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**FINE TIME UPLINK SYNCHRONIZATION
FOR FREQUENCY-HOPPED SPREAD SPECTRUM
SATELLITE COMMUNICATION SYSTEMS**

by

Rolands Eriks Ezers, B. A. Sc.

**A thesis submitted to
the Faculty of Graduate Studies and Research
in partial fulfillment of
the requirements for the degree of**

Master of Engineering

**Ottawa-Carleton Institute for Electrical Engineering
Department of Electronics
Faculty of Engineering
Carleton University
Ottawa, Ontario, Canada
August, 1989
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FOR FREQUENCY-HOPPED SPREAD SPECTRUM
SATELLITE COMMUNICATION SYSTEMS"**

submitted by Rolands Eriks Ezers in partial fulfillment of the requirements
for the degree of Master of Engineering.

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August, 1989

Abstract

A fine time uplink synchronization scheme for a frequency-hopped spread spectrum satellite communication system has been implemented in a hardware-based laboratory experiment. The experiment is implemented using an analog surface acoustic wave spectrum analyzer. The synchronization scheme employed is the two-tone method. Measurements are taken using the two-tone method with and without bias reduction. Experimental results taken using the method without bias reduction lead to an inconsistent estimator for the actual timing error. The method with bias reduction produces a consistent estimator for the actual timing error. The experimental results from the method with bias reduction have been compared with simulated results generated from theory. The two results agree very well. Thus, the two-tone method is a viable method for fine time uplink synchronization.

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List of Symbols

A	– amplitude
a_n	– data sequence
α	– channel attenuation
c	– threshold
c_n	– pseudorandom sequence
$c(t)$	– spreading waveform
ΔT	– timing error
E_b	– energy per bit
E_h	– energy per hop period
E_s	– symbol energy
ϵ	– actual timing error
$\hat{\epsilon}$	– timing error estimate
f_j	– hop period frequency
$g(\cdot)$	– time window
$G(f)$	– Fourier transform of $g(t)$
J	– jammer power
L	– number of hop periods
M	– number of frequency tones in a FSK modulation scheme
μ	– chirp slope
$n_j(t)$	– noise signal
n_j	– noise component of sample
$n_{c,j}$	– inphase component of n_j
$n_{s,j}$	– quadrature component of n_j
$N_j(t)$	– Fourier transform of noise
N_0	– noise power spectral density
N_j	– spectral density of jammer power
PG	– processing gain
P_s	– power of signal
ϕ_j	– hop period phase
$r_j(t)$	– received signal
R_b	– uncoded data rate
$R_j(t)$	– output of spectrum analyzer
$s_m(t)$	– transmitted signal

- S – signal power
- $S_j(t)$ – Fourier transform of signal
- σ^2 – variance of sample
- T_c – coded signal's period
- T_h – hop period
- T_s – symbol period
- θ – actual spectral parameter
- θ_a – actual advanced spectral parameter
- θ_d – actual delayed spectral parameter
- θ_e – actual early filter spectral parameter
- θ_l – actual late filter spectral parameter
- θ_n – actual normal spectral parameter
- θ_1 – actual spectral parameter for first tone
- θ_2 – actual spectral parameter for second tone
- $\widehat{\theta}_a$ – advanced spectral estimate
- $\widehat{\theta}_d$ – delayed spectral estimate
- $\widehat{\theta}_e$ – early filter spectral estimate
- $\widehat{\theta}_l$ – late filter spectral estimate
- $\widehat{\theta}_n$ – normal spectral estimate
- $\widehat{\theta}_1$ – spectral estimate for first tone
- $\widehat{\theta}_2$ – spectral estimate for second tone
- $u_m(t)$ – complex baseband representation of transmitted signal
- U_m – decision variables
- $v(t)$ – received signal
- $w(t)$ – time windowing function
- $W(f)$ – Fourier transform of $w(t)$
- W_{ss} – spread spectrum bandwidth
- z_j – spectral samples
- z_{aj} – advanced spectral samples
- z_{dj} – delayed spectral samples
- z_{ej} – early filter spectral samples
- z_{lj} – late filter spectral samples
- z_{1j} – spectral samples for first tone
- z_{2j} – spectral samples for second tone
- z_{0j} – noise spectral sample
- z_{01j} – noise spectral samples for first tone
- z_{02j} – noise spectral samples for second tone

Glossary of Terms

ASK – amplitude – shift – keying

coarse time synchronization – alignment of timing to within half of a hop period

consistent estimator – an estimator whose value can be made arbitrarily close to a parameter by choosing the sample size sufficiently large

EHF – extremely high frequency

fine time synchronization – alignment of timing to within required tolerance

FLTSAT – an American military satellite

FSK – frequency – shift – keying

NCFSK – noncoherent frequency – shift – keying

NRZ – non – return to zero

PSK – phase – shift – keying

SAW – surface acoustic wave

SHF – super high frequency

SNR – signal – to – noise ratio

SWDS – TMS320C25 Software Development System

TTL – transistor – transistor logic

UHF – ultra high frequency

unbiased estimator – an estimator whose expected value is equal to the parameter for all sample sizes

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Chapter 1

Introduction

1.1 Motivation

The demand for communications and information transfer is ever increasing. It is desirable to be able to send more information to more and more locations more often. This demand is especially prevalent in the military world. Here, a user wishes to be able to communicate with any other in his armed forces who may be on a submarine, a battleship, in an aircraft, at headquarters, or in the field at any time. He wants his communication to be reliable, secure and robust. A communication system which achieves the above objectives is the goal of today's military communication researchers.

The desire for this type of communication system has led to sophisticated military satellite communication systems. In order to achieve reliable, secure and robust communication, the system must incorporate many elaborate system concepts such as: frequency hopping, onboard signal processing, frequency-division multiple access, and antenna-beam hopping. These concepts have greatly increased the complexity of the system. This added complexity has created many new problems.

One such problem is that of time synchronization. Time synchronization is the process by which the transmitter, the satellite, and the receiver agree on the time. Consider a system with frequency hopping, onboard processing and frequency-division multiple access, where the satellite maintains the master clock. The transmitter must set its clock so that its information is received by the satellite at the correct time. Also, the receiver must set its clock so that it receives the information from the satellite at the correct time. The synchronization process occurs in many steps. The scope of this thesis is to look at part of the total synchronization process - that of fine time uplink synchronization. This is the mechanism which synchronizes the transmitter with the satellite when they are misaligned by a small time difference.

Early attempts[1],[2] at fine time uplink synchronization reported in the literature led to inadequate methods. These methods achieved poor performance in low signal-to-noise

ratios. More recent approaches[3]-[5] show the possibility for improved performance. However, the majority of the approaches[1],[2],[4],[5] in the literature present only theoretical analyses of the problem. A paper by Hodsdon[3] presents some experimental results, but does not give results for the performance in the presence of noise. Such experimental results are necessary to validate theoretical analyses. Thus, the motivation for this thesis is the lack of experimental results presented in the literature for the performance in the presence of noise of fine time uplink synchronization schemes for frequency-hopped spread spectrum satellite communication systems.

1.2 Thesis Objectives

The thesis objective is to propose and test a fine time synchronization scheme for the uplink of a frequency-hopped spread spectrum satellite communication system. The analyses reported in the literature of several methods are presented. Simulation results for these methods are also presented, and their performances are compared. For one particular method - the two-tone method - a hardware-based laboratory experimental system has been constructed. This experimental system contains all the pertinent elements of an actual frequency-hopped spread spectrum satellite communication system. This allows the experimental results to reflect actual performance. The system has been tested in an additive noise environment, and the experimental results are compared with simulation results for this method. Conclusions are then drawn about the various methods of fine time uplink synchronization for frequency-hopped spread spectrum communication systems.

1.3 Thesis Outline

Chapter 2 discusses background material in satellite communication and spread spectrum systems necessary for the thesis. It also discusses particular aspects of frequency-hopped spread spectrum satellite communication systems. In Chapter 3, synchronization in satellite communication systems is discussed, and the analyses and simulated results reported in the literature for various schemes employed to achieve fine time uplink synchronization are presented and compared. Chapter 4 describes the hardware for the experimental work, while Chapter 5 describes the experimental results and compares them with

the simulated results. Chapter 6 gives a summary of the thesis and recommendations for future work in this area.

Chapter 2

Background

2.1 Introduction

This chapter discusses the background information necessary to understand the role of fine time uplink synchronization in frequency-hopped spread spectrum communication systems. The chapter discusses satellite communication systems, spread spectrum communication systems, and frequency-hopped satellite communication systems.

2.2 Satellite Communication

2.2.1 Introduction

Satellite systems provide many important services in our lives today. These services include: fixed point-to-point communication, broadcast communication, mobile communication, meteorological information collection, earth-exploration information collection, satellite-to-satellite communication, radionavigation, radio-astronomical information collection, radiodetermination, spacecraft tracking, and frequency and time standards. In this section, background information on the communication services of satellite systems is presented.

Satellite communication systems offer the ability to send information over long or moderate distances. The types of information that can be sent include: voice, video, facsimile, teletype, and data. This information must be encoded into a signal which is suitable for transmission. There are principally two types of these signals: analog and digital. Analog signals are signals in which information is represented by the level or phase of a continuous signal, whereas digital signals are signals in which information is represented by discrete levels or phases of a signal. Digital signal transmission has several advantages. For example, digital transmission is much less sensitive to noise. Hence, a received information signal is much less corrupted by channel noise and interference. Also, the handling of digital signals is much more convenient for multiplexing and switching. Further, digital hardware provides a great deal of flexibility for implementing large complicated systems. However,

most information exists naturally in analog form. Hence, to use digital transmission, analog information must be converted to digital form at the transmitter and converted back to analog form at the receiver. This is done using analog-to-digital converters and digital-to-analog converters. Even with these added conversions, digital transmission is still the preferred method for modern satellite communication systems. In all further discussions, it is assumed that information is transmitted in digital form.

Satellite communication falls into three categories: fixed point-to-point communication, broadcast communication, and mobile communication. Fixed point-to-point communication involves an immobile earth station transmitting information to another immobile earth station via the satellite. This type of communication usually involves a terrestrial link from the user to the earth station at each end of the network. An example of this sort of communication is a private long distance telephone call. Broadcast communication involves an immobile earth station sending information to many receivers in a region via the satellite. The information is usually transmitted in only one direction. An example of this kind of communication is a satellite television station. Mobile communication involves two earth stations communicating to each other where one or both of the earth stations are moving. An example of this type of communication is ship-to-ship communication via satellite. This type of communication presents many technical problems. All three of these types of communication can be operated over different distances. Some satellite communications systems are operated internationally or globally. This type of system must involve several satellites and perhaps several earth stations. Other systems operate regionally and require only a single satellite.

2.2.2 Basic System

A basic satellite system has two segments: a space segment and a ground segment. The space segment consists of the satellites and the control stations. Usually, there is a main satellite with one or several spare satellites. The control stations are located on the earth and are used to check the satellite subsystems' status, to monitor outputs, to provide ranging data, to act as a test facility, and to update the satellite configuration via the telemetry links. The satellite itself has two major components: a communication payload and a spacecraft bus. The communication payload consists of the antennas and the transponders. The payload may have one or two antennas and many transponders.

The transponders are of two types: transparent or processing. Transparent transponders simply transmit an amplified and frequency-converted version of the received signal. This is analogous to a "bent pipe". Processing transponders receive the signal and demodulate it. Then, they modulate and transmit the signal. Thus, the processing transponder creates two communication links. The processing transponders are much more immune to noise, fading, and jamming and allow for much easier and more efficient switching and multiplexing of signals. Most transponders currently in use are of the transparent type. The other component of the satellite - the spacecraft bus - provides the subsystems to support the satellite and its payload. These subsystems are: the spacecraft structure, the electrical power subsystem, the altitude control subsystem, the thermal control subsystem, and the telemetry, tracking, and control subsystem.

The ground segment of the satellite system consists of the earth stations which are needed to perform communications through the satellite. These earth stations contain: antennas, feed systems, high-power amplifiers, low-noise amplifiers, up and down converters, ground communication equipment, control and monitoring equipment, and power supplies. The earth stations must be connected to the user either directly or through a network. Thus, the complete communication system includes a connection from a user to the transmitting earth station, a earth station to satellite to earth station link, and a connection from the receiving earth station to another user.

2.2.3 Satellite Orbits and Frequency Bands

In order to have a satellite communication system, it is necessary to have a satellite orbiting the earth. There are many possible orbits; however, only some of these are good for satellite communication. The one that is most commonly used for satellite communication is the 24 hour circular geostationary orbit. The satellite is located approximately 35 786 kilometers above the earth's surface. The motion of a geostationary satellite is such that its angular speed is the same as that of the earth. Thus, the satellite is always located above the same position on the earth. There are small perturbations in the orbit caused by the gravitational forces of the sun, the moon, and the planets and by the fact that the earth is not a perfect sphere. Still, the geostationary orbit imposes minimal tracking requirements on the earth stations. There are two main problems with the geostationary orbit. Firstly, because the orbit has many advantages, it is very popular, and this has led

to congestion in the orbit. Secondly, the orbit is not visible at the extreme north or south latitudes of the earth. Thus, other orbits are used where coverage of the extreme latitudes is needed, or where congestion is too great. Examples of other orbitals are: low circular, inclined circular, Molnya, and tundra. These orbits differ from the geostationary orbits in that they are elliptical, have a different altitude, or are inclined to the equator. They do not appear stationary to earth stations and suffer from Doppler shifts. Hence, in order to achieve communications, several satellites must be employed and earth stations must track these satellites.

A satellite communication system requires some bandwidth in the frequency spectrum to operate. International frequency spectrum regulatory agencies have assigned certain frequency bands for certain regions. The uplink frequency and the downlink frequency must be sufficiently far apart for a given satellite so that they do not interfere with each other. Also, an effort must be made to ensure that satellite communication does not interfere with other communication and that other communication does not interfere with satellite communication.

2.2.4 Propagation, Noise and Interference

A chief concern in satellite communication is noise. A certain signal-to-noise ratio is required to maintain reliable communication. Either increases in noise or decreases in signal can cause the signal-to-noise ratio to decrease below an acceptable level. Sources of noise and sources of signal attenuation are discussed below.

In nature, there are many sources of noise. Noise caused by the random motion of molecules is called thermal noise. It tends to have a flat frequency spectrum. There is also noise caused by radiation from atoms and molecules. This noise tends to have a spectral line structure. Some sources of the above noise are: the cosmos, the atmosphere, the sun, and man.

In a satellite system, there are effects upon the signal caused by the transmission path. The effects include: attenuation, propagation time delay, refraction, polarization rotation, group delay, dispersion, depolarization and fading. These effects are caused by: the presence of free electrons, the earth's magnetic field, gases, clouds, precipitation, sand and dust storms, and multipath propagation.

For reliable communication, it is necessary to consider interference effects also. There

is interference between the satellite system and other satellite systems, between the satellite system and terrestrial systems, and between the satellite system and aircraft and spacecraft.

