_{n-1} and θ_n is the phase at T_{n-1} and T_n respectively; $\Delta T = T_n - T_{n-1}$.

The typical method to compute the frequency f_n is through the function of $\text{atan2}()$, that is,

$$f_n = \frac{a \tan 2[y_I(n), y_Q(n)] - a \tan 2[y_I(n-1), y_Q(n-1)]}{2\pi \cdot \Delta T} \quad (4.1)$$

where y_I and y_Q is the output of the in-phase arm and the quadrature arm in Fig. 4-8. There are some other discriminator functions [18], which are with different pull-in ranges and properties.

The discriminator function adopted in our implementation is

$$\text{discrimina tor} = \frac{\text{ATAN2}(\text{cross}, \text{dot})}{(t_2 - t_1) \cdot 2\pi} \quad (4.2)$$

Where $\text{dot} = I_{t1} \cdot I_{t2} + Q_{t1} \cdot Q_{t2}$, $\text{cross} = I_{t1} \cdot Q_{t2} - I_{t2} \cdot Q_{t1}$, here I and Q is the output of the integrate and dump filter in I-arm and Q-arm respectively in Fig.4-8. The above discriminator is optimal at high and low SNR [18], and the slope is not amplitude dependent. The other advantage of this discriminator is that it has the largest pull-in range when compared with other discriminator functions specified in [18]. In fact, the discriminators in equation (4.1) and (4.2) are equivalent in mathematics, and the demonstration can be found in [33], except that the discriminator (4.2) has lower computation load.

The loop filter in Fig. 4-8 determines the characteristics of the FLL. There are three types of low-pass filters presented in [18], and the first-order loop filter given in Fig. 4-9 is adopted in the SGR (the last integrator in Fig. 4-9 actually represents the NCO).

Fig. 4-9 First-order loop filter architecture

In Fig. 4-9, the ω_0 is natural radian frequency and a_2 is a constant. As illustrated in [18], ω_0 and a_2 can be computed once the equivalent noise bandwidth B_n is determined. In the SGR, B_n is set to 10 Hz.

Fig. 4-10 Architecture of digital integrator

The digital form of the analog integrator in Fig. 4-9 is given in Fig. 4-10, where T is the time interval between two neighboring samples, and $y(n)$ is the integration of $x(n)$.

A simulation result of the implemented FLL is given in Fig.4-11, where the input signal is a pure sinusoidal waveform with a frequency of 4.309 MHz and

SNR of -20dB; the local frequency is intentionally set to 4.3092 MHz; the pre-detection integration time used in the simulation is 1ms. Fig.4-11 shows the variation of the local frequency, which indicates that the local frequency is able to adjust itself and converges to the correct input frequency.

Fig. 4-11 Simulation of FLL

Fig. 4-12 FLL lock detector outputs

Table 4.3 Loop filter specifications

	k_0k_1	ω_n (Hz)	ζ	t_s (s)
DLL	50	2	0.707	0.001
Costas	400π	20	0.707	0.001

A FLL lock detector [103] is also implemented to adjudge whether the FLL is locked or not, and its output corresponding to the simulation of Fig.4-11 is shown in Fig.4-12. The dotted line in Fig.4-12 is the output from the lock detector, but it is not appropriate for use because it is too noisy. The result after it is passed through a low-pass filter [103] is much smoother, and can indicate whether the FLL is locked. In the implemented SGR, once the filtered detector output reaches 0.9, FLL is looked as locked.

Similarly, in the Costas loop an accumulator (integrate and dump filter) is applied in both I-arm and Q-arm. In the locked state, the output from the I-arm accumulator is the navigation data bits, and the output from the Q-arm is noise. The details on Costas loop can be referenced to Fig.2-16.

The loop filters used in both the DLL and the Costas loop are first-order low pass filter, whose transfer function is,

$$H(z) = \frac{(C_1 + C_2) - C_1 \cdot z^{-1}}{1 - z^{-1}} \quad (4.3)$$

where

$$C_1 = \frac{1}{k_0k_1} \cdot \frac{8\xi\omega_n t_s}{4 + 4\xi\omega_n t_s + (\omega_n t_s)^2}$$

$$C_2 = \frac{1}{k_0k_1} \cdot \frac{4(\omega_n t_s)^2}{4 + 4\xi\omega_n t_s + (\omega_n t_s)^2}$$

Table 4.3 lists the settings of the loop filter for the DLL and the Costas loop respectively, and more detailed mathematic derivation can be found in [34]. The filter expressed by equation (4.3) has the same form as illustrated in Fig.4-9, and

only the methods to determine the filter's parameters are slightly different.

4.3.4 First Navigation Data Bit Transition

The demodulated data usually are with good SNR, which can be further binarized to +1 if the data amplitudes are positive and -1 if negative. One navigation data bit is 20 ms long and hence should exactly contain 20 C/A codes. In other words, one navigation data bit consists of 20 correlation peaks. In software GPS receivers, the entire navigation data (e.g., 30sec) are usually demodulated first. In these data, the first several or tens of correlation peaks usually are the results when the tracking loop has not entered 'locked' state, as shown in Fig.4-13, so one navigation data bit could contain more or less than 20 correlation peaks. Additionally, if the tracking loop is not in 'locked' state, the demodulated data are usually with a low SNR, and these data are invalid and must be removed.

Fig. 4-13 Navigation data in the first second

Due to the above reasons, the first navigation data bit transition must be

detected. A transition could be from +1 to -1 or from -1 to +1 and represents the exact beginning of a navigation data bit. Only when the first transition is found, can the subframe matching be performed.

In the implemented SGR, a threshold is set to judge if a transition has occurred, and this threshold simply equals the amplitude of the demodulated navigation data, which can be obtained by averaging multiple navigation data samples to reduce noise. The reason that the threshold can be selected by this method is based on the fact that the entire or at least 30sec navigation data were already demodulated and saved on hard disk. Once the threshold is calculated, the first data bit transition can then be detected.

4.3.5 User Position Calculation

The demodulated data from the Costas loop are processed by the navigation solution module in the SGR to obtain the user position. Fig.4-14 briefly illustrates the three key blocks for calculating user position [34].

Fig. 4-14 Brief illustrations on navigation solution

Fig.4-15 gives the detailed flow chart for the user position calculation in the implemented software receiver.

a) Determine the First Data Bit Transition

This block has been discussed in section III, and it is used to determine the beginning of the effective navigation data.

Fig. 4-15 Flowchart for calculating user position

b) Check Sample Number per Navigation Data Bit

The integration time period used in the SGR is one millisecond, while one navigation data bit is 20ms long, so one navigation data bit must exactly consist of 20 correlation peaks, otherwise there is an error occurred in the signal tracking and the data must be discarded.

c) Parity and Polarity Check

Parity check is to check whether there is any error occurred in the demodulated navigation data bits, and if the parity check failed, the data cannot be used. Due to the Costas loop, the polarity of the data is ambiguous, and must be corrected based on the polarity bits in the data, otherwise they will be converted to wrong decimal satellite information.

d) Subframe Matching

Subframe matching is used to find each subframe from the demodulated data according to the preamble in each subframe. To further guarantee that the demodulated data are effective, the IDs from five subframes and their order are checked in the SGR.

e) Convert Navigation Data to Decimal

Once all the above steps are passed successfully, the demodulated data can be converted to decimal number for further processing, and detailed information can be found in [20].

f) Calculate Satellite Position

In the SGR, it is the relative pseudoranges or relative signal transit time that can be obtained, because the digitized data at the receiver side do not contain any absolute time reference. Relative pseudoranges do not affect the calculation of user position, because for a certain satellite signal, the difference between the relative pseudorange and the true pseudorange is a constant which is same for all the satellites being tracked and hence can be looked as the bias between the satellite clock and receiver clock. The implementation of calculating the relative pseudoranges will be presented in Section III-F. Meanwhile, assume that they are

known.

The satellite position is calculated based on the signal coarse transmission time, i.e., Time of Week (TOW) corrected by the relative signal transit time or relative pseudorange. That is, once the relative delay is found, the signal transmission time can be estimated by:

$$t_c = \text{TOW} - \text{'relative delay'}$$

and the total time difference between t_c and the epoch t_{oe} after accounting for the beginning or the end of the week crossovers is,

$$\text{if } t_k = t_c - t_{oe} > 302400 \text{ then } t_k = t_k - 604800 \text{ or } t_c = t_c + 604800$$

$$\text{if } t_k = t_c - t_{oe} < -302400 \text{ then } t_k = t_k + 604800 \text{ or } t_c = t_c + 604800$$

Once the GPS system time at time of transmission is found, the mean anomaly can be found from

$$M = M_0 + n \times (t_c - t_{oe})$$

where M_0 is the mean anomaly at reference time obtained from subframe 2 of the demodulated navigation data.

The eccentric anomaly E can be found from,

$$E = M + e_s \times \sin E$$

where e_s is the eccentricity of satellite orbit obtained from subframe 2.

The relativistic correction term can be obtained from

$$\Delta t_r = F \times e_s \times \sqrt{a_s} \times \sin E$$

where $F = -4.442807633 \times 10^{-10} \text{ sec/m}^{1/2}$ is a constant.

The overall time correction term is

$$\Delta t = a_{f0} + a_{f1}(t_c - t_{oc}) + a_{f2}(t_c - t_{oc}) + \Delta t_r - T_{GD}$$

where a_{f0} , a_{f1} , a_{f2} , t_{oc} and T_{GD} can be read from the subframe 1 of the navigation data.

The GPS time at time transmission can be corrected again by

$$t = t_c - \Delta t$$

The true anomaly can be found from

$$v_1 = \cos^{-1} \left(\frac{\cos E - e_s}{1 - e_s \cos E} \right)$$

$$v_2 = \sin^{-1} \left(\frac{\sqrt{1 - e_s^2} \sin E}{1 - e_s \cos E} \right)$$

$$v = v_1 \text{sign}(v_2)$$

and angle ϕ can be found as

$$\phi = v + \omega$$

where ω is the argument of perigee in subframe 3.

The following correction terms are also needed,

$$\delta\phi = C_{us} \sin 2\phi + C_{uc} \cos 2\phi$$

$$\delta r = C_{rs} \sin 2\phi + C_{rc} \cos 2\phi$$

$$\delta i = C_{is} \sin 2\phi + C_{ic} \cos 2\phi$$

where C_{us} , C_{uc} , C_{rs} , C_{rc} , C_{is} and C_{ic} can all be obtained from subframe 3 of the navigation data. And these three terms are used to correct the following terms,

$$\phi = \phi + \delta\phi$$

$$r = r + \delta r$$

$$i = i + \delta i + \dot{i} \text{dot}(t - t_{oe})$$

here $\dot{i}$ is the rate of inclination angle and can be obtained from navigation data.

The angle between the ascending node and the Greenwich meridian Ω_{er} can be found from

$$\Omega_{er} = \Omega_e + \dot{\Omega}(t - t_{oe}) - \dot{\Omega}_{ie} t$$

The final two steps are to find the position of the satellite from the above equations and adjust the pseudorange by the overall clock correction terms as

$$\begin{bmatrix} x \\ y \\ z \end{bmatrix} = \begin{bmatrix} r \cos \Omega_{er} \cos \phi - r \sin \Omega_{er} \cos i \sin \phi \\ r \sin \Omega_{er} \cos \phi + r \cos \Omega_{er} \cos i \sin \phi \\ r \sin i \sin \phi \end{bmatrix}$$

$$\rho_i = \rho_i + c \times \Delta t$$

Once the satellite position is obtained, the user position can be calculated based on the equation (2.1), and next section will present the detailed method on solving the equations in (2.1).

Additionally, to reduce the effect of earth rotation, the satellite signal transit time must be taken into account [34]. In other words, the time used to calculate the satellite position must be adjusted to be the same time for calculating user position. That is,

$$\rho = \sqrt{(x - x_u)^2 + (y - y_u)^2 + (z - z_u)^2}$$

$$t_t = \sqrt{(x - x_u)^2 + (y - y_u)^2 + (z - z_u)^2} / c$$

where x, y, z and x_u, y_u, z_u are coordinates of the satellite and the user respectively, c is the speed of light. Use this transit time to modify the angle Ω_{er} in the above equation as,

$$\Omega_{er} = \Omega_{er} - \dot{\Omega}_{ie} t_t$$

Use this new Ω_{er} in the first portion of equation to calculate the satellite position again, based on which the new user position can be calculated. This is an iteration, which will stop when the changes of the satellite position or user position are below a predetermined value, and the final position is the desired user position.

g) Calculate User Position

The equations in (2.1) are nonlinear simultaneous equations, and in order for a computer to be able to solve them automatically, the equations (2.1) must be linearized.

A general representation of the equations in (2.1) is shown below,

$$\rho_i = \sqrt{(x_i - x_u)^2 + (y_i - y_u)^2 + (z_i - z_u)^2} + b_u$$

where $i = 1, 2, 3$ and 4 , x_u, y_u, z_u and b_u are the user positions to be calculated; the pseudorange ρ_i and the satellite position x_i, y_i, z_i are known.

The result after this equation is differentiated is,

$$\begin{aligned}\delta\rho_i &= \frac{(x_i - x_u)\delta x_u + (y_i - y_u)\delta y_u + (z_i - z_u)\delta z_u}{\sqrt{(x_i - x_u)^2 + (y_i - y_u)^2 + (z_i - z_u)^2}} + \delta b_u \\ &= \frac{(x_i - x_u)\delta x_u + (y_i - y_u)\delta y_u + (z_i - z_u)\delta z_u}{\rho_i - b_u} + \delta b_u\end{aligned}$$

In this differentiated equation, δx_u , δy_u , δz_u and δb_u are the only unknowns. The user position (x_u, y_u, z_u) is known, because some initial values can be assumed for these quantities.

After linearized, the above equation can be written as,

$$\begin{bmatrix} \delta\rho_1 \\ \delta\rho_2 \\ \delta\rho_3 \\ \delta\rho_4 \end{bmatrix} = \begin{bmatrix} \alpha_{11} & \alpha_{12} & \alpha_{13} & 1 \\ \alpha_{21} & \alpha_{22} & \alpha_{23} & 1 \\ \alpha_{31} & \alpha_{32} & \alpha_{33} & 1 \\ \alpha_{41} & \alpha_{42} & \alpha_{43} & 1 \end{bmatrix}^{-1} \begin{bmatrix} \delta\rho_1 \\ \delta\rho_2 \\ \delta\rho_3 \\ \delta\rho_4 \end{bmatrix}$$

where

$$\alpha_{i1} = -\frac{x_i - x_u}{\rho_i - b_u} \quad \alpha_{i2} = -\frac{y_i - y_u}{\rho_i - b_u} \quad \alpha_{i3} = -\frac{z_i - z_u}{\rho_i - b_u}.$$

The solution of δx_u , δy_u , δz_u and δb_u is,

$$\begin{bmatrix} \delta x_u \\ \delta y_u \\ \delta z_u \\ \delta b_u \end{bmatrix} = \begin{bmatrix} \alpha_{11} & \alpha_{12} & \alpha_{13} & 1 \\ \alpha_{21} & \alpha_{22} & \alpha_{23} & 1 \\ \alpha_{31} & \alpha_{32} & \alpha_{33} & 1 \\ \alpha_{41} & \alpha_{42} & \alpha_{43} & 1 \end{bmatrix}^{-1} \begin{bmatrix} \delta\rho_1 \\ \delta\rho_2 \\ \delta\rho_3 \\ \delta\rho_4 \end{bmatrix}$$

The above equation will run iteratively, and the quantity used to stop the iteration is,

$$\delta v = \sqrt{\delta x_u^2 + \delta y_u^2 + \delta z_u^2 + \delta b_u^2}.$$

In the implemented software receiver, δv is set to be 10^{-7} based on simulations.

For an initial user position and pseudorange, a new set of δx_u , δy_u , δz_u and δb_u can be calculated based on the above equations, which is then used to correct the initial user position and hence generate a new user position. This calculation will repeat until δx_u , δy_u , δz_u , δb_u are very small and within a certain predetermined limit, and the final x_u, y_u, z_u are the desired solution.

The user position obtained from the above equations is in a Cartesian coordinate system, but in daily life the geodetic longitude, latitude and altitude is used, so the conversion is needed.

The geodetic longitude can be calculated as,

$$\text{Longi} = \begin{cases} \arctan\left(\frac{y_u}{x_u}\right), & x_u > 0 \\ 180^\circ + \arctan\left(\frac{y_u}{x_u}\right), & x_u < 0, \quad y_u \geq 0 \\ -180^\circ + \arctan\left(\frac{y_u}{x_u}\right), & x_u < 0, \quad y_u < 0 \end{cases}$$

where x_u, y_u is the calculated user position above. If $\text{Longi} > 0$, the user is at the east of the Meridian, and if $\text{Longi} < 0$, user is at the west of the Meridian.

The geodetic latitude can be obtained by,

$$\text{Lati} = \text{Lc} + e_p \times \sin(2 \times \text{Lati})$$

where

$$\text{Lc} = \tan^{-1}\left(\frac{z_u}{\sqrt{x_u^2 + y_u^2}}\right).$$

The calculation of latitude is iteration, and usually converges very fast. If the calculated $\text{Lati} > 0$, the user is in the north hemisphere of the Earth, otherwise the user is in the south hemisphere of the Earth.

The altitude can be found from,

$$h = \sqrt{x_u^2 + y_u^2 + z_u^2} - a_e \times (1 - e_p \times \sin^2(\text{Lati}))$$

The receiver must keep updating the user position, typically once per second. This is implemented by choosing a different reference point for calculating the relative delay. If an update rate of one second is desired, the new relative delay must be estimated based on the reference correlation peak which is one second later. A new user position is then obtained based on the new relative delay and the above equations.

4.3.6 Determination of the Relative Pseudoranges

As mentioned above, the digitized data do not contain any absolute time, and only the relative pseudoranges can be calculated. To do this, a reference must be determined first. Generally, the delay of one satellite is randomly selected as the reference, to which the delays of other satellites are compared and the relative delays are then obtained. The reason behind this is that all the GPS signals are emitted from the satellites simultaneously. In other words, the corresponding subframes from every satellite are synchronized.

However, the demodulated data from the tracking loop typically are with a resolution of one millisecond, which is not appropriate for positioning. The relative delay must be calculated in units of sample points, which are output of the high speed digitizer. To accomplish this, a mapping from the demodulated data to the raw sample points are needed, which is illustrated in Fig. 4-16.

Fig. 4-16 Calculation of relative pseudorange

In Fig.4-16, 'RawData' refers to the discrete sample points directly obtained from digitizer; the empty circles represent the correlation peaks with a resolution of one

millisecond. Note that Fig.4-16 is just for illustration, and it does not represent the real case. In practice, one navigation data bit should contain 20 correlation peaks, and one subframe contains 300 navigation data bits, while in Fig.4-16, only five correlation peaks are plotted in one subframe.

Assume that the top figure in Fig.4-16 is the demodulated data corresponding to C/A code 1. After subframe matching, the subframe 1 can be found from the data, and the index of the first sample point or correlation peak in subframe 1 will be selected as the reference point. Since C/A code 1 is selected as reference code, its relative pseudorange equals to zero. For the same set of raw GPS data, another set of navigation data can be demodulated after changing the local C/A code, say C/A code 2, and are plotted in the bottom figure in Fig.4-16. Similarly, the subframe 1 corresponding to the C/A code 2 can be found, and assume that the index of the first correlation peak in subframe 1 is *index2* in units of sample points as that shown in Fig.4-16. Therefore, the relative delay of the C/A code 2 will be equal to the difference between *index1* and *index2*, i.e., $index2 - index1$.

If the C/A code 2 is selected as the reference code, the relative delay of C/A code 1 can be calculated by the same method, except that the relative delay could be negative, which is allowed in the software GPS receiver.

To calculate the relative pseudorange correctly, the time for comparing must be the same. For instance, in the above example, the index of the first correlation peak of the subframe 1 of C/A code 1 is selected as reference, so only the first correlation peak in the navigation data from C/A code 2 can be used for calculating relative delay. If the reference is changed, the correlation peak in the navigation data by C/A code 2 must be changed correspondingly.

In the implemented SGR, the tracking loops output two important data arrays. One is the navigation data in a form of individual correlation peaks, and the other array is the indices (in units of sample points) of these peaks. Therefore, it is very convenient to calculate the relative pseudorange.

4.3.7 Estimation of Positioning Error

Estimation of the position error is not the function of a GPS receiver, but in order to quantify the position error and hence evaluate the performance of the software GPS receiver, this step is necessary.

The position error is the difference between the calculated result and the exact user position, and is usually represented by East-West (EW) error, North-South (NS) error and height error. The error in height is very easy to calculate, and it just equals the result of the calculated height minus the exact height. However, the calculation of the EW and the NS error must be done carefully; otherwise some artificial error may be induced. This is because the user is on the surface of the Earth which is a sphere, and the error must be represented in terms of the angular arc length.

Fig. 4-17 Calculation of NS and EW error

Fig.4-17 illustrates the method on how to estimate the EW error and the NS error, where r_e is the earth radius, θ represents the latitude, r_s is the radius of the small circle which has the same latitude as the user and is parallel with the equator.

The NS position error is the error between the calculated position and the exact position in latitude, and if in units of meters, it equals to the latitude error in radians multiplied by the radius (in meters) of the great circle. A great circle is any circle on the earth surface which passes through the Earth's center, so its radius is constant, i.e., r_e . So no matter where the user is, the NS error in meters can be calculated by this method.

For EW position error, the situation is different. The EW position error is the error between the calculated position and the exact position in longitude. But for the same longitude error, the EW error may be different if the user is in certain places. An instance is that for one degree difference in longitude, the EW error is around 110 km when the user is on the equator, while it equals to zero when the user is at the North or the South Pole of the earth. One must calculate the EW error in meters by the radius of the small circle, i.e., r_s in Fig.4-17, and r_s can be derived by the earth radius and the user latitude, that is,

$$r_s = r_e \times \cos(\theta)$$

where θ is the user latitude and shown in Fig.4-17.

4. 4 SGR Test Results

The implemented software GPS receiver is tested first based on the set-up shown in Fig.4-1, except that the signal source is GS1010 GPS simulator from WelNavigate Inc. The RF front end shown in Fig.4-2 is used to down convert the 1575.42 MHz signal to 4.309 MHz, which is then sampled by the NI-5112

digitizer under a sample rate of 20 MSamples/sec, and totally 30-second signal is captured and saved on a PC hard disk.

The above signal is processed by the FFT based acquisition algorithm presented in section 7.3.2, and one typical result is shown in Fig.4-18.

Fig. 4-18 Acquisition result

The signal used in the acquisition is one millisecond long, and the frequency is searched in the range of ± 10 KHz around 4.309 MHz with a step size of 1 KHz [30] to determine Doppler shift.

If the input GPS signal contains a certain C/A code, a high correlation peak will be produced by the acquisition program, as shown in Fig.4-18, and then the peak value is compared with the computed threshold to determine whether a satellite is acquired or not. If a successful acquisition was reported by the program, the code phase and the frequency of the maximum value are the desired result, which will be passed to the tracking program.

As introduced already, acquisition is used to estimate the rough Doppler and code phase. Tracking, once the acquisition was accomplished, is to track the

variations of carrier phase and code offset mainly due to the movement between satellites and the receiver. The conventional DLL/PLL/FLL based tracking loop illustrated in Fig.4-6 is implemented in the SGR.

Fig. 4-19 Demodulated navigation data (I-Arm)

Fig. 4-20 Q-Arm output of Costas loop

Fig.4-19 and Fig.4-20 show the output of the in-phase arm and the quadrature

arm of the Costas loop respectively. In the steady or locked state, the I-arm output is the demodulated navigation data, while the Q-arm output should be noise, which can be confirmed by the above results.