In order to compensate for the above effects, it is necessary to incorporate enough transmit power so that the received signal-to-noise ratio is adequate for reliable communications. Some detrimental effects can be reduced by the choice of modulation scheme.

2.2.5 Modulation

There are principally two reasons for modulation. The first is to permit transmission of electromagnetic waves in the allocated frequency band. The second is to represent digital data. The signal is placed in the allocated frequency band by upconverting the baseband signal at the transmitter. At the receiver, the signal is downconverted back to baseband.

In satellite communication, there are several ways to represent data. Some of these are: amplitude-shift keying, phase-shift keying, and frequency-shift keying. There are also hybrid schemes which combine the above methods. To compare the various methods, binary signalling schemes are considered. Binary signalling is the transmission of data represented in two states. Usually, zeroes and ones are used to represent the data. The signal that represents each piece of information or bit takes a certain time duration called the bit period.

Binary amplitude-shift-keyed signals represent data using a signal with one frequency and two different amplitudes. For example, to represent a one, the signal might be present with a given amplitude. A zero would be represented by the absence of the signal. Binary phase-shift-keyed signals represent data using a signal with one frequency and two different relative phases. A one might be represented by a signal with a phase of 0 degrees and a zero might be represented by a phase of 180 degrees. Binary frequency-shift-keyed signals represent data using a signal with two frequencies. A one is represented with a given frequency. A zero is represented with another frequency.

There are two principal ways of detecting signals: coherent and incoherent. Coherent detection requires prior knowledge of the received carrier phase. If this knowledge is not known or cannot be obtained, then, incoherent detection must be used. In the presence of additive Gaussian noise, coherent detection has better tolerance to the noise. Table 2.1, from Evans[6], shows the relative performance of various binary signalling schemes in the presence of Gaussian noise. The numerical results quoted are the relative performance of

Modulation Method	Coherent Detection	Incoherent Detection
PSK	0	
ASK	-3	-4
FSK(h=1)	-3	-4
FSK(n=0.7)	-2	-5

Performance in dB relative to coherently detected PSK

Table 2-1: Tolerance to Noise of Binary Signals

each scheme compared with coherent binary phase-shift-keyed signalling. The numbers are in decibels and are rounded to the nearest integer. For the frequency-shift-keyed scheme, the parameter h is the product of the bit period and the frequency separation of the two tones. From the table, it is clear that binary phase-shift keying with coherent detection offers the best performance.

In order to increase the transmission rate, it is necessary with binary signalling to decrease the bit period. However, this also increases the bandwidth. A multilevel scheme may help in this regard. For multilevel amplitude-shift keying, several different amplitudes of the signal are used. For multilevel phase-shift keying, several different phases of the signal are used. For multilevel frequency-shift keying, several different frequencies are used. Table 2.2, from Evans[6], shows the relative performance of various multilevel signalling schemes. The numerical results are once again relative to coherent binary phase-shift-keyed signalling and are in decibels rounded to the nearest integer. All of the methods in Table 2.2 use coherent reception. It would seem that multilevel frequency-shift-keyed signalling offers the best performance. However, the amplitude-shift-keyed and phase-shift-keyed signals do not increase in bandwidth. The frequency-shift-keyed scheme does increase the bandwidth.

An important consideration in satellite communication systems is constant-envelope modulation schemes. This occurs because in order to transmit a signal from a given earth station to a satellite and from a satellite to another earth station, it is necessary to use high-power amplifiers. These amplifiers are nonlinear. Their output amplitude is not a linear function of their input amplitude. Also, their output phase is not independent from their input amplitude. Thus, if the input signal amplitude varies, the output signal has phase and amplitude distortion. This fact makes amplitude-shift-keyed signals unpopular for satellite communication. Constant-envelope modulation schemes are necessary to control this sort of distortion.

2.2.6 Multiple Access

Satellites typically have more users than transponders. This means that users must share transponders. The sharing of transponders is accomplished using multiple-access techniques. Some multiple-access schemes are: frequency-division multiple access, time-division multiple access, packet access, code-division multiple access, and hybrids. The schemes are compared on the basis of a single transponder.

Number of Levels	2	4	8	16
Suppressed Carrier ASK	0	-7	-13	-19
PSK	0	-3	-8	-14
FSK(h=1)	-3	-3	-3	-3

Performance in dB relative to coherently detected PSK

Table 2-2: Multilevel Scheme Performance

There are two types of frequency-division multiple-access schemes: multiple channels per carrier and single channel per carrier. In a multiple channels per carrier scheme, the transponder bandwidth is divided up into a few subbands. Each subband is assigned to an earth station. The earth station combines the users' signals and transmits them in this subband. This scheme is good when there are a few high capacity users. In a single channel per carrier scheme, the bandwidth is divided into a large number of small subbands. There is also usually a channel allocated for assigning of channels. The subbands can be dynamically assigned. Thus, each user is assigned his own channel and uses it to transmit data. The allocation of channels is done by demand assignment using the demand-assignment signalling channel. Thus, a channel is only assigned when communication is requested. A serious drawback to frequency-division multiple access is intermodulation distortion. This is caused by nonlinear high-power amplifiers. The problem can be reduced by operating the amplifiers with backoff.

There are two types of time-division multiple access: classical and satellite-switched. In classical time-division multiple access, each transmitting earth station sends its signal in short bursts of time. For all earth stations using a given transponder, these bursts do not overlap. The receiving earth stations receive all bursts, but they only demodulate the information they wish to have. Information is usually transmitted in a frame. Each frame begins with a reference burst and is followed by the bursts of information from all of the transmitting earth stations. Satellite-switched time-division multiple access is used with satellites that have multiple spot beams for different locations. The satellite receives a time-division multiple-access signal and sends this signal to the appropriate spot beam. This method makes use of space-division multiplexing as well as time-division multiplexing. This method does not require backoff of amplifiers since only one signal is present in the transponder at a given time. The method is excellent for very variable traffic.

A random-access scheme is good when the traffic is sporadic. Here, a transmitter sends his signal and waits to get a reply. The reply tells him whether the signal has been received. If it has not, the transmitter sends the signal again. This is repeated until the message is received. The method has large user delays. However, it allows for high link usage, many users, and relatively simple and cheap earth stations.

Code-division multiple access uses direct-sequence spread spectrum communication techniques to achieve multiple access. Since direct-sequence codes are designed to reject

interference, several users can transmit on the same channel without interfering with each other. This is the basis of this scheme. Each user is assigned a different direct-sequence spread spectrum code; the user can then communicate on that channel without interfering with other users.

Many networks use a combination of the above schemes to do multiple access. Often, a hybrid scheme gives the best performance.

2.2.7 Military Satellite Communication

Military satellite communication systems provide the military with enhanced communication services. Military satellite communications systems have a difficult task. They must provide communications for a wide variety of users and traffic in potentially hostile environments.

Satellite communication systems have certain inherent advantages. They have a large coverage area. They can provide communications at short notice anywhere in the world. They have high availability and bandwidth compared with many terrestrial systems.

Military satellite communication systems generally use geostationary orbits. From Evans[6], the allocated frequency bands are: UHF(220-400 MHz), SHF(uplink 7.9-8.4 GHz, downlink 7.25-7.75 GHz), and EHF(uplink 43.5-45.5 GHz, downlink 20.2-21.2 GHz). Current systems use transparent transponders. Special antennas, spread spectrum techniques and nuclear-hardened electronics are necessary for a successful system. Systems employ both wide coverage beams and narrow spot beams. Good linear power amplifiers are a necessity.

The main disadvantage of military satellite communication systems is that they can easily be subjected to hostile jamming. For this reason, recent research is being conducted in building jamming-tolerant systems. One key element in this is the use of spread spectrum techniques. Background in spread spectrum is presented in the next section.

2.3 Spread Spectrum Communication

2.3.1 Introduction

Spread spectrum techniques are very important to both military and civilian communication systems. In spread spectrum systems, the bandwidth of the transmitted signal is

greater than the bandwidth of the source signal. The bandwidth of the source signal is increased by modulating the source signal with a spreading signal at the transmitter. The bandwidth of the spread signal is collapsed at the receiver. The transmitter and the receiver both have knowledge of the spreading signal. Others such as interceptors and jammers do not. This fact gives this technique many advantages.

2.3.2 Advantages of Spread Spectrum Techniques

There are many characteristics of spread spectrum communication systems that make them advantageous to use for certain applications. Various characteristics are now discussed.

One of the major benefits is interference rejection. In a system which is not using spread spectrum techniques, a jammer must jam only the bandwidth of the signal. However, the use of spread spectrum modulation causes the signal to have a larger bandwidth. Now, the jammer is forced to spread his jamming power over this larger bandwidth. The receiver can use his knowledge of the spreading signal to combine the signal in such a way that the signal-to-jammer power ratio is increased. The jammer tries to combat this by employing some of the following jamming strategies: broadband jamming, partial-band jamming, tone jamming, pulse jamming, and repeat-back jamming. The relative effectiveness of each of these jamming strategies depends on the signal-to-jammer power ratio and the spreading signal. Thus, the use of spread spectrum techniques provides interference rejection.

Another benefit of spread spectrum techniques is low probability of detection. If the enemy is able to detect the signal, he may be able to do two things: locate the communicator and intercept the signal. Neither of these is desirable. Spread spectrum techniques make it difficult to detect signals in the following way: the enemy must monitor a wide frequency range over which the signal has a low power spectral density. The fact that the signal has a low power spectral density has an added benefit in civilian applications. In areas where high power spectral densities cannot be tolerated, spread spectrum techniques can be used to reduce the power spectral density.

The interference rejection capabilities of spread spectrum techniques make them suitable for multiple-access systems. By assigning each user with its own spreading signal, the same channel bandwidth can be reutilized by the users. The spreading signals can be chosen to minimize interference between users.

Certain types of spreading codes can be used to provide very accurate information about timing or range of a system. This sort of application is necessary in satellite communication systems and radar systems.

Spread spectrum techniques have many desirable properties for certain types of applications.

2.3.3 Methods of Spread Spectrum

Spread spectrum communication is achieved by modulating the source signal by a spreading signal. Various systems are discussed in Simon et al.[7]. A typical spread spectrum system from Simon et al.[7] is shown in Figure 2.1. Different types of spread spectrum systems are distinguished by their spreading signals. There are three main types of spread spectrum techniques: frequency hopping, time hopping, and direct sequence.

For frequency hopping, the spreading signal consists of a sum of waveforms $w(t)$ of duration T_h multiplied by a sine wave of a given frequency. The frequency is changed every T_h where T_h is called the hop time. Thus, the spreading waveform $c(t)$ is:

$$c(t) = \sum_n \exp[j(2\pi f_n t + \phi_n)] w(t - nT_h) \quad (2.1)$$

The frequency f_n for any given hop period is determined by a pseudorandom sequence. If the hop period is shorter than the symbol period of the source's signal, then the system is a fast-hopping system. If the hop period is longer than the symbol period, then it is a slow-hopping system. One important point is that the phase of the frequency hop, ϕ_n , is random. Thus, coherent reception of a frequency-hopped signal is not possible. Frequency hopping is often used in spread spectrum systems.

For time hopping, the spreading signal consists of a sum of waveforms $w(t)$ of duration T_h/M . The waveform is placed pseudorandomly within each frame T_h . There are M segments within the frame where the waveform may be placed. Thus, the spreading waveform $c(t)$ is:

$$c(t) = \sum_n w(t - (n + \frac{a_n}{M})T_h) \quad (2.2)$$

The frame's length relative to the symbol period of the source signal determines whether it is fast or slow hopping. Time hopping is not a commonly used method on its own.

Figure 2-1: A Typical Spread Spectrum System

For direct sequence, the spreading signal consists of a sum of waveforms $w(t)$ of duration T_c multiplied by the output sequence c_n of a pseudorandom number generator. The spreading waveform $c(t)$ is:

$$c(t) = \sum_n c_n w(t - nT_c) \quad (2.3)$$

T_c is called the chip time. The chip time is less than the source signal's symbol time. Direct-sequence spreading is often used.

Often, a combination of the above techniques is used. This is called hybrid modulation. This allows the assets of each waveform to be exploited. The form of the spreading waveform is:

$$c(t) = \prod_i c^i(t) \quad (2.4)$$

$c^i(t)$ are spreading signals which have been mentioned above.

The spread spectrum communication signal is generated by modulating the source signal with the spreading signal. The modulated signal is then transmitted. Of the spreading methods mentioned above, the two most commonly used methods are frequency hopping and direct sequence.

2.3.4 Processing Gain

Processing gain is a quantitative measure of the effectiveness of spread spectrum techniques. Processing gain provides a theoretical upper limit for this performance. Figure 2.2 show a basic spread spectrum communication system that is presented in Simon et al.[7]. The following quantities need to be defined:

W_{ss} = total spread spectrum signal bandwidth available

R_b = uncoded data rate in bits per second

S = signal power at the input of the receiver

J = jammer power at the input of the receiver

E_b = energy per bit at the receiver

N_j = spectral density of jammer power

PG = processing gain

Figure 2-2: A Basic Spread Spectrum Communication System

Note that both the signal and jammer powers are time-averaged values. From Simon et al.[7], the equivalent bit energy-to-jammer noise ratio is:

$$\frac{E_b}{N_j} = \frac{W_{ss}S}{R_bJ} \quad (2.5)$$

Thus, the processing gain can be defined as:

$$PG = \frac{W_{ss}}{R_b} \quad (2.6)$$

Thus,

$$\frac{E_b}{N_j} = PG \frac{S}{J} \quad (2.7)$$

There are several key assumptions which have been made in the above analysis. It is assumed that the communicators have knowledge of the pseudorandom sequence used in the system, but the jammer does not. However, the jammer does have knowledge of all other parameters of the system. Another key assumption is that the jammer's power limits the performance of the system. Thus, it is assumed that the receiver noise floor is much lower than the jammer's power. In most interesting cases, this is valid. Also, the above analysis assumes a broadband jammer. With pulse and tone jammers, processing gains may be lower.

Now, there is a measure with which to compare systems. Achievable processing gains are technology limited. With direct-sequence spreading, chip rates tend to be the limiting factor. With frequency-hopping systems, hopping bandwidths tend to limit the processing gain. With current technology, frequency-hopping systems tend to achieve higher processing gains.

2.3.5 Coding

There is a serious degradation in performance of the system when pulse or tone jamming is used. The performance is much below the expected performance predicted by the processing gain. This degradation can be neutralized with the use of coding. Coding can force the worst case jammer to be a constant broadband jammer.

Simple repeat coding is often used. If the symbol is repeated m times, then the system is said to have diversity of order m . The combination of these symbols can be done in two fashions: soft-decision decoding and hard-decision decoding. Hard-decision decoding

makes a decision on each of the m symbols transmitted and then decides on which bit has been transmitted. Soft-decision decoding examines the information received by each of the m symbols, but does not make a decision. It decides only after all m symbols have been received. A comparison of the two methods in Simon et al.[7] shows that neither is optimal for all signal-to-jammer power ratios. Also, the use of coding must include interleaving to be effective. This means that symbols resulting from one bit are not adjacent to one and other. This is necessary to maintain independence between symbols.