The variation in the first 20 ms in Fig.4-20 is due to the unlocked state of the Costas loop. In other words, the loop is trying to lock the input signal. Correspondingly, the navigation data in the first 20 ms in Fig.4-19 have smaller amplitude.

Fig. 4-21 PLL discriminator output

Fig. 4-22 DLL discriminator output

Additionally, the tracking performance can be further confirmed by the discriminator output of the DLL and the Costas loop, which are shown in Fig.4-21 and Fig.4-22.

In the locked state, the local signal phase varies with that of the incoming signal, so the difference between them is ideally zero or the zero-mean noise in real life, as shown in Fig.4-21. Similarly, the DLL is to synchronize the local C/A code with the incoming C/A code, but some other factors, e.g., noise, relative motion between satellite and receiver, etc., can cause small misalignment between the incoming code and the local code, as shown in Fig.4-22, which is fed back to adjust the local code phase.

The demodulated data are then processed by the navigation module to obtain user position with an update rate of once per second. The calculated user position is compared with that set in the GS1010 simulator, and the errors are plotted in Fig. 4-23.

Fig. 4-23 Position errors in 30 seconds based on GS1010 simulator

The Root Mean Square (RMS) error in East-West direction is 15.5 meters, and

in North-South direction, it is 12.3 meters, which indicate that the implemented software GPS receiver has acceptable positioning performance.

Testing is re-conducted based on the set-up in Fig.4-1, and this time a rooftop Motorola antenna in Fig.4-24 is used to receive real GPS signal, which is then processed by the implemented SGR. The positions obtained by the SGR are compared with those from the GPS reference station (with NovAtel OEM4 receiver) in PWTC, and the errors are depicted in Fig.4-25.

Fig. 4-24 Motorola GPS antenna

Fig. 4-25 Position errors in 30 seconds based on real GPS signal

The RMS errors in East-West and North-South direction are 12.3 meters and 9.7 meters respectively, which indicate that the implemented SGR is able to work in the real world.

Summary

The implementation of a software GPS receiver is presented in this chapter. The implemented SGR is based on the typical receiver architecture and uses conventional positioning algorithms. The testing results show that it is able to achieve user position errors from 10 to 15 meters. The software receiver allows users to thoroughly understand the principles behind it, and is very flexible in adopting new techniques, so its implementation will be helpful to the future work relevant to global navigation in the Positioning and Wireless Technology Centre.

Chapter 5 Multipath Mitigation Based on Partial Autocorrelation Function

The Chapter 2 and Chapter 3 introduced the basic knowledge to understand GPS and the issue of multipath. This chapter will present the first technique on mitigating short-delay multipath signals for GPS proposed in this research work. The performances of the proposed techniques will be compared with the typical published technique through simulations based on both the ideal and the real signals.

5.1 Introduction

As has been mentioned, two steps in a standard GPS receiver are used to determine the time delay of the GPS signal. The first step is acquisition, in which the coarse delay is estimated in units of integer number of chips, and in the second step, the fraction of the delay (FOD), which is within one PRN code chip, is usually tracked by the code delay lock loop [31]. One GPS PRN code chip corresponds to approximately 300 meters in pseudorange, so the FOD must be estimated accurately. However, in indoor or urban canyons environments, multipath signals can distort the autocorrelation function of the PRN code and induce a non-negligible tracking error in the code delay lock loop, which degrades the positioning accuracy.

In this chapter, the first side lobe of the autocorrelation function of the pseudorandom noise code is to be studied, and the technique E1&E2 Tracker, which was introduced in Section 3.3.2, will be re-studied and the tracking error will be presented analytically here rather than empirically as done in the original paper [66]. In the technique presented below, the partial autocorrelation function

of the LOS signal which is slightly affected by multipath signals is to be detected based on linear interpolation algorithm. The pseudorange error in terms of multipath relative delay is calculated analytically, and the simulation results indicate that the proposed technique has superior performance compared with the 0.1-chip-spacing narrow correlator on mitigating multipath signals, especially the short-delay multipath signals.

The idea to be discussed in the following looks similar to the existing technique E1&E2 Tracker. However, there are three major differences: first, the conclusion presented in the E1&E2 Tracker is empirical, while analytical expressions are provided in this work discussed below; second, the linear interpolation method is used in our simulation when detecting the leading edge of the autocorrelation function; third, the property of the first side lobes of the autocorrelation functions of the 32 C/A codes and their effect on the multipath tracking error are discussed in details and grouped into seven groups in our work, which is not involved in the E1&E2 Tracker.

5.2 Property of Side lobes of PRN Code Autocorrelation Function

An infinite-length code of truly random chips has an autocorrelation function [78]:

$$R(\tau) = \begin{cases} 1-|\tau| & \text{for } |\tau| \leq 1 \\ 0 & \text{otherwise} \end{cases},$$

where $R(\tau)$ represents the autocorrelation function, and τ is the lag value in units of chips.

A m -sequence has an autocorrelation function slightly different from the above expression [78].

$$R(\tau) = \begin{cases} 1 - |\tau|(1/N + 1), & \text{for } |\tau| \leq 1 \\ -1/N, & \text{for } 1 \leq \tau \leq (N-1) \end{cases}$$

where N is the length of the m -sequence and τ is the lag value.

For GPS, the C/A codes for all navigation satellites are taken from a family of codes known as Gold codes [22]. Gold codes are generated by forming the product of two m -sequences with the same period. They are characterized by 4-level autocorrelation functions [55]:

$$R(\tau) = \begin{cases} 1 - |\tau|(1 - k/N), & \text{for } \tau \leq 1 \\ x/N, -1/N, \text{ or } -y/N, & \text{for } \tau = \pm 2, \dots, \pm[N-1] \end{cases}$$

where $k = 1, x$, or y ; $x = 2^{(n+2)/2} - 1$; $y = 2^{(n+2)/2} + 1$ and $N = 2^n - 1$. In GPS, $n = 10$, so $N = 1023$, and the side lobes of autocorrelation function of the PRN code is 3-level, i.e., $63/1023$, $-1/1023$ and $-65/1023$.

The C/A codes in GPS repeat themselves every one millisecond and one period includes 1023 chips, so one C/A code chip is roughly 997.5 nanoseconds and corresponds to about 300 meters in range. In urban areas, reflectors are usually tens of meters away from the receiver, so in the following, only the side lobe between -1 and -2 (or between 1 and 2) in chips of the autocorrelation function, called the first side lobe, will be discussed based on simulations.

(1)

(2)

(3)

(4)

(5)

(6)

(7)

Fig.5-1 Autocorrelation functions and their first side lobes

We found that the 32 navigation satellites in GPS can be categorized as seven groups according to the correlation values at -1 and -2 chip of their autocorrelation functions (Table 5.1), where N equals 1023, and the autocorrelation functions are shown in Fig.5-1.

Table 5.1 Navigation satellite groups and their first side lobes

	Group (1)	Group (2)	Group (3)	Group (4)	Group (5)	Group (6)	Group (7)
Satellites included	#1, #2, #3, #5, #9, #11, #12, #13, #14, #20, #23, #25, #26, #27, #29, #31, #32	#10, #18, #30	#4, #6, #16, #19, #28,	#7, #15, #17, #21	#8	#22	#24
V_A	$-1/N$	$-1/N$	$-1/N$	$63/N$	$-65/N$	$-65/N$	$63/N$
V_B	$-1/N$	$63/N$	$-65/N$	$-1/N$	$63/N$	$-1/N$	$-65/N$

In Table 5.1, V_A and V_B are the amplitude of the autocorrelation function at -1 and -2 chip respectively. Table 5.1 and Fig.5-1 indicate that all the peak values of the autocorrelation functions are always one. The first side lobe of the PRN code autocorrelation functions in Group (1) is flat and equals to $-1/1023$, which is quite small and hence can be neglected. However, the first side lobes in other groups are nonzero, linear with different slopes, and they will have different effects on the tracking error caused by multipath signals.

5.3 FOD Estimation Technique under Multipath

A new technique for mitigating multipath will be presented in this section, and

for ease of discussions, it is assumed that the GPS signal is shifted to baseband and the PRN code in Group (1) is selected as an example in the illustration of the proposed technique. The multipath caused tracking errors corresponding to the PRN codes in other groups of Table 5.1 will be given based on mathematic analysis.

5.3.1 Methodology

In GPS, each satellite transmits a unique PRN code, which can be denoted by,

$$s_1(t) = p(t - \tau_0 - \tau_e) \quad (5.1)$$

where τ_0 and τ_e are the coarse time delay and the FOD in chips respectively, and $p(t)$ is the PRN code. In a GPS receiver, τ_0 can be estimated in units of integer number of chips by acquisition, so only τ_e , i.e., FOD, remains unknown. The following discussion will take the PRN code in Group (1) of Table 5.1 as an example to introduce the proposal and only focus on the estimation of τ_e .

Fig.5-2 ACF and the shifted ACF

In Fig.5-2, the dotted line represents the ideal normalized autocorrelation function (ACF) of the PRN code in Group (1), and the solid line is the crosscorrelation function of the incoming code, $s_1(t)$, and the locally generated

code, i.e., $p(t - \tau_0)$. τ_e is not included in the local replica, because it is unknown. In Fig.5-2, τ_e can be represented as the distance between the ideal ACF and the shifted ACF. The value corresponding to the side lobes of the ACF is $-1/N$, where N equals 1023. Ideally, the property of the PRN code determines that the coordinates of P_1 are roughly $(-1, 0)$ in units of chips. If the coordinates of P_2 is obtained, e.g. $(x, 0)$, the FOD of the LOS thus can be calculated by:

$$\hat{\tau}_e = x + 1 \quad (5.2)$$

where $\hat{\tau}_e$ is the estimate of τ_e in chips.

For ease of discussion, only one multipath signal will be considered in the following, and in order to focus on multipath, noise is ignored. Only the range error caused by the multipath code which is in-phase with the LOS signal is to be studied in detail. The error caused by an out-of-phase multipath code can be studied in the same way and will be mentioned briefly below.

Fig.5-3 LOS, multipath and the resulting correlation functions

When a LOS code is accompanied by an in-phase multipath code, the incoming code can be expressed as,

$$s_2(t) = p(t - \tau_0 - \tau_e) + \alpha \cdot p(t - \tau_0 - \tau_1)$$

where τ_e is the value to be estimated, τ_1 is the FOD of the multipath code, and

α is defined as the multipath-to-direct (M/D) ratio in amplitude. Fig.5-3 shows the correlation functions for the LOS signal, the multipath signal, and the resultant composite signal, where the first side lobes of the correlation functions are neglected, because they equals $-1/1023$ and hence are very small. The important information in Fig.5-3 is that the linear segment from point A to point B of the composite crosscorrelation function is not affected by multipath, because the multipath signal is later than the LOS signal. In other words, the linear segment from A to B does not contain any contribution from multipath signals ideally. So the coordinates of point A can be approximately regarded as that of the point P_1 in Fig.5-2, i.e., $(-1,0)$.

Fig.5-4 Correlation function and shifted correlation function

Similar to the case when there is no multipath, if a multipath signal is present, a shifted composite correlation function in Fig.5-4 will be obtained when performing the crosscorrelation between the local PRN code, i.e., $p(t - \tau_0)$, and the incoming code, i.e., $s_2(t)$. The coordinates of point A' in Fig.5-4 are not affected by the multipath signal, so the previous discussions of estimating the FOD of the LOS signal with no multipath and equation (5.2) are still applicable, even if multipath signals now exist.

5.3.2 Discussions on an Out-of-phase Multipath Signal

When the multipath signal is out-of-phase with the direct signal, the composite correlation function equals the direct correlation function minus the multipath correlation function, as shown in Fig.5-5.

Fig.5-5 Composite correlation function caused by a negative multipath signal

It can be seen from Fig.5-5 that the composite correlation function has a smaller peak amplitude than that of the direct ACF, but again the line segment AB is not affected by the multipath signal, so the previous analysis on an in-phase multipath signal is still applicable here.

5.3.3 Algorithm

The above proposal can be illustrated by the algorithm below and Fig.5-7. Denote the incoming PRN code as $s(t)$ and the local PRN code as $c(t)$ which equals $p(t - \tau_0)$, where $s(t)$ includes one LOS code with/without multipath codes and τ_0 is estimated by the process of acquisition, then the FOD of the LOS signal can be estimated by the following steps:

- i) Calculate the crosscorrelation value of $c(t)$ and $s(t)$, e.g., v_0 , and record it

as $(0, v_0)$;

- ii) Shift $c(t)$ by a certain spacing, e.g., $-\Delta$, then calculate the crosscorrelation value of $c(t+\Delta)$ and $s(t)$, and record it as $(-\Delta, v_1)$. Repeat this step and obtain enough crosscorrelation samples and record them as a vector, i.e., $\mathbf{V} = [(0, v_0), (-\Delta, v_1), \dots, (-n_i\Delta, v_i)]$;
- iii) Use a linear interpolation algorithm on the vector $\mathbf{V}$ to obtain the line segment AB; An example is illustrated in Fig.5-6, where the black dots represent the crosscorrelation values or the vector $\mathbf{V}$ in step (ii). By using a typical linear interpolation algorithm, the line segment AB can be obtained from these discrete cross correlation values.
- iv) Find the intersection of the line segment AB and the horizontal axis, i.e., $(x, 0)$;
- v) Use equation (5.2) to estimate the FOD of the LOS signal.

Fig.5-6 Linear interpolation on discrete correlation values

Fig.5-7 Block diagram of the proposed algorithm

5.4 Mathematic Analysis

In the above illustrations of the proposed technique, the first side lobe of the autocorrelation function of the PRN codes in Group (1) is neglected, because it is very small. However, the first side lobes of the autocorrelation functions in other groups must be taken into considerations before the above methodology is applied. In the following, a generic mathematical analysis on the effect of the first side lobe will be presented.

Fig.5-8 The autocorrelation function of the PRN codes in Group (3)

As an example, the autocorrelation function of the PRN codes in Group (3) is shown in Fig.5-8. The line segment OP and OQ can be expressed respectively as:

$$y_1 = k_1x + b_1 \quad (5.3)$$

$$y_2 = k_2x + b_2 \quad (5.4)$$

Where, k_1 and k_2 is the slope of the line segment OP and OQ; b_1 and b_2 is their intercept respectively. Actually, the intercept of OP is always equal to one for all the PRN codes, i.e., $b_1=1$, because the coordinates of P is always at (0, 1); k_1 , k_2 and b_2 can be calculated easily based on the values in Table 5.1. From the

expression of OP, i.e., equation (5.3), we can obtain the reference point, i.e. $(x_r, 0)$, which is the intersection of the early edge of the autocorrelation function and the x -axis when multipath is not present. So,

$$x_r = -\frac{1}{k_1}$$

If a single multipath signal exists, its first side lobe of autocorrelation function will be shifted into the range of the main lobe of the LOS autocorrelation function and added onto its early edge, because multipath signals are always later than the LOS signal. Therefore, the expression of the resultant partial composite correlation function is:

$$y = y_1 + M \cdot y_2(x - d) = (k_1 + Mk_2)x + 1 - Mk_2d + Mb_2 \quad (5.5)$$

where M is the M/D ratio in amplitude, and d is the relative multipath delay. The intersection of this composite expression and x -axis can be calculated by setting y in (5.5) to zero:

$$x = -\frac{1 - Mk_2d + Mb_2}{k_1 + Mk_2} \quad (5.6)$$

The FOD of the LOS signal is thus estimated by the distance between this intersection and the reference point, i.e.

$$\hat{\tau}_e = x - x_r = \frac{1}{k_1} - \frac{1 - Mk_2d + Mb_2}{k_1 + Mk_2} \quad (5.7)$$

where $\hat{\tau}_e$ is the estimate of τ_e .

Equation (5.7) shows that $\hat{\tau}_e$ is a function of relative multipath delay, once the M/D, i.e., M , is fixed. Under ideal conditions, i.e. $x_r = -1$, equation (5.7) will be equivalent to equation (5.2).

5.5 Simulation Results and Discussions

5.5.1 Ideal Simulations

Fig.5-9 Range errors as a function of relative multipath delay

Simulations are conducted under the assumption that only a single multipath signal exists and the M/D equals 0.5. In order to focus on multipath, noise is neglected at the moment. Range error of the proposed technique and that of the 0.1-chip-spacing narrow correlator are shown in Fig.5-9, where the multipath signal is in-phase with the LOS signal. As has been mentioned, Group (1) includes 17 satellites out of the total 32 satellites. From Fig.5-9, it can be seen that the range error of Group (1) is approximately zero, which is desired. The absolute range errors in Group (2) and Group (3) are much smaller than that of narrow correlator as well in the range from 0 to 0.8 chips. 0.8 chips correspond to approximately 240 meters for C/A code, which includes our interested range in general, especially for indoor environments. However, the reverse is observed in Group (4), Group (5), Group (6) and Group (7). The reason is that the value of V_A in the Group (4-7) in Table 5.1 is either $63/N$ or $-65/N$, and when the point of V_A

from the multipath correlation function is shifted into the main lobe of the direct correlation function due to multipath delay, it will have a more significant effect than the point of V_A from the groups (1-3) which is with the value of $-1/N$. Fortunately, these four groups only include 7 satellites from the entire constellation of 32 satellites. Furthermore, in many practical GPS receivers, a narrow correlator with an early-late spacing greater than 0.1 chips is usually used, which will give worse performance than that shown in Fig.5-9. For multipath signal that is out-of-phase with the LOS signal, simulations can be similarly conducted.

The range errors from the maximum likelihood based method are also given in Fig.5-9. Within the range of short multipath delay, the multipath range errors from the groups (1-3) in the proposed technique have similar performance as those from the maximum likelihood based method. But the overall performance of the maximum likelihood based method is much better than the proposed technique here, because the maximum likelihood based method uses a number of correlators to estimate the amplitude, delay and carrier phase of each multipath signal, and its computation load is very high. The proposed technique here uses relatively simple processing on the correlation function, so it is different from the maximum likelihood based estimator in principle, and therefore they are not comparable in fact.

5.5.2 Measurement Set-up Descriptions

Fig.5-10 GPS signal measurement set-up

Fig.5-11 Instance of the measurement set-up

Fig.5-12 Interconnection between HP8593E, HP89411A and HP89410A

The proposed technique is further verified by using the GPS signal from GS1010 L1 signal simulator. The measurement setup block diagram is shown in Fig.5-10 and a photo of the setup is given in Fig.5-11.

In order to make the HP8593E, HP89411A and HP89410A work together, additional cabling is needed, which is illustrated in Fig.5-12.

The reference signal in Fig.5-12 is a 10 MHz signal for synchronizing the three pieces of equipment; otherwise the relative frequency drift may cause the measurement to be inaccurate. The GPIB bus is for the communications between HP8593E and HP89410A so that the HP8593E can be controlled by the HP89410A if needed.

Table 5.2 Configurations of the GS1010 Simulator

Date	Year = 2004, Month = 1, Day = 8.
Starting time	Hour = 2, min = 15, second = 0.
Latitude	N 1°20'34.102"
Longitude	E 103°40'45.714"
Altitude	100 meters
speed	0
Heading	0

The GS1010 simulator is from WelNavigate Inc., and is able to output the customized GPS signal. By this simulator, the user can build a test scenario including date, time, route and speed etc. to dynamically simulate the signal received by a receiver. The configurations of the simulator used in the measurements are shown in Table 5.2, which shows that the GS1010 is simulating the GPS signal received by a static receiver at the specified time and location.

The amplifier configuration used is two ZHL-42 amplifiers from Mini Circuits Inc., each of which is with a frequency range from 700 MHz to 4200 MHz and a

typical gain of 30 dB.

The HP8593E spectrum analyzer is not used for observing the signal spectrum here, but is used for down converting the input signal to an Intermediate Frequency (IF) of 21.4 MHz, so in this measurement, it is used as a mixer. To make it a mixer, some special configurations must be set, which are listed in Table 5.3.

Table 5.3 Configurations for HP 8593E Spectrum Analyzer

Center Frequency	1575.42 MHz
Resolution Bandwidth	3 MHz (max)
Span	0 Hz
Attenuation	0 dB

Fig.5-13 Basic mixer architecture

In these settings, 1575.42 MHz is the GPS L1 band frequency. The span and resolution bandwidth can be understood through the general mixer block diagram, shown in Fig.5-13. The span must be set to zero; otherwise the oscillator in the spectrum analyzer will sweep its signal frequency in the range of span around the center frequency, which will generate a cluster of IF signals rather than the desired single signal of 21.4 MHz. The resolution bandwidth acts as the bandwidth of the low-pass filter, and it must be larger than the GPS signal bandwidth of 2.046 MHz (double-sided). In the measurement, the resolution

bandwidth is set as 3 MHz. In addition, the spectrum analyzer usually attenuates the input signal by 10 dB by default to protect itself. For GPS signal measurement, setting the span to zero will be helpful to obtain a stronger signal.

The HP89411A down converter has the fixed input and output signal frequency, which is 21.4 MHz and 5.6 MHz respectively. It has only one adjustable parameter, conversion gain. The GPS signal is very weak, so this conversion gain should be set to its maximum value (+5 dB) to increase the signal power.

The HP89410A vector signal analyzer is used to collect the down converted signal and save it into its memory. To capture signal, this equipment must work in the vector mode; to save signal in the real data format rather than the complex data, the start frequency in HP89410A must be set to zero and the capture mode must be set to 'baseband' rather than 'zoom'. The sampling rate of HP89410A is determined by its span. In this measurement, the span is set to 8 MHz, and then the sampling rate is automatically configured as 25.6 MHz. The HP89410A can be accessed through network, and the data can be shifted to PC hard disk through File Transfer Protocol (FTP). Usually, the initial data are in Standard Data Format (SDF). By the software tool coming with the HP89410A analyzer, the SDF can be converted to Matlab or other file formats.

By the above set-up, the 1575.42 MHz GPS signal from GS1010 GPS Simulator will be downconverted to an intermediate frequency of 21.4 MHz after passing through the HP8593E spectrum analyzer, and further down-converted to 5.6 MHz by HP89411A, which is then captured by HP89410A vector signal analyzer under a sampling rate of 25.6 MHz and stored into PC for post processing.

5.5.3 Measurement Results and Discussions

The captured signal from the above set-up acts as a LOS signal, and an in phase multipath signal with a M/D of 0.5 is constructed in a PC by delaying the LOS signal by a certain delay, i.e., the relative multipath delay, as shown in Fig.5-14.

Fig.5-14 Construct the composite signal

It should be noted that the GS1010 GPS simulator shown in Fig.5-11 simulates the real satellite constellation and the output GPS signal has the same noise level as the real navigation signal. For a received GPS signal with a low signal-to-noise ratio (SNR), averaging can be further adopted to reduce noise when calculating the correlation values, because the mean of noise is zero ideally.

The GS1010 simulator is with a 3MHz filter, so the generated GPS signal is band limited. Typically, a band limited signal will cause the peak of the C/A code autocorrelation function rounded. However, in the technique proposed here, the linear interpolation algorithm is used, which weakened the filtering effect.

Fig.5-15 shows the range error caused by the multipath signal in the seven groups respectively. Take the signal in Fig.5-15(1) as an example: it shows that the absolute range error caused by the constructed multipath signal is within 5 meters roughly, which is better than that of the narrow correlator.

(1)

(2)

(3)

(4)

(5)

(6)

(7)

Fig.5-15 Range errors as a function of relative multipath delay for different groups

5.5 Work with Delay Lock Loop

The proposed technique is based on the autocorrelation function, so it is very easy to combine the technique with the conventional delay lock loop. The discrimination function which is distorted by multipath signal is plotted in Fig.5-16.