Coding is important for protection against certain jamming strategies.

2.3.6 Pseudonoise Generators

Pseudonoise sequences are necessary in spread spectrum communication systems. They provide the modulating sequence for direct-sequence spreading and the hopping-pattern sources for frequency-hopped spreading. Pseudorandom sequences are binary sequences that appear to be random, but are generated by deterministic means. These sequences usually are generated from a key. This key is given to all communicators so that they can transmit and receive information in the communication system. The mathematics of these sequences is beyond the scope of this thesis. However, it is important to note some characteristics of the sequences. In order to prevent an enemy from attaining the sequence, it is necessary that the sequence be long. Sequences should be longer than the required lifetime of the system. From Simon et al.[7], a system operating at a chip rate of 10^9 bits per second would need a key of 62 bits to have a period of a century. Thus, for practical purposes, it is possible to achieve a non-repetitive sequence. However, this creates another problem. The communicators must have knowledge of where in the code one is at a given time in order to use the system. Usually this information is available to enable the user to synchronize to the system in a short period of time.

2.3.7 Military Spread Spectrum Satellite Communication

Military satellite communication systems employ spread spectrum for two reasons: anti-jamming protection and low probability of signal interception. To achieve sizable processing gains, frequency hopping must be used. Also, to maintain performance under heavy jamming, demodulation must be done onboard the satellite. This involves onboard

processing. This problem is quite technologically challenging. A frequency-hopped spread spectrum satellite communication system is discussed in the next section.

2.4 Frequency-Hopped Satellite Communication Systems

2.4.1 Introduction

The need for a military satellite communication system has been discussed earlier. Such a communication system should employ spread spectrum techniques to achieve robust communications. Frequency hopping is the preferred technique for two reasons: large processing gains can be achieved and synchronization requirements are much less stringent. Onboard processing should be used for anti-jamming protection. Because it is difficult for frequency-hopped signals to maintain phase coherence, the system should use noncoherent reception. Then, noncoherent M-ary frequency-shift-keyed signalling is used. Also, in order that only one dehopper is on the satellite, frequency-division multiple access should be used. Figure 2.3 shows the system. The earth station transmits its information at some uplink frequency. The satellite receives this transmission and demodulates the information. It then sends the information to another earth station at a downlink frequency. The uplink and downlink frequencies are spaced sufficiently far apart so that they do not interfere with each other. A processing satellite has two independent communication links. The design of these links can be treated separately. The structure of these links is discussed in the following sections.

2.4.2 Transmitter

A possible transmitter structure is suggested by Shaw[8]. It is shown in Figure 2.4. The information source is transmitting binary information with a period of T_b . The encoder encodes this information in an error-correction code of period T_c such that:

$$T_c = rT_b \quad (2.8)$$

The binary information is then changed to M-ary form. The symbol period T_s is such that:

$$T_s = T_c \log_2 M \quad (2.9)$$

Figure 2-3: A Frequency-Hopped Satellite Communication System

Figure 2-4: A Transmitter Structure

The symbols then are modulated by the frequency-shift-keyed modulator. Other users then are added in a frequency-division multiple-access scheme. The output of the frequency-shift-keyed modulator and the output of the hopping-pattern generator are fed to the mixer. The output of the mixer goes to the antenna where it is transmitted into the channel. The hopping-pattern generator consists of a frequency synthesizer driven by a pseudorandom code generator. This determines the hopping pattern. The hop period T_h is given by:

$$T_h = \frac{T_s}{L} \quad (2.10)$$

For fast hopping, L must be an integer. If L is less than one, then, the system is a slow-hopping one.

An alternate transmitter structure is given by Proakis[9]. It is shown in Figure 2.5. Here, m bits from the encoder and k bits from the pseudorandom code generator are combined at the frequency synthesizer to produce the transmitted tone's frequency. In general, k is much greater than m .

2.4.3 Receiver

The problem of reception has been investigated by Shaw[8]. The receiver structure is shown in Figure 2.6. The salient feature of the receiver is the processing of the dehopped signal.

The information signal consists of one of M frequency tones each symbol period. For orthogonal signalling, the frequency tones should be spaced at integer multiples of $1/T_s$. From Shaw[8], the frequency-shift-keyed information signal for each symbol period before hopping has the form:

$$s_m(t) = \sqrt{\frac{2E_s}{T_s}} \cos[2\pi(f_c + a_m \Delta f)t] \quad (2.11)$$

E_s is the information symbol energy. T_s is the symbol period. $2\Delta f$ is the frequency spacing of adjacent tones. f_c is the intermediate frequency carrier. The data is represented in a_m . It can have the values $\pm 1, \pm 3, \dots, \pm(M-1)$. The information signal can also be expressed as:

$$s_m(t) = \text{Re}\{u_m(t) \exp(j2\pi f_c t)\} \quad (2.12)$$

where $u_m(t)$ is:

$$u_m(t) = \sqrt{\frac{2E_s}{T_s}} \exp(j2\pi a_m \Delta f t) \quad (2.13)$$

Figure 2-5: Another Transmitter Structure

Figure 2-6: A Receiver for a Frequency-Hopped M-ary FSK System

The total information signal is:

$$s(t) = \sum_{n=-\infty}^{\infty} g(t - nT_s) \text{Re}[u_m(t - nT_s) \exp(j2\pi f_c t + j\Theta_n)] \quad (2.14)$$

where $g(t)$ is given by:

$$g(t) = \begin{cases} 1 & , 0 \leq t \leq T_s; \\ 0 & , \text{otherwise} \end{cases} \quad (2.15)$$

and Θ_n is a random phase uniformly distributed over $(0, 2\pi)$ and is independent from symbol to symbol. The magnitude of the power spectral density of $s(t)$ is:

$$S_s(f) = \frac{E_s}{2MT_s^2} \sum_{i=1}^M |G[f - (M - 2i + 1)\Delta f - f_c]|^2 + |G[f + (M - 2i + 1)\Delta f + f_c]|^2 \quad (2.16)$$

where $G(f)$ is the Fourier transform of $g(t)$. In this case, $|G(f)|^2$ is given by:

$$|G(f)|^2 = T_s^2 \text{sinc}^2(T_s f) \quad (2.17)$$

The power spectral density of the noncoherent frequency-shift-keyed signal is the sum of the spectral densities of $g(t)$ centered at the frequency-shift-keyed tones. There is an optimum receiver when the received signal is a uniformly delayed and attenuated version of the transmitted signal corrupted by white Gaussian noise.

Without any time diversity, the optimum receiver is arrived at as follows. The received signal $v(t)$ is:

$$\begin{aligned} v(t) &= \text{Re}[(u_m(t)\alpha \exp(j\Theta) + z(t)) \exp(j2\pi f_c t)] \\ &= \text{Re}[r(t) \exp(j2\pi f_c t)] \end{aligned} \quad (2.18)$$

where α is the channel attenuation, Θ is a random phase uniformly distributed over $(0, 2\pi)$, and $z(t)$ is the equivalent lowpass additive white Gaussian noise with zero mean and variance N_0 . From Shaw[8], the receiver forms the decision variables U_m given by:

$$U_m = \left| \int_0^{T_s} v(t) \exp(-j2\pi f_m t) dt \right| \quad (2.19)$$

where f_m is a frequency-shift-keyed tone given by:

$$f_m = f_c - (M + 1 - 2m)\Delta f \quad (2.20)$$

for $m = 1, 2, \dots, M$. The largest value of U_m for all m is chosen as the transmitted tone.

In order to achieve desired performance, time diversity can be employed. This means that the symbol is repeated. Diversity of order L implies that the symbol is repeated L

times. Thus, decision variables U_{mk} are formed for $m = 1, 2, \dots, M$ and for $k = 1, 2, \dots, L$ from (2.19). The receiver structure for this, given by Shaw[8], is shown in Figure 2.7. The received signal is passed into M energy detectors. Each energy detector detects energy at a given frequency f_m . An energy detector is shown in Figure 2.8. The output of the energy detectors is sampled at intervals of the hop period, T_h . There is a nonlinearity $f(\bullet)$ applied to the samples U_{mk} . Two forms of $f(\bullet)$ are:

$$U'_{mk} = U_{mk} \quad (2.21)$$

which is called soft-decision detection and:

$$U'_{mk} = \begin{cases} 1 & , U_{mk} > c \\ 0 & , U_{mk} \leq c \end{cases} \quad (2.22)$$

which is called hard-decision detection. The threshold level in hard-decision detection is c . Then, the samples U'_{mk} are summed as follows:

$$U_m = \sum_{k=1}^L U'_{mk} \quad (2.23)$$

The outputs of the summers are sampled at intervals of symbol periods T_s . The largest U_m is chosen as the transmitted symbol. When the only interference is additive white Gaussian noise, then, the optimum demodulator uses soft-decision detection. However, in other cases, the optimum receiver might use hard-decision detection or some other nonlinearity $f(\bullet)$.

The above method works well as long as the number of energy detectors M is small. If M is large, then, a spectral analysis receiver should be used. It is equivalent to the receiver in Figure 2.7. The decision variables can be formed by observing the frequency spectrum of the received signal at the frequencies corresponding to the M tones of the frequency-shift-keyed signal. Thus, decision variables can be formed as follows:

$$\begin{aligned} U_m &= \left| \int_0^{T_s} v(t) \exp(-j2\pi f_m t) dt \right| \\ &= \left| \int_{-\infty}^{\infty} v(t) w(t) \exp(-j2\pi f_m t) dt \right| \\ &= |F[v(t)w(t)]|_{f=f_m} \end{aligned} \quad (2.24)$$

where $w(t)$ is the time windowing function and $F(\bullet)$ denotes the Fourier transform operation. The decision variables are the sampled magnitudes of the frequency spectrum at the M tones. The largest sample U_m is chosen to be the transmitted signal. Surface acoustic

Figure 2-7: Receiver Structure

Figure 2-8: Energy Detector at Frequency f_m

wave devices can be used to implement a spectral analysis receiver. These devices use a chirp transform which represents input spectral content by an output time waveform. The output time waveform can be sampled at the instants of time which reflect the M tones. In this fashion, the decision variables U_m can be obtained. This kind of spectral analysis receiver has been investigated by Felstead, Pearce, and Selin[10]. According to the report, this receiver is quite practical. An important aspect of spectral analysis receivers is the time windowing function $w(t)$. The most basic window is a rectangular window. However, it does not always provide the best performance.

The rectangular window is defined as:

$$w(t) = \begin{cases} 1 & , 0 \leq t \leq T_s \\ 0 & , \text{otherwise} \end{cases} \quad (2.25)$$

The magnitude of the Fourier transform of $w(t)$ is:

$$|W(f)| = T_s \frac{\sin \pi f T_s}{\pi f T_s} \quad (2.26)$$

This is a sinc function. The magnitude is zero for integer multiples of $1/T_s$ except for $f=0$. In a M-ary frequency-shift-keyed communication scheme, the M tones are placed at these nulls. Also, other users in a frequency-division multiple-access scheme place their tones at these nulls. In a system with perfect frequency and time synchronization, there would be no interference between these tones. However, perfect frequency and time synchronization cannot be achieved between users in practice. Thus, there is interference between users. This interference can be minimized by the use of time windowing.

Time windowing causes a decrease in sidelobe levels. However, there is also an increase in width of the main lobe and a decrease in the height of the main lobe. This suggests that the use of windowing causes a suboptimal receiver. The signal is smaller and thus, is more difficult to detect. This decrease in performance is called processing loss. Also, users must be spaced farther apart to accommodate the spreading of the main lobe. However, the interference from other tones decreases. Knowing the exact configuration of the system, it is possible to choose an optimum window. However, if the parameters are slightly changed, then, a different window becomes optimum. A family of windows that has good sidelobe properties is the Kaiser-Bessel family. This family of windows is defined as:

$$w(t) = \frac{I_0[\alpha \pi \sqrt{1 - (1 - \frac{2t}{T_s})^2}]}{I_0[\alpha \pi]} \quad (2.27)$$

for $0 \leq t \leq T_s$. The parameter α can be varied to control the shape of the frequency spectrum. The use of windows is important in reducing interference between tones.

2.5 Summary

This chapter has discussed the background in satellite communication and spread spectrum communication necessary to understand frequency-hopped spread spectrum satellite communication systems. The next chapter discusses fine time uplink synchronization for such systems.

Chapter 3

Theory

3.1 Introduction

Synchronization is an important aspect in any communication system. There are three main types of synchronization: spatial, frequency, and time. The main goal of this thesis is to investigate one aspect of time synchronization for one particular system - fine time uplink synchronization for a frequency-hopped spread spectrum satellite communication system. The lack of time synchronization causes signal attenuation and loss of orthogonality at the receiver which degrades system error probability performance. This chapter discusses the details of the system to be synchronized and the entire synchronization process for the system. It also presents analyses and simulation results, reported in the literature, for various fine time uplink synchronization methods.

3.2 The Frequency-Hopped Communication System

The frequency-hopped spread spectrum satellite communication system consists of several ground terminals and a satellite. The ground terminals send signals to the satellite using an uplink frequency band, and the satellite sends signals to the ground terminals using a downlink frequency band. These two bands are not overlapping. The link of interest in this thesis is the uplink.

The uplink signalling scheme uses frequency-hopping with noncoherent frequency-shift-keying. The frequency-hopped signal is generated by a frequency synthesizer controlled by a pseudorandom number generator. The same frequency-hopped signal is common to all terminals. The frequency-hopped signal modulates the information-carrying noncoherent frequency-shift-keyed signal. The system employs fast hopping. This means that there is one or more hops per data symbol.

The generation of the information-carrying signal is discussed below. The information-carrying signal is different for each terminal. Each ground terminal represents its transmitted information in M-ary form. The satellite has an intermediate frequency band available

for reception of all transmitted information signals. Each ground terminal is assigned its own continuous frequency band of equal width within the total intermediate frequency band. This type of ground terminal signal separation is called frequency-division multiple access. The ground terminal can represent its M-ary information using M tones within its frequency band. Thus, during each hop period, the ground terminal represents a piece of information using one of the M tones. The given tone is its intermediate signal for that hop period. The intermediate tone is multiplied by the hopping pattern. The product is the transmitted uplink signal.

The transmitted uplink signal from each ground terminal is received at the satellite and demodulated. The first step is to dehop the received signal. This is done by multiplying the received signal by the frequency-hopping pattern. The product leaves only the transmitted intermediate frequency signal. A Fourier transform is performed on the intermediate frequency signal using an analog surface acoustic wave spectrum analyzer. The output of the surface acoustic wave spectrum analyzer is a time waveform which represents the Fourier transform of the input intermediate signal. By sampling at the appropriate time, it is possible to obtain the spectral power at a given frequency. Thus, by sampling at the times corresponding to the transmitted communication tones, it is possible to determine the spectral power of these tones. The largest sample is chosen as the transmitted tone. Thus, the signal has been demodulated. The satellite would then send the signal to the appropriate receiving ground terminal using the downlink signalling scheme.