Fig.5-16 Multipath-contaminated discrimination function

Because the multipath signal has less effect on the leading portion of the discrimination function, i.e., AB in Fig.5-16, the proposed technique thus can be applied into it, which means to use the portion of AB as an assisted method to determine the location of the direct C/A code

Summary

The property of the first side lobe of the autocorrelation functions of the PRN codes used in GPS and a technique aiming at mitigating multipath are investigated in this section. For 25 satellites out of the total 32 navigation satellites, the proposed technique has a superior performance over the 0.1-chip-spacing narrow correlator on multipath mitigation. In the case of hardware implementations, the proposed technique can be implemented economically, because only one correlator is needed for the FOD estimation of the LOS signal. If fast processing is desired, multiple and parallel correlators should be adopted.

Chapter 6 Short-Delay Multipath Mitigation Technique Based on Virtual Multipath

This chapter will introduce the second proposed technique on multipath mitigation, i.e., a virtual multipath based technique focusing on mitigating short-delay multipath signals. Analysis under ideal conditions as well as simulations based on real GPS signals indicate that the proposed technique is able to mitigate short-delay multipath signals by determining the peak location of the direct ACF from the peak of the Virtual Line-of-Sight function, and the maximum range error of this technique is within 2 meters for a single multipath signal with a typical multipath-to-direct ratio (M/D) of 0.5. The main advantages of the proposed technique include significantly improved ranging accuracy and insensitivity to the M/D ratio when compared with other typical multipath mitigation techniques. In addition, these performances do not require wideband filters in the RF front-end and is not dependant on environments.

6.1 Removing Multipath under Ideal Conditions

6.1.1 Signal model and assumptions

The baseband signal model to be used is shown in (6.1).

$$s(t) = A(t)c(t - \tau) + \sum_{k=1}^N \alpha_k \cdot c(t - \tau_k) \cdot e^{j\phi_k} + n(t) \quad (6.1)$$

where,

$A(t)$ or α_k	LOS or multipath signal amplitude;
$c(t)$	C/A code;
τ or τ_k	delay of the direct or multipath code;

N	the Number of multipath signals;
ϕ_k	phase shift of the k -th multipath signal;
$n(t)$	noise;

In real life, the phase shifts are uniformly distributed in $[0, 2\pi]$, and the worse case for the delay estimation of the line-of-sight signal is when the multipath signals are in phase or out of phase ($\phi_k = 0, \pi$) with the LOS signal [79, 80]. Multipath components with other phases will cause less distortion in the composite correlation function at the receiver side when compared with the in-phase or out-of-phase signal.

Similar to that mentioned in section 5.1 in Chapter 5, the proposed method is used to only estimate the fraction of delay, which is less than one C/A code chip.

For ease of discussion, noise will be neglected and the range errors caused by a single multipath signal which is in-phase or out-of-phase with the direct signal will be introduced, followed by the discussions on multiple reflections.

6.1.2 Mitigate a Single Multipath

In baseband, when a direct C/A code is accompanied by a single in-phase multipath, the received code at a GPS receiver can be expressed as,

$$s(t) = \alpha_0 \cdot p(t - \tau_0 - \tau_e) + \alpha_1 \cdot p(t - \tau_0 - \tau_1) \quad (6.2)$$

where τ_e and τ_1 is the fraction of the delay; α_0 and α_1 is the amplitude of the direct and the multipath code respectively.

At the receiver side, the obtained composite correlation function could be positive or negative. Table 6.1 lists all the combinations of a LOS signal and a multipath signal, where ‘+’ represents a positive correlation function and ‘-’ represents a negative correlation function. The following discussion will focus on the strong multipath signal, because the strong multipath signal is dominant in

Table 6.1 Combination Possibilities of a LOS signal and a Multipath signal

Case	1	2	3	4	5	6	7	8
+LOS	√	√	√	√				
-LOS					√	√	√	√
+MP(strong)	√				√			
+MP(weak)		√				√		
-MP(strong)			√				√	
-MP(weak)				√				√
Resulting Composite	+	+	-	+	+	-	-	-

indoor environment and causes larger range error when compared with a weak multipath signal. As for the weak multipath signal, it can be analyzed in the same way.

Case A: A positive direct ACF with a positive strong multipath ACF

Fig.6-1 shows the correlation functions for the LOS signal, the multipath signal, and the resulting composite signal. This case corresponds to the Case 1 and the Case 7 in Table 6.1 (Note: the composite correlation function in the Case 7 must be multiplied by -1 first).

Fig.6-1 LOS, multipath and the resulting composite functions

Under multipath, the composite correlation function is neither triangular nor symmetric, but the two line segments, i.e., PA and PB in Fig.6-1, are still linear. The linear nature here is due to the contribution of two linear functions. The major difference between the line segment PA and PB is that their slopes are different, due to the slope difference between the early and the late edge of the direct ACF. Accurate estimation of pseudorange needs to find the beginning point of the direct C/A code, which is represented by the peak position of the direct ACF. Fig.6-1 indicates that the peak position of the direct ACF is further represented by the turning point, i.e., P , because the peak of the direct ACF is aligned with the point of P . However, it is not easy to determine the location of P , because the slope of PA and PB may both be positive, especially when the multipath signal is much stronger than the direct one. This will occur when the GPS receiver is located indoors.

Assume that the amplitude of the direct signal is known, and then the mathematic expression for the early edge and the late edge of the direct ACF can be obtained easily as,

$$y_E(\tau) = \alpha_0 \tau + \alpha_0, \quad (-1 < \tau < 0) \quad (6.3)$$

$$y_L(\tau) = -\alpha_0 \tau + \alpha_0, \quad (0 < \tau < 1) \quad (6.4)$$

Fig.6-2 Concept of virtual multipath

In Fig.6-2, the concept of Virtual Multipath (VM) correlation function, i.e., CBD, is illustrated. The virtual multipath correlation function has peak amplitude equal to the maximum value of the composite correlation function, and has a width of two C/A code chips on the horizontal axis.

The slope of the line segment PA in Fig.6-2 equals to the slope of the early edge of the direct ACF plus the slope of the early edge of the multipath ACF, i.e.,

$$k_{PA} = \alpha_0 + \alpha_1 \quad (6.5)$$

And the slope of PB is,

$$k_{PB} = -\alpha_0 + \alpha_1 \quad (6.6)$$

where α_0 and α_1 is the peak amplitude of the direct and the multipath ACF respectively, as shown in Fig.6-1.

Similarly, the slope of the line segment CB in Fig.6-2 is

$$k_{CB} = \alpha_x + \alpha_1 \quad (6.7)$$

where α_x is the amplitude of the point Z on the late edge of the direct ACF, which has a delay of τ_1 and here τ_1 is the relative delay of the multipath code, as shown in Fig.6-2, i.e.,

$$\alpha_x = y_L \Big|_{\tau=\tau_1} = -\alpha_0 \tau + \alpha_0 \Big|_{\tau=\tau_1} = \alpha_0(1 - \tau_1)$$

Obviously, the following inequalities exist.

$$k_{VM} = \alpha_x + \alpha_1 < \alpha_0 + \alpha_1 = k_{PA} \quad (6.8)$$

$$k_{VM} = \alpha_x + \alpha_1 > -\alpha_0 + \alpha_1 = k_{PB} \quad (6.9)$$

Now, we assume that the composite correlation function in Fig.6-2 does contain such a virtual multipath correlation function. Since the amplitude and the location of the VM correlation function, i.e., CBD, is known, the resultant correlation function obtained after the VM correlation function is subtracted from the composite signal is shown in Fig.6-3, and this correlation function, i.e., A'P'B'D in Fig.6-3, will be named as the Virtual Line-Of-Sight (VLOS)

correlation function in the following. From equation (6.8) and (6.9), the following can be obtained,

$$\begin{aligned} k_{P'A'} &= k_{PA} - k_{VM} > 0 \\ k_{P'B'} &= k_{PB} - k_{VM} < 0 \end{aligned} \quad (6.10)$$

Fig.6-3 The resulting VLOS after the VM is subtracted

The inequalities of (6.10) indicate that the point of P' represents a peak on the VLOS correlation function. Fig.6-3 also indicates that the position of the turning point P can be represented by P' , because P is aligned with P' . It is known that P is also aligned with the peak of the direct ACF as mentioned in Fig.6-2, so once the peak position of the VLOS is found, the position of the direct signal is determined.

It is worth noting that the location of the virtual multipath correlation function can be at another location (for example due to a longer delayed multi-path given by points $C''B''D''$ in Fig.6-3). However, the location of the peak P' remains unchanged if $C''B''D''$ instead of CBD in the above discussion is subtracted from the original composite correlation function. This feature implies that the location of P' can be determined by a VM correlation function with different relative delays, which is helpful to determine the peak position of the direct ACF when multipath signal is weaker.

Case B: A negative direct ACF with a positive strong multipath ACF

This case corresponds to the Case 5 and the Case 3 in Table 6.1 (Note: the composite correlation function in the Case 3 must be multiplied by -1 first).

The obtained direct ACF could be negative due to the phase change in propagation. For this case, the VM correlation function can be constructed in the same way as that in Fig.6-2 and the VLOS obtained as in Fig.6-3. However, mathematic analysis in (6.8) and (6.9) indicates that the slope of P'A' is negative and the slope of P'B' is positive, which means that the negative peak on the VLOS function represents the peak position of the direct ACF. To be consistent, the entire VLOS function can be multiplied by -1 first once it is detected that the slope of the P'A' is negative, and then the positive peak of the resultant VLOS function will represent the peak position of the direct ACF. Since only the short-delay multipath signals are concerned, the P'A' is the dominant line segment on the VLOS function and can be detected easily.

6.1.3 Discussions on N multipath signals

Fig.6-4 The case of two multipath signals

Fig.6-4 gives a typical example for discussing the case of N multipath signals, where there is a direct signal with two multipath signals. The relationships of the amplitudes and the delays are

$$\alpha_2 < \alpha_0 < \alpha_1 \quad (6.11)$$

$$\tau_1 < \tau_2 \quad (6.12)$$

According to the definition, the peak amplitude or the slope of the early edge of the VM correlation function equals the maximum value of the composite correlation function, so,

$$k_{VM} = V_B = \alpha_1 + V_1 + V_2 \quad (6.13)$$

where α_1 , V_1 , V_2 is the peak amplitude of the ACF of MP1, the amplitude of point Z1 and the amplitude of point Z2 respectively in Fig.6-4. It is seen that the point of Z1 is on the late edge of the direct ACF and the point of Z2 is on the early edge of the ACF of multipath2. From (6.3) and (6.4), it can be obtained that,

$$\begin{aligned} k_{VM} &= \alpha_1 + y_L \Big|_{\tau=\tau_1} + \frac{\alpha_2}{\alpha_0} \cdot y_E(\tau - \tau_2) \Big|_{\tau=\tau_1} \\ &= \alpha_1 + \alpha_2 + \alpha_0(1 - \tau_1) + \alpha_2(\tau_1 - \tau_2) \end{aligned} \quad (6.14)$$

Since $\tau_1 < \tau_2$ and $\alpha_2 < \alpha_0$,

$$\begin{aligned} k_{VM} &> \alpha_1 + \alpha_2 + \alpha_0(1 - \tau_1) + \alpha_0(\tau_1 - \tau_2) \\ &= \alpha_1 + \alpha_2 + \alpha_0(1 - \tau_2) \end{aligned} \quad (6.15)$$

While, the slope of the line segment PA and PB is,

$$k_{PA} = \alpha_1 + \alpha_2 + \alpha_0 \quad (6.16)$$

$$k_{PB} = \alpha_1 + \alpha_2 - \alpha_0 \quad (6.17)$$

Compare (6.14) with (6.16) and (6.15) with (6.17) will obtain

$$k_{PB} < k_{VM} < k_{PA} \quad (6.18)$$

which is the same as the result in (6.8) and (6.9).

Fig.6-5 The resulting VLOS after the VM is subtracted

The inequality (6.18) indicates that the peak position of the direct ACF can still be represented by the peak position of the VLOS correlation function, as shown in Fig.6-5, which is obtained by subtracting the VM correlation function from the composite function in Fig.6-4. Furthermore, the LOS correlation function peak should be always earlier than that of the VM signal, because the VM signal is later than the LOS signal, and therefore only the VLOS peak at the left of the VM correlation function peak is effective.