In order to function, the communication system must be synchronized. The following section discusses the entire synchronization process for this communication system.

3.3 The Synchronization Process

In order to understand the role of fine time uplink synchronization in a satellite communication system, it is necessary to understand where this type of synchronization fits in the entire synchronization process. There are three main parts to synchronization: spatial synchronization, frequency synchronization, and timing synchronization. It is important to note that in this system, most of the synchronization is done by the ground terminal. The ground terminal must locate and track the satellite. It must also match its transmit and receive frequencies and its transmit and receive timing to those of the satellite's communica-

tion payload. This means that the ground terminal must point its antenna at the satellite. Also, the ground terminal must match its receiver's frequency and timing to the received signal's frequency and timing. Also, the ground terminal must change its transmitter's frequency and timing so that the transmitted signal arrives at the satellite's communication payload aligned with the payload receiver's frequency and timing. The satellite's communication payload does provide some aids to the ground terminal to achieve the above synchronizations.

Synchronization is one of the first tasks prior to communication. The first time that a given ground terminal wishes to use the satellite, it must go through the entire synchronization process. For subsequent usages, it may only be necessary to do part of the entire process. How much of the process must be done depends on how badly the synchronization has degraded. The ground terminal should check its synchronization periodically to avoid performance degradation. This check usually involves correction for small misalignments. If the misalignment is large, then, it may be necessary to do part or all of the synchronization process again. This check of synchronization is called tracking. Different users may start at different points in the synchronization process. The starting point is determined by how much knowledge the ground terminal has about the state of the system. A user that has not been using the system for a few minutes or seconds may have to do only a few steps in the process, whereas a user who has been idle for several days without any contact with the system may have to go through the entire process. The entire synchronization process is discussed in what follows. It is assumed in the following discussion that the synchronization process is being undertaken by a first-time user.

The first part of the synchronization process is spatial synchronization. This involves pointing the ground terminal's antenna in the direction of the satellite. If the terminal's antenna is not pointed correctly, then, the system may not work at all or may have its performance degraded. For a first-time user, it is necessary to have some information about the location of the satellite. This information is obtained by using ephemeris data to compute the satellite's orbit. The antenna is then pointed in the direction given by the orbital computation. The satellite's communication payload is now sending a frequency-hopped signal. If the antenna is correctly pointed, then, a wideband energy detector detects a signal. The antenna can be slightly steered to maximize the detected signal. However, if a signal is not detected, then, it is necessary to search for the satellite. This is done

by steering the antenna and observing the output of the wideband energy detector until a detected signal is observed. When this happens, the antenna pointing is adjusted until a maximum occurs in the detected signal. At this point, spatial synchronization has occurred. Spatial synchronization is particularly difficult at high frequencies such as EHF. Here, narrow beamwidths - a fraction of a degree - make pointing difficult. Also, atmospheric conditions can cause deep fades in the signals - especially at low elevation pointing angles - due to changes in refractive indices. Thus, accurate pointing and constant updating make spatial synchronization very important for good system performance.

The second step in the entire synchronization process is the downlink frequency and time acquisition. Time acquisition is usually done first. Here, the goal is to align the downlink received signal's timing with the timing of the ground terminal's hopping pattern. This is usually done in two steps: coarse and fine acquisition. The coarse acquisition brings the two hopping pattern to within a half of a hop period. The fine acquisition aligns the two patterns to within the required tolerances. The fine acquisition process is also used to track the downlink timing synchronization. If the misalignment exceeds half of a hop period, then, it is necessary to synchronize the downlink timing again. After the hopping patterns have been aligned, symbol and frame synchronization is initiated. Frequency synchronization follows in order to match the received signal's frequencies with the ground terminal's frequencies. Once time and frequency synchronization have been achieved, then, the downlink can be used as a communication link.

The third step in the entire synchronization process is the uplink time and frequency synchronization. Uplink frequency synchronization is done by using the information obtained from the downlink frequency synchronization. The ground terminal's uplink transmitted frequencies are corrected accordingly. Time synchronization is then initiated. It is also done in two steps: coarse and fine. The coarse synchronization aligns the uplink received signal's hopping pattern with the satellite's communication payload receiver's hopping pattern to within a half of a hop period. Then, the fine synchronization aligns the two hopping patterns to within the required tolerance. Symbol and frame synchronization follow this. The fine time synchronization is also used for tracking of the uplink timing.

At this point, the system is synchronized and ready for use. The objective of this thesis is the study of the uplink fine time synchronization. Methods for this are discussed in the next section.

3.4 Methods for Fine Time Uplink Synchronization

3.4.1 Introduction

Fine time uplink synchronization for a frequency-hopped spread spectrum satellite communication system presents a problem. In many communication systems, time synchronization is done by the receiver. However, for the uplink of this type of satellite communication system, this is not possible. The system uses a single dehopper on the satellite and has many users per satellite transponder. Also, the use of a processing transponder makes it necessary to demodulate signals onboard the satellite. For these reasons, the satellite's clock must not change. Hence, the ground terminals must change their timing to accommodate the satellite.

Fine time uplink synchronization is achieved by having the ground terminal transmit a series of synchronization probes to the satellite. The satellite processes these probes and returns an estimate of the timing error. The ground terminal, then, corrects its timing and sends another set of probes. This process is repeated until the system is synchronized within the required tolerances. Note that at the time of fine time uplink synchronization, the downlink has been synchronized and can be used as a communication link for information transmission from the satellite to the ground terminal. Also, coarse time uplink synchronization has occurred, and the uplink timing is correct to within a half of a hop period.

Recent publications have presented analyses of the problem of fine time uplink synchronization and of possible methods for achieving it. These are described in this section. The presented analyses follow work reported in the literature.

3.4.2 Receiver Processing

The processing of the uplink synchronization probes is at the heart of the fine time uplink synchronization process. It is, therefore, important to understand the processing done by the satellite. A simplified model of the satellite receiver is shown in Figure 3.1. The receiver consists of a dehopper which converts the frequency-hopped input signal to an intermediate signal consisting of the data tones. The dehopper also truncates the intermediate signal so that it fits within the surface acoustic wave spectrum analyzer's window.

Figure 3-1: Satellite Receiver

This gating is shown in Figure 3.2. The output of the dehopper goes to the surface acoustic wave spectrum analyzer. This device performs a Fourier transform on its input. The output of the surface acoustic wave spectrum analyzer is sampled at the appropriate places corresponding to the tones used for communication. The samples are sent to the processor. Here, a decision is made concerning the transmitted signal.

A mathematical description of the receiver processing is required in order to understand the synchronization process. The description follows the work of Mason and Felstead[5]. The ground terminal transmits signals of length T_h , where T_h is the hop period. Each signal is a pulsed microwave tone. The signal is the data tone modulated by the frequency-hopping pattern. The satellite receives this signal. The signal is passed through the dehopper. For the data tone to appear at the output of the dehopper, each of the satellite's hop periods must overlap with the corresponding hop period of the received signal. Only the overlapped portions appear at the output of the dehopper. In the case of perfect synchronization, a data tone appears for the entire hop period. At the commencement of fine synchronization, the received signal and the dehopper's hopping pattern are aligned to within half of a hop period. Therefore, there is at least half of a hop period of data tone signal at the dehopper's output. However, if the system is out of synchronization by more than a hop period, only noise appears at the output of the spectrum analyzer.

A mathematical description of the signals at the input of the surface acoustic wave spectrum analyzer is necessary to understand the receiver processing. The timing is referred to the input. The satellite receiver expects to receive a particular signal during the following hop period:

$$-\frac{T_h}{2} < t < \frac{T_h}{2} \quad (3.1)$$

The actual received signal arrives at:

$$-\frac{T_h}{2} + \Delta T < t < \frac{T_h}{2} + \Delta T \quad (3.2)$$

ΔT is the timing error. Since coarse synchronization has occurred,

$$|\Delta T| < \frac{T_h}{2} \quad (3.3)$$

Now, consider the output of the dehopper. Output signals only occur for overlapping corresponding hop periods of received signal and satellite's hopping pattern. Therefore, for

Figure 3-2: The Gating by the Dehopper

$\Delta T > 0$, the dehopper output is:

$$r_j(t) = \begin{cases} n_j(t) & -\frac{T_h}{2} < t < -\frac{T_h}{2} + \Delta T \\ A\cos(2\pi f_j t + \phi_j) + n_j(t) & -\frac{T_h}{2} + \Delta T < t < \frac{T_h}{2} \end{cases} \quad (3.4)$$

For $\Delta T < 0$, the dehopper output is:

$$r_j(t) = \begin{cases} A\cos(2\pi f_j t + \phi_j) + n_j(t) & -\frac{T_h}{2} < t < \frac{T_h}{2} + \Delta T \\ n_j(t) & \frac{T_h}{2} + \Delta T < t < \frac{T_h}{2} \end{cases} \quad (3.5)$$

where $r_j(t)$ is the received signal and noise, $n_j(t)$ is zero mean white Gaussian noise with power spectral density $N_0/2$, A is the amplitude of the signal, f_j is the data tone frequency, and ϕ_j is the phase of the received signal,

The surface acoustic wave spectrum analyzer performs a Fourier transform on the dehopper output as follows:

$$R_j(t) = \left| \int_{-\frac{T_h}{2}}^{\frac{T_h}{2}} r_j(\tau) \exp(-j2\pi\mu t\tau) d\tau \right| \quad (3.6)$$

where $R_j(t)$ is the output signal from the surface acoustic wave device, t is the output time, τ is the input time, and μ is the chirp slope. Note that μt has the dimension of frequency. Therefore, an output from the surface acoustic wave spectrum analyzer at time t is due to an input signal of frequency μt . For the given $r_j(t)$, $R_j(t)$ is:

$$R_j(t) = |S_j(t) + N_j(t)| \quad (3.7)$$

where

$$S_j(t) = \frac{A}{2}(1 - |\epsilon|)T_h \exp(j(\phi_j + \pi(f_j - \mu t)\epsilon T_h)) \frac{\sin(\pi(f_j - \mu t)(1 - |\epsilon|)T_h)}{\pi(f_j - \mu t)(1 - |\epsilon|)T_h} \quad (3.8)$$

and

$$N_j(t) = \int_{-\frac{T_h}{2}}^{\frac{T_h}{2}} n_j(\tau) \exp(-j2\pi\mu t\tau) d\tau \quad (3.9)$$

ϵ is the normalized timing error given by:

$$\epsilon = \frac{\Delta T}{T_h} \quad (3.10)$$

The sampler samples $R_j(t)$ at:

$$t = \frac{f_j}{\mu} \quad (3.11)$$

The sample z_j is:

$$z_j = |R_j(\frac{f_j}{\mu})| = |\frac{A}{2} \exp(j\phi_j)(1 - |\epsilon|)T_h + n_j| \quad (3.12)$$

where:

$$n_j = N_j(\frac{f_j}{\mu}) = n_{c,j} + j n_{s,j} \quad (3.13)$$

Define θ to be:

$$\theta = \frac{A}{2}(1 - |\epsilon|)T_h \quad (3.14)$$

Then,

$$z_j = |\theta \cos(\phi_j) + n_{c,j} + j(\theta \sin(\phi_j) + n_{s,j})| \quad (3.15)$$

The variances of $n_{c,j}$ and $n_{s,j}$ are the same and equal:

$$\sigma^2 = \frac{N_0 T_h}{4} \quad (3.16)$$

From Benedict and Soong[11], the probability distribution of z_j is Rician and is given by:

$$p(z_j) = \frac{z_j}{2\sigma^2} \exp\left(\frac{z_j^2 + \theta^2}{-2\sigma^2}\right) I_0\left(\frac{\theta z_j}{\sigma^2}\right) \quad (3.17)$$

for $z_j > 0$ and:

$$p(z_j) = 0 \quad (3.18)$$

for $z_j \leq 0$. Note that I_0 is the modified zeroth Bessel function of the first kind. Moments play an important role in the formation of an estimator. From Benedict and Soong[11], the first and second moments of z_j are:

$$E[z_j] = \sqrt{\frac{\pi}{2}} \sigma \exp\left(-\frac{\theta^2}{2\sigma^2}\right) F_{11}(3/2; 1; \theta^2/2\sigma^2) \quad (3.19)$$

and

$$E[z_j^2] = \theta^2 + 2\sigma^2 \quad (3.20)$$

where F_{11} is the confluent hypergeometric function. These moments are the key to estimation of the timing error.

Determining ϵ is based on estimating θ . However, a single measurement of θ is not sufficient to determine ϵ since A and the sign of ϵ are not known. Thus, by estimating θ under two different conditions, ϵ is able to be determined. The samples, z_j , are used to

estimate θ . Before discussing how this can be done, it is important to mention some of the characteristics of estimators.

3.4.3 Estimator

It is important to understand certain characteristics of the estimate of the timing error. The estimate is statistical in nature. Hence, the estimate has a probability density and from this, properties of the estimator such as mean and standard deviation can be computed. Ideally, the estimator should have a mean equal to the timing error and a relatively small standard deviation. At this point, it is important to discuss two statistical properties of estimators - consistency and bias. From Kendall and Stuart [12], a consistent estimator, t_n , of a parameter θ formed from a sample size n is an estimator whose value can be made arbitrarily close to the parameter θ by choosing n sufficiently large. An estimator is unbiased if the mean of the estimator has an expected value equal to the parameter for all n . For a consistent estimator, the bias - if any - reduces as the number of samples increases.

The following sections discuss various ways of forming the fine time uplink timing error estimator in order to achieve a consistent estimator.

3.4.4 Delay-Advance Method

This method was first discussed in a paper by Chadwick and Springett[1]. They used this method for fine time synchronization of a frequency-shift-keyed communication system which did not employ frequency hopping. The method has been modified by Simon et al.[2] for use with frequency-hopped systems. The method has further been modified by Mason and Felstead[5] for improved performance. It is worthwhile to understand the Simon approach in order to see why an improved method is necessary. Thus, Simon's method is discussed first followed by Mason and Felstead's method.

In order to obtain an estimate for the timing error, it is necessary to make measurements on the system under two different conditions. Simon et al.[2] have suggested that a series of synchronization probes be sent which are intentionally delayed by half of a hop period relative to the current synchronization position and that then a series be sent which are intentionally advanced by half of a hop period. This allows the determination of the sign of ϵ and also, allows the determination of its magnitude. The sampled outputs of the surface

acoustic wave device are:

$$z_{dj} = \left| \frac{A}{2} \exp(j\phi_j) \left(\frac{1}{2} - \epsilon \right) T_h + n_{dj} \right| \quad (3.21)$$

$$z_{aj} = \left| \frac{A}{2} \exp(j\phi_j) \left(\frac{1}{2} + \epsilon \right) T_h + n_{aj} \right| \quad (3.22)$$

Equation (3.21) applies to the delayed transmitter situation and (3.22) applies to the advanced transmitter situation. The desired parameters to be estimated are:

$$\theta_d = \frac{A}{2} \left(\frac{1}{2} - \epsilon \right) T_h \quad (3.23)$$

and

$$\theta_a = \frac{A}{2} \left(\frac{1}{2} + \epsilon \right) T_h \quad (3.24)$$

If $\widehat{\theta}_d$, $\widehat{\theta}_a$, and $\widehat{\epsilon}$ are the estimates for θ_d , θ_a , and ϵ , then,

$$\widehat{\epsilon} = \frac{1}{2} \frac{\widehat{\theta}_a - \widehat{\theta}_d}{\widehat{\theta}_a + \widehat{\theta}_d} \quad (3.25)$$

The estimates are formed from the samples of L hop periods. In Simon's method, $\widehat{\theta}_d$ is formed from $L/2$ hop periods when the transmitter is delayed as follows:

$$\widehat{\theta}_d = \sqrt{\frac{2}{L} \sum_{j=1}^{L/2} z_{dj}^2} \quad (3.26)$$

$\widehat{\theta}_a$ is formed from $L/2$ hop periods when the transmitter is advanced as follows:

$$\widehat{\theta}_a = \sqrt{\frac{2}{L} \sum_{j=1}^{L/2} z_{aj}^2} \quad (3.27)$$

The problem with this method is that the estimator $\widehat{\epsilon}$ is inconsistent. In low signal-to-noise ratios, the estimate $\widehat{\epsilon}$ tends to be much less than the actual error ϵ . This is caused by the fact that the z_{ij} is a Rician random variable, and the expected value of z_{ij}^2 is given by (3.20). This causes the expected value of $\widehat{\theta}_i$ to be:

$$E[\widehat{\theta}_i] = \sqrt{2\sigma_i^2 + \theta_i^2} \quad (3.28)$$

where $2\sigma_i^2$ is the noise variance term. Thus, $\hat{\theta}_i$ is expected to be higher than θ_i . This causes the estimate $\hat{\epsilon}$ to be low in low signal-to-noise cases.

Simulation results from Mason[13] are shown in Figures 3.3 and 3.4. The results are obtained for $L=256$. This means that 128 delayed and 128 advanced hop periods are used. Figure 3.3 shows that the method results in estimates which are lower than they should be for both 10 dB and 0 dB signal-to-noise ratio cases. Mason[4] shows that increasing L does not cause the estimates to increase to the actual timing error value. Thus, the estimate is inconsistent. However, this method does produce an estimate with fairly low standard deviation relative to other methods discussed later. However, because of its inconsistency, the method is not useful.

In Mason[4], an analysis is done using maximum likelihood estimation. It shows that the maximum likelihood estimators for θ_i are:

$$\hat{\theta}_i = \frac{2}{L} \sum_{j=1}^{\frac{L}{2}} z_{ij} \quad (3.29)$$

However, because z_{ij} is a Rician random variable, its expected value is given by (3.19). Thus,

$$E[\hat{\theta}_i] > \theta_i \quad (3.30)$$

Once again, this causes the estimator $\hat{\theta}_i$ to be high in low signal-to-noise ratio cases. Thus, $\hat{\epsilon}$ is low. This method also leads to an inconsistent estimator.

A consistent estimator results from a method using noise estimation. This is described in Mason and Felstead[5]. Let z_{0j} be a measurement made of the surface acoustic wave device's output with no input signal. Its expected value is:

$$E[z_{0j}^2] = 2\sigma^2 \quad (3.31)$$

Recall that:

$$E[z_{ij}^2] = \theta_i^2 + 2\sigma^2 \quad (3.32)$$

where z_{ij} is a measurement made with a signal present. Thus,

$$E[z_{ij}^2 - z_{0j}^2] = \theta_i^2 \quad (3.33)$$

Thus, a consistent estimate can be obtained by:

$$\hat{\theta}_i = \sqrt{\left(\frac{2}{L} \sum_{j=1}^{L/2} z_{ij}^2\right) - \left(\frac{1}{L} \sum_{j=1}^L z_{0j}^2\right)} \quad (3.34)$$

Figure 3-3: Delay-Advance Method without Bias Reduction

Figure 3-4: Delay-Advance Method without Bias Reduction

Note that if the quantity under the square root sign is negative, then, set:

$$\hat{\theta}_i = 0 \quad (3.35)$$

This method leads to a consistent estimator.

Simulation results are presented in Figures 3.5 and 3.6 from Mason[13]. $L=256$ is used. Note that there is some bias in the mean of the estimator in Figure 3.5. The bias is particularly noticeable near $\epsilon = 0.5T_h$. However, Mason[4] shows that the bias disappears as L is increased. The standard deviation of $\hat{\epsilon}$ increases considerably from the results of the method with bias reduction.

3.4.5 Early-Late Filter Method

The early-late filter method was first discussed in a paper by Hodsdon[3]. The method has been modified by Mason and Felstead[5]. The method involves filtering the output of the surface acoustic wave device instead of sampling it. The amount of energy in the filter determines the estimates.

The key to this method is the spectrum of the signal at the output of the surface acoustic wave device. Consider the following hop period:

$$-\frac{T_h}{2} < t < \frac{T_h}{2} \quad (3.36)$$

Assume that the received signal arrives at:

$$-\frac{T_h}{2} + \Delta T < t < \frac{T_h}{2} + \Delta T \quad (3.37)$$

The output of the surface acoustic wave spectrum analyzer is at an intermediate frequency f_v . It can be shown that the magnitude of the Fourier transform of the spectrum analyzer's output has a rectangular response with a bandwidth equal to:

$$B = \mu T_h (1 - |\epsilon|) \quad (3.38)$$

and a centre frequency equal to:

$$f_{\text{centre}} = f_v + \frac{\mu T_h \epsilon}{2} \quad (3.39)$$

and a magnitude of $A\mu/2$. Two filters - designated early and late - are placed at the following locations. The early filter has its passband from $(f_c - \frac{\mu T_h}{2})$ to (f_c) . The late

Figure 3-5: Delay-Advance Method with Bias Reduction

Figure 3-6: Delay-Advance Method with Bias Reduction

filter has its passband from (f_c) to $(f_c + \frac{\mu T_h}{2})$. The filters are assumed to be ideal with a unity passband, infinite stopband, and linear phase. The time outputs of the two filters are sampled and envelope detected. From Mason[4], the outputs are for $\epsilon > 0$:

$$z_{e_j} = |\frac{A}{4} \exp(j\phi_j) T_h + n_{e_j}| \quad (3.40)$$

$$z_{l_j} = |\frac{A}{2}(\frac{1}{2} - \epsilon) \exp(j\phi_j) T_h + n_{l_j}| \quad (3.41)$$

and for $\epsilon < 0$:

$$z_{l_j} = |\frac{A}{4} \exp(j\phi_j) T_h + n_{l_j}| \quad (3.42)$$

$$z_{e_j} = |\frac{A}{2}(\frac{1}{2} + \epsilon) \exp(j\phi_j) T_h + n_{e_j}| \quad (3.43)$$

z_{e_j} is the sampled output of the early filter. z_{l_j} is the sampled output of the late filter. Note that z_{l_j} are Rician random variables. σ^2 is:

$$\sigma^2 = \frac{N_0 T_h}{8} \quad (3.44)$$

Thus, in order to get a consistent estimate, it is necessary to use noise estimation. Let

$$z_{e_j} = |\theta_e \exp(j\phi_j) + n_{e_j}| \quad (3.45)$$

$$z_{l_j} = |\theta_l \exp(j\phi_j) + n_{l_j}| \quad (3.46)$$

where for $\epsilon > 0$,

$$\theta_e = \frac{A}{4} T_h \quad (3.47)$$

$$\theta_l = \frac{A}{2}(\frac{1}{2} - \epsilon) T_h \quad (3.48)$$

and for $\epsilon < 0$,

$$\theta_l = \frac{A}{4} T_h \quad (3.49)$$

$$\theta_e = \frac{A}{2}(\frac{1}{2} + \epsilon) T_h \quad (3.50)$$

Estimates can be formed by transmitting L hops. The early and late filters are sampled at the appropriate time to produce z_{e_j} and z_{l_j} . An unoccupied channel for an output which is not passed through either filter is sampled to get z_{0j} . θ_e is estimated by:

$$\hat{\theta}_e = \sqrt{\frac{1}{L} \sum_{j=1}^L z_{e_j}^2 - \frac{1}{2L} \sum_{j=1}^L z_{0j}^2} \quad (3.51)$$

θ_l is estimated by:

$$\hat{\theta}_l = \sqrt{\frac{1}{L} \sum_{j=1}^L z_{lj}^2 - \frac{1}{2L} \sum_{j=1}^L z_{0j}^2} \quad (3.52)$$

Define β as:

$$\beta = \frac{\hat{\theta}_l}{\theta_e} \quad (3.53)$$

An estimate of timing error can now be determined. If $\beta < 1$, then,

$$\hat{\epsilon} = \frac{1}{2}(1 - \beta) \quad (3.54)$$

If $\beta > 1$, then,

$$\hat{\epsilon} = \frac{1}{2}\left(\frac{1}{\beta} - 1\right) \quad (3.55)$$

Simulation results are presented in Figure 3.7 and 3.8 from Mason[13]. The results are for $L=256$. Figure 3.7 shows that the method leads to a consistent estimator. The bias is less than in the delay-advance method with bias reduction. The standard deviation is also slightly smaller. This is caused by the fact that for the early-late filter method, 256 hop periods generate 256 timing error estimates. For the delay-advance method, 256 hop periods generate only 128 estimates. This occurs because a delayed and an advanced hop period are required to form an estimate. Thus, the early-late filter method performs better than the delay-advance method.

A form of this method has been implemented in the FLTSAT system and is described by the paper by Hodsdon[3]. The experimental system allows for the reception and demodulation of uplink frequency-hopped noncoherent frequency-shift-keyed signals. The system uses a surface acoustic wave spectrum analyzer with an analog-to-digital converter and a digital signal processor to demodulate the uplink signals. The system uses overlapping early and late filters. The paper provides a graph showing the estimate versus the actual timing error. However, facts such as signal-to-noise ratio and number of hop periods are omitted. Thus, it is not possible to compare its performance with the simulation results.

3.4.6 Modified Early-Late Filter Method

This method is described in a paper by Mason and Felstead[5]. This method uses three estimates of the signal. It uses the two estimates employed by the early-late filter method

Figure 3-7: Early-Late Filter Method with $L=256$

Figure 3-8: Early-Late Filter Method with $L=256$

and also an extra estimate from the signal that is not passed through either filter. The third estimate is:

$$\theta_n = \frac{A}{2}(1 - |\epsilon|)T_h \quad (3.56)$$

Bias reduction is again used to form $\widehat{\theta}_e$, $\widehat{\theta}_l$, and $\widehat{\theta}_n$. Only the two largest of these estimates is used. $\widehat{\theta}_n$ is always one of them. The time estimate is formed as follows. For $\epsilon > 0$,

$$\alpha = \frac{\widehat{\theta}_n}{\widehat{\theta}_l} \quad (3.57)$$

and

$$\hat{\epsilon} = 1 - \frac{\alpha}{2} \quad (3.58)$$

and for $\epsilon < 0$,

$$\alpha = \frac{\widehat{\theta}_n}{\widehat{\theta}_e} \quad (3.59)$$

and

$$\hat{\epsilon} = \frac{\alpha}{2} - 1 \quad (3.60)$$

Simulation results are shown in Figures 3.9 and 3.10 from Mason[13]. The bias is considerably reduced, especially near $|\epsilon| = 0.5$. The mean of the estimate is very close to the actual error. The standard deviation is about the same as for the early-late filter method. This method gives the best results.

3.4.7 Two-Tone Method

The two-tone method was first discussed in Mason and Felstead[5]. The synchronization probes that are sent contain two frequencies. For the first half of the hop period, one frequency is transmitted, and for the second half, a second frequency is transmitted. The two-tone probe and the effect of the dehopper gating is shown in Figure 3.11.

The transmitter transmits its signal during the following hop period:

$$-\frac{T_h}{2} + \Delta T < t < \frac{T_h}{2} + \Delta T \quad (3.61)$$

The receiver expects to receive the signal during the following hop period:

$$-\frac{T_h}{2} < t < \frac{T_h}{2} \quad (3.62)$$

Figure 3-9: Modified Early-Late Method with $L=256$

Figure 3-10: Modified Early-Late Method with $L=256$

Figure 3-11: The Two-Tone Probe Gating

The received signal at the surface acoustic wave device for $\Delta T > 0$ is:

$$r_j(t) = \begin{cases} n_j(t) & -\frac{T_h}{2} < t < -\frac{T_h}{2} + \Delta T \\ A \cos(2\pi f_1 t + \phi_{1j}) + n_j(t) & -\frac{T_h}{2} + \Delta T < t < \Delta T \\ A \cos(2\pi f_2 t + \phi_{2j}) + n_j(t) & \Delta T < t < \frac{T_h}{2} \end{cases} \quad (3.63)$$

For $\Delta T < 0$, the received signal is:

$$r_j(t) = \begin{cases} A \cos(2\pi f_1 t + \phi_{1j}) + n_j(t) & -\frac{T_h}{2} < t < \Delta T \\ A \cos(2\pi f_2 t + \phi_{2j}) + n_j(t) & \Delta T < t < \frac{T_h}{2} + \Delta T \\ n_j(t) & \frac{T_h}{2} + \Delta T < t < \frac{T_h}{2} \end{cases} \quad (3.64)$$

where $r_j(t)$ is the received signal, $n_j(t)$ is zero mean white Gaussian noise with power spectral density $N_0/2$. A is the amplitude of the received signal, f_1 and f_2 are the frequencies of the two tones, and ϕ_{1j} and ϕ_{2j} are the received phases.