It will be very difficult to discuss analytically when the number of multipath signals are larger than two, because there are numerous combinations of multipath delay, amplitude and polarization, so in the thesis, simulations are used to verify the performance of the proposed technique. Please refer to the Section of 6.3.

6.2 Real-life Multipath Mitigation

The above section introduced the proposed technique when the correlation functions are ideally triangular and are with sharp peaks. However, the peak of the autocorrelation function of the PRN code is actually rounded due to the finite-bandwidth filter in a GPS receiver's RF front-end, as shown in Fig.6-6, where the plot is zoomed in.

Fig.6-6 The rounded autocorrelation function peak (zoomed in)

The rounded ACF will bring two additional issues when estimating the peak position of the direct ACF by the above technique. First, the peak on the VLOS function, which is obtained after the virtual multipath correlation function is subtracted from the composite correlation function, does not represent the location of the actual peak of the direct ACF. In fact, the peak now on the VLOS function represents the point after which the slope of the early edge of the original composite correlation function is smaller than the slope of the early edge of the VM. In other words, there will be an offset between the peak location of the VLOS and the peak location of the actual direct ACF. Second, if the location of the virtual multipath changes, the peak location of the VLOS function could be affected slightly due to that the actual correlation function peak is rounded, and this imply that an average location of the VLOS function peak should be used.

In order to solve these two issues, a step called learning, i.e., estimating the offset mentioned above, needs to be performed based on the direct peak-rounded ACF. This is intended to find the maximum value on the direct ACF, construct a

VM based on this maximum value, and then subtract the VM from the direct signal. However, the peak location of the VLOS is not obvious, because there is no linear line segment on the VLOS function when a real-world GPS signal is used, as shown in Fig.6-7. In order to determine the peak position of the VLOS function, linear curve fitting can be used to find the segment P'A' and P'B' in Fig.6-7, and the peak thus can be found at the intersection of P'A' and P'B'. The offset is achieved by calculating the distance between this intersection and the actual peak of the LOS autocorrelation function. For a virtual multipath correlation function with different delays, the obtained offsets could be slightly different, and the mean of all the offsets is selected as the final offset.

The goal of learning is to minimize the effect caused by finite-bandwidth filters, so for a certain GPS receiver, learning needs to be conducted only once. As for the LOS signal used in learning, it must be a real GPS signal captured in a multipath-free environment.

Fig.6-7 Determine the peak location of VLOS function by linear curve fitting

6.3 Simulations and Discussions

Fig.6-8 Construction of the composite signal

Simulations have been conducted to test the proposed algorithm by using a direct signal and nine deterministic multipath signals constructed by Fig.6-8, which is deduced from the two-ray model given in [81]. The signals in the simulations below do not contain navigation data or it is assumed that the navigation data bit transitions are known. The LOS signal in Fig.6-8 is with a 1575.42 MHz carrier modulated by a C/A code. In each path, the LOS signal is delayed according to Table 5.2, and then passes through a Rayleigh fading channel [82-84], in which the amplitude α_i is Rayleigh distributed and the phase φ_i is uniformly distributed in $[0, 2\pi]$. A constant gain is assigned in each path according to Table 6.2 to determine whether the corresponding multipath signal is attenuated or not, and a higher path gain will result in a stronger multipath signal statistically. The maximum Doppler shift applied in the Rayleigh fading channel is 10 KHz [23], and the AWGN noise channel is with a typical Signal-to-Noise Ratio of -20 dB for GPS signal [19].

Table 6.2 Relative Delays (in chips) and the Corresponding Path Gain

	Path0	Path1	Path2	Path3	Path4	Path5	Path6	Path7	Path8	Path9
Path Delay (d_i)	0	0.01	0.02	0.03	0.04	0.05	0.06	0.07	0.08	0.09
Path Gain (k_i)	1.0	0.8	0.7	0.6	0.5	2	0.4	0.3	0.2	0.1

Fig.6-9 Range errors by VLOS-based method when a strong multipath exists

Fig.6-10 Range errors by narrow correlator when a strong multipath exists

The correlations and range errors are computed every one millisecond (cycle length of C/A code) based on the real part of the composite complex signal, and the simulation runs one thousand milliseconds to evaluate the performance of the proposed method under fading and a noisy channel. The range errors in terms of time are plotted in Fig.6-9, in which the average absolute range error is 5.5 meters, and 78.2% of the entire absolute range errors are within 10 meters. Even for the large range errors in Fig.6-9, which are around 15 meters, they are acceptable as well, because multiple reflections, including one strong multipath at 0.05 chip delay (around 15 meters), are applied in the simulation. The range errors obtained by the 0.1-chip-spacing narrow correlator under the same conditions are shown in Fig.6-10, which are with a mean of 22.7 meters. Comparison indicates that the VLOS based method achieved an improvement of 17.2 meters on the mean range error.

The simulation is re-run based on the signals in Table 6.2 and Fig.6-8, except that the LOS signal is disabled, to simulate the scenario when the direct signal is totally blocked, and the range errors are shown in Fig.6-11, in which 66.3% of the absolute range errors are within 10 meters. The average absolute range error is 7.8 meters, which is larger than the previous simulation, and the reason is that there is no LOS signal now and the proposed method will track the first arrival multipath signal. In some indoor or urban environments, the direct signal may be blocked by metal structures or building, so tracking the first-arrival multipath signal will help reduce the range errors. The range errors by the 0.1-chip-spacing narrow correlator under the conditions are given in Fig.6-12 which is with a mean of 24.3 meters, and the VLOS based method has an improvement of 16.5 meters on the mean range error.

Fig.6-11 Range errors by VLOS based method for NLOS

Fig.6-12 Range errors by narrow correlator for NLOS

Fig.6-13 Average range errors variation with strongest multipath delay

The third simulation is conducted to evaluate the effect of the delay of the strongest multipath signal on the average range error when the LOS signal is present. Again, the simulation is based on Table 6.2 and Fig.6-8, except that the relative delay of the strongest multipath, i.e., Path5, changes from 0.01 chips to 0.09 chips while the path gain of other multi-paths and the direct path remains unchanged. For each relative delay, the simulation runs one thousand milliseconds and the mean of the 1000 range errors is recorded and plotted in Fig.6-13. As depicted in Fig.6-13, the minimum error is around 4.1 meters and the maximum error is around 6.1 meters when the strongest multipath signal varies from 0.01 chip (around 3 meters) delay to 0.09 chip (around 27 meters) delays. This indicates that the delay of the strong multipath signal has a weak influence on the average range errors.

Fig.6-14 Set-up for capturing LOS GPS signal

The proposal is further tested based on the real GPS signal. The multipath-free GPS signal is received by a Motorola GPS antenna which is set up on the rooftop of Research Techno Plaza, Nanyang Technological University, and then the signal is downconverted to the intermediate frequency of 4.309 MHz by the RF front-end which is with a chipset of GP2010 from Zarlink Inc., and sampled by Agilent Infiniium Scope under a sampling frequency of 1 GHz so that a short-delay multipath can be constructed from the captured signal. The entire set-up is shown in Fig.6-14. The satellite signal of PRN28 is selected from the received GPS signal for our simulation.

Fig.6-15 Construct the composite signal by a LOS signal

A multipath signal is constructed by delaying the LOS signal by a certain relative delay, as shown in Fig.6-15, where M/D represents the multipath-to-direct ratio in amplitude, and the composite signal is obtained by adding the multipath signal onto the direct signal in Matlab. In order to compare the performance of the proposal with other typical techniques, only a single multipath signal is constructed in the simulation.

Fig.6-16 shows the range errors in terms of the relative multipath delay, where the results of other techniques are obtained under ideal conditions of no noise and a single multipath with a M/D of 0.5 [60]. It indicates that the proposed technique has superior performance in terms of absolute range error. For M/D of 0.5, the proposed technique has a maximum range error of -2 meters. As for the strong multipath, it is shown that the maximum range error is around -6.5 meters for M/D of 10, which is acceptable. Unlike the 0.1-chip-spacing Narrow Correlator and Strobe Correlator which is able to work only when the full-bandwidth GPS

signal is acquired [60], the proposed method here is able to work with normal low-end GPS receivers, because the GP2010 RF front-end is with a Surface Acoustic Wave (SAW) filter with 1 dB bandwidth of 1.9 MHz [29]. The other advantage of the proposed technique is that it works based on the region close to the correlation peak, which represents high SNR, and thus the effect of noise is mitigated significantly.

Fig.6-16 Range errors for real signal

The computation load can be mainly represented by the number of correlators needed. To estimate the signal delay, the Narrow Correlator technique needs two correlators and the Strobe Correlator technique needs four correlators, so their computation load is quite low. To determine the peak position of the VLOS function, the proposed technique here needs to perform linear interpolation on both early edge and late edge of the obtained VLOS function, so multiple correlation values and hence multiple correlators are needed. The comparison on the computation load is shown in Fig.6-17. However, in the proposed technique only the correlation values near the peak of the VLOS correlation function are

needed, as illustrated in Fig.6-7, and the simulations in the section of 6.3 are based on 10 correlators, so the computation load of the proposed technique is acceptable.

(b) Correlators needed in the Narrow Correlator technique

(c) Correlators needed in the proposed technique

Fig.6-17 Comparison of computation load between the Narrow Correlator and the proposed technique

Summary

In this chapter, a new concept of virtual multipath is introduced and a novel technique for mitigating short-delay multipath is proposed, which has superior performance on mitigating both weak and strong short-delay multipath signals, and has no special requirements on the bandwidth of the filters in receiver's RF front-end. Although the discussions are based on GPS, the proposed technique is also applicable to other correlation based ranging systems.

Chapter 7 Multipath Mitigation in WLAN-based Indoor Positioning

As has been mentioned, the power of the typical outdoor GPS signal is -130 dBm and is very weak. Therefore, in most indoor environments, GPS receivers usually are unable to work due to the absence of signal.

In recent years, the research efforts on WLAN based indoor positioning have drawn great attention [85-87], and two WLAN based positioning products [88, 89] are also available in the market. There are the two main reasons pushing this application forward. First, the navigation satellite signals are too weak to be detected in indoor environments; second, WLAN has been widely deployed in many places all over the world.

To do positioning, ranging is the first step, and there are three typical solutions which can be applied into the WLAN based ranging. The first one is based on the prior-knowledge of the signal power profile in a certain area, which is stored in a database. Each time when performing positioning, the receiver first estimates the power of each received signal transmitted by four or more Access Points (APs) respectively, and then searches the signal powers in the database to find the user position. The method of [88] is based on this solution. The second solution is based on the propagation model or the degradation of the signal strength. The degradation is modeled by path loss equations, based on which the distances from the receiver to the access points can be estimated. This solution is adopted by [89]. The third solution is based on Time of Arrival (TOA), that is, the receiver measures the signal transit time and then multiplies it by the speed of light to obtain the distance between the receiver and the AP.

The disadvantage of the first solution above is that measurement must be conducted first to obtain the signal power profiles. If the layout of the

environment is changed, the measurement must be made again. The second solution mainly suffers from multipath propagation, which makes it difficult in obtaining an accurate power loss equation. The TOA based method needs to detect the direct or the first-arrival signal, because only the transit time of the direct signal can represent the desired distance correctly. But again, the receiver usually tracks the composite signal, including the direct and the multiple reflected signals, rather than the direct signal only, which results in significant positioning error.

This chapter will introduce two proposed algorithms for multipath mitigation with a target to improve the WLAN based ranging accuracy, and the performance on mitigating multipath signals of the two proposals will be shown by ideal simulations as well as real measurements in indoor office environment based on the 802.11b WLAN signal.

7.1 Multipath Mitigation Based on APOC

The proposed Asymmetric Phase Only Correlation (APOC) will be introduced in this section, and for the first time it is applied into the area of ranging by separating the direct signal from multipath signals.