The surface acoustic wave chirp transformer performs a Fourier transform on the de-hopped signal to give:

$$R_j(t) = \left| \int_{-\frac{T_h}{2}}^{\frac{T_h}{2}} r_j(\tau) \exp(-j2\pi\mu t\tau) d\tau \right| \quad (3.65)$$

where $R_j(t)$ is the output signal from the surface acoustic wave device, t is output time, τ is input time, and μ is the chirp slope. For the given $r_j(t)$, $R_j(t)$ can be expressed as:

$$R_j(t) = |S_j(t) + N_j(t)| \quad (3.66)$$

where

$$S_j(t) = S_{1j}(t) + S_{2j}(t) \quad (3.67)$$

For $\epsilon > 0$,

$$S_{1j}(t) = \frac{AT_h}{4} \exp[j\phi_{1j} + j(\frac{\pi}{2})(\mu t - f_1)T_h(1 - 4\epsilon)] \frac{\sin(\frac{\pi}{2})(\mu t - f_1)T_h}{\frac{\pi}{2}(\mu t - f_1)T_h} \quad (3.68)$$

and

$$S_{2j}(t) = \frac{AT_h}{2} (\frac{1}{2} - \epsilon) \exp[j\phi_{2j} - j\pi(\mu t - f_2)T_h(\frac{1}{2} + \epsilon)] \frac{\sin(\pi(\mu t - f_2)(\frac{1}{2} - \epsilon)T_h)}{\pi(\mu t - f_2)(\frac{1}{2} - \epsilon)T_h} \quad (3.69)$$

In order to minimize sidelobe interference, the two tones should have an orthogonal spacing. Therefore,

$$|f_2 - f_1|T_h = 2n \quad (3.70)$$

where n is an integer. $R_j(t)$ should be sampled at:

$$t_1 = \frac{f_1}{\mu} \quad (3.71)$$

and at:

$$t_2 = \frac{f_2}{\mu} \quad (3.72)$$

The output samples then are:

$$z_{1j} = R_j(t_1) = \left| \frac{AT_h}{4} \exp(j\phi_{1j}) + \frac{AT_h}{2} \left(\frac{1}{2} - \epsilon \right) \exp(j\phi_{2j} + jn\pi(1 + 2\epsilon)) \frac{\sin(n\pi(1 - 2\epsilon))}{n\pi(1 - 2\epsilon)} + n_{1j} \right| \quad (3.73)$$

and

$$z_{2j} = R_j(t_2) = \left| \frac{AT_h}{2} \left(\frac{1}{2} - \epsilon \right) \exp(j\phi_{2j}) + n_{2j} \right| \quad (3.74)$$

Note that the second term in z_{1j} tends to be small compared to the first term. It is caused by sidelobe interference. Thus, z_{1j} is assumed to be:

$$z_{1j} = \left| \frac{AT_h}{4} \exp(j\phi_{1j}) + n_{1j} \right| \quad (3.75)$$

From this, it can be seen that both z_{1j} and z_{2j} are Rician variables. If $\epsilon < 0$, a similar analysis can be undertaken. Then, z_{1j} is given by:

$$z_{1j} = \left| \frac{AT_h}{2} \left(\frac{1}{2} + \epsilon \right) \exp(j\phi_{1j}) + n_{1j} \right| \quad (3.76)$$

and z_{2j} is approximately given by:

$$z_{2j} = \left| \frac{AT_h}{4} \exp(j\phi_{2j}) + n_{2j} \right| \quad (3.77)$$

Since the two samples have a Rician distribution, it is useful to correct the sample estimates by subtracting the noise variance in order that the timing error estimate be consistent. A third sample, z_{0j} , is taken at a location where there is no signal. For $\epsilon > 0$, the parameters to be estimated are:

$$\theta_1 = \frac{AT_h}{4} \quad (3.78)$$

$$\theta_2 = \frac{AT_h}{2} \left(\frac{1}{2} - \epsilon \right) \quad (3.79)$$

For $\epsilon < 0$, the parameters to be estimated are:

$$\theta_1 = \frac{AT_h}{2} \left(\frac{1}{2} + \epsilon \right) \quad (3.80)$$

$$\theta_2 = \frac{AT_h}{4} \quad (3.81)$$

θ_1 and θ_2 are estimated as follows:

$$\hat{\theta}_1 = \sqrt{\frac{1}{L} \sum_{j=1}^L z_{1j}^2 - \frac{1}{L} \sum_{j=1}^L z_{0j}^2} \quad (3.82)$$

$$\hat{\theta}_2 = \sqrt{\frac{1}{L} \sum_{j=1}^L z_{2j}^2 - \frac{1}{L} \sum_{j=1}^L z_{0j}^2} \quad (3.83)$$

If the quantity under the square root sign is negative, then, set:

$$\hat{\theta}_i = 0 \quad (3.84)$$

for $i = 1, 2$. For $\hat{\theta}_1 > \hat{\theta}_2$, the estimate for ϵ is:

$$\hat{\epsilon} = \frac{1}{2} \left(1 - \frac{\hat{\theta}_2}{\hat{\theta}_1} \right) \quad (3.85)$$

For $\hat{\theta}_1 < \hat{\theta}_2$, the estimate for ϵ is:

$$\hat{\epsilon} = \frac{1}{2} \left(\frac{\hat{\theta}_1}{\hat{\theta}_2} - 1 \right) \quad (3.86)$$

For $\hat{\theta}_1 = \hat{\theta}_2$, the estimate for ϵ is:

$$\hat{\epsilon} = 0 \quad (3.87)$$

Simulation results from Mason[13] are shown in Figures 3.12 and 3.13. The results show a slightly greater bias than the early-late filter method. Also, the standard deviation is slightly higher.

3.4.8 A Comparison of Methods

The various methods are now compared. In each case, the method employing bias reduction is used for comparison since it is the one that gives the best performance.

The delay-advance method has the disadvantage of forcing the transmitter to change its symbol timing by half of a hop period. Usually, $L/2$ probes are sent up delayed and then, $L/2$ probes are sent up advanced. Thus, this scheme disrupts the transmitter's timing. Another disadvantage is that the method forms an estimate from two hop periods. There could be a systematic error introduced by variations in the receiver's hop band response. Also, the method gives the largest bias of all the methods.

Figure 3-12: The Two-Tone Method with $L=256$

Figure 3-13: The Two-Tone Method with $L=256$

The modified early-late filter gives the best performance by far. There is no change needed in signal structure. The only drawback is the added hardware needed in the satellite. The modified early-late filter method provides better performance than the early-late filter method without any added disadvantages. Hence, there is no reason not to use the modified method.

The two-tone method also does not change the protocols. This method has been chosen as the one to implement in a laboratory experiment because of its simplicity. It requires no additional hardware. The experimental implementation is described in the next chapter.

3.5 Summary

This chapter has presented a description of the synchronization process in frequency-hopped satellite communication systems. It also has presented analyses of various methods of fine time uplink synchronization that are reported in the literature. Simulation results from Mason[13] indicate that the modified early-late filter method give the best performance.

There is a noticeable lack of experimental data in the literature on these methods. Only Hodsdon[3] reports experimental results, but does not give results for the performance in the presence of noise. Experimental results are imperative in order to validate theoretical analyses. Clearly, there is a need for experimental results. Thus, the goal of the next two chapters is to obtain experimental results on the performance of the two-tone method.

Chapter 4

Experimental Hardware

4.1 Introduction

Proof of a theoretical concept is achieved through experiment. The two-tone method presents a plausible technique for fine time uplink synchronization as shown by the theoretical analyses presented in the last chapter. This chapter endeavours to provide experimental verification for the two-tone method as a viable method for fine time uplink synchronization. Obtaining this kind of verification is an involved process. This chapter describes the hardware for this involved process.

To test the performance of the two-tone method, a hardware-based experiment should be conducted. This hardware-based experiment should be designed to reflect the performance of the two-tone method in an actual frequency-hopped spread spectrum satellite communication system. Since a frequency-hopped satellite communication channel is unavailable, a simplified system that could be implemented on a laboratory bench has been built. This experiment could not possibly incorporate all the elements of an actual satellite system. However, all the essential elements of the satellite system pertinent to the fine time synchronization process have been included.

Consider an actual frequency-hopped spread spectrum satellite communication system that consists of a ground terminal, the satellite's communication payload, and the channel between them. This simplified system is to undergo the fine time uplink synchronization process using the two-tone method. This implies that spatial synchronization and frequency synchronization have already occurred. Also, the downlink time synchronization and the coarse time uplink synchronization have occurred.

Figure 4.1 shows the simplified satellite system. The fine time synchronization process begins with the ground terminal generating a series of uplink synchronization probes. In this case, they are two-tone probes. These probes are frequency-hopped and transmitted. The channel corrupts the signal by adding noise. In fact, there are many effects caused by the channel; however, only additive noise is shown here. The satellite's payload receives

Figure 4-1: Simplified Satellite System Configuration

the series of probes and sends the dehopped signals to the surface acoustic wave spectrum analyzer. The spectrum analyzer Fourier transforms the signal. The analog-to-digital converter samples the output and sends the samples to the microprocessor. The microprocessor averages the spectral samples over the series of probes that have been sent. The averaged spectral samples are sent to the ground terminal. These include the spectral estimates at the frequencies of the two tones of the probe and at least one spectral estimate of noise. The ground terminal receives the spectral estimates and calculates the timing error estimate, ϵ . It then corrects its timing for future transmissions.

It is important to note that the ground terminal's microprocessor must control the timing of the ground terminal's components - in particular, the probe generator. The satellite's payload's microprocessor must control the satellite's components - in particular, the spectrum analyzer and the analog-to-digital converter. The control of the various components is the main function of each microprocessor.

4.2 Experimental System

The simplifications made in the hardware-based laboratory experiment are discussed here. The system does not employ frequency hopping. Frequency hopping and dehopping of the signal would add considerable complexity to the experimental system without providing any additional information on the performance of the two-tone estimator. The signal is gated at the input of the surface acoustic wave spectrum analyzer in the same manner as if it were frequency-hopped. Also, the switching of the oscillators at the probe generator causes the two-tone probe to be phase incoherent. This is precisely the case in a frequency-hopping system.

In the experimental system, it is assumed that the downlink has been established and that information can be transferred perfectly from the satellite to the ground terminal. This assumption is valid because the uplink time synchronization occurs after the downlink time synchronization. Thus, the downlink should be available. Downlink transmission errors are not considered.

In an actual system, synchronization probes are transmitted L at a time. In the experimental system, synchronization probes are transmitted one at a time. The experimental setup has pauses between pulses. This is necessary because the surface acoustic wave spec-

trum analyzer has only a 50 % duty cycle. The experimental setup takes into account the channel by adding Gaussian noise.

The experimental system is shown in Figure 4.2. There are two major differences between this system and the satellite system configuration shown in Figure 4.1. The two microprocessors in Figure 4.1 have been replaced by a single microprocessor to do the control functions. Also, a personal computer has been added to calculate statistics of the estimator.

The two synchronization processes are quite similar. The probe generator creates the two-tone probe. Noise is added to the probe. The probe plus noise is applied to the input of the surface acoustic wave spectrum analyzer. The output of the spectrum analyzer is sampled by the analog-to-digital converter. Its output is then transferred to the computer. The above is repeated for L hop periods. The microprocessor averages the L hops to obtain averaged signal spectral estimates. The averaged signal spectral estimates are stored in the microprocessor's memory. The above procedure is repeated many times to obtain many averaged signal spectral estimates. Then, the microprocessor's memory is read into a file, and the file is stored in the personal computer. Then, only noise is input to the surface acoustic wave spectrum analyzer, and L noise spectral estimates are obtained. These are averaged to obtain averaged noise spectral estimates. A program is run to calculate the timing error estimates from the averaged spectral estimates. The mean and standard deviation of the timing error are, then, computed by the program.

The remainder of the chapter describes the various components of the system in more detail.

4.3 Probe Generator

The probe generator creates the two-tone probe. The two-tone probe, shown in Figure 4.3, consists of two pulsed frequency tones. The two pulses, each lasting $12.5\mu s$, are placed one right after the other. Thus, the total probe lasts $25\mu s$, the duration of the hop period. It is desirable to have sharp edges on the pulses. In order to achieve this, two oscillators are switched - one for the first half and one for the second half. The two frequencies are 70.16MHz and 70.32MHz and have a separation of 160kHz. Since $T_h = 25\mu s$, then,

$$\frac{1}{T_h} = 40kHz \quad (4.1)$$

Figure 4-2: Experimental System Configuration

Figure 4-3: The Two-Tone Probe

Since the frequency separation should be an even integer multiple of $1/T_h$ to minimize sidelobe interference, the spacing of 60 kHz has been chosen. The two-tone spacing of 160 kHz corresponds to the M-ary frequency-shift-keying signal spacing when $M=8$. For $M=8$, a continuous intermediate frequency band of 320 kHz would be allocated to each user. The user spaces 8 frequency slots or bins within the band with a frequency separation of 40 kHz. For the two-tone probe method, the two tones could then be placed in bins 2 and 6, or bins 3 and 7. Bins 1 and 8 should not be used because the two-tone probe could interfere with the adjacent users with this placement.

The probe generator is shown in Figure 4.4 where the switches are labelled "SW". The switches are Mini-Circuits ZFSWA-2-46's having 5 terminals: two input terminals, two control terminals, and one output terminal. The input terminals are labelled "IN1" and "IN2", the control terminals are labelled "C1" and "C2", and the output terminal is labelled "OUT". A control voltage of -5 Volts applied at one of the control terminals, while keeping the other control terminal at 0 Volts, connects one of the input terminals to the output. The insertion loss is less than 1.1 dB and the isolation is at least 50 dB. However, if both control terminals are at 0 Volts, or at -5 Volts, then the isolation from input to output of each path is only about 25 dB. This is not acceptable for the experimental setup since there is a time when both control voltages are at 0 Volts. This allows too much signal feedthrough at a time when only noise is supposed to be present. Thus, two switches are put in series in each path as shown in Figure 4.4. These Mini-Circuits switches were chosen since it is desirable to have a two-tone probe with sharp edges. They have rise and fall times of 3 ns. Thus, the switching speed is limited by the speed of the control signals.

4.4 Noise Generator

Noise is added to the two-tone probe in order to investigate its effect on the timing error estimator. The noise is generated as shown in Figure 4.5. The noise source is a Digital Communications Corporation IF Noise Source Model 3002. It creates noise which is flat from dc to 140MHz and then tails off. This noise is passed through a bandpass filter in order not to exceed the input power limitations of the surface acoustic wave spectral analyzer and the amplifier. The filter is a K & L FH070-2 B5B51-70/X.1-0/0 filter, having a 70 MHz centre frequency and a 4 MHz 3 dB bandwidth. The noise signal is, then, passed through an attenuator. The attenuator, an Alan 10 dB attenuator, reduces the power level so that

Figure 4-4: Probe Generator

Figure 4-5: Noise Generator

it can be applied to the input of the amplifier. The amplifier increases the noise power to the desired level. The amplifier is an Avantek UTO-521 30 dB amplifier. The output noise signal from the amplifier is applied to a combiner where it adds to the two-tone probe.

4.5 Surface Acoustic Wave Spectrum Analyzer

The surface acoustic wave spectrum analyzer provides the Fourier transform necessary for the demodulation of signals. The unit used in this experiment is the Racal-MESL Real Time SAW Spectrum Analyzer W1741-70. The device is described in a document by Lee and Miller[14].

The spectrum analyzer operates on signals with a centre frequency of 70 MHz and a bandwidth of 4 MHz. Figure 4.6 shows the multiply-convolve type structure of the spectrum analyzer.

The input signal preparation does two things. First, it downconverts the input signal of center frequency of 70.0 MHz to a signal with a center frequency 57.3 MHz. Second, it passes the signal through a time varying amplifier which allows the user to put a temporal window on the input signal. For the purposes of this thesis, a simple rectangular window has been used.

The prepared signal is applied to a mixer. Here, it is multiplied by an upchirp. The upchirp is $25\mu\text{s}$ long going from 20.0 MHz to 24.0 MHz. The frequency-time diagram of the signal is shown in Figure 4.7.

The output of the mixer is applied to a convolver. The output of the convolver gives the frequency content of the input signal as a function of output time. The output is multiplied by an intermediate frequency of 69.3 MHz.

The envelope detector removes the intermediate frequency and leaves only the magnitude of the input signal's spectrum. This can be sampled at the appropriate times to give the spectral estimates.

The surface acoustic wave spectrum analyzer requires a trigger signal $5\mu\text{s}$ before the signal to be analyzed is received. The trigger signal is the leading edge of a positive-going TTL signal of $1.0\mu\text{s}$ duration and is supplied by the microprocessor. It, then, outputs the frequency spectrum $44.525\mu\text{s}$ after the trigger pulse. The chirp slope is $0.16\text{MHz}/\mu\text{s}$.

Figure 4-6: Surface Acoustic Wave Spectrum Analyzer

Figure 4-7: Chirp Generator's Output Signal

4.6 Analog-to-Digital Converter

An Analog Devices 12-Bit Video Analog-to-Digital Converter model MOD-1205 is used as the experimental system's analog-to-digital converter. The converter is shown in Figure 4.8. It is a twelve bit analog-to-digital converter having a resolution of 1 mV. It can digitize input signals with swings from -2.047 to 2.047 Volts. It produces a 12 parallel bit output in NRZ form compatible with TTL. The output is in 2's complement notation. The analog-to-digital converter requires an encode signal to tell it when to sample. The encode signal is a 100 ns wide positive-going TTL pulse. The data resulting from an encode command is available two conversion cycles plus 275 ns after the encode command. This implies that three encode commands are required to get the data to the output. The data remains on the output lines until the next set of data is ready. The analog-to-digital converter can encode every 200 ns periodically or randomly. The encode commands are supplied by the microprocessor.

4.7 Microprocessor

The microprocessor system used in this experiment is the TMS320C25 system, although an actual TMS320C25 microprocessor with associated memory is not used. Instead, a special board has been installed in a Compaq 386 personal computer allowing the Compaq to emulate the operation of the TMS320C25. The board and its associated software are called the TMS320C25 Software Development System(SWDS). The SWDS interfaces to the target system through a 68 pin socket. The target system perceives the SWDS to be the TMS320C25 microprocessor. The SWDS performs all microprocessor operations plus reads and writes to the memory which is located inside the Compaq 386 computer. The reason for using the development system is that it allows easy software manipulation and also, simplifies the hardware design by not having to implement the memory and other components. The microprocessor's hardware and software are described below beginning with the hardware.

The TMS320C25 microprocessor is shown in Figure 4.9. VSS and VCC are the power supply connections; VSS is tied to 0 Volts and VCC is tied to +5 Volts. The READY line is tied to +5 Volts to indicate that external devices are prepared for bus transfer of

Figure 4-8: Analog-to-Digital Converter

Figure 4-9: The TMS320C25 Microprocessor

information. In the experimental system, all devices are fast enough that a wait cycle is not required. The HOLD input can be used to put data, address, and control lines in a high impedance state although this is not needed for the experiment. Hence, the HOLD line is also tied to +5 Volts. The SYNC input allows clock synchronization of two or more TMS320C25 microprocessors. Since only one processor is used here, the SYNC line is tied to +5 Volts. If the BIO line is low, the TMS320C25 executes a branch. This is not needed in the experiment; hence, it is tied to +5 Volts. INT0, INT1, and INT2 are external user interrupt inputs. Since external interrupts are not needed, these lines are also tied to +5 Volts. D0 through D15 are parallel data bus lines used for data input and data output to and from the microprocessor. D0 is the least significant bit, and D15 is the most significant bit. A0 through A3 are the four least significant parallel address lines. A0 is the least significant line. These address lines are needed to address the I/O space. IS is the I/O space select signal. It is always high except when it is asserted low for communicating to the I/O space. The D0-D15, A0-A3, and IS lines are used to send control signals to the rest of the system and to receive data from the rest of the system. The TMS320C25 has a 100 ns instruction cycle. The interfacing between these lines and the various components in the system are discussed in the following.

The interfacing to the probe generator and the surface acoustic wave spectrum analyzer is shown in Figure 4.10. The TMS320C25 interfaces to the I/O space through two 74AS138 integrated circuits. One of the 74AS138's is used for the interface to the probe generator and the spectrum analyzer; the other is used for the analog-to-digital converter interface. The 74AS138 is a 3-line to 8-line decoder. The three input lines - A, B, and C - are used to select one of eight output lines - Y0 through Y7. There are three enable lines - G1, G2A, and G2B. When G1 is high, and G2A and G2B are low, the three input lines select one output line to go low. The remaining output lines are high. Other combinations of enable lines cause all the output lines to be high. The microprocessor when given the software command "OUT d,n" causes an address to be put on the address lines A0 through A3 corresponding to the number n and causes data to be put on the data lines which is located in memory location d. In general,

$$0 < n < 15 \quad (1.2)$$

With the arrangement shown in Figure 4.10, an n such that

$$0 < n < 7 \quad (4.3)$$

Figure 4-10: Probe Generator and Spectrum Analyzer Interfacing

causes one of the output lines of the 74AS138 to go low. Other n's cause the other 74AS138 to have one of its outputs selected. The Y0, Y1, and Y2 outputs of the 74AS138 are connected to one of the 74AS109 integrated circuits. The Y3 and Y4 outputs are connected to another 74AS109. The Y5, Y6, and Y7 outputs are not connected and are not used.

The 74AS109 integrated circuit is a dual J-K positive-edge-triggered flip-flop with clear and preset. The CLK line is tied to ground. This means that the R and S lines control the output line, Q, of the flip-flop. The setting of J and K do not matter. When R and S are high, then, the output line holds its value. When S is low, and R is high, then, Q is high. When R is low, and S is high, then, Q is low. When both are low, then, the output is unknown. Note that there are two flip-flops on each integrated circuit.

The flip-flops are used to generate the pulses necessary to control the switches for the probe generator and the pulse necessary to trigger the surface acoustic wave spectrum analyzer. Consider the switch controls first. There are two switches to control - one for the 70.16 MHz path and one for the 70.32 MHz path. For the two-tone probe, each control signal must be $12.5\mu\text{s}$ long. The 70.16 MHz control signal must end as the 70.32 MHz control signal begins. This can be achieved as follows using a 74AS109. The outputs, 1Q and 2Q, of the 74AS109 are set to be low. The output line Y0 is selected and goes low. This causes 1Q to go high. $12.5\mu\text{s}$ later, Y1 is selected. This causes 1Q to go low and 2Q to go high. $12.5\mu\text{s}$ later, Y2 is selected. This causes 2Q to go low. Thus, the outputs, 1Q and 2Q, are two control signals as required. Consider the surface acoustic wave spectrum analyzer trigger signal now. The trigger signal is $1.0\mu\text{s}$ in duration. The other 74AS109 is now used. 1Q is originally low. The selection of Y3 causes 1Q to go high. After $1.0\mu\text{s}$, the selection of Y4 causes 1Q to go low. Thus, the trigger signal is generated. The trigger signal for the spectrum analyzer is TTL compatible. Hence, the output of the flip-flop can be used directly. However, the switches of the probe generator require a -5 Volt to ground signal. Hence, there must be a voltage change.

The output of the 74AS109 flip-flop can be made suitable for the probe generator switch control signal by the use of the operational amplifier arrangement shown in Figure 4.10. Harris HA1-2540-5 wideband, fast-settling operational amplifiers are used. The operational amplifiers are arranged with the positive input terminal grounded, a feedback resistor of $13\text{k}\Omega$ connected between the negative input terminal and the output terminal, and an input resistor of $10\text{k}\Omega$ connected to the negative input. This arrangement converts a positive-

2

OF/DE

2

Figure 4-11: Analog-to-Digital Converter Interfacing

Figure 4-11: Analog-to-Digital Converter Interfacing

Figure 4-12: The Simplified TMS320C25 Program

The second step is to initialize the address registers and the experimental system. The address registers are used to count the number of hops that have been analyzed, to count the number of trials, and to store and keep track of the location of data in memory. Then, the program sends the trigger signal to the surface acoustic wave spectrum analyzer. Note that all timing is relative to this signal. The program, then, waits and sends the control signals to the probe generator to create the two-tone probe at the appropriate time. The program then waits again until it is time to send the encode commands to the analog-to-digital converter. It waits again and reads in the spectral samples from the analog-to-digital converter. The program continues to send trigger signals to the spectrum analyzer until L hops have been analyzed. It, then averages the L hop spectral estimates. For some methods, it calculates the mean as the average and in other methods, it calculates the mean square average. The average is stored in memory. The whole procedure repeats until N averaged spectral estimates have been accumulated. These N estimates are read into a file which is further processed by a personal computer program.

The program is run twice - once to collect the signal spectral estimates and once to collect the noise spectral estimates. The two-tone probe plus noise is applied to the input of the surface acoustic wave spectrum analyzer in order to collect the signal spectral estimates. Only noise is applied to collect the noise spectral estimates.

The TMS320C25 microprocessor is used to perform two functions in the experimental system. The first is to control the various parts of the system; the second is to compute averaged spectral estimates.

4.8 Personal Computer

The personal computer receives a file containing the spectral estimates from many trials of L hops. It calculates the timing error estimate, ϵ , from them. It also calculates the mean and standard deviation of the timing estimates from the various trials. There are two programs that are needed. One does the computation with bias reduction and one does the computation without bias reduction. The programs are written in QuickBasic.

The QuickBasic program is shown in Figure 4.13 for the bias reduction method. It requires two input files. One contains the averaged signal spectral estimates and the other contains the averaged noise spectral estimates. These estimates are stored in hexadeci-

Figure 4-13: The Simplified QuickBasic Bias Reduction Program

mal form. The first step is to change the values to decimal form. The averaged signal estimates, $\frac{1}{L} \sum_{j=1}^L z_{1j}^2$ and $\frac{1}{L} \sum_{j=1}^L z_{2j}^2$, are stored in one array and the averaged noise estimates, $\frac{1}{L} \sum_{j=1}^L z_{01j}^2$ and $\frac{1}{L} \sum_{j=1}^L z_{02j}^2$, are stored in another array. The corrected spectral estimates, $\hat{\theta}_1$ and $\hat{\theta}_2$, are formed as follows:

$$\hat{\theta}_1 = \sqrt{\frac{1}{L} \sum_{j=1}^L z_{1j}^2 - \frac{1}{L} \sum_{j=1}^L z_{01j}^2} \quad (4.4)$$

$$\hat{\theta}_2 = \sqrt{\frac{1}{L} \sum_{j=1}^L z_{2j}^2 - \frac{1}{L} \sum_{j=1}^L z_{02j}^2} \quad (4.5)$$

If the quantity under the square root sign is negative, then,

$$\hat{\theta}_i = 0 \quad (4.6)$$

The timing error estimates, $\hat{\epsilon}$, are formed using equations (3.85), (3.86), or (3.87). Then, the mean and standard deviation of the $\hat{\epsilon}$ are computed and are printed out.

The QuickBasic program for the method without bias reduction requires only one input file, containing the averaged signal spectral estimates. The program proceeds in a similar manner as the one in Figure 4.13. The file containing the averaged signal spectral estimates is translated from hexadecimal to decimal. The averaged signal estimates, $\frac{1}{L} \sum_{j=1}^L z_{1j}$ and $\frac{1}{L} \sum_{j=1}^L z_{2j}$, are stored in an array. Then,

$$\hat{\theta}_1 = \frac{1}{L} \sum_{j=1}^L z_{1j} \quad (4.7)$$

$$\hat{\theta}_2 = \frac{1}{L} \sum_{j=1}^L z_{2j} \quad (4.8)$$

These are used to form $\hat{\epsilon}$ using equations (3.85), (3.86), or (3.87). The mean and standard deviation of $\hat{\epsilon}$ are computed, and the results are printed out.

4.9 Summary

The experimental system has been discussed in this chapter. All of the parts have been described in detail. The next chapter presents the testing of this system. The testing procedure is described, and the experimental results are presented.

Chapter 5

Experimental Results

5.1 Introduction

Experiments have been run on the system described in Chapter 4 to determine the performance of the two-tone method in an additive noise environment. These experiments involve producing a timing error estimate, $\hat{\epsilon}$, for various different actual timing errors, ϵ . The mean and standard deviation of the 128 $\hat{\epsilon}$'s for various different ϵ 's are calculated. The estimates are obtained for different numbers of hop periods, L , and for different signal-to-noise ratios. Also, results are obtained for two different algorithms - one without bias reduction and one with it. The results are compared with results obtained through simulation. Sources of error are also discussed.

The signal-to-noise used here is:

$$SNR = \frac{E_h}{N_0} \quad (5.1)$$

where E_h is the energy per hop period and N_0 is the spectral noise density. E_h has units of Joules and is equal to:

$$E_h = P_s T_h \quad (5.2)$$

where P_s is the power of the signal and T_h is the hop period. In this case,

$$T_h = 25\mu s \quad (5.3)$$

Using a power meter, P_s is measured at the input of the surface acoustic wave spectrum analyzer. Using a spectrum analyzer, N_0 is measured at the same place. The units of N_0 are Watts per Hertz.

The surface acoustic wave spectrum analyzer is used as the timing reference. It processes a $25\mu s$ long pulse. In a synchronized position, the received signal - also $25\mu s$ long - must be aligned with the processing period. If the two are not aligned, then, there is a timing error. The timing error can be either early or late. It is, therefore, necessary to

define a convention for denoting the direction of the misalignment. Thus, if the received signal is too early, then, the estimate is positive. If it is too late, the estimate is negative. In this chapter, it is important to note that the estimator, $\hat{\epsilon}$, is in units of μs . It is not normalized to the hop period.

5.2 Estimate Without Bias Reduction

This method obtains the timing error estimate without using bias reduction. These measurements show that this method forms an inconsistent estimator.

The method proceeds as follows. L two-tone probes are generated. They produce L pairs of spectral estimates, z_{1j} and z_{2j} . Two averaged spectral estimates are produced as follows:

$$\hat{\theta}_i = \frac{1}{L} \sum_{j=1}^L z_{ij} \quad (5.4)$$

for $i=1,2$. The timing error estimate is produced as follows. For $\hat{\theta}_1 > \hat{\theta}_2$,

$$\hat{\epsilon} = \frac{T_h}{2} \left(\frac{\hat{\theta}_2}{\hat{\theta}_1} - 1 \right) \quad (5.5)$$

For $\hat{\theta}_1 < \hat{\theta}_2$,

$$\hat{\epsilon} = \frac{T_h}{2} \left(1 - \frac{\hat{\theta}_1}{\hat{\theta}_2} \right) \quad (5.6)$$

Here, 128 $\hat{\epsilon}$'s are produced from 128 trials. The mean and standard deviation of these estimates are calculated.