7.1.1 Proposal of APOC

Phase only correlation (POC) [90], which was reported in 1975 and is still under investigation [91, 92], gained popular use in image processing and pattern recognition [93-97], due to its good property of much sharper correlation peak when compared with the conventional correlation method. POC is derived from the conventional Fourier Transform based convolution for two signals, $f(x)$ and

$h(x)$, which is shown below.

$$R(\tau) = \text{IFFT}[F(\omega) \cdot H^*(\omega)] \quad (7.1)$$

$$F(\omega) = \text{FFT}[f(x)] = A_1(\omega) \cdot e^{j\phi_1(\omega)} \quad (7.2)$$

$$H(\omega) = \text{FFT}[h(x)] = A_2(\omega) \cdot e^{j\phi_2(\omega)}, \quad (7.3)$$

where $R(\tau)$ is the time-domain convolution function; FFT, IFFT represents Fast Fourier Transform and Inverse Fast Fourier Transform respectively; the asterisk in (7.1) denotes complex conjugation. By setting both the $A_1(\omega)$ in (7.2) and the $A_2(\omega)$ in (7.3) to unity in frequency domain, phase only correlation is obtained.

Ranging systems, aiming at determining the distance between transmitter and receiver, is usually based on Time of Arrival (TOA). In order to achieve a high positioning accuracy, the direct signal needs to be detected. Conventionally, correlation is the most popular method to compute TOA. However, in indoor environments, numerous multipath signals usually can distort the correlation function obtained at the receiver side and induce a non-negligible range error, which degrades the positioning accuracy of the system.

The most important property of POC is that the correlation function obtained from it is much narrower than the conventional one, so it is expected that multipath signals can be separated from the direct signal. However, it is difficult to directly apply the original POC definition to process communication signals, mainly due to the heavy noise caused by environments and nearby hardware, which are unlike images that have less noise.

Fortunately, it is found that by assigning $A_1(\omega)$ in (7.2) as the amplitude spectrum of the received signal and setting it to unity, and assigning $A_2(\omega)$ in (7.3) as the amplitude spectrum of the local/reference signal and keeping it unchanged before performing the IFFT operation in (7.1), a higher correlation peak can be achieved in practice, and this method is named as Asymmetric Phase Only Correlation (APOC) in the following.

7.1.2 Mathematic Analysis

Similarly to the equations (7.1-7.3), the proposed APOC can be expressed in (7.4).

$$\begin{aligned}
 R'(\tau) &= \text{IFFT}[F'(\omega) \cdot H^*(\omega)] \\
 &= \text{IFFT}[e^{j\phi_1(\omega)} \cdot e^{-j\phi_2(\omega)} \cdot A_2(\omega)] \\
 &= \text{IFFT}\{e^{j[\phi_1(\omega) - \phi_2(\omega)]} \cdot A_2(\omega)\} \\
 &= f'(x) \otimes h'(x)
 \end{aligned} \tag{7.4}$$

where,

$$F'(\omega) = \text{FFT}[f(x)] / |\text{FFT}[f(x)]| = e^{j\phi_1(\omega)};$$

$$H(\omega) = \text{FFT}[h(x)] = A_2(\omega) \cdot e^{j\phi_2(\omega)};$$

$$f'(x) = \text{IFFT}\{e^{j[\phi_1(\omega) - \phi_2(\omega)]}\};$$

$$h'(x) = \text{IFFT}[A_2(\omega)];$$

and $\otimes$ represents time domain convolution.

Since the frequency domain amplitude spectrum of $f'(x)$ is unity, so ideally $f'(x)$ is a delta function in time domain, whose position is determined by the phase information in frequency domain, i.e., $\phi_1(\omega) - \phi_2(\omega)$, which represents the phase difference between the received and the local signal in communication systems. The function $h'(x)$ is determined by $A_2(\omega)$.

The fundamental property of delta function gives that,

$$\delta(x - \alpha) \otimes h'(x) = h'(x - \alpha). \tag{7.5}$$

where α is the delay of the delta function. So, the obtained time-domain function in (7.4) is the shifted $h'(x)$, and its position represents the relative delay between $f(x)$ and $h(x)$, i.e., the received and the local signal.

7.1.3 Ideal Simulations

Fig.7-1 Comparison between conventional correlation and APOC

To ascertain the above discussions, a two-path model, i.e., a direct signal plus a multipath signal, is constructed and the comparison between the conventional correlation and the APOC is shown in Fig.7-1, where the 11-bit Barker Code [98], widely used in 802.11b based WLAN, is generated as the LOS code with a typical chip rate of 11Mbps and a sampling rate of 500 MSamples/sec in Matlab; a multipath code with a relative delay of 10 sample points and a relative amplitude of 1.1 is constructed and added to the LOS code to represent the received composite signal, which is correlated with a local Barker Code based on the APOC and the conventional method respectively. The correlation functions are obtained from the magnitudes of the complex array $R'(\tau)$ in (7.4) and plotted in Fig.7-1, which indicates that the conventional correlation function is distorted by the multipath code, while two obvious peaks are produced by APOC and the first peak thus can be regarded as the location of the LOS code, because the multipath code is later than the LOS code.

7.1.4 Measurement Method and Set-up

Fig.7-2 Measurement set-up

The above expectation on locating the direct signal is verified based on real signal collected by the set-up shown in Fig.7-2, where an Agilent E4438C signal generator generates a 2.45 GHz carrier with a power of 17dBm, the typical transmitting power for 802.11b based WLAN, which is modulated by the 11-bit Barker Code with a chip rate of 11 Mbps (rectangular pulse shape) and transmitted out via an antenna centered at 2.45 GHz. Another antenna of the same type is connected to the Channel 1 of the Agilent DSO8452D Infiniium Scope to receive the signal. Additionally, a cable is utilized to connect the second output of the power splitter directly to the Channel 2 of the Infiniium Scope as a reference to the signal received in Channel 1. Each time, the two channels are captured simultaneously and are directly down sampled under a sampling rate of 500 MSamples/sec in the Scope. By this set-up, the obtained signals in the two channels are at the intermediate frequency of 50 MHz, which are then processed in Matlab.

The first set of data is used for calibration to evaluate the excess delay between Channel 1 and Channel 2 caused by both the instruments and the cable. For the rest of the data, the delay of the signal in Channel 1 and hence the distance between the two antennas, are processed by the conventional correlation and the APOC respectively. The range errors at different distances are obtained by comparing with the actual line-of-sight distance.

Fig.7-3 Layout of PWTC and measurement route

7.1.5 Averaging Method

With the moving of the receiving antenna, the received signal power will decrease, and noise could cause some pseudo peaks on the obtained correlation function. Fig.7-5 shows the blocks of APOC with averaging to mitigate the effect of noise, where APOC is performed on each buffer of the signal that is shown in Fig.7-4.

Fig.7-4 Received signal saved in buffers

Fig.7-5 APOC with averaging

In the above illustration, assume that each buffer length in Fig.7-4 is L , and then the result after the IFFT in Fig.7-5 is

$$\mathbf{R}_i(n) = \mathbf{M}_i(n) \cdot e^{j\theta_i(n)} \quad (n = 1, \dots, L),$$

where i represents the i _{th} buffer; the obtained $\mathbf{R}_i(n)$ is a complex array with the length of L , and $\mathbf{M}_i$ is its amplitude, θ_i is the corresponding phase.

Due to the clock drift in the digitizer and variation of signal propagation channel, the phase of the received signal is uniformly distributed in $[0, 2\pi]$, which determines that the result of IFFT, i.e., $\mathbf{R}_i$, cannot be summed together directly, because the correlation functions $\mathbf{R}_i$ from different buffer could be destructive to each other.

To address this problem, a block called phase adjuster is added into Fig.7-5, whose function is to adjust the phase of the $\mathbf{R}_i$ obtained from the i -th buffer. Recall that the $\mathbf{M}_i$ and θ_i are all with a length of L . We assume that $\mathbf{M}_i(k)$ is the maximum value in $\mathbf{M}_i$, so $\theta_i(k)$ will be the phase of the maximum value. By the phase adjuster in Fig.7-5, $\theta_i(k)$ will be adjusted to zero, i.e.,

$$\mathbf{Y}_i = \mathbf{R}_i * \mathbf{F}_i \quad (7.6)$$

where,

$$\mathbf{F}_i = e^{-j\theta_i(k)}$$

By equation (7.6), all the phase elements in array θ_i are changed, and particularly the element corresponding to the maximum value in $\mathbf{M}_i$ is adjusted to zero. After adjusting the phase, all the correlation functions obtained from different buffers in Fig.7-4 are summed together constructively and thus noise can be reduced.

7.1.6 Results and Discussions

Fig.7-6 Range errors in terms of LOS distance

Range errors obtained from the above experiment are plotted in Fig.7-6, where the highest peak is regarded as the location of the direct signal in the conventional correlation method; in APOC, a threshold of 0.45 on the normalized correlation function, similar to that discussed in [99], is set to distinguish the useful received signals and the first peak is regarded as the direct signal. Here, the techniques in [85-89] are not compared, because these techniques and hence their results are obtained based on signal power, which is totally different from APOC

and not comparable.

In Fig.7-3, the measurement within the first 8 meters is with LOS signal and after that there is no LOS signal. As anticipated, larger range errors are observed in the conventional correlation method, while the errors by APOC are much smaller. The average range error in the conventional correlation is 7.9 meters, while the average error in APOC is around 2.3 meters. Especially in the first 10 meters when the measurement is with LOS signal (see Fig.7-3), the average range error in APOC is 0.5 meters. Only at two distances, i.e., 19m and 23m, the range errors from APOC are quite large. Correlations at these two distances indicate that the received signal is very weak, the reason of which could be that the direct signal is totally blocked by wall or cancelled by multipath signals. The other point observed from the figure is that the errors from the conventional correlation vary more severely than that of the errors in APOC, which is reasonable due to the fact that multipath signals are different in amounts and amplitudes at different distances.

Fig.7-7 Layout of PWTC and measurement route

A measurement was conducted again by following another route shown in Fig.7-7, and the corresponding range errors are plotted in Fig.7-8. In Fig.7-7, the length of the route segment AB, BC, CD and DE is 8m, 10.7m, 9.5 and 8m respectively, among which AB is with LOS signal.

Fig.7-8 Range errors in terms of LOS distance

The range errors in Fig.7-8 have similar features as that in Fig.7-6, i.e., when the received signal is strong, the range errors from APOC are usually very small, but if the received signal is weak, the APOC will not have significant improvement compared with the conventional correlation method. The reason is that, as has been mentioned, APOC relies on the detection of the direct signal or the first arrival signal, but if the direct signal is absent or too weak to be detected, APOC will lose its advantage.

7.2 Multipath Mitigation Based on BA-POC

This section will introduce the Binary Asymmetric Phase Only Correlation (BA-POC), which is modified from APOC and has advantages on hardware cost when compared with APOC, but BA-POC is only applicable when the strong LOS signal is available.

7.2.1 Concept of BA-POC

The work in [100] indicates that binarizing the amplitude of the input signal is helpful to give a higher Signal-to-Noise Ratio. Based on this conclusion, the amplitudes of the received signal $f(x)$ in equation (7.2) are first binarized to +1 if they are positive and -1 if negative, after which the APOC is performed. This method is named as Binary Asymmetric Phase Only Correlation (BA-POC). Binarizing the signal amplitudes is equivalent to two-bit quantization, and can be implemented easily in practice.

7.2.2 Ideal Simulation

Fig.7-9 Conventional correlation and BA-POC

A comparison between the conventional correlation and the BA-POC is shown in Fig.7-9, where again the 11-bit Barker Code is generated as the LOS code with a chip rate of 11Mbps and a sampling rate of 500 MSamples/sec in Matlab; a multipath code with a relative delay of 10 sampling points and the same amplitude as that of the LOS code is constructed and added to the LOS code to represent the received composite signal, which is then correlated with a local Barker Code based on the BA-POC. The correlation functions are obtained from the magnitudes of the complex array $R(\tau)$ in (7.1) and plotted in Fig.7-9, which indicates that the conventional correlation function is distorted by the multipath code, while two distinct peaks are produced by BA-POC, and the first peak thus can be regarded as the location of the LOS code, because the multipath code is later than the LOS code.