The experimental results for this method are shown in Figures 5.1, 5.2, 5.3, and 5.4. Two values - 16 and 256 - for the number of hop periods, L , have been used. Three signal-to-noise ratios - 20 dB, 10 dB, and 0 dB - have been used. For the three different signal-to-noise ratios, the signal power is held constant at -23 dBm. For the 0 dB and 10 dB cases, the externally input noise is -69 dBm/Hz and -79 dBm/Hz respectively. For the 20 dB case, all the noise is internally generated in the surface acoustic wave spectrum analyzer. It produces an output noise that is approximately 10 dB less than the output noise when N_0 is -79 dBm/Hz. Thus, the input referred noise is about -89 dBm/Hz for this case, and yields a signal-to-noise ratio of 20 dB.

The mean of the estimator is shown in Figures 5.1 and 5.3. The desired curve is:

$$\hat{\epsilon} = \epsilon \quad (5.7)$$

Figure 5-1: Estimates without Bias Reduction and $L=16$

Figure 5-2: Estimates without Bias Reduction and $L=16$

Figure 5-3: Estimates without Bias Reduction and $L=256$

Figure 5-4: Estimate without Bias Reduction and $L=256$

The 20 dB curve in both cases is close to this. There is some bending in it near $|\epsilon| = 12.5\mu s$. However, reasonable error estimates can be achieved using this curve. This is not the case for the 10 dB or 0 dB curves. Both curves bend down severely from the expected curve of (5.7) due to the inconsistency of the estimator. Hence, the method performs poorly. The 0 dB curve is so close to being horizontal that estimates made at this signal-to-noise ratio are totally unreliable. Note that there is little change in the curves if L is increased from 16 to 256. Hence, no improvement can be expected if L is made even larger. This conclusion concurs with the theory presented earlier. The goal now is to find a consistent estimator.

5.3 Estimate with Bias Reduction

Bias reduction is used in order to find a consistent estimator. The experimental results presented here verify that this method produces a consistent estimator.

The method proceeds as follows. L two-tone probes are generated. Then, L probes without any signal are generated. The first L probes give L pairs of signal spectral estimates, z_{1j} and z_{2j} . The next L probes give L pairs of noise spectral estimates, z_{01j} and z_{02j} . These estimates are made in the same bin as the signal in order to minimize systematic error. The averaged spectral estimates are formed as follows:

$$\hat{\theta}_i = \sqrt{\frac{1}{L} \sum_{j=1}^L z_{ij}^2 - \frac{1}{L} \sum_{j=1}^L z_{0ij}^2} \quad (5.8)$$

for $i = 1, 2$. If the quantity under the square root sign in (5.8) is negative, $\hat{\theta}_i$ is set to zero. The estimate, $\hat{\epsilon}$, is formed the same way as in the method without bias reduction from the $\hat{\theta}_i$. The equations are given in (5.5) and (5.6). Again, 128 $\hat{\epsilon}$'s are produced from 128 trials. From these trials, the mean and standard deviation of the estimates are calculated.

The experimental results are shown in Figures 5.5, 5.6, 5.7, and 5.8. The measurements are done with the same signal and noise powers as in the method without bias reduction. Note that the curves of the estimator mean in Figures 5.5 and 5.7 for the different signal-to-noise ratios are much closer together. In Figure 5.5, the 20 dB curve is almost exactly on the desired curve of (5.7). The 10 dB curve is on top of the 20 dB curve near the middle. However, the 10 dB curve shows a bias near the ends. The 0 dB curve still shows a bias throughout. However, the bending of the 0 dB curve in Figure 5.5 is much less than the bending of the 0 dB curve in Figure 5.1. Thus, with bias reduction techniques, the bias of

Figure 5-5: Estimates with Bias Reduction and $L=16$

Figure 5-6: Estimates with Bias Reduction and $L=16$

Figure 5-7: Estimates with Bias Reduction and $L=256$

Figure 5-8: Estimates with Bias Reduction and $L=256$

the estimator is reduced as expected. The estimator is expected to be consistent because the bias should reduce as L increases.

In Figure 5.7, measurements are done with L equal to 256. The three curves are almost on top of one another. They also are very close to the expected curve of (5.7). There is still some bias in the 0 dB curve near its ends. However, this bias could be reduced by increasing L . Thus, it can be concluded that the estimator is consistent. However, the disadvantage of the bias reduction method is that the estimator standard deviation increases. This can be observed by comparing Figures 5.6 and 5.8 with Figures 5.2 and 5.4 respectively. The increase is particularly large for the 0 dB case.

5.4 Comparison of Theory with Experiment

This section compares the experimental results with the simulated results generated by Mason[13]. The simulated results have been generated for L 's of 16 and 256 and for signal-to-noise ratios of 0 dB and 10 dB. Some of the same simulation results have been previously presented in Chapter 3. The difference is that the results presented in this chapter are expressed in μs . Also, the simulated results presented results for only positive timing error. The negative timing error has been obtained by simply taking the negative of the positive result for the same magnitude.

The experimental results are compared with the simulation in Figures 5.9 through 5.16. Figures 5.9 and 5.10 show the $L=16$ and 0 dB signal-to-noise ratio case. Because of the small number of samples and the large noise power, the two curves of the mean have large statistical deviation. Both curves have a large bias. The experimental standard deviation is larger than the theoretical standard deviation.

Figures 5.11 and 5.12 show the $L=16$ and 10 dB signal-to-noise ratio case. Here, both mean curves are smooth and have low bias. The two mean curves are not exactly aligned. The experimental standard deviation is slightly larger than the theoretical one.

Figures 5.13 and 5.14 show the $L=256$ and 0 dB signal-to-noise ratio case. Here, the two mean curves are fairly smooth and are closely aligned. There is some bias shown near the ends. Again, the experimental standard deviation is larger than the theoretical standard deviation.

Figures 5.15 and 5.16 show the $L=256$ and 10 dB signal-to-noise ratio case. The two

Figure 5-9: Estimates with $L=16$ and $\text{SNR}=0\text{dB}$

Figure 5-10: Estimates with $L=16$ and $\text{SNR}=0\text{dB}$

Figure 5-11: Estimates with $L=16$ and $\text{SNR}=10\text{dB}$

Figure 5-12: Estimates with $L=16$ and $\text{SNR}=10\text{dB}$

Figure 5-13: Estimates with $L=256$ and $\text{SNR}=0\text{dB}$

Figure 5-14: Estimates with $L=256$ and $\text{SNR}=0\text{dB}$

Figure 5-15: Estimates with $L=256$ and $\text{SNR}=10\text{dB}$

Figure 5-16: Estimates with $L=256$ and $SNR=10dB$

mean curves align very closely. They show a slight bias at the ends. The standard deviation of the experimental curve is slightly higher than the theoretical curve.

The discrepancies between the experimental results and the theoretical results are caused by various factors as discussed in the next section.

5.5 Sources of Error

There are several sources of error introduced by the experimental system which cause the estimator to have imperfect performance. Ideally, the curve of estimated timing error versus actual timing error should be a straight line with a slope of one as given by (5.7). However, it is not. One of the causes is the additive noise.

One of the principal sources of performance degradation arises from the time of arrival of the half-pulse at the surface acoustic wave spectrum analyzer. The half-pulse is a pulse of width $12.5\mu\text{s}$. The spectrum analyzer has a window of $25\mu\text{s}$. The output of the spectrum analyzer varies depending on the location of the half-pulse within the window. Two main effects are observed. First, a half-pulse located in the second half of the window causes the spectrum analyzer to have a greater peak output voltage than a half-pulse located in the first half of the window causes. Thus, in the synchronized position, the second tone in the two-tone probe has a greater spectral estimate than the first tone in the probe. This effect causes the estimated timing error to be greater than zero when the actual timing error is zero. For example, for $L=256$ and signal-to-noise ratio of 20dB, the timing offset is 604 ns. A second effect caused by the location of the half-pulse is a time shift in the output voltage peak of the spectrum analyzer. This time shift causes the analog-to-digital converter not to sample at the peak of the output spectrum. This sampling error causes the spectral estimates to be low for $\epsilon \neq 0$. Here, the effect is to cause offsets in the estimator for unsynchronized positions.

The measurements of the time of arrival error are shown in Figures 5.17 and 5.18. Figure 5.17 shows a time-shifted half-pulse consisting of a 70.32 MHz signal with a power of -23 dBm within the surface acoustic wave spectrum analyzer's window. At various time shifts, 256 samples have been taken of the spectrum analyzer's output. Figure 5.18 plots the mean of the spectrum analyzer's output voltage samples as a function of the position of the half-pulse within the spectrum analyzer's window.

Figure 5-17: The Half-Pulse

Figure 5-18: Time of Arrival Error

There are two possible corrections for the performance degradation introduced by the time of arrival of the half-pulse. One is the use of a calibration curve. The curve is used as a "look-up" table in order to synchronize. The other possible correction is the use of a correction factor. The correction factor, γ , would be applied as:

$$\hat{\theta}_1' = \gamma \hat{\theta}_1 \quad (5.9)$$

Then, $\hat{\epsilon}$ is formed using equations (5.5) and (5.6) with $\hat{\theta}_1$ replaced by $\hat{\theta}_1'$. For the case of $L=256$ and signal-to-noise ratio of 20 dB, γ is 1.0508. No "look-up" table is required with this method of correction.

Another principal source of performance degradation is sidelobe interference. Sidelobe interference is caused by imperfect time or frequency synchronization. When the system is perfectly synchronized, there is no interference with the spectrum analyzer's output peak voltages due to the two tones of the two-tone probe because the two tones are orthogonally spaced. However, when the system is out of synchronization, part of the two-tone probe is chopped off. This causes the two half-pulses of the two-tone probe to be of different time duration. One half-pulse is $12.5\mu s$ long, whereas the other is $(12.5 - |\epsilon|)\mu s$ long. The output of the shorter pulse interferes with the peak output voltage of the longer pulse. However, the output of the longer pulse has no effect on the peak output voltage of the shorter pulse. The sidelobe interference acts as an added noise and causes the estimator to have a larger bias and a larger standard deviation.

Another error is caused by frequency weighting of the spectrum analyzer. The analyzer weights different frequencies with different weights. This causes two different tones of the same power to have different output voltages. However, this effect has been measured to be small.

5.6 Summary

In this chapter, the experimental results have been presented and compared with theory. Experimental results for the method without bias reduction verify that this method produces an inconsistent estimator. Experimental results for the method with bias reduction show that this method produces a consistent estimator. There is excellent agreement between experimental results and simulated results[13] in most cases. The standard deviations of the estimators for the experimental results tend to be higher than those for the

simulated results. Some systematic errors cause discrepancies between the means of the simulated and experimental results. Sources of error for these cases have been discussed. The experimental results show that the theoretical model predicts the performance of the estimator very well. They also show that the two-tone method is a viable method for implementation in a frequency-hopped spread spectrum satellite communication system.

Chapter 6

Conclusion

6.1 Summary

This thesis has investigated the problem of fine time uplink synchronization for frequency-hopped spread spectrum satellite communication systems. The background material in satellite communication and spread spectrum communication has been presented. The role of fine time uplink synchronization in a frequency-hopped spread spectrum satellite communication system has been discussed. Theoretical analyses have been presented for various methods of fine time uplink synchronization. One particular method has been implemented in a hardware-based laboratory experiment. Experimental results have been obtained from this hardware.

There are two major conclusions that can be drawn from the work. One concerns the theoretical analyses. The other involves the experimental results.

The theoretical analyses showed that the best performance in the presence of additive noise is achieved by the modified early-late filter method. This method produces a consistent estimator. The modified early-late filter method has the smallest bias and the lowest estimator standard deviation of all the consistent methods. The method's only disadvantage is that it requires extra hardware in the satellite's communication payload. If this extra weight and complexity can be tolerated in the satellite, then the modified early-late filter method is the preferred method. If the additional weight and complexity cannot be tolerated, one of the other methods should be chosen. The early-late filter method requires the same additional hardware as the modified early-late method and has poorer performance. The delay-advance method forces the transmitter to alter its timing to create the probes. It also introduces errors by obtaining signal spectral estimates from two different hop periods. The two-tone method has the same amount of bias as the delay-advance method, does not alter transmitter timing, and obtains signal spectral estimates from a single hop period. Thus, the two-tone method is the preferred method if the satellite's weight and payload complexity are a concern.

There have been a number of publications[3]-[5] in the last two years which describe new methods for fine time uplink synchronization of frequency-hopped spread spectrum satellite communication systems. These publications do not provide experimental results for the performance in noise of any of the fine time uplink synchronization methods. One of the new methods which is described in the publications[4]-[5] is the two-tone method. In this thesis, the two-tone method has been implemented in a hardware-based laboratory experiment, and measurements of noise performance of the method have been presented. The first set of measurements has been performed on the two-tone method without bias reduction. These measurements show an inconsistent estimator. Thus, this method shows the need for bias reduction. The second set of measurements has been performed on the two-tone method with bias reduction. Here, the estimator is consistent. A bias exists for timing errors near half of a hop period; however, this can be reduced by increasing L , the number of hop periods. Also, the estimator has a substantial standard deviation for low signal-to-noise ratios. Again, this can be reduced by increasing L . Thus, system specifications for the accuracy of the timing error estimate determine L . The experimental results demonstrate that increasing L reduces the bias and standard deviation of the estimator. The experimental results are compared with simulated results generated from theory[13]. The estimator means are almost exactly the same for both results. There are small discrepancies caused by systematic error and statistical deviation. The estimator standard deviation of the experimental results are slightly higher than the theoretical standard deviation. This is thought to be caused by sidelobe interference. Thus, it can be concluded that the theoretical model used to generate the simulated results provides a fairly accurate indicator of the actual performance of the system. In addition, the experimental results show that the two-tone method is indeed a viable approach.

There are areas of interest that should still be investigated regarding fine time uplink synchronization. These are discussed in the next section.

6.2 Future Work

Future studies in fine time uplink synchronization for frequency-hopped spread spectrum communication systems can be conducted in many areas. These are described below.

One area of interest is the single-hop two-tone method. In this thesis, the bias reduction

method uses two hop periods to generate an estimate. The first hop period is used for signal spectral estimates, and the second is used for noise spectral estimates. A method using a single hop period to obtain both estimates is of interest since it would decrease the number of probes needed. The performance would also degrade; however, the amount of degradation is of interest.

A second area of interest concerns tone jamming. The system should be tested in the presence of a hostile tone jammer. Various jamming strategies should be tried.

A third area is windowing. The surface acoustic wave spectrum analyzer uses windowing for communication. The synchronization should be attempted using windows other than rectangular. This may be necessary if the window cannot be changed.

A fourth area is the implementation of the modified early-late filter method. This method gives the best performance in simulated results. It would be valuable to see if the method has as good a performance for experimental results.

A fifth area is the implementation of a modified two-tone method. This method would require three hop periods. The first would be a conventional two-tone probe used to collect two spectral samples. The second would be a single frequency normal communication's signal used to collect one spectral sample. The third would be a probe without a signal used to collect noise samples. The smallest of the three signal samples would be discarded, and the other two signal samples and the noise samples would be used to form spectral estimates. These spectral estimates would be placed in equations similar to those employed by the modified early-late filter method. This method is expected to be consistent and may perform as well as the modified early-late filter method. It has the additional benefit of not requiring extra hardware in the satellite's communication payload.

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