Fig.7-10 Effect of binarizing

If the strength of the LOS code is weaker than that of the multipath code, BA-POC can only produce one narrow correlation peak in the above simulation, which represents the location of the stronger code. To make it clearer, a two-path model as shown in Fig.7-10 is chosen where the multipath code is weaker than the

LOS code. The plot of the composite code, equal to the LOS plus the multipath, is intentionally shifted left a little bit for convenience of viewing. It can be seen that the binary version (not plotted in Fig.7-10) of the composite code, i.e., reset the chips of the composite code to +1 or -1, will be exactly equivalent to the original LOS code which is stronger than the multipath code, and hence BA-POC is helpful to mitigate multipath signal when strong LOS signal exists. Similarly, the example in Fig.7-10 can also explain the statement in [100], i.e., binarizing is helpful to suppress noise.

7.2.3 Measurement and Results

The measurement set-up and method for testing BA-POC is exactly the same as that described in section 7.1.4, except that the measurement here is with the LOS signal.

Fig.7-11 gives the photo of the measurement environment.

Fig.7-11 Measurement environment (PWTC)

Fig.7-12 Range errors in terms of LOS distance

Range errors obtained from the above experiment are plotted in Fig.7-12, where the highest peak is regarded as the location of the LOS signal in the conventional correlation method while in BA-POC, a threshold of 0.6 on the normalized correlation function is set to distinguish the useful received signals from noise, and if multiple peaks are observed, the first peak is regarded as the LOS signal. As anticipated, larger range errors are observed in the conventional correlation, while the errors by BA-POC are much smaller at most distances. The maximum error in the conventional correlation is 26 meters, while the maximum error in BA-POC is only around 4 meters. Again, the errors from the conventional correlation vary more severely than that of the errors in BA-POC, as has been explained in section 7.1.6, due to multipath signals adding constructively or destructively over local variations hence the conventional correlation is affected more easily.

The range errors in BA-POC mainly result from noise as well as multipath. With the receiving antenna moving away from the transmitting antenna, the LOS signal power will degrade and noise could cause pseudo peaks on the obtained correlation function and hence will induce range error if any pseudo peak is selected as the LOS signal. In the measurement, averaging is used to reduce noise until the peak amplitude is 10 times higher than the average side lobe amplitude. Most multipath signals can be separated from the LOS signal by BA-POC, but those with very short relative delay can still be mixed with the LOS signal and cause range errors.

Fig.7-13 Correlation function comparison between the BA-POC and the conventional method

The normalized correlation functions corresponding to the 14th meter location in Fig.7-12 are given in Fig.7-13, from which one can clearly see the advantage of BA-POC. The conventional correlation function is obviously distorted by multipath signals and has a peak located at the 129th sample point, while BA-POC produces several narrow and sharp correlation peaks with the first one located at

the 115th sample point. As mentioned, the LOS signal travels the shortest path compared with multipath signals, so the first peak above the threshold can be selected as the LOS signal. The other aspect observed from Fig.7-13 is that the leading edge of the correlation function from BA-POC is much smoother than the late edge. This is because multipath signals arrive later than the LOS signal, and hence their correlation peaks affect the late edge of the composite correlation function more significantly.

7.2.4 Discussion on Threshold

Fig.7-14 Range errors by two different thresholds in APOC

The range errors under two different thresholds are shown in Fig.7-14, which corresponds to the measurement in Fig.7-3. It is found in Fig.7-14 that increase or decrease the threshold by 10% in the measurements can cause an additional range error up to 5 meters. Therefore, a suitable threshold must be selected to minimize the range errors in both APOC and BA-POC.

In different environments, the threshold can significantly vary due to different physical environment structures. Even in the same environment, the threshold can be different if the signal transmitter is moved to another location. This is because

the signal passes different path before reaching the receiving antenna, and therefore the signal power loss is different.

When the receiver changes its position, different combinations and numbers of multipath signals can cause the received composite signal stronger or weaker, which results in different signal to noise ratio and hence affects the threshold.

The simulation result under ideal conditions shown in Fig.7-1 indicates that the correlation function produced by APOC has side lobes. In the real measurements, if any side lobe is identified as the first arrival signal, big range error will occur. Therefore, the threshold must be higher than the side lobe. According to the result in Fig.7-1, the bottom limit for the threshold is 0.3.

Based on the above facts, it is difficult to analytically obtain the threshold. In this thesis, the threshold is statistically obtained based on a number of measurements [99]. Furthermore, BA-POC has different performance on noise reduction from APOC, because it binarizes the amplitude of the received signal to ± 1 , by which the noise value is also changed, and therefore the threshold in BA-POC is different from that in APOC.

Summary

Two techniques were introduced in this chapter, and their performances on mitigating multipath were illustrated by ideal simulations and real measurements in indoor office environment. APOC is applicable to both Non-Line-of-Sight (NLOS) and LOS environment, while BA-POC is useful only when the strong LOS signal exists. The expected advantage of BA-POC is on saving hardware cost, because it uses less storage. Although the measurements conducted are based on WLAN carrier, both techniques have potential use in other correlation based ranging systems.

Chapter 8 Conclusions and Future Work

This chapter is to conclude the research works presented in the previous chapters and gives some recommendations on the future work related to the works conducted already in this thesis.

8.1 Conclusions

In this thesis, three ideas are proposed on mitigating short-delay multipath signals which are numerous in indoor positioning applications, and one practical work, namely the implementation of the software GPS receiver in Matlab, is accomplished.

The first research work on mitigating multipath signals is to make use of the partial autocorrelation function of the C/A code to detect the direct signal. The idea behind this is that multipath signals have a very weak effect on the early edge of the autocorrelation function of the C/A code, because multipath signals are always later than the direct signal. The simulation results indicate that for the 25 C/A codes of the total 32, the multipath caused code delay error is smaller than that from the 0.1-chip-spacing narrow correlator under the conditions of a single multipath with a multipath-to-direct ratio of 0.5 in amplitude. The reason that multipath signals have a different effect on the code delay error for different C/A codes is that the side lobes of different C/A code autocorrelation functions are different, which can be grouped into seven categories, as that presented in Chapter 5. Although the mathematic expressions of the side lobes have been very well known for a long time, their effect on the code delay error caused by multipath has not been seen in other literatures. This work is further tested based on the measurement of the signal from the GS1010 GPS Signal Simulator from

WelNavigate Inc. The results from the captured signal also indicate that the proposed technique has superior performance on mitigating the multipath caused code delay errors and confirmed that the code delay errors are different for different C/A codes. The mathematic analysis made in this work gives a clearer understanding than that in the E1&E2 Tracker which looks more like an empirical technique.

The second proposal is a virtual multipath based technique aiming at mitigating short-delay multipath. For the first time, the concept of virtual multipath is defined according to the obtained autocorrelation function at the receiver side, and the overall effects of the multipath signals are then represented by the virtual multipath. The position of the direct or LOS signal can be determined from the resultant function which is obtained by subtracting the virtual multipath correlation function from the original composite correlation function. Simulations based on both the Matlab generated signals and the captured real GPS signals reveal that this technique is able to achieve good ranging accuracy by mitigating the multipath. Furthermore, simulations also show that this technique can mitigate the multipath signals which are stronger than the direct signal. In other words, it is insensitive to the multipath-to-direct ratio in amplitude, which is a very important feature for indoor positioning, because in indoor environments, multipath signals can be stronger. For the real band-limited GPS signals, a method called learning is proposed to solve the issue of rounded correlation peak. By learning, the filtering effect from the RF front end of a GPS receiver is weakened, and hence the virtual multipath based technique can be applied into the low-end GPS receivers.

Phase only correlator, which has been widely used in image processing and pattern recognition, is investigated and modified in this thesis, which is then used to improve the ranging accuracy by mitigating multipath signals in indoor environments. The results in Chapter 7 reveal that the correlation function from

the modified phase only correlator, named as APOC, has very sharp and narrow peak which is with a higher signal-to-noise ratio when compared with that from the original phase only correlator. It is because of this feature that the correlation peaks can be separated from each other, and thus if there are multipath signals, the first peak can be regarded as the direct signal. The conducted measurements are based on the 2.45 GHz carrier modulated by Barker code, and the results indicate that high ranging accuracy can be obtained in indoor office environment. BA-POC, which aims to save hardware cost, is the further improvement of APOC, but it is only applicable to the situation when the LOS signal is present. BA-POC can achieve a ranging accuracy of the same order as that of APOC. However, because the received signal amplitudes are quantized to ± 1 only, the BA-POC thus needs less storage when compared with that of APOC. The measurement results of BA-POC also indicate that it is a promising technique for mitigating multipath signals in indoor environments.

8.2 Recommendations on Future Work

This section will present some recommendations for the future work related to this thesis.

The partial autocorrelation function based method was introduced in Chapter 5, and both the simulations and the mathematic analysis indicate that this technique indeed is able to mitigate multipath. However, two challenges have to be faced before applying it to the real GPS receiver, that is, effective detection and noise reduction. The proposed technique is to use the linear interpolation algorithm to represent the multipath-free portion of the correlation function, and the linear interpolation is based on discrete correlation values. To be a feasible technique, these correlation values must be computed effectively, otherwise the

portion being detected will be wrong and not the desired multipath-free portion. In addition, as mentioned in Chapter 2, GPS signal is very weak and is buried in noise, so these discrete correlation values are with very low signal-to-noise ratio. If the noise cannot be reduced effectively, range errors will be induced. Averaging is the typical method for reducing noise, but is not good enough here and needs to seek more robust technique for noise reduction.

The current results from the virtual multipath based technique are based on the high sampling rate, and thus the computation load is relatively high, so to make it to be a feasible technique, the computation burden must be reduced.

APOC and BA-POC are quite promising techniques applicable to indoor positioning applications. However, to determine the first arrival signal, a flexible but effective scheme for determining the threshold must be explored. This is because on the obtained correlation function, there could be some pseudo correlation peaks existing, and therefore if the threshold is inappropriate, the first arrival signal will be determined incorrectly, which can cause large range error. It can be imagined that this threshold is different for different environments, so the investigation on the threshold may be based on multiple measurements and signal propagation channel modeling. Ray tracing technique may also help on the threshold determination. The other important issue is how to synchronize the transmitter and the receiver. In the current measurement set-up, a long coaxial cable is used to connect the transmitter and the receiver, which provides a reference signal for the range error estimation. For a real system, this method is unfeasible, and therefore a scheme for synchronizing the transmitter and the receiver must be explored so that the signal transit time can be estimated. Additionally, the work on APOC and BA-POC are based on WLAN signal in this thesis, and according to our measurements, they are unable to process GPS signal so far. This is because the GPS signal is too weak, and the obtained correlation function is similar to the conventional one. Further modifications on APOC or

BA-POC and make them applicable to GPS or Pseudolite signal is also an interesting work.

The key function blocks of the software GPS receiver has been implemented, and the positioning error is from 10 meters to 15 meters according to the testing based on the rooftop antenna. Although it is able to run, some additional work is needed to complement it. For instance, if tracking failed, the receiver must be able to re-track the signal, while the current program assumes that the tracking loop will not lose tracking. To accomplish this, a state machine needs to be developed for the receiver to transit between different states. The accuracy also needs to be improved, because for the current receiver, the root mean square error in height is around 40 meters, which is not acceptable. To improve the accuracy, the study on the ionospheric and tropospheric delay model will be helpful. The receiver also needs more realistic testing in different environments to evaluate its performance and reliability. Furthermore, a software GPS receiver, which is able to run in real time, will be a very promising product for market due to that it uses minimum hardware, and nowadays some universities or institutes are working toward this target, so re-write the current software receiver in other programming language and make it run in real time in PC is another valuable future work.

In addition, the US is to launch new navigation satellites and to use a more advanced navigation system, i.e., GPS modernization; Europe is also developing their global navigation satellite system Galileo. Therefore, attention should be given on the progress and the research on these new navigation systems in future.

